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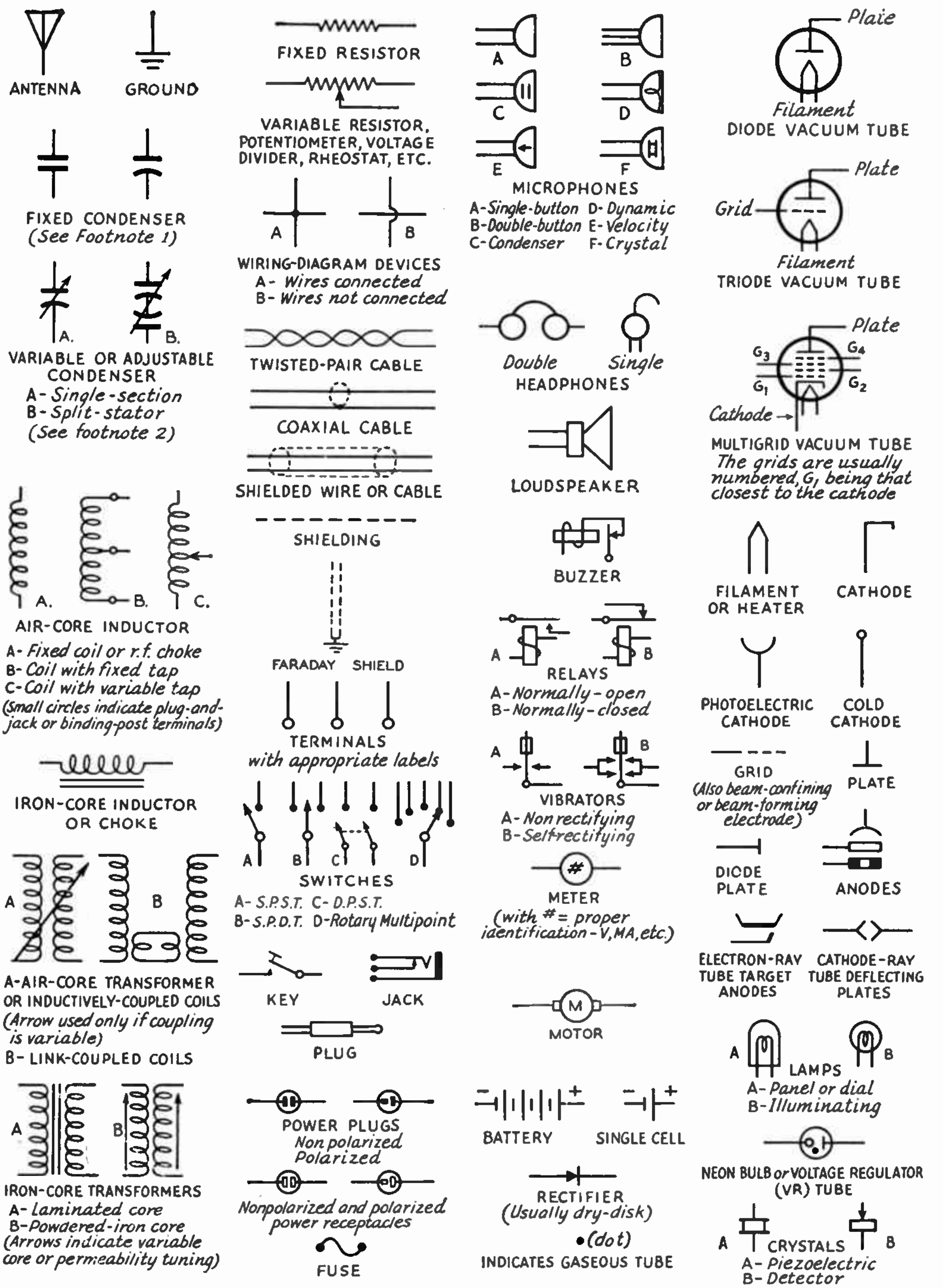
The radio amateur's handbook

THE STANDARD MANUAL OF AMATEUR RADIO COMMUNICATION

$3.00
U.S.A. Proper

PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

SCHEMATIC SYMBOLS USED IN CIRCUIT DIAGRAMS



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

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

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

THE RADIO AMATEUR'S HANDBOOK

By the HEADQUARTERS STAFF
of the
AMERICAN RADIO RELAY LEAGUE
WEST HARTFORD, CONN., U.S.A.

1954

Thirty-first Edition

Thirty-first Edition

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Foreword

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

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

In contrast to most publications of a comparable nature, the *Handbook* is printed in the format of the League's monthly magazine, *QST*. This, together with extensive and usefully-appropriate catalog advertising by manufacturers producing equipment for the radio amateur 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.

This thirty-first edition takes note of the changes in technical practice that have occurred in the past year. A considerable amount of new equipment in all categories appears throughout the book. The chapter on high-frequency transmitters includes new units for the Novice as well as more elaborate units for the accomplished amateur. Continuing the trend of recent years, all transmitting equipment has been designed with the reduction of harmonics in the telecasting bands as a primary feature. The chapter on v.h.f. receivers has been extensively redone, making use of the newest circuits and tubes. And the always informative data chapter on vacuum tubes and semiconductors has been expanded to include over 150 new tube types plus transistors.

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.

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THE AMATEUR'S CODE

• ONE •

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

• TWO •

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

• THREE •

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

• FOUR •

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

• FIVE •

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

• SIX •

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

— *Paul M. Segal*

Amateur Radio

Amateur radio is a scientific hobby, a means of gaining personal skill in the fascinating art of electronics and an opportunity to communicate with fellow citizens by private short-wave radio. Scattered over the globe are more than 150,000 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 100,000, 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 Army and Navy seek the coöperation of the amateur in developing communications reserves. Amateur radio supports a manufacturing industry which, by the very demands of amateurs for the latest and best equipment, is always up-to-date in its designs and production techniques — in itself a national asset. Amateurs have won the gratitude of the nation for their heroic performances in times of natural disaster. Through their organization, amateurs have coöperative working agreements with such agencies as the United Nations and the Red Cross. Amateur radio is, indeed, a magnificently useful institution.

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

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

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

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

Today, few amateurs realize that World

HIRAM PERCY MAXIM
President ARRL, 1914–1936

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

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

TRANS-ATLANTICS

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

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

Finally, in November, 1923, after some months of careful preparation, two-way amateur trans-Atlantic communication was accomplished, when Schnell, 1MO, and Reinartz, 1XAM (now W9UZ and K6BJ, respectively) worked for several hours with Deloy, 8AB, in France, with all three stations on 110 meters! Additional stations dropped down to 100 meters and found that they, too, could easily work two-way across the Atlantic. The exodus from the 200-meter region had started. The "short-wave" era had begun!

By 1924 dozens of commercial companies had rushed stations into the 100-meter region. Chaos threatened, until the first of a series of national and international radio conferences partitioned off various bands of frequencies for the different services. Although thought still centered around 100 meters, League officials at the first of these frequency-determining conferences, in 1924, wisely obtained amateur bands not only at 80 meters but at 40, 20, 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 Army and Navy of the value of the amateur as a source of skilled radio personnel in time of war. Another asset is best described as "public service."

About 4000 amateurs had contributed their skill and ability in '17–'18. After the war it was only natural that cordial relations should prevail between the Army and Navy and the amateur. These relations strengthened in the next few years and, in gradual steps, grew into coöperative activities which resulted, in 1925, in the establishment of the Naval Communications Reserve and the Army-Amateur Radio System (now the Military 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 coöperation with expeditions began in 1923 when a League member, Don Mix, 1TS, of Bristol, Conn. (now assistant technical editor of *QST*), accompanied MacMillan to the Arctic on the schooner *Bowdoin* with an amateur station. Amateurs in Canada and the U.S. provided the home contacts. The success of this venture was such that other explorers followed suit. During subsequent years a total of perhaps two hundred voyages and expeditions were assisted by amateur radio, and for many years no expedition has taken the field without such plans.

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

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.

TECHNICAL DEVELOPMENTS

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

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

A corner of the ARRL laboratory.

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

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

THE AMERICAN RADIO RELAY LEAGUE

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

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

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

The operating room at W1AW.

The operating territory of ARRL is divided into 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-two sections. There are appointments for qualified members as Official Relay Station or Official 'Phone Station for traffic handling; as Official Observer for monitoring frequencies and the quality of signals; as Route Manager and 'Phone Activities Manager for the establishment of trunk lines and networks; as Emergency Coördinator for the promotion of amateur preparedness to cope with natural disasters; and as Official Experimental Station for those pioneering the frequencies above 50 Mc. Mimeographed bulletins keep appointees informed of the latest developments. Special activities and contests promote operating skill. A special section is reserved each month in *QST* for amateur news from every section of the country.

AMATEUR LICENSING IN THE UNITED STATES

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. Examinations for all classes but the Amateur Extra may be taken by mail where the applicant lives further than a specified distance from the examining centers. Station licenses are granted only to licensed operators and permit communication between such stations for amateur purposes, i.e., for personal noncommercial aims flowing from an interest in radio technique. An amateur station may not be used for material compensation of any sort nor for broadcasting. Narrow bands of frequencies are allocated exclusively for use by amateur stations. Transmissions may be on any frequency within the assigned bands. All the frequencies may be used for c.w. telegraphy and some are available for radiotelephony by any amateur, while others are reserved for radiotelephone use by persons holding higher grades of license. The input to the final stage of amateur stations is limited to 1000 watts and on frequencies below 144 Mc. must be adequately-filtered direct current. Emissions must be free from spurious radiations. The licensee must provide for measurement of the transmitter frequency and establish a procedure for checking it regularly. A complete log of station operation must be maintained, with specified data. The station license also authorizes the holder to operate portable and mobile stations subject to further regulations. An amateur station may be operated only by an amateur operator licensee, but any licensed amateur operator may operate any amateur station within the scope of privileges conveyed by the licenses. 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 at a local district office or examining points in most of our larger cities, through FCC at Washington. A complete up-to-the-minute discussion of license requirements, and 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	dahdididit
2	dididahdahdah	7	dahdahdididit
3	didididahdah	8	dahdahdahdidit
4	dididididah	9	dahdahdahdahdit
5	dididididit	0	dahdahdahdahdah

Period: didahdidahdidah. Comma: dahdahdididahdah. Question mark: dididahdahdidit. Error: dididididididit. 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*, 25¢ postpaid.

THE AMATEUR BANDS

Amateurs are assigned bands of frequencies at approximate octave intervals throughout the spectrum. Like assignments to all services, they are subject to modification to fit the changing picture of world communications needs. 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. AØ means an unmodulated carrier, A1 means c.w. telegraphy, A2 is tone-modulated c.w. telegraphy, A3 is amplitude-modulated 'phone, A4 is facsimile, AØ is television, n.f.m. designates narrow-band frequency- or phase-modulated radiotelephony, and f.m. means frequency modulation, 'phone (including n.f.m.) or telegraphy, 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.
11 m.	26.960–27.230	AØ, A1, A2, A3, A4, 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.
	52.5–54	f.m.
2 m.	144–148	AØ, A1, A2, A3, A4, f.m.
1¼ m.	220–225	AØ, A1, A2, A3, A4, f.m.
	420–450[1]	AØ, A1, A2, A3, A4, A5, f.m.
	1,215–1,300	AØ, A1, A2, A3, A4, A5, f.m.
	2,300– 2,450	AØ, A1, A2, A3, A4, A5, f.m., pulse
	3,300– 3,500	AØ, A1, A2, A3, A4, A5, f.m., pulse
	5,650– 5,925	AØ, A1, A2, A3, A4, A5, f.m., pulse
	10,000–10,500	AØ, A1, A2, A3, A4, A5, f.m., pulse
	21,000–22,000	AØ, A1, A2, A3, A4, A5, f.m., pulse
	All above 30,000	AØ, A1, A2, A3, A4, A5, f.m., pulse

[1] Peak antenna power must not exceed 50 watts.

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

Area	*Band, kc.*	*Power (watts) Day*	*Night*
Minn., Iowa, Mo., Ark., La. and states east, plus Puerto Rico and Virgin Ids.	1800–1825 1875–1900	500	200
N. and S. Dak., Neb., Colo., N. Mex., and states west, plus Hawaiian Ids.	1900–1925 1975–2000	500*	200*
Texas, Okla., Kansas	1800–1825 1875–1900	200	75

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

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
7.175–7.200	A1	145–147	A1, A2, A3, f.m.

Technician licensees are permitted all amateur privileges 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.** 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 square inch, or per 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.

While in a normal atom the positive charge on the nucleus is exactly balanced by the negative charges on the electrons, 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 (that is, 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. 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

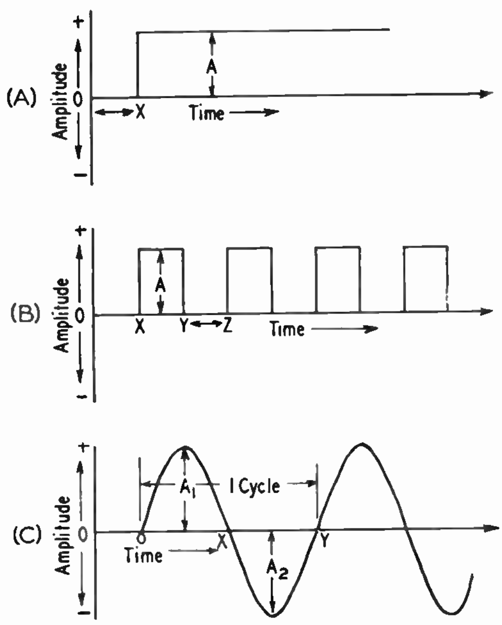


Fig. 2-1 — Three types of current flow. A — direct current; B — intermittent direct current; C — alternating current.

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. 2-1 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 equals 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

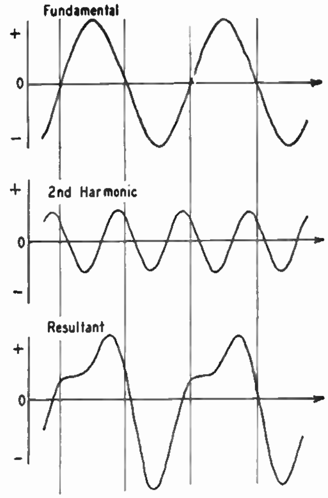


Fig. 2-2 — **A complex waveform. A fundamental (top) and second harmonic (center) added together, point by point at each instant, result in the waveform shown at the bottom. When the two components have the same polarity at a selected instant, the resultant is the simple sum of the two. When they have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

that the current (or voltage) has at any selected instant in the cycle.

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	11.3
Nickel	6.25 to 8.33
Phosphor Bronze	2.78
Silver	0.91
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 in the Miscellaneous Data chapter. This table gives the resistance, in ohms per thousand feet, of each standard wire size.

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

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

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

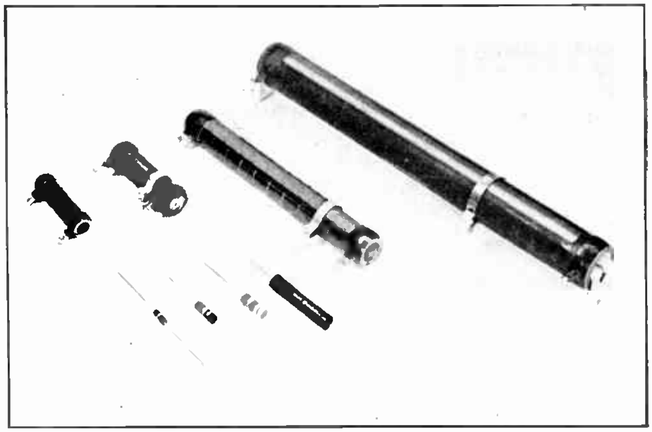
$$\frac{14}{1000} \times R = 0.05 \text{ ohm}$$

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

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

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

When the wire is not copper, the resistance values given in the wire table should be 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 the adjustable type, using 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

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

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.

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

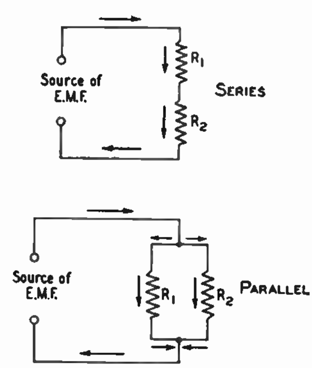


Fig. 2-4 — Resistors connected in series and in parallel.

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

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

Resistors in Series

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

R (total) $= R_1 + R_2 + R_3 + R_4 + \ldots$

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

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

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

The current flowing in the circuit is then

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

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

Voltage Drop

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

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

The 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.

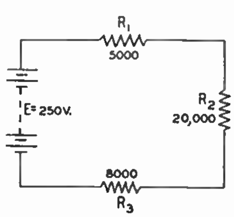


Fig. 2-5 — An example of resistors in series. The solution of the circuit is worked out in the text.

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

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

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

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

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

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

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

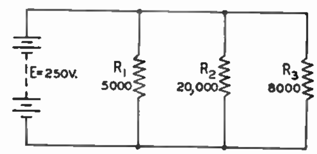


Fig. 2-6 — An example of resistors in parallel. The solution is worked out in the text.

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 such 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.

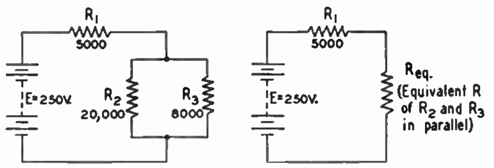


Fig. 2-7 — An example of resistors in series-parallel. The solution is worked out in the text.

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

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

The total resistance in the circuit is then

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

The current is

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

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

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

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

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

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

where I_2 = Current through R_2
I_3 = Current through R_3

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

POWER AND ENERGY

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

$$P = EI$$

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

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

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

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

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

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

$$P = I^2R$$

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

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

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

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

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

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

Electrical power in a resistance is turned into heat. The greater the power the more rapidly the heat is generated. 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 and Condensers

Suppose two flat metal plates are placed close to each other (but not touching) as shown in Fig. 2-8. Normally, the plates will be electrically "neutral"; that is, no electrical charge will be evident on either plate.

Now suppose that the plates are connected to a battery through a switch, as shown. At the instant the switch is closed, electrons will be attracted from the upper plate to the positive terminal of the battery, and the same number will be repelled into the lower plate from the negative battery terminal. This electron movement will continue until enough electrons move into one plate and out of the other to make the e.m.f. between them the same as the e.m.f. of the battery.

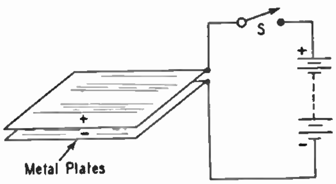


Fig. 2-8 — A simple condenser.

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

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

The charge or quantity of electricity that can be placed on a condenser is proportional to the applied voltage and to the **capacitance** or **capacity** of the condenser. The larger the plate area and the smaller the spacing between the plates 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 of a condenser many times. The ratio of the capacitance of a condenser with some material other than air between the plates, to the capacitance of the same condenser with air insulation, is called the **specific inductive capacity** or **dielectric constant** of that particular insulating material. The material itself is called a **dielectric.** The dielectric constants of a number of materials

TABLE 2-III
Dielectric Constants and Breakdown Voltages

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

* In volts per mil (0.001 inch).

commonly used as dielectrics in condensers are given in Table 2-III. If a sheet of photographic glass is substituted for air between the plates of a condenser, for example, the capacitance of the condenser will be increased 7.5 times.

Units

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

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

The formula for calculating the capacitance of a condenser is:

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

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

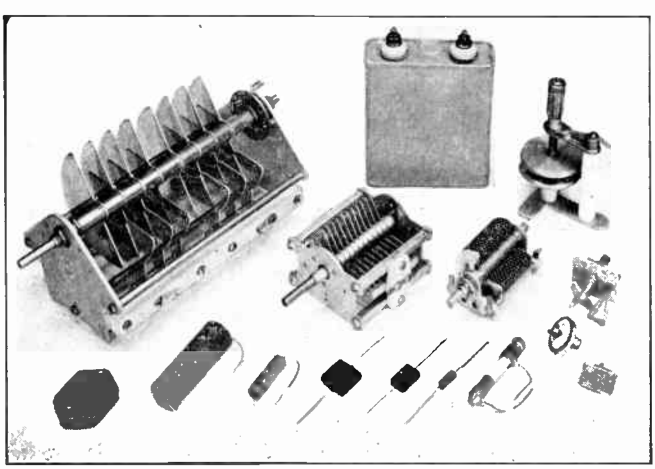
If the plates in one group do not have the same area as the plates in the other, use the area of the *smaller* plates.

> Example: A "variable" condenser has 7 semicircular plates on its rotor, the diameter of the semicircle being 2 inches. The stator has 6 rectangular plates, with a semicircular cut-out to clear the rotor shaft, but otherwise large enough to face the entire area of a rotor plate. The diameter of the cut-out is ½ inch. The distance between the adjacent surfaces of rotor and stator plates is ⅛ inch. The dielectric is air. What is the capacitance of the condenser with the plates fully meshed?
>
> In this case, the "effective" area is the area of the rotor plate minus the area of the cut-out in the stator plate. The area of either semicircle is $\pi r^2/2$, where r is the radius. The area of the rotor plate is $\pi/2$, or 1.57 square inches (the radius is 1 inch). The area of the cut-out is $\pi(\frac{1}{4})^2/2 = \pi/32 = 0.10$ square inch, approximately. The "effective" area is therefore $1.57 - 0.10 = 1.47$ square inches. The capacitance is therefore
>
> $$C = 0.224 \frac{KA}{d} (n - 1) = 0.224 \frac{1 \times 1.47}{0.125} (13 - 1)$$
> $$= 0.224 \times 11.76 \times 12 = 31.6 \ \mu\mu\text{fd.}$$
>
> (The answer is only approximate, because of the difficulty of accurate measurement, plus a "fringing" effect at the edges of the plates that makes the actual capacitance a little higher.)

The usefulness of a condenser in electrical circuits lies in the fact that it can be charged with electricity at one time and then discharged at a later time. In other words, it is capable of storing electrical energy that can be released later when it is needed; it is an "electrical reservoir."

Condensers in Radio

The types of condensers used in radio work differ considerably in physical size, construction, and capacitance. Some representative types are shown in the photograph. In **variable** condensers (almost always constructed with air for the dielectric) one set of plates is made movable with respect to the other set so that the capacitance can be varied. **Fixed** condensers — that is, having fixed capacitance — also can be made with metal plates and with air as the dielectric, but usually



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

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** condenser uses aluminum-foil plates with a semiliquid conducting chemical compound between them; the actual dielectric is a very thin film of insulating material that forms on one set of plates through electrochemical action when a d.c. voltage is applied to the condenser. The capacitance obtained with a given plate area in an electrolytic condenser is very large, compared with condensers having other dielectrics, because the film is so extremely thin — much less than any thickness that is practicable with a solid dielectric.

Voltage Breakdown

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

Since the dielectric must be thick to withstand high voltages, and since the thicker the dielectric the smaller the capacitance for a given plate area, a high-voltage condenser must have more plate area than a low-voltage condenser of the same capacitance. High-voltage high-capacitance condensers are physically large.

CONDENSERS IN SERIES AND PARALLEL

The terms "parallel" and "series" when used with reference to condensers have the same circuit meaning as with resistances. When a number of condensers are connected in parallel, as in Fig. 2-10, the total capacitance of the group is equal to the sum of the individual capacitances, so

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

However, if two or more condensers are connected in series, as in the second drawing,

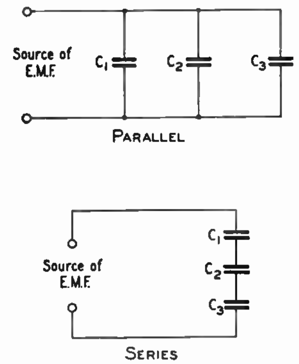


Fig. 2-10 — Condensers in series and parallel.

the total capacitance is less than that of the smallest condenser in the group. The rule for finding the capacitance of a number of series-connected condensers is the same as that for finding the resistance of a number of *parallel*-connected resistors. That is,

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

and, for only two condensers in series,

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

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

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

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

Example: Three condensers having capacitances of 1, 2 and 4 μfd., respectively, are con-

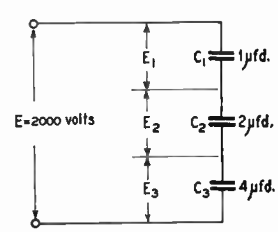


Fig. 2-11 — An example of condensers connected in series. The solution to this arrangement is worked out in the text.

nected in series as shown in Fig. 2-11. The total capacitance is

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

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

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

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

Similarly, the voltages across C_2 and C_3 are

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

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

totaling approximately 2000 volts, the applied voltage.

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

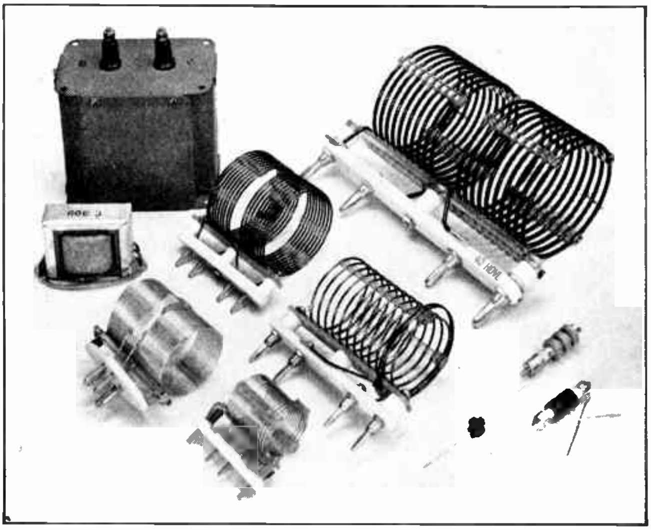
Inductance

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.

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

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

The induced e.m.f. (sometimes called **back e.m.f.**) tends to send a current through the circuit in the *opposite* direction to the current that flows because of the external e.m.f. so long as the latter current is *increasing.* However, if the current caused by the applied e.m.f. *decreases*, the induced e.m.f. tends to send current through the circuit in the *same* direction as the current from the applied e.m.f. The effect of inductance, therefore, is to oppose any *change* in the current flowing in the circuit, regardless of the nature of the change. It accomplishes this by storing energy in its magnetic field when the current in the circuit is being increased, and by releasing the stored energy when the current is being decreased.



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

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

Inductance Formula

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

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

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

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

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

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

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

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

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.

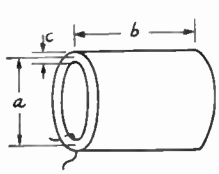


Fig. 2-12 — Coil dimensions used in the inductance formula.

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

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

IRON-CORE COILS

Permeability

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

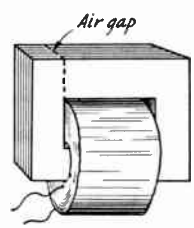


Fig. 2-13 — Typical construction of an iron-core coil. The small air gap prevents magnetic saturation of the iron and increases the inductance at high currents.

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 coil is highly dependent upon the current flowing in the coil. In an air-core coil, the inductance is independent of current because air does not "saturate."

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

Eddy Currents and Hysteresis

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

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

Eddy-current and hysteresis losses in iron increase rapidly as the frequency of the alternating current is increased. For this reason, we can use ordinary iron cores only at power and audio frequencies — up to, say, 15,000 cycles. Even so, a very good grade 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 the air surrounding the coil, the slug is quite effective in increasing the coil inductance. By pushing the slug in and out of the coil the inductance can be varied over a considerable range.

INDUCTANCES IN SERIES AND PARALLEL

When two or more inductance coils (or **inductors**, as they are frequently called) are connected in series (Fig. 2-14, left) the total induc-

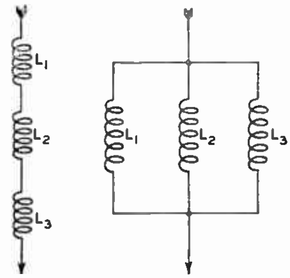


Fig. 2-14 — **Inductances in series and parallel.**

tance is equal to the sum of the individual inductances, *provided the coils are sufficiently separated so that no coil is in the magnetic field of another.* That is,

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

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

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

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

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

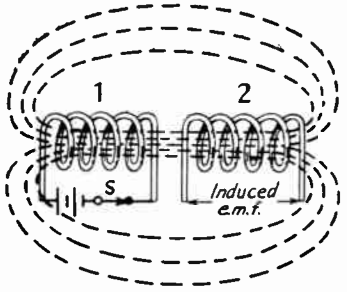


Fig. 2-15 — Mutual inductance. When the switch, *S*, is closed current flows through coil No. 1, setting up a magnetic field that induces an e.m.f. in the turns of coil No. 2.

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. Coils that have nearly the maximum possible mutual inductance are said to be **closely,** or **tightly,** coupled, but if the mutual inductance is relatively small the coils are said to be **loosely** coupled. The degree of coupling depends upon the physical spacing between the coils and how they are placed with respect to each other. Maximum coupling exists when they have a common axis 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

In Fig. 2-16A a battery having an e.m.f., *E*, a switch, *S*, a resistor, *R*, and condenser, *C*, are connected in series. Suppose for the moment that *R* is short-circuited and that there is no other resistance in the circuit. If *S* is now closed, condenser *C* will charge *instantly* to the battery voltage; that is, the electrons that constitute the charge redistribute themselves in a time interval so small that it can be considered to be zero. For just this instant, therefore, a very large current flows in the circuit, because all the electricity needed to charge the condenser has moved from the battery to the condenser at an extremely high rate.

When the resistance *R* is put into the circuit the condenser no longer can be charged instantaneously. If the battery e.m.f. is 100 volts, for example, and *R* is 10 ohms, the maximum current that can flow is 10 amperes, and even this much can flow only at the instant the switch is closed. But as soon as *any* current flows, condenser *C* begins to acquire a charge, which means that the voltage between the condenser plates rises. Since the upper plate (in Fig. 2-16A) will be positive and the lower negative, the voltage on the condenser tries to send a current through the circuit in the opposite direction to the current from the battery. Immediately after the switch is closed, therefore, the current drops below its

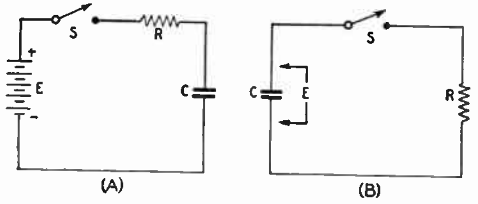


Fig. 2-16 — Schematics illustrating the time constant of an *RC* circuit.

initial Ohm's Law value, and as the condenser continues to acquire charge and its potential or e.m.f. rises, the current becomes smaller and smaller.

The length of time required to complete the charging process depends upon the capacitance of the condenser and the resistance in the circuit. Theoretically, the charging process is never really finished, but eventually the 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 condenser to reach 63 per cent of the applied e.m.f. (this figure is chosen for mathematical reasons). The voltage across the condenser rises logarithmically, as shown by Fig. 2-17.

The formula for time constant is

$$T = CR$$

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

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

Example: The time constant of a 2-μfd. condenser and a 250,000-ohm resistor is

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

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

If a charged condenser is *discharged* through a resistor, as indicated in Fig. 2-16B, the same time constant applies. If there were no resistance, the condenser would discharge instantly when *S* was closed. However, since *R* limits the current flow the condenser voltage cannot instantly go to zero, but it will decrease just as rapidly as

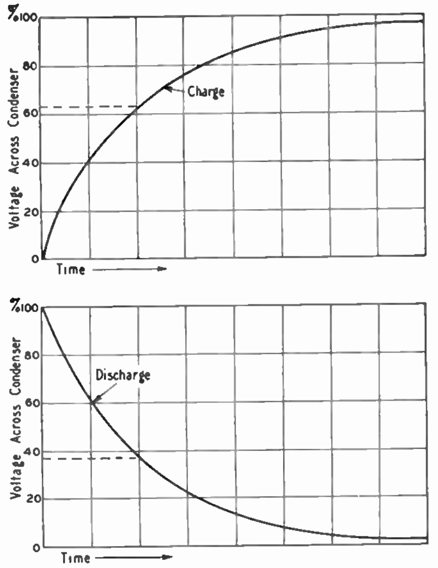


Fig. 2-17 — How the voltage across a condenser rises, with time, when a condenser is charged through a resistor. The lower curve shows the way in which the voltage decreases across the condenser terminals on discharging through the same resistor.

the condenser can rid itself of its charge through *R*. When the condenser is discharging through a resistance, the time constant (calculated in the same way as above) is the time, in seconds, that it takes for the condenser to *lose* 63 per cent of its voltage; that is, for the voltage to drop to 37 per cent of its initial value.

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

Inductance and Resistance

A comparable situation exists when resistance and inductance are in series. In Fig. 2-18, first consider *L* to have no resistance 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. In such a circuit the current is small at first, just as in the case without resistance. But as the current increases the voltage drop across *R* becomes larger. The back e.m.f. generated in *L* has only to equal the *difference* between *E* and the drop across *R*, because that difference is the voltage actually applied to *L*. This difference becomes smaller as the current approaches the final Ohm's Law value. Theoretically, the back e.m.f. never quite disappears (that is, the current never quite reaches the Ohm's Law value) but practically it becomes unmeasurable after a time. The difference between the actual current and the Ohm's Law value also becomes undetectable. The time 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.

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

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

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

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

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

An inductor cannot be discharged in the same way as a condenser, because the magnetic field disappears as soon as current flow ceases. Opening *S* does not leave the inductor

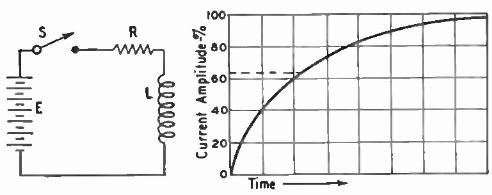


Fig. 2-18 — Time constant of an *LR* circuit.

"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.

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. When a baseball pitcher throws the ball to the catcher there is a definite interval, represented by the time of flight of the ball, between the act of throwing and the act of catching. The throwing and catching are "out of phase" because they do not occur at exactly the same time.

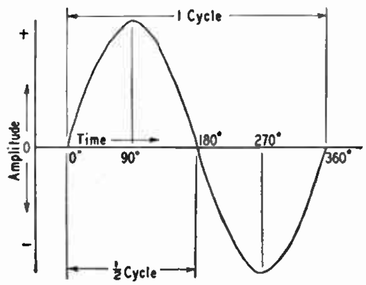


Fig. 2-19 — An a.c. cycle is divided off into 360 degrees that are used as a measure of time or phase.

Simply saying that two events are out of phase does not tell us which one occurred first. To give this information, the later event is said to **lag** the earlier, while the one that occurs first is said to **lead**. Thus, throwing the ball "leads" the catch, or the catch "lags" the throw.

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. If there are two or more frequencies, the measurement of phase has to be modified just as the measurements of two lengths must be reconciled if one is given in feet and the other in meters.

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. 2-19 should help make this method of measurement clear.

Measuring Phase

To compare the phase of two currents of the same frequency, we measure between corresponding parts of cycles of the two currents. This is shown in Fig. 2-20. The current labeled *A* leads the one marked *B* by 45 degrees, since *A*'s cycles begin 45 degrees sooner in time. It is equally correct to say that *B lags A* by 45 degrees.

Two important special cases are shown in Fig. 2-21. In the upper drawing *B* lags 90 degrees behind *A*; that is, its cycle begins just one-quarter cycle later than that of *A*. When one wave is passing through zero, the other is just at its maximum point.

In the lower drawing *A* and *B* are 180 degrees out of phase. In this case it does not matter which one is 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. 2-20 and 2-21 could represent current, voltage, or both. *A* and *B* might be two currents in separate circuits, or *A* might represent voltage while *B* represented current in the same circuit. If *A* and *B* represent two currents in the *same* circuit (or two voltages in the same circuit) the 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

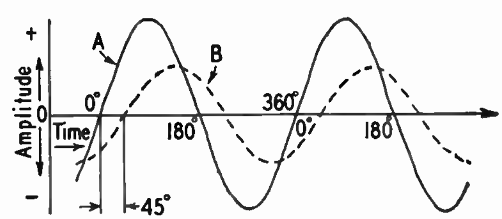


Fig. 2-20 — When two waves of the same frequency start their cycles at slightly different times, the time difference or phase difference is measured in degrees. In this drawing wave *B* starts 45 degrees (one-eighth cycle) later than wave *A*, and so lags 45 degrees behind *A*.

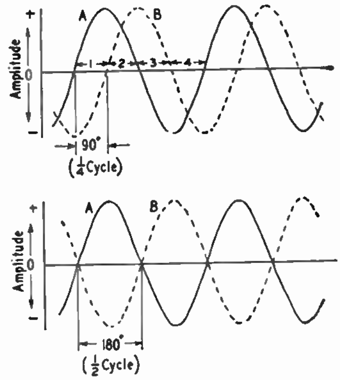


Fig. 2-21 — Two important special cases of phase difference. In the upper drawing, the phase difference between *A* and *B* is 90 degrees; in the lower drawing the phase difference is 180 degrees.

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 Condensers

Suppose a sine-wave a.c. voltage is applied to a condenser in a circuit containing no resistance, as indicated in Fig. 2-22. In the period *OA*, the applied voltage increases from zero to 38 volts; at the end of this period the condenser 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.

Thus as the instantaneous value of the applied voltage increases the current decreases.

By dividing the first quarter cycle into a very large number of intervals it could be shown that the current charging the condenser has the shape of a sine wave, just as the applied voltage does. The current is largest at the beginning of the cycle and becomes zero at the maximum value of the voltage (the condenser cannot be charged to a higher voltage than the maximum applied, so no further current can flow) so there is a phase difference of 90 degrees between the voltage and current. During the first quarter cycle of the applied voltage the current is flowing in the normal way through the circuit, since the condenser is being charged. Hence the current is positive during this first quarter cycle, as indicated by the dashed line in Fig. 2-22.

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

The third and fourth quarter cycles repeat the events of the first and second, respectively, with this difference — the polarity of the applied voltage has reversed, and the current changes to correspond. In other words, *an alternating current flows "through" a condenser when an a.c. voltage is applied to it.* (Actually, current never flows "through" a condenser. It flows in the associated circuit because of the alternate charging and discharging of the capacitance.) As shown by Fig. 2-22, the current starts its cycle 90 degrees before the voltage, so the current in a condenser *leads the applied voltage by 90 degrees.*

Capacitive Reactance

The amount of charge that is alternately stored in and released from the condenser is proportional to the applied voltage and the capacitance. Consequently, the current in the circuit will be proportional to both these quantities, since current is simply the rate at which charge is moved. The current also will be proportional to the frequency

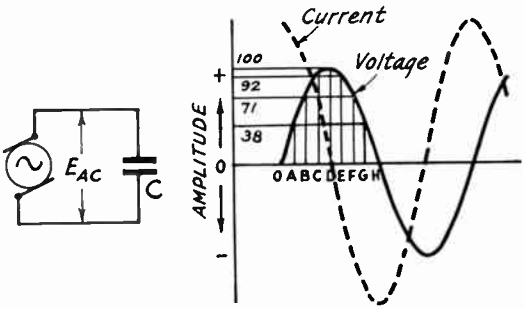


Fig. 2-22 — Voltage and current phase relationships when an alternating voltage is applied to a condenser.

of the a.c. voltage, because the same charge is being moved back and forth at a rate that is proportional to the number of cycles per second.

The fact that the current is proportional to the applied voltage is important, because it is the same thing that Ohm's Law says about current flow in a resistive circuit. That being the case, there must be something in the condenser that corresponds in a general way to resistance — something that tends to limit the current that can flow when a given voltage is applied. The "something" clearly must include the effect of capacitance and frequency, since these also affect the amount of current that flows. It is called **reactance**, and its relationship to capacitance and frequency is given by the formula

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

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

Reactance and resistance are not the same thing, but because they have a similar current-limiting effect the same unit, the ohm, is used for both. Unlike resistance, reactance does not consume or dissipate power. The energy stored in the condenser 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 condenser of 470 μμfd. (0.00047 μfd.) at a frequency of 7150 kc. (7.15 Mc.) is

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

Inductive Reactance

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

Since the value of the induced e.m.f. is proportional to the rate at which the current changes, a small current changing rapidly (that is, at a high frequency) can generate a large back e.m.f. in a given inductance just as well as a large current changing slowly (low frequency). Consequently, the current that flows through a given inductance will decrease as the frequency is raised, if the applied e.m.f. is held constant. Also, when the applied voltage and frequency are fixed, the value of current required becomes less as the inductance is made larger, because the induced e.m.f. also is proportional to inductance.

When the frequency and inductance are constant but the applied e.m.f. is varied, the necessary rate of current change (to induce the proper back e.m.f.) can be obtained only if the amplitude of the current is directly proportional to the voltage. This is Ohm's Law again, and again the current-limiting effect is similar to, but not identical with, the effect of resistance. It is called **inductive reactance** and, like capacitive reactance, is measured in ohms. There is no energy loss in inductive reactance; the energy is stored in the magnetic field in one quarter cycle and then returned to the circuit in the next.

The formula for inductive reactance is

$$X_L = 2\pi fL$$

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

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

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

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

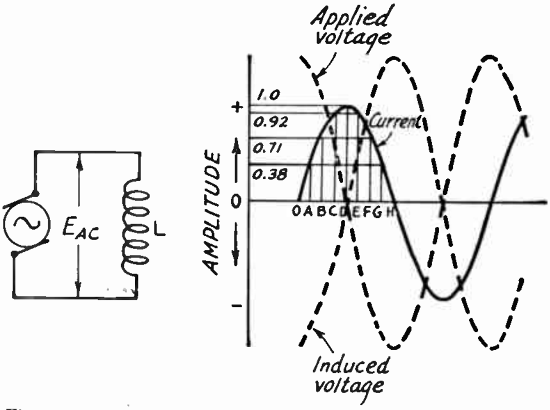


Fig. 2-23 — Phase relationships between voltage and current when an alternating voltage is applied to an inductance.

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

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

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

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

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

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

When the circuit consists of an inductance in series with a capacitance, the same current flows through both reactances. However, the voltage across the coil *leads* the current by 90 degrees, and the voltage across the condenser *lags* behind the current by 90 degrees. The coil and condenser voltages therefore are 180 degrees out of phase.

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

The "equivalent" or total reactance of any circuit containing inductive and capacitive reactances in series is equal to $X_L - X_C$. If there are several coils and condensers in series, simply add up all the inductive reactances, then add up all the capacitive reactances, and then subtract the latter from the former. It is customary to call inductive reactance "positive" and capacitive reactance "negative." If the equivalent or net reactance is positive, the voltage leads the current by 90 degrees; if the net reactance is negative, the voltage lags the current by 90 degrees.

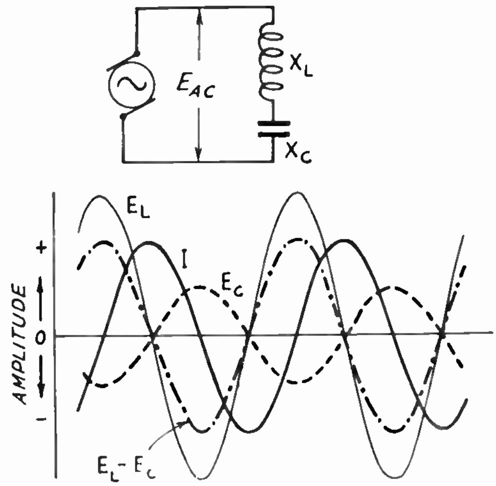


Fig. 2-24 — Current and voltages in a circuit having inductive and capacitive reactances in series.

Reactive Power

In Fig. 2-24 the voltage drop across the coil 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 coil's magnetic field, energy is being returned to the circuit from the condenser's electric field, and vice versa. This stored energy is responsible for the fact that the voltages across reactances in series can be larger than the voltage applied to them.

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** instead of the watt. Reactive power is sometimes called "wattless" power.

● IMPEDANCE

The fact that resistance, inductive reactance and capacitive reactance all are measured in ohms does not indicate that they can be combined indiscriminately. Voltage and current are in phase in resistance, but differ in phase by a quarter cycle in reactance. In the simple circuit shown

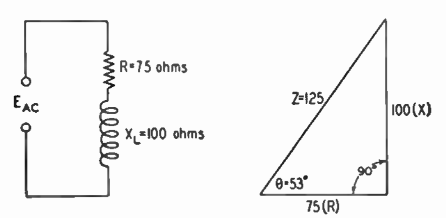


Fig. 2-25 — Resistance and inductive reactance connected in series.

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

The term "impedance" also is generalized to include any quantity that can be expressed as a ratio of voltage to current. Pure resistance and pure reactance are both included in "impedance" in this sense. A circuit with **resistive impedance** is either one with resistance alone or one in which the effects of any reactance present have been eliminated. Similarly, a **reactive impedance** is one having reactance only. A **complex impedance** is one in which both resistance and reactance effects are observable.

It can be shown that resistance and reactance can be combined in the same way that a right-angled triangle is constructed, if the resistance is laid off to proper scale as the base of the triangle and the reactance is laid off as the altitude to the same scale. This is also indicated in Fig. 2-25. When this is done the hypotenuse of the triangle represents the impedance of the circuit,

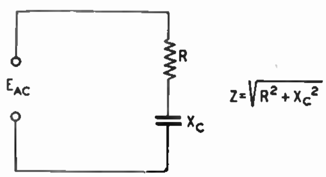


Fig. 2-26 — Resistance and capacitive reactance in series.

to the same scale, and the angle between Z and R (usually called θ and so indicated in the drawing) is equal to the phase angle between the applied e.m.f. and the current. By geometry,

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

In the case shown in the drawing,

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

The phase angle can be found from simple trigonometry. Its tangent is equal to X/R; in this case $X/R = 100/75 = 1.33$. From trigonometric tables it can be determined that the angle having a tangent equal to 1.33 is approximately 53 degrees. In ordinary amateur work it is seldom necessary to give much consideration to the phase angle.

A circuit containing resistance and capacitance in series (Fig. 2-26) can be treated in the same way. The difference is that in this case the current *leads* the applied e.m.f., while in the resistance-inductance case it *lags* behind the voltage.

If either X or R is small compared with the other (say 1/10 or less) the impedance is very nearly equal to the larger of the two quantities. For example, if $R = 1$ ohm and $X = 10$ ohms,

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

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

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

Ohm's Law for Impedance

Ohm's Law can be applied to circuits containing impedance just as readily as to circuits having resistance or reactance only. The formulas are

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

$$E = IZ$$

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

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

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

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

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

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

The arithmetical sum of these voltages is greater than the applied voltage. However, the *actual* sum of the two when the phase relationship is taken into account is equal to 250 volts r.m.s., as shown by Fig. 2-27, where the instantaneous values are added throughout the cycle. Whenever resistance and reactance are in series, the

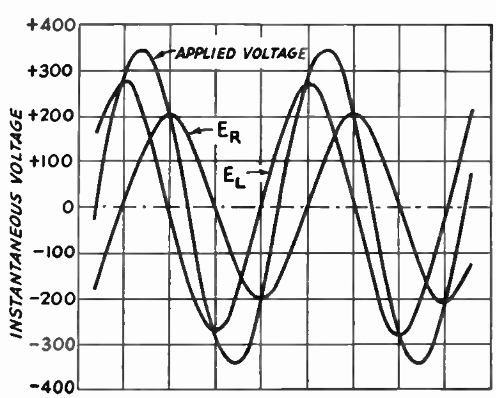


Fig. 2-27 — Voltage drops around the circuit of Fig. 2-25. Because of the phase relationships, the applied voltage is less than the arithmetical sum of the drops across the resistor and inductor.

individual voltage drops always add up, arithmetically, to more than the applied voltage. There is nothing fictitious about these voltage drops; they can be measured readily by suitable instruments. It is simply an illustration of the importance of phase in a.c. circuits.

A more complex series circuit, containing resistance, inductive reactance and capacitive reactance, is shown in Fig. 2-28. In this case it is necessary to take into account the fact that the phase angles between current and voltage differ

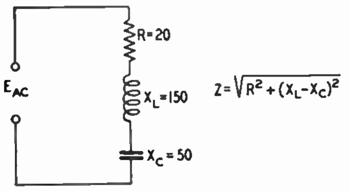


Fig. 2-28 — Resistance, inductive reactance, and capacitive reactance in series.

in all three elements. Since it is a series circuit, the current is the same throughout. Considering first just the inductance and capacitance and neglecting the resistance, the net reactance is

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

Thus the impedance of a circuit containing resistance, inductance and capacitance in series is

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

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

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

$$= \sqrt{(20)^2 + (150 - 50)^2} = \sqrt{(20)^2 + (100)^2}$$

$$= \sqrt{10{,}400} = 102 \text{ ohms}$$

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

Parallel Circuits

Suppose that a resistor, condenser and coil are connected in parallel as shown in Fig. 2-29 and an a.c. voltage is applied to the combination. In any one branch, the current will be unchanged if one or both of the other two branches is disconnected, so long as the applied voltage remains unchanged. Hence the current in each branch can be calculated quite simply by the Ohm's Law formulas given in the preceding sections. The total current, I, is the sum of the currents through all three branches — not the arithmetical sum, but the sum when phase is taken into account.

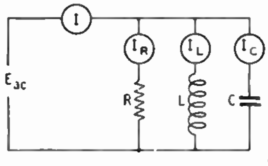


Fig. 2-29 — Resistance, inductance and capacitance in parallel. Instruments connected as shown will read the total current, I, and the individual currents in the three branches of the circuit.

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

The impedance looking into the parallel circuit from the source of voltage is equal to the applied

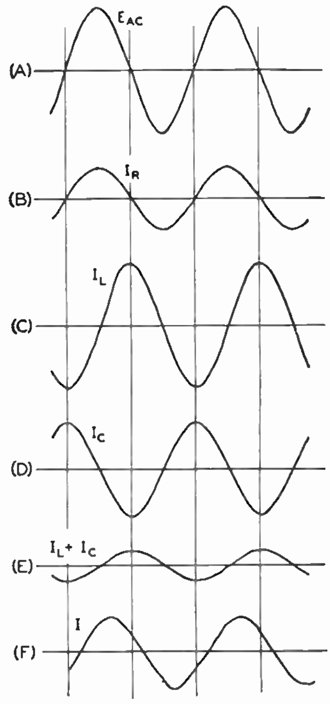


Fig. 2-30 — Phase relationships between branch currents and applied voltage for the circuit of Fig. 2-29. The total current through L and C in parallel ($I_L + I_C$) and the total current in the entire circuit (I) also are shown.

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

X_C are quite different the net reactive current can be relatively large and the total current also will be appreciably larger than I_R. In such a case the circuit impedance will be lower than the resistance of R alone.

Power Factor

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

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

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

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

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

Complex Waves

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

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

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

Transformers

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

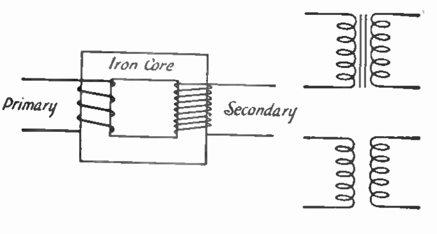


Fig. 2-31 — The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

Voltage and Turns Ratio

For a given varying magnetic field, the voltage induced in a coil in the field will be proportional to the number of turns on the coil. If the two coils of a transformer are in the same field (which is the case when both are wound on the same closed core) it follows that the induced voltages will be proportional to the number of turns on each coil. In the primary the induced voltage is practically equal to, and opposes, the applied voltage. 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 **turns ratio** of the transformer.

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

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

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

Either winding of a transformer can be used as the primary, *providing* the winding has enough turns (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 of 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.

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

$$P_o = nP_i$$

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

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

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

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

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 and raise the operating temperature. There is a limit to the temperature rise that can be tolerated, because too-high

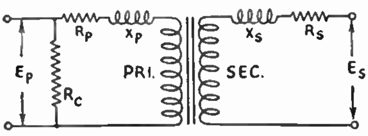


Fig. 2-32 — The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance *Rc* is an equivalent resistance representing the constant core losses. Since these are comparatively small, their effect may be neglected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

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.**

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 looks 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 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 transform the actual load into an impedance of the desired value. This is called **impedance matching.** From the preceding,

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

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

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

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

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

Impedance matching means, in general, adjusting the load impedance — by means of a transformer or otherwise — to a desired value. However, there is also another meaning. It is possible to show that any source of power will have its maximum possible output when the impedance of the load is equal to the internal impedance of the source. The impedance of the source is said to be "matched" under this condition. 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.

Transformer Construction

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

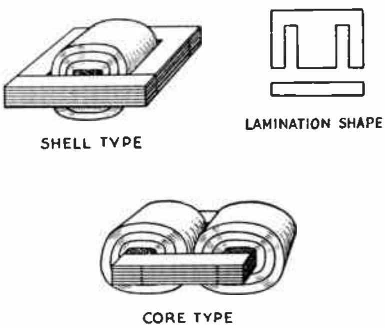


Fig. 2-33 — Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path with as low reluctance as possible.

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

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

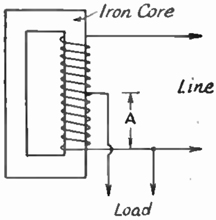


Fig. 2-34 — The autotransformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding (*A*) flow in opposite directions, so that the resultant current is the difference between them. The voltage across *A* is proportional to the turns ratio.

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

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

Autotransformers

The transformer principle can be utilized with only one winding instead of two, as shown in Fig. 2-34; the principles just discussed apply equally well. A one-winding transformer is called an **autotransformer.** The current in the common section (A) of the winding is the difference between the line (primary) and the load (secondary) currents, since these currents are out of phase. Hence if the line and load currents are nearly equal the common section of the winding may be wound with comparatively small wire. This 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.

Radio-Frequency Circuits

● RESONANCE

Fig. 2-35 shows a resistor, condenser and coil 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 condenser reactance will be much larger than the resistance of *R*, and the inductive reactance will be small compared with either the reactance of *C* or the resistance of *R*. (*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 case the current will be small, because the reactance is large at either low or high frequencies.

At some intermediate frequency, the reactances of *C* and *L* will be equal and the voltage drops across the coil and condenser 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 can occur 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 ($\mu\mu$fd.)
π = 3.14

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

$$= \frac{10^6}{2\pi\sqrt{LC}} = \frac{10^6}{6.28 \times \sqrt{5 \times 35}}$$
$$= \frac{10^6}{6.28 \times 13.2} = \frac{10^9}{83} = 12{,}050 \text{ kc.}$$

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

Resonance Curves

If a plot is drawn of the current flowing in the circuit of Fig. 2-35 as the frequency is varied (the applied voltage being constant) it would look like one of the curves in Fig. 2-36. The shape of the **resonance curve** at frequencies near resonance is determined by the ratio of reactance to resistance at the particular frequency considered.

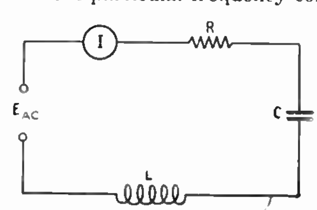


Fig. 2-35 — A series circuit containing *L*, *C* and *R* is "resonant" at the applied frequency when the reactance of *C* is equal to the reactance of *L*.

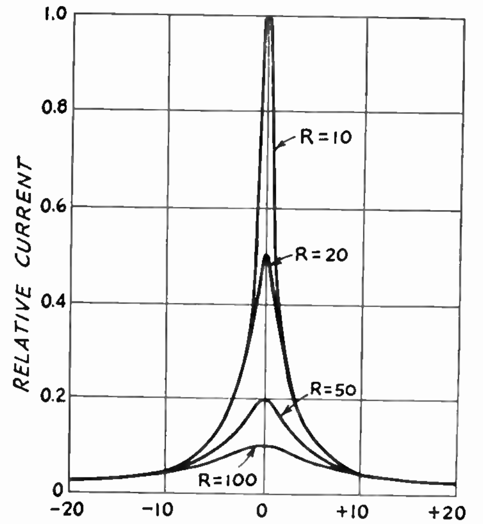


Fig. 2-36 — Current in a series-resonant circuit with various values of series resistance. The values are arbitrary and would not apply to all circuits, but represent a typical case. It is assumed that the reactances (at the resonant frequency) are 1000 ohms (minimum Q = 10). Note that at frequencies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

If the reactance of either the coil or condenser is of the same order of magnitude as the resistance, the current decreases rather slowly as the frequency is moved in either direction away from resonance. Such a curve is said to be **broad.** On the other hand, if the reactance is considerably larger than the resistance the current decreases rapidly as the 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.

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.

Q

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

The value of the reactance of either the coil or condenser at the resonant frequency, divided by the resistance in the circuit, is called the Q (quality factor) of the circuit, or

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

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

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

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

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

Voltage Rise

When a voltage of the resonant frequency is inserted in series in a resonant circuit, the voltage that appears across either the coil or condenser is considerably higher than the applied voltage. The current in the circuit is limited only by the actual resistance of the coil-condenser combination in the circuit and may have a relatively high value; however, the same current

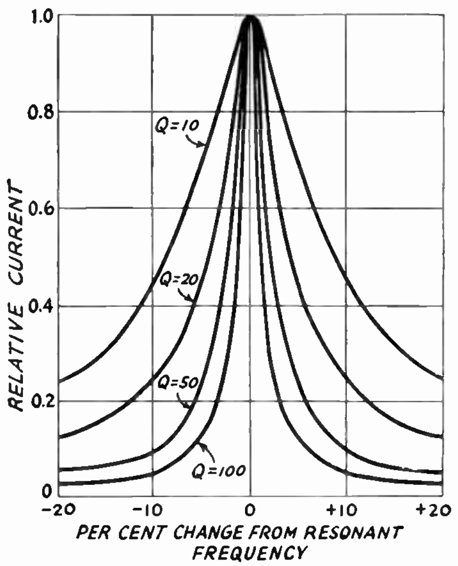


Fig. 2-37 — Current in series-resonant circuits having different Qs. In this graph the current at resonance is assumed to be the same in all cases. The lower the Q, the more slowly the current decreases as the applied frequency is moved away from resonance.

flows through the high reactances of the coil and condenser 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 the Q of the circuit. Therefore, the voltage across either the coil or condenser is equal to Q times the voltage inserted in series with the circuit.

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

Parallel Resonance

When a variable-frequency source of constant voltage is applied to a parallel circuit of the type shown in Fig. 2-38 there is a resonance effect

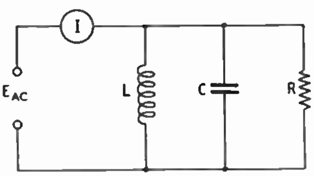


Fig. 2-38 — Circuit illustrating parallel resonance.

similar to that in a series circuit. However, in this case the current (measured at the point indicated) is *smallest* at the frequency for which the coil and condenser reactances are equal. At that frequency the current through L is exactly canceled by the out-of-phase current through C, 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.

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

Parallel and series resonant circuits are quite alike in some respects. For instance, the circuits given at A and B in Fig. 2-39 will behave identically, when an external voltage is applied, if (1) L and C are the same in both cases; and (2) R_p

multiplied by R_s equals the square of the reactance (at resonance) of either L or C. When these conditions are met the two circuits will have the same Qs. (These statements are approximate, but are quite accurate if the Q is 10 or more.) 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. 2-39A has an equivalent **parallel impedance** (at resonance) equal to R_p, the relationship between R_s and R_p being as explained above. Although R_p is not an actual resistor, to the source of voltage the parallel-resonant circuit "looks like" a pure resistance of that value. It is "pure" resistance because the coil and condenser currents are 180 degrees out of phase and are equal; thus there is no reactive current in the line. At the resonant frequency the parallel impedance of a resonant circuit is

$$Z_r = QX$$

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

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

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

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

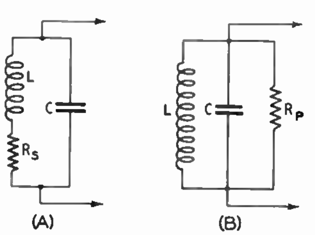


Fig. 2-39 — Series and parallel equivalents when the two circuits are resonant. The series resistor, R_s, in A can be replaced by an equivalent parallel resistor, R_p, in B, and vice versa.

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

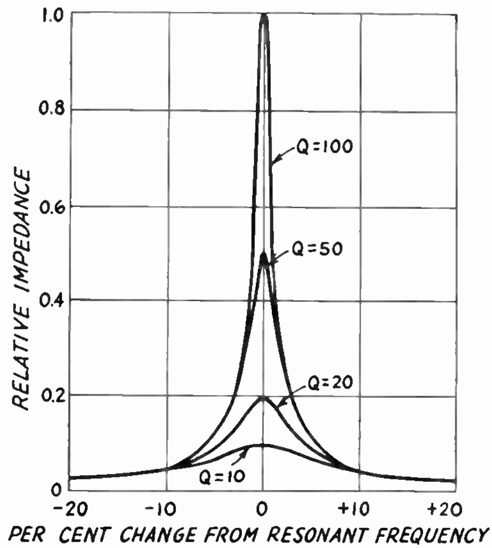


Fig. 2-40 — Relative impedance of parallel-resonant circuits with different Qs. These curves are similar to those in Fig. 2-37 for current in a series-resonant circuit. The effect of Q on impedance is most marked near the resonant frequency.

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. 2-39A, 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 in tuning 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 on 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.

However, when the circuit delivers energy to a load (as in the case of the resonant circuits used in transmitters) the energy consumed in the circuit itself is usually negligible compared with that consumed by the load. The equivalent of such a circuit is shown in Fig. 2-41A, where the parallel resistor represents the load to which power is delivered. If the power dissipated in the

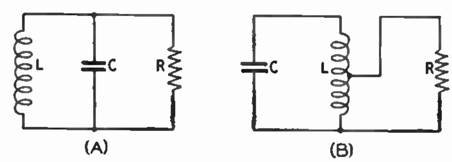


Fig. 2-41 — The equivalent circuit of a resonant circuit delivering power to a load. The resistor R represents the load resistance. At B the load is tapped across part of L, which by transformer action is equivalent to using a higher load resistance across the whole circuit.

load is at least ten times as great as the power lost in the coil and condenser, the parallel impedance of the resonant circuit itself will be so high compared with the resistance of the load that for all practical purposes the impedance of the combined circuit is equal to the load resistance. Under these conditions the Q of a parallel-resonant circuit loaded by a resistive impedance is

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

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

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

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

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

Impedance Transformation

An important application of the parallel-resonant circuit is an impedance-matching device in the output circuit of a vacuum-tube r.f. power amplifier. As described in the chapter 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 resistance to the desired value the load may be tapped across part of the coil, as shown in Fig. 2-41B. This is equivalent to connecting a higher value of load resistance across the whole circuit, and is similar in principal 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. 2-39A, 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 impedance at resonance
X = Reactance (in ohms) of either the coil or condenser
R = Load resistance inserted in series

If the Q is lower than 10 the reactance will have to be adjusted somewhat, as described previously, to obtain a resistive impedance of the desired value.

L/C Ratio

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

LC Constants

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 (μμfd.)
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 μμfd. The LC constant is

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

With 25 μμfd. $L = 1900/C = 1900/25 = 76$ μh.
50 μμfd. $L = 1900/C = 1900/50 = 38$ μh.
100 μμfd. $L = 1900/C = 1900/100 = 19$ μh.
500 μμfd. $L = 1900/C = 1900/500 = 3.8$ 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 variations of this type of coupling shown at A, B and C of Fig. 2-42, utilize a common inductance, capacitance and resistance, respectively. Current circulating in one LC branch flows through the common element (L_c, C_c or R_c) and the voltage developed across this element causes current to flow in the other LC branch.

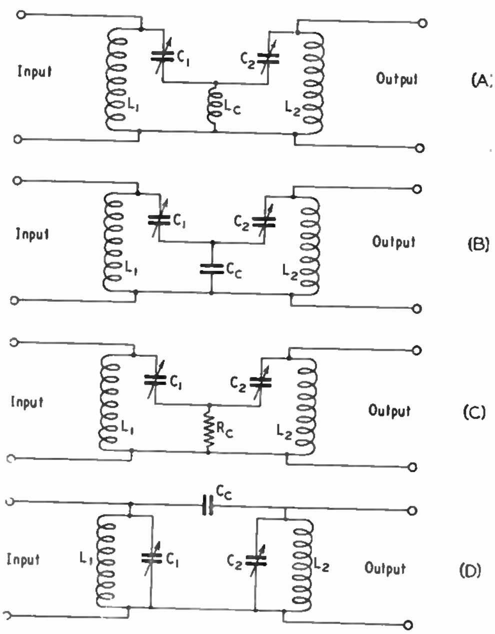


Fig. 2-42 — Four methods of circuit coupling.

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

Capacitive Coupling

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

Inductive Coupling

Figs. 2-43 and 2-44 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 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. 2-43. 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. 2-41B.

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. 2-41B.

Coupled Resonant Circuits

When the primary and secondary circuits are both tuned, as in Fig. 2-44, 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

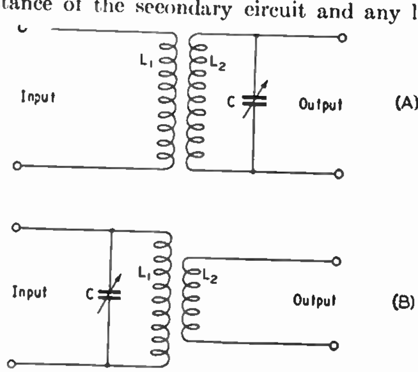


Fig. 2-43 — Single-tuned inductively-coupled circuits.

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 **critical coupling,** but then decreases if the coupling is tightened still more (still without changing the tuning).

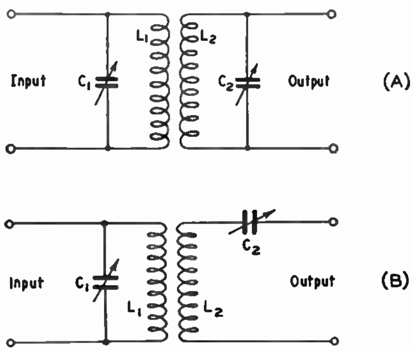


Fig. 2-44 — **Inductively-coupled resonant circuits. Circuit A is used for high-resistance loads (at least several times the reactance of either L_2 or C_2 at the resonant frequency). Circuit B is suitable for low resistance loads where the reactance of either L_2 or C_2 is at least several times the load resistance.**

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

With loaded circuits 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. 2-39, this circuit is in effect loaded by a parallel resistance (effect of coupled-in resistance). In the parallel-tuned secondary circuit, Fig. 2-44A, 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. 2-41). In the series-tuned secondary circuit, Fig. 2-44B, 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 Q of each circuit is at least 10. Smaller values will

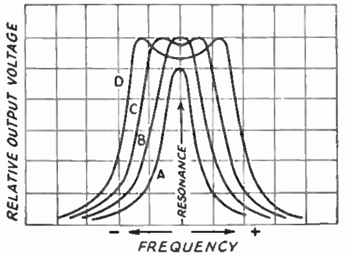


Fig. 2-45 — **Showing the effect on the output voltage from the secondary circuit of changing the coefficient of coupling between two resonant circuits independently tuned to the same frequency. The voltage applied to the primary is held constant in amplitude while the frequency is varied, and the output voltage is measured across the secondary.**

suffice if the coil construction permits tight coupling.

Selectivity

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

In Fig. 2-44, the selectivity is the same as that of a single tuned circuit having a Q equal to the *product* of the Qs of the individual circuits — *if* the coupling is well below critical and both circuits are tuned to resonance. The Qs of the individual circuits are affected by the degree of coupling, because each couples resistance into the other; the tighter the coupling, the lower the individual Qs and therefore the lower the over-all selectivity.

If both circuits are independently tuned to resonance, the over-all selectivity will vary about as shown in Fig. 2-45 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 the humps represent a new condition of critical coupling, at a frequency to which the primary is detuned by the 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 *Q*s 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 *Q*s 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 adjust the two circuits, by experiment, to give the desired performance.

Link Coupling

A modification of inductive coupling, called **link coupling,** is shown in Fig. 2-46. 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.

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.

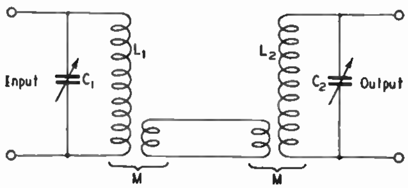


Fig. 2-46 — **Link coupling. The mutual inductances at both ends of the link are equivalent to mutual inductance between the tuned circuits, and serve the same purpose.**

Piezoelectric Crystals

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

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

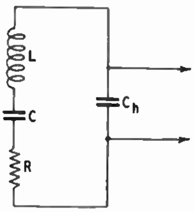


Fig. 2-47 — **Equivalent circuit of a crystal resonator.** ***L***, ***C*** **and** ***R*** **are the electrical equivalents of mechanical properties of the crystal;** C_h **is the capacitance of the electrodes with the crystal plate between them.**

Crystalline plates also are mechanical vibrators that have natural frequencies of vibration ranging from a few thousand cycles to several megacycles per second. The vibration frequency depends on the kind of crystal, the way the plate is cut from the natural crystal, and on the dimensions of the plate. Because of the piezoelectric effect, the crystal plate can be coupled to an electrical circuit and made to substitute for a coil-and-condenser resonant circuit. The thing that makes the **crystal resonator** valuable is that it has extremely high *Q*, ranging from 5 to 10 times the *Q*s 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 electrodes between which it is sandwiched; these electrodes form, with the crystal as the dielectric, a small condenser like any other condenser constructed of two plates with a dielectric between. The crystal itself is equivalent to a series-resonant circuit, and together with the capacitance of the electrodes forms the equivalent circuit shown in Fig. 2-47. The equivalent inductance of the crystal is extremely large and the series capacitance, *C*, is correspondingly low; this is the reason for the high *Q* of a crystal. The electrode capacitance, C_h, is so very large compared with the series capacitance of the crystal that it has only a very small effect on the resonant frequency.

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

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 super-high 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. 2-18. 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.

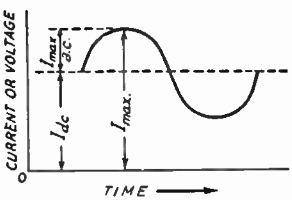


Fig. 2-48 — Pulsating, composed of an alternating current or voltage superimposed on a steady direct current or voltage.

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 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.

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

Series and Parallel Feed

Fig. 2-49 shows in simplified form how d.c. and a.c. may be combined in a vacuum-tube circuit. (The tube is shown only in bare outline; so far as the d.c. is concerned, it can be looked upon as a resistance of rather high value. On the other hand, the tube may be looked upon as the *generator* of the a.c. The mechanism of tube operation is described in the next chapter.) In this case, it is assumed that the a.c. is at radio frequency, as suggested by the coil-and-condenser 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 condenser,** *C*, is placed in the circuit to prevent it. (Without *C*, the d.c. supply would be short-circuited by the low resistance of *L*.) On the other hand, the r.f. current generated by the tube can easily flow through *C* to the tuned circuit because the capacitance of *C* is intentionally chosen to have low reactance (compared with the impedance of the tuned circuit) at the radio frequency. The r.f. current cannot flow through the d.c. supply because the inductance of *RFC* is intentionally made so large that it has a very high reactance at the radio frequency. The resistance of *RFC*, however, is too low to have an appre-

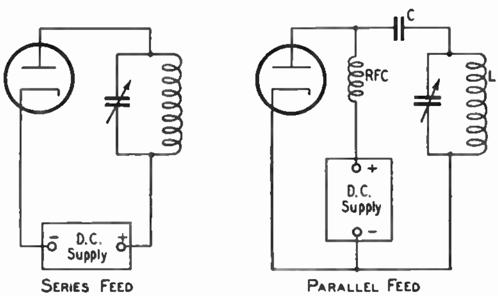


Fig. 2-49 — Illustrating series and parallel feed.

ciable 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 usually preferred, therefore, because it is relatively easy to keep the impedance between the a.c. circuit and the tube low.

By-Passing

In the series-feed circuit just discussed, it was assumed that the d.c. supply had very low impedance at radio frequencies. This is not likely to be true in a practical power supply, partly

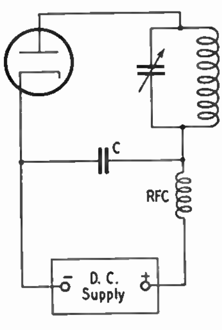


Fig. 2-50 — Typical use of a by-pass condenser in a series-feed circuit.

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 **by-pass condenser,** as shown in Fig. 2-50. Condenser *C* is chosen to have low reactance at the operating frequency, and is installed right in the circuit where it can be wired to the other parts with quite short connecting wires. 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 by-pass condenser should not be more than one-tenth of the impedance of the by-passed part of the circuit. Very often the latter impedance is not known, in which case it is desirable to use the largest capacitance in the by-pass that circumstances permit. To make doubly sure that r.f. current will not flow through a non-r.f. circuit such as a power supply, an r.f. choke may be connected in the lead to the latter, as shown in Fig. 2-50.

The same type of by-passing is used when audio frequencies are present in addition to r.f. Because the reactance of a condenser changes with frequency, it is readily possible to choose a capacitance that will represent a very low reactance at radio frequencies but that will have such high reactance at audio frequencies that it is practically an open circuit. A capacitance of 0.001 μfd. is practically a short circuit for r.f., for example, but is almost an open circuit at audio frequencies. (The actual value of capacitance that is usable will be modified by the impedances concerned.) By-pass condensers also are used in audio circuits to carry the audio frequencies around a d.c. supply.

Distributed Capacitance and Inductance

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

This **distributed inductance** in a condenser and the **distributed capacitance** in a coil have important practical effects. Actually, every condenser is a tuned circuit, resonant at the frequency where its capacitance and distributed inductance have the same reactance. The same thing is true of a coil and its distributed capacitance. At frequencies well below these **natural resonances,** the condenser will act like a normal capacitance and the coil will act like a normal inductance. Near the natural resonant points, the coil and condenser act like self-tuned circuits. Above resonance, the condenser acts like an inductance and the coil acts like a condenser. Thus there is a limit to the amount of capacitance that can be used at a given frequency. There is a similar limit to the inductance that can be used. At audio frequencies, capacitances measured in microfarads and inductances measured in henrys are practicable. At low and medium radio frequencies, inductances of a few millihenrys and capacitances of a few thousand micromicrofarads are the largest practicable. At high radio frequencies, usable inductance values drop to a few microhenrys and capacitances to a few hundred micromicrofarads.

Distributed capacitance and inductance are important not only in r.f. tuned circuits, but in by-passing and choking as well. It will be appreciated that a by-pass condenser that actually acts like an inductance, or an r.f. choke that acts like a condenser, cannot work as it is intended they should.

Grounds

Throughout this book there are frequent references to **ground** and **ground potential.** When a connection is said to be "grounded" it does not mean that it actually goes to earth (although in many cases such earth connections *are* used). What it means is that an actual earth connection *could* be made to that point in the circuit without disturbing the operation of the circuit in any way. The term also is used to indicate a "common" point in the circuit where power supplies and metallic supports (such as a metal chassis) are electrically tied together. It is customary, for example, to "ground" the negative terminal of a d.c. power supply, and to "ground" the filament or heater power supplies for vacuum tubes. Since the cathode of a vacuum tube is a junction point for grid and plate voltage supplies, it is a natural point to "ground." Also, since the various circuits connected to the tube elements have at least one point connected to cathode, these points also are "returned to ground." "Ground" is therefore a common reference point in the radio circuit. "Ground potential" means that there is no "difference of potential" — that is, no voltage — between the circuit point and the earth.

Single-Ended and Balanced Circuits

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

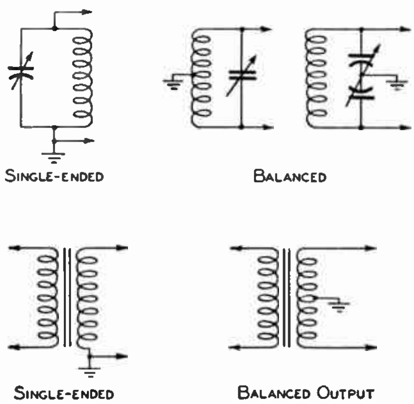


Fig. 2-51 — Single-ended and balanced circuits.

Typical single-ended and balanced circuits are shown in Fig. 2-51. 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" condenser and connecting the condenser rotor to ground. In the iron-core transformer, one or both windings may be tapped at the center of the winding to provide the ground connection.

In the single-ended circuit, only one side of the circuit is "hot" — that is, has a voltage that differs from ground potential. In the balanced circuit, both ends are "hot" and the grounded center point is at ground potential.

Shielding

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

Capacitive coupling may readily be prevented by enclosing one or both of the circuits in grounded low-resistance metallic containers, called **shields.** The electric field from the circuit components does not penetrate the shield. A metallic plate, called a **baffle shield,** inserted between two components also may suffice to prevent electrostatic coupling between them. It should be large enough to make the components invisible to each other.

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

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. The reduction in inductance and Q will be small if the shield is sufficiently far away from the coil; the spacing between the sides of the coil and the shield should be at least half the coil diameter, and the spacing at the ends of the coil should at least equal the coil diameter. The higher the conductivity of the shield material, the less the effect on the inductance and Q. Copper is the best material, but aluminum is quite satisfactory.

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

Modulation, Heterodyning and Beats

Since one of the most widespread uses of radio frequencies is the 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 some frequency around 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. Unfortunately, no simple method for doing such a thing directly has ever been devised, although the *effect* is obtained and used in some 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. Fig. 2-52A 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

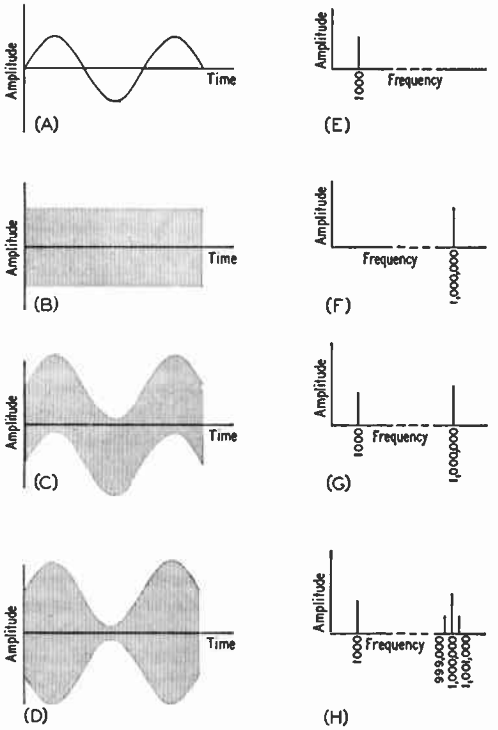


Fig. 2-52 — Amplitude-*vs.*-time and amplitude-*vs.*-frequency plots of various signals. (A) 1½ cycles of a 1000-cycle signal. (B) A 1,000,000-cycle signal plotted to the same scale as A. Because there are 1500 cycles during this time, they cannot be shown accurately. (C) The signals of A and B flowing in the same circuit. (D) The signals of A and B combined in a circuit where A can control the amplitude of B. The 1,000,000-cycle signal is *modulated* by the 1000-cycle signal. (E), (F), (G), (H) Amplitude-*vs.*-frequency plots of the signals in A, B, C and D.

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 one peak and the minimum will occur at its other peak. The process is called **amplitude modulation,** and the effect is shown in Fig. 2-52D. 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) frequencies, and hence the radio frequencies appearing in the circuit after modulation are 999,000, 1,000,000 and 1,001,000 cycles.

Many circuits have been devised for obtaining amplitude modulation, and they will be treated in detail in later chapters. When an audio frequency is used to control the amplitude of a radio frequency, the process is generally called "amplitude modulation," as mentioned previously, but when a radio frequency modulates another radio frequency it is called **heterodyning.** However, 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.

In A, B, C and D of Fig. 2-52, 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. 2-52 show the signals of Fig. 2-52A, B, C and D on an amplitude-*vs.*-frequency basis. Any one frequency is, of course, represented by a vertical line. Fig. 2-52H shows the side frequencies appearing as a result of the modulation process.

Amplitude modulation **(AM)** is not the only possible type nor is it the only one in use. This and other 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. The miniature, metal-envelope and small glass tubes in the foreground are receiving types. The two tubes with connections at the top of the bulb, lying down, are transmitting triodes of moderate power ratings. Those in the rear are transmitting-type beam tetrodes.

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

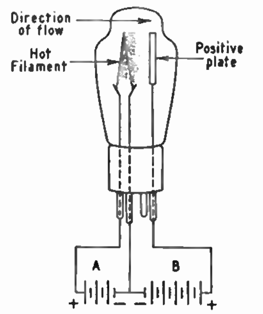


Fig. 3-1 — Conduction by thermionic emission in a vacuum tube. One battery is used to heat the filament to a temperature that will cause it to emit electrons. The other battery makes the plate positive with respect to the filament, thereby causing the emitted electrons to be attracted to the plate. Electrons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

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.

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-

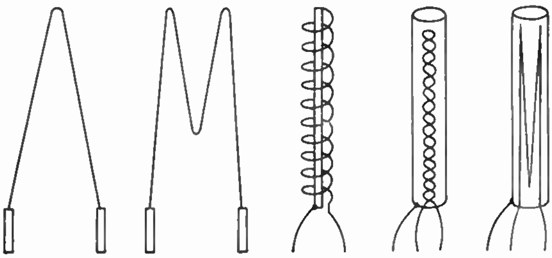


Fig. 3-2 — Types of cathode construction. Directly-heated cathodes or filaments are shown at A, B, and C. The inverted V filament is used in small receiving tubes, the M in both receiving and transmitting tubes. The spiral filament is a transmitting-tube type. The indirectly-heated cathodes at D and 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 into a load of one type or another; in this respect a tube is much like a generator or transformer. A circuit that did not provide a load for the tube would be like a short-circuit across a transformer; no useful purpose would be accomplished and the only result would be the generation of heat in the transformer. So it is with vacuum tubes; they must deliver power to a load in order to serve a useful purpose. Also, to be *efficient* most of the power must do useful work in the load and not be used in heating the plate of the tube. This means that most of the voltage should appear as a drop across the load rather than as a drop between the plate and cathode.

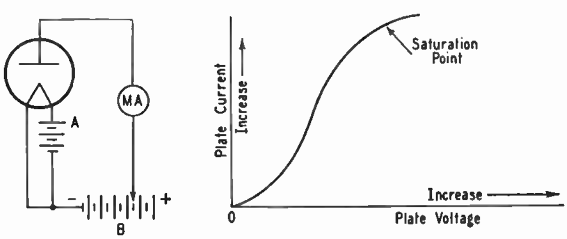


Fig. 3-3 — The diode, or two-element tube, and a typical curve showing how the plate current depends upon the voltage applied to the plate.

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.

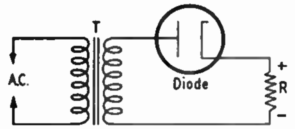
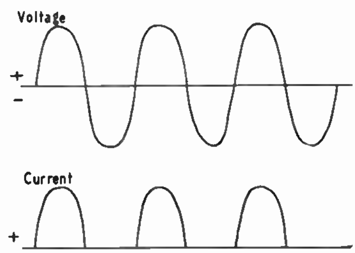


Fig. 3-4 — Rectification in a diode. Current flows only when the plate is positive with respect to the cathode, so that only half-cycles of current flow through the load resistor, *R*.



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

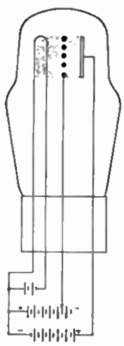


Fig. 3-5 — Construction of an elementary triode vacuum tube, showing the filament, grid (with an end view of the grid wires) and plate. The relative density of the space charge is indicated roughly by the dot density.

result is that, at any selected plate voltage, more electrons will flow to the plate than if the grid were not present. On the other hand, if the grid is made negative with respect to the cathode the negative charge on the grid will *add* to the space charge. This will *reduce* the number of electrons that can reach the plate at any selected plate voltage.

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

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

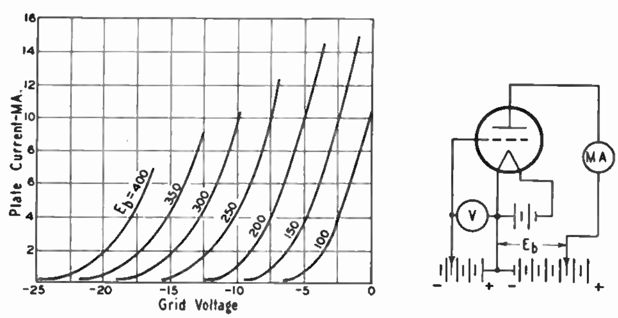


Fig 3-6 — Grid-voltage-*vs.*-plate-current curves at various fixed values of plate voltage (E_b) for a typical small triode. Characteristic curves of this type can be taken by varying the battery voltages in the circuit at the right.

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 **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.*

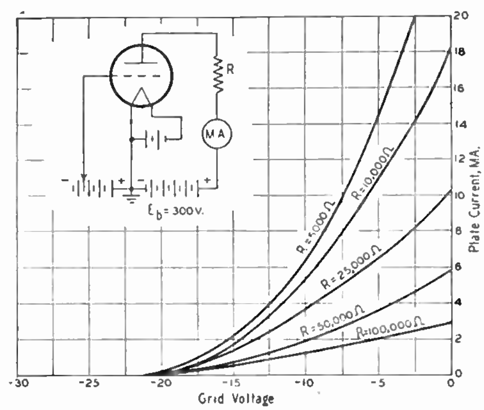


Fig. 3-7 — Dynamic characteristics of a small triode with various load resistances from 5000 to 100,000 ohms.

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, so 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 $50{,}000 \times 0.002 = 100$ 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

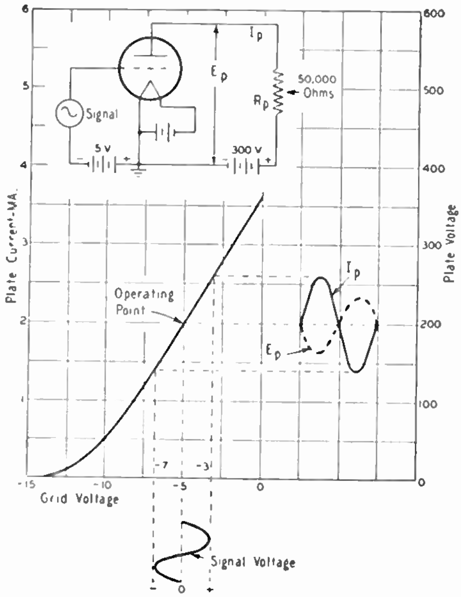


Fig. 3-8 — Amplifier operation. When the plate current varies in response to the signal applied to the grid, a varying voltage drop appears across the load, R_p, as shown by the dashed curve, E_p. I_p is the plate current.

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

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

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 waveshape 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 waveshape 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 waveshape that results from working over a part of the curve that is not straight (that is, a **nonlinear** part of the curve) has the effect of transforming a sine-wave grid signal into a more complex waveform. As explained in an earlier chapter, a complex wave can be resolved into a fundamental and a series of harmonics. In other words, distortion from nonlinearity causes the generation of harmonic frequencies — frequencies that are not present in the signal applied to the grid. Harmonic distortion is undesirable in most amplifiers, although

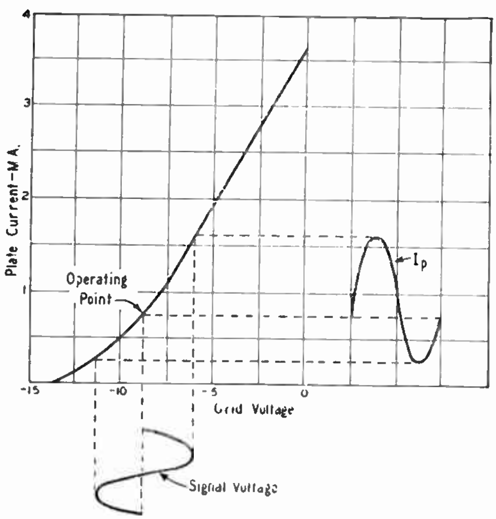


Fig. 3-9 — Harmonic distortion resulting from choice of an operating point on the curved part of the tube characteristic. The lower half-cycle of plate current does not have the same shape as the upper half-cycle.

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, between the plate and cathode of the tube) is applied to a second resistor, R_g, through a **coupling condenser,** C_c. The condenser "blocks off" the 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 condenser, 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-μfd. condenser 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.

The transformer-coupled amplifier uses a transformer with its primary connected in the plate

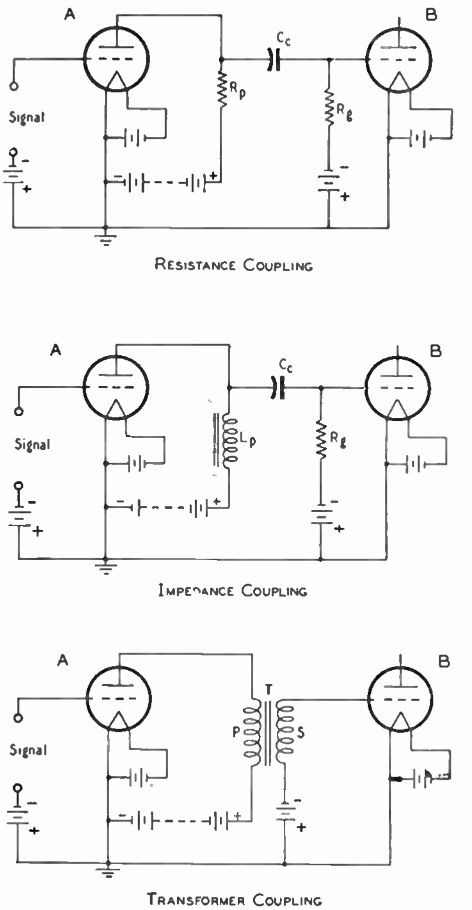


Fig. 3-10 — Three basic forms of coupling between vacuum-tube amplifiers.

circuit of the tube and its secondary connected to the load (in the circuit shown, a following amplifier). There is no direct connection between the two windings, so the plate voltage on tube *A* is isolated from the grid of tube *B*. The transformer-coupled amplifier has the same advantage as the impedance-coupled circuit with respect to loss of voltage from the plate supply. 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 10 or less, for the reason that the primary inductance of a practicable transformer cannot be made large enough to work well with a tube having high plate resistance.

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

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

Power Amplifiers

The end result of any amplification is that the amplified signal does some *work*. 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.

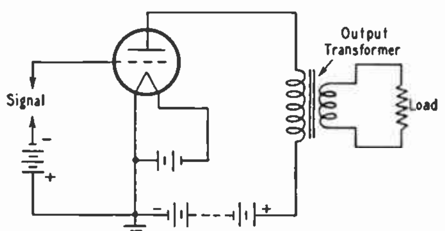


Fig. 3-11 — An elementary power-amplifier circuit in which the power-consuming load is coupled to the plate circuit through an impedance-matching transformer.

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 fixed plate-dissipation rating.

Parallel and Push-Pull

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

If the amplifier operates in such a way as to consume power in the grid circuit, the grid power required 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.

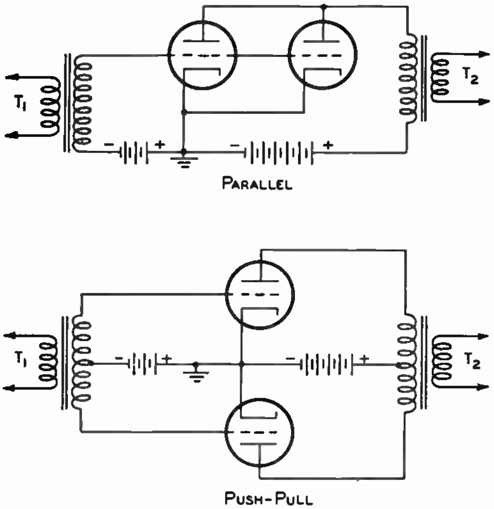


Fig. 3-12 — Parallel and push-pull a.f. amplifier circuits.

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 a number of 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. 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 is considerably more efficient than the Class A amplifier. Furthermore, the d.c. plate current of a Class B amplifier is proportional to the signal voltage on the grids, so the power input is small with small signals. The d.c. plate power input to a Class A amplifier is the same whether the signal is large, small, or absent altogether; therefore the maximum input that can be applied to a Class A amplifier is 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 grids must be driven positive with respect to the cathode during at least part of the cycle, so grid current flows and the grid circuit consumes power. While the power requirements are fairly low (as compared with the power output), the fact that the grids are positive during only *part* of the cycle means that the load on the 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

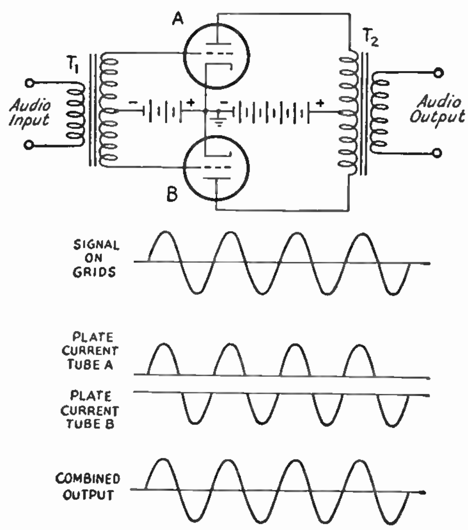


Fig. 3-13 — Class B amplifier operation.

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. waveform 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 from 150 to 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.

FEED-BACK

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 **feed-back.**

If the voltage that is inserted in the grid circuit is 180 degrees out of phase with the signal voltage acting on the grid, the feed-back is called **negative,** or **degenerative.** On the other hand, if the voltage is fed back *in* phase with the grid signal, the feed-back is called **positive,** or **regenerative.**

Negative Feed-Back

With negative feed-back the voltage that is fed back *opposes* the signal voltage. This decreases the amplitude of the voltage acting between the grid and cathode 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 feed-back (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 feed-back are therefore comparatively free from harmonic distortion. These advantages are worth while if the amplifier otherwise has enough voltage gain for its intended use.

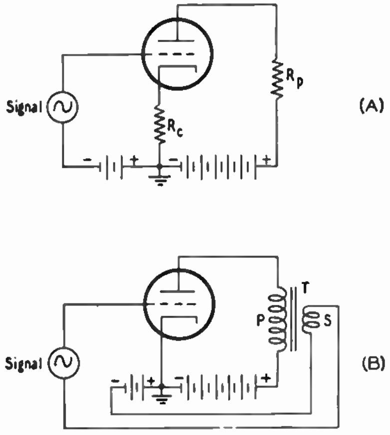


Fig. 3-14 — Simple circuits for producing feed-back.

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 feed-back. The secondary of a transformer is connected back into the grid circuit to insert a desired amount of feed-back voltage. Reversing the terminals of either transformer winding (but not both simultaneously) will reverse the phase.

Positive Feed-Back

Positive feed-back *increases* the amplification because the feed-back 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 (depending 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. Oscillations obviously would be undesirable in an ordinary audio-frequency amplifier, and for that reason (as well as the others mentioned above) the use of positive feed-back is confined principally to "oscillators."

● INTERELECTRODE CAPACITANCES

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

Input Capacitance

It was explained previously that the a.c. grid voltage and a.c. plate voltage of an amplifier 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. The greater the voltage amplification the greater this effective input capacitance. The input capaci-

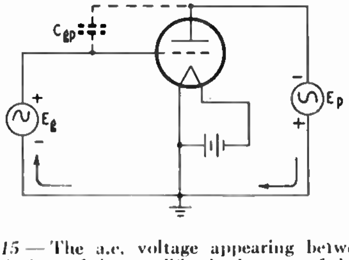


Fig. 3-15 — The a.c. voltage appearing between the grid and plate of the amplifier is the sum of the signal voltage and the output voltage, as shown by this simplified circuit. Instantaneous polarities are indicated.

tance of a resistance-coupled amplifier is given by the formula

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

where C_{gk} is the grid-to-cathode capacitance, C_{gp} is the grid-to-plate capacitance, and A is the voltage amplification. The 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 cannot be used at r.f., for example, because the reactances of the interelectrode "condensers" 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 it is, in effect, a coupling condenser between the grid and plate circuits. Since its reactance is relatively low at r.f., it offers a path over which energy can be fed back from the plate to the grid. In practically every case the feed-back is in the right phase and of sufficient amplitude to cause oscillation, so the circuit becomes useless as an amplifier.

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

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 also 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.

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 by-pass condenser 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.**

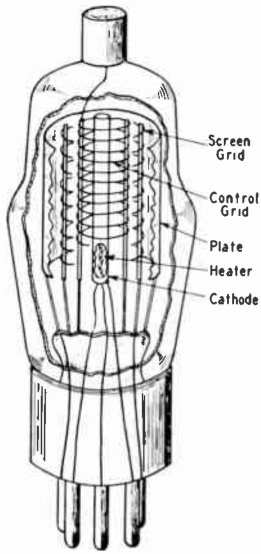


Fig. 3-16 — Representative arrangement of elements in a screen-grid tube, with front part of plate and screen grid cut away. In this drawing the control-grid connection is made through a cap on the top of the tube, thus eliminating the capacitance that would exist between the plate- and grid-lead wires if both passed through the base. "Single-ended" tubes that have both leads going through the base use special shielding and construction to eliminate interlead capacitance.

Pentodes

When an electron traveling at appreciable velocity through a tube strikes the plate it dislodges other electrons which "splash" from the plate into the interelement space. This is called **secondary emission.** In a triode the negative grid repels the secondary electrons back into the plate and they cause no disturbance. In the screen-grid tube, however, the positively-charged screen *attracts* the secondary electrons, causing a reverse current to flow between screen and plate.

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

Although the screen grid in either the tetrode or pentode greatly reduces the influence of the plate upon plate-current flow, the control grid still can control the plate current in essentially the same way that it does in a triode. Consequently, the grid-plate transconductance (or mutual conductance) of a tetrode or pentode will be of the same order of value as in a triode of corresponding structure. On the other hand, since 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 micro-microfarad. 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.

Pentode R.F. Amplifier

Fig. 3-17 shows a simplified form of r.f. amplifier circuit using a pentode tube. Radio-frequency energy in the small coil coupled to L_1 is built up in voltage in the tuned circuit, L_1C_1, when L_1C_1 is tuned to resonance with the frequency of the incoming signal. The voltage that appears across L_1C_1 is applied to the grid and cathode of the tube and is amplified by the tube. A second resonant circuit, L_2C_2, is the load for the plate of the tube, its parallel impedance being high because it is tuned to resonance with the frequency applied to the grid. R.f. output can be taken from the coil coupled to L_2. The screen-grid voltage is obtained from a tap on the plate battery; most tubes are designed for operation with the screen voltage considerably lower than the plate voltage. In this circuit the batteries are assumed to have low impedance for the r.f. current; in a practical circuit, by-pass condensers would be used to make sure that the impedances of the return paths are so low as to be negligible.

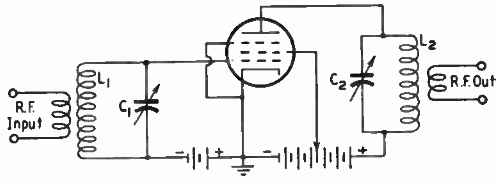


Fig. 3-17 — Simplified pentode r.f.-amplifier circuit. L_1C_1 and L_2C_2 are tuned to the same frequency.

Audio Amplification

In addition to their applications as radio-frequency amplifiers, pentode or tetrode screen-grid tubes also can be constructed for audio-frequency power amplification. In tubes designed for this purpose the 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 pentode type because large power outputs can be secured with very small amounts of grid driving power. The circuits with which they are used are practically identical with those used for pentodes.

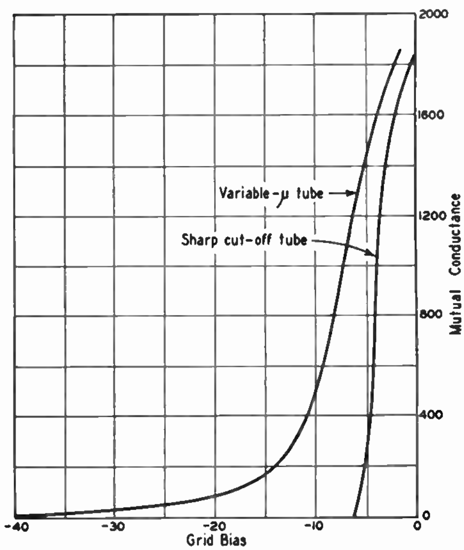


Fig. 3-18 — Curves showing the relationship between mutual conductance and negative grid bias for two small receiving pentodes, one a sharp cut-off type and the other a variable-μ type.

Variable-μ Tubes

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

The ordinary type of tube has what is known as a **sharp cut-off** characteristic. The mutual conductance decreases at a uniform rate as the negative bias is increased, as shown in Fig. 3-18. The amount of signal voltage that such a tube can handle without causing distortion is not sufficient to take care of very strong signals. To overcome this, some tubes are made with a **variable-μ** characteristic (that is, the amplification factor changes with the grid bias), resulting in the type of curve shown in Fig. 3-18. The variable-μ tube can handle a much larger signal than the sharp cut-off type before the signal swings either beyond the zero grid-bias point or the plate-current cut-off point.

OTHER TYPES OF AMPLIFIERS

In the amplifier circuits so far discussed, the signal has been applied between the grid and cathode and the amplified output has been taken from the plate-to-cathode circuit. That is, the *cathode* has been the 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 different kinds of amplifiers, commonly called the **grounded-grid amplifier** (or **grid-separation** circuit) and the **cathode follower.**

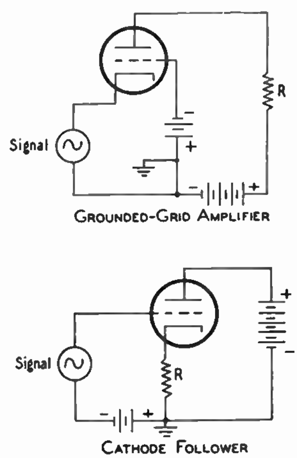


Fig. 3-19 — **In the upper circuit, the grid is the junction point between the input and output circuits. In the lower drawing, the plate is the junction. In either case the output is developed in the load resistor, *R*, and may be coupled to a following amplifier by the usual methods.**

These two circuits are shown in simplified form in Fig. 3-19. In both circuits the resistor *R* represents the load into which the amplifier works; the actual load may be resistance-capacitance-coupled, transformer-coupled, may be a tuned circuit if the amplifier operates at radio frequencies, and so on. Also, in both circuits the batteries that supply grid bias and plate power are assumed to have such negligible impedance that they do not enter into the operation of the circuits.

Grounded-Grid Amplifier

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

The input impedance of the grounded-grid amplifier consists of a capacitance, calculated in a similar way as for the grounded-cathode amplifier, in parallel with an equivalent resistance representing the power furnished by the driving source to the load. 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 finds its chief application at v.h.f. and u.h.f., where the more conventional amplifier circuit fails to work properly. With a triode tube designed for this type of operation, an r.f. amplifier can be built that is free from the type of feed-back that causes oscillation. This requires that the grid act as a shield between the cathode and plate, reducing the plate-cathode capacitance to a very low value.

Cathode Follower

The cathode follower uses the plate of the tube as the common element. The input signal is applied between the grid and plate (assuming negligible impedance in the batteries) and the output is taken from between cathode and plate. This circuit, like the grounded-grid amplifier, is degenerative; in fact, *all* of the output voltage is fed back into the input circuit. 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.

An important feature of the cathode follower is its low output impedance, which is given by the formula (neglecting the grid-to-cathode capacitance)

$$Z_{output} = \frac{r_p}{1 + \mu}$$

where r_p is the tube plate resistance and μ is the amplification factor. This 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)

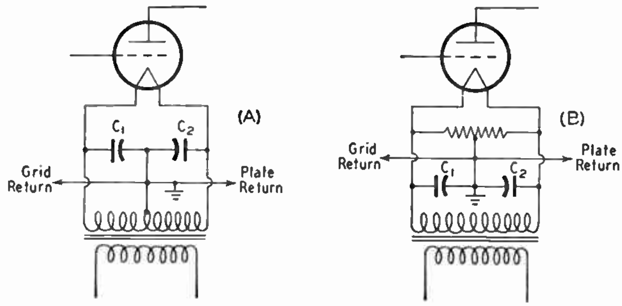


Fig. 3-20 — Filament center-tapping methods for use with directly-heated tubes.

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. 3-20. In both cases the grid- and plate-return circuits are connected to the electrical midpoint **(center-tap)** of the filament supply. Thus, so far as the grid and plate are concerned, the voltage and current on one side of the filament are balanced by an equal and opposite voltage and current on the other side. The balance is never quite perfect, however, so filament-type tubes are never completely hum-free. For this reason directly-heated filaments are employed for the most part in power tubes, where the amount of hum introduced is extremely small in comparison to 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. 3-20.

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 the type commonly used.

The cathode-bias method uses a resistor (cathode resistor) connected in series with the cathode, as shown at R in Fig. 3-21. The direction of plate-current flow is such that the end of the resistor nearest the cathode is positive. The voltage drop across R therefore places a *negative* voltage on the grid. This negative bias is obtained from the steady d.c. plate current.

If the alternating component of plate current flows through R when the tube is amplifying, the voltage drop caused by the a.c. will be degenerative (note the similarity between this circuit and that of Fig. 3-14A). To prevent this the resistor is by-passed by a condenser, C, that has very low reactance compared 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 by-passing at the low audio frequencies, C should be 10 to 50 microfarads (electrolytic condensers are used for this purpose). At radio frequencies, capacitances of about 100 $\mu\mu$fd. to 0.1 μfd. are used; the small values are sufficient at very high frequencies and the largest at low and medium frequencies. In the range 3 to 30 megacycles a capacitance of 0.01 μfd. is satisfactory.

The value of cathode resistor 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

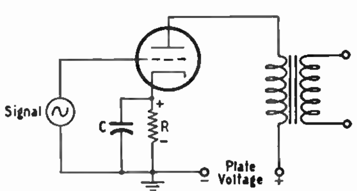


Fig. 3-21 — Cathode biasing. R is the cathode resistor and C is the cathode by-pass condenser.

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.

Screen Supply

In practical circuits using tetrodes and pentodes the voltage for the screen frequently is taken from the plate supply through a resistor. A typical circuit for an r.f. amplifier is shown in Fig. 3-22. Resistor *R* is the **screen dropping resistor,** and *C* is the **screen by-pass condenser.** In flowing through *R*, the screen current causes a voltage drop in *R* that reduces the plate-supply voltage to the proper value for the screen. When the plate-supply voltage and the screen current are known, the value of *R* can be calculated from Ohm's Law.

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

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

The power to be dissipated in the resistor is

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

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

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

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

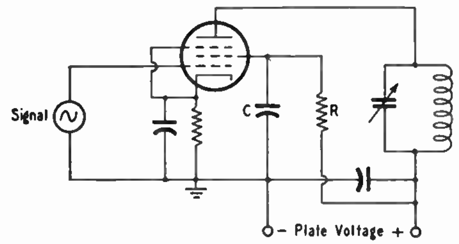


Fig. 3-22 — Screen-voltage supply for a pentode tube through a dropping resistor, *R*. The screen by-pass condenser, *C*, must have low enough reactance to bring the screen to ground potential for the frequency or frequencies being amplified.

SPECIAL TUBE TYPES

Multipurpose Tubes

"Combination" tubes are available to perform more than one function, particularly in receiver circuits. For the most part these are simply multiunit tubes made up of individual tube-element structures, combined in a single bulb for compactness and economy.

Among the simplest multipurpose types are full-wave rectifiers, combining two diodes in one envelope, and twin triodes, consisting of two triodes in one bulb. More complex types include duplex-diode triodes (two diodes and a triode in one structure), duplex-diode pentodes, converters and mixers (for superheterodyne receivers), combination power tubes and rectifiers, and so on.

Mercury-Vapor Rectifiers

For a given value of plate current, the power lost in a diode rectifier will be reduced if it is possible to decrease the voltage drop from plate to cathode. A small amount of mercury in the tube will vaporize when the cathode is heated and, further, will ionize when plate voltage at least equal to a certain minimum value **(ionizing voltage)** is applied. The positive ions neutralize the space charge and reduce the plate-cathode voltage drop to a practically constant value of about 15 volts, regardless of the value of plate current.

Since this voltage drop is smaller than can be attained with purely thermionic conduction, there is less power loss in a mercury-vapor rectifier than in a vacuum rectifier. Also, the voltage drop in the tube is constant despite variations in load current. Mercury-vapor tubes are widely used in rectifiers built to deliver large power outputs.

Grid-Control Rectifiers

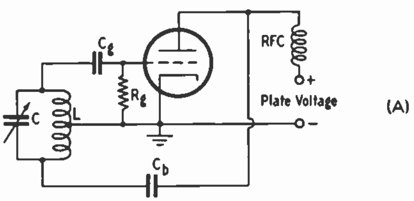
If a grid is inserted in a mercury-vapor rectifier it is found that, with sufficient negative grid bias, it is possible to prevent plate current from flowing. However, this is true *only if the bias is present before plate voltage is applied.* If, after applying plate voltage, the bias is lowered to the point where plate current can flow, the mercury vapor will ionize and the grid will lose control of plate current, because the space charge disappears when ionization occurs. The grid can assume control again only after the plate voltage is reduced below the **deionizing voltage,** which is somewhat less than the plate-cathode voltage drop during plate-current flow.

The same phenomenon also occurs in triodes filled with other gases that ionize at low pressure. **Grid-control rectifiers** or **thyratrons** find considerable application in "electronic switching," and in timing devices. Both triode and tetrode types are manufactured.

Oscillators

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

Oscillations normally take place at only one frequency, and a desired frequency of oscillation can be obtained by using a resonant circuit tuned to that frequency. For example, in Fig. 3-23A 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 if the top end of L is positive at some instant the bottom end will be negative, and the point at which the tap is connected will be at an intermediate potential. 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 feed-back.



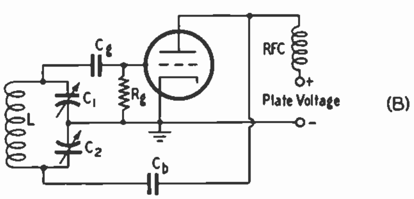


Fig. 3-23 — Basic oscillator circuits. Feed-back voltage is obtained by tapping the grid and cathode across a portion of the tuned circuit. In the Hartley circuit the tap is on the coil, but in the Colpitts circuit the voltage is obtained from the drop across a condenser.

The amount of feed-back 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 feed-back 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 feed-back usually is obtained when the tap is somewhere near the center of the coil.

The circuit of Fig. 3-23A is parallel-fed, C_b being the blocking condenser. The value of C_b is not critical so long as its reactance is low (a few hundred ohms) at the operating frequency.

Condenser C_g is the **grid condenser.** It and R_g (the **grid leak**) are used for the purpose of obtaining grid bias for the tube. In practically all 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. 3-23 uses the voltage drops across two condensers in series in the tuned circuit to supply the feed-back. Other than this, the operation is the same as just described. The feed-back can be varied by varying the ratio of the reactances of C_1 and C_2 (that is, by varying the ratio of their capacitances).

Another type of oscillator, called the **tuned-plate tuned-grid** circuit, is shown in Fig. 3-24. Resonant circuits tuned approximately to the same frequency are connected between grid and cathode and between plate and cathode. The two coils, L_1 and L_2 are not magnetically coupled. The feed-back is through the grid-plate capacitance of the tube, and will be in the right phase to be positive when the plate circuit, C_2L_2, is tuned to a slightly higher frequency than the grid circuit, L_1C_1. The amount of feed-back can be adjusted by varying the tuning of either circuit. The frequency of oscillation is determined by the tuned circuit that has the higher Q. The grid leak and grid condenser have the same functions as in the other circuits. In this case it is convenient to use series feed for the plate circuit, so C_b is a by-pass condenser to guide the r.f. current around the plate supply.

There are many oscillator circuits, some using two or more tubes, but the basic feature of all of them is that there is positive feed-back in the proper amplitude to sustain oscillation.

Oscillator Operating Characteristics

When an oscillator is delivering power to a load, the adjustment for proper feed-back will depend on how heavily the oscillator is loaded — that is, how much power is being taken from the circuit. If the feed-back 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 feed-back will make the grid current excessively high, with the result that the power loss in the grid circuit is larger than necessary. Since the oscillator itself supplies this grid power, excessive feed-back lowers the over-all efficiency because whatever power is used in the grid circuit is not available as useful output.

One of the most important considerations in oscillator design is frequency stability. The principal factors that cause a change in frequency are (1) temperature, (2) plate voltage, (3) loading, (4) mechanical variations of circuit elements. Temperature changes will cause vacuum-tube elements to expand or contract slightly, thus causing variations in the interelectrode capacitances. Since these are unavoidably part of the tuned circuit, the frequency will change correspondingly. Temperature changes in the coil or condenser will alter their inductance or capacitance slightly, again causing a shift in the resonant frequency. These effects are relatively slow in operation, and the frequency change caused by them is called **drift.**

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. Since the tube and load

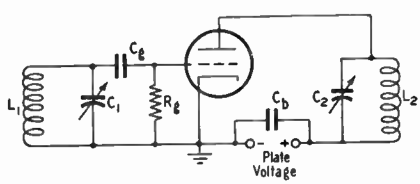


Fig. 3-24 — The tuned-plate tuned-grid oscillator.

represent a relatively low resistance in parallel with the circuit, this means that a low L/C ratio **(high-C)** must be used and that the circuit should be lightly loaded. A high value of grid leak resistance also is helpful because, by increasing the grid bias without increasing grid current, it raises the effective tube grid and plate resistances as seen by the tank circuit. Using relatively high plate voltage and low plate current also is desirable.

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. 3-23 and 3-24 the cathode is connected to ground. It is not actually essential that the radio-frequency circuit should be grounded at the cathode; in fact, there are many times when an *r.f.* ground on some other point in the circuit

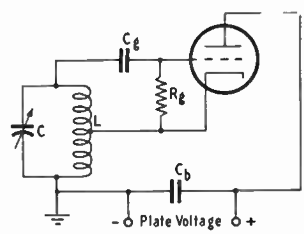


Fig. 3-25 — Showing how the plate may be grounded for r.f. in a typical oscillator circuit (Hartley)

is desirable. The r.f. ground can be placed at any point so long as proper provisions are made for feeding the supply voltages to the tube elements.

Fig. 3-25 shows the Hartley circuit with 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 condenser, 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 condenser can be grounded.

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.

Semiconductor Devices

Although not vacuum tubes, there is another group of rectifying devices that can perform similar functions. These include the crystal diode and the transistor. They make use of the peculiar properties of certain crystals, particularly germanium, called **semiconductors.**

CRYSTAL DIODES

As the name implies, the **crystal diode** is a two-element rectifying device comparable with a tube diode. In its common form it consists of a small piece of an appropriate crystalline substance with one contact made through a fine pointed wire or catwhisker. The other contact is through the metal mounting, as shown in Fig. 3-26. Such a device will conduct current much more readily in one direction than the other.

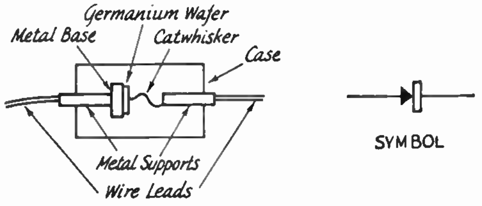


Fig. 3-26 — The germanium crystal and circuit symbol. The arrow points in the direction of minimum resistance.

As compared with a tube diode, the crystal diode has the advantages of very small size, very low interelectrode capacitance (less than one micromicrofarad), and requiring no heater or filament power. Its **forward resistance** — in the favored direction of current flow — is a few hundred ohms, comparable with that of a tube diode. Its disadvantage is a relatively low inverse peak voltage rating (see Power Supply chapter) and a **back resistance** (in the direction of least current flow) that may be as low as 20,000 ohms, although in some types the back resistance may be as high as a megohm. The tube diode, in contrast, simply does not conduct in the reverse direction, and so has infinite back resistance for all practical purposes.

The crystal diode is widely used in measuring equipment and as a detector and mixer in receivers.

TRANSISTORS

If two catwhiskers are placed very close together on a germanium crystal and a positive voltage applied to one while a negative voltage is applied to the other, both with respect to a common connection called the **base**, it is found that a change in current through the first (the **emitter**) will cause a corresponding change in the current through the second (the **collector**), and vice versa. Such a device, shown in Fig. 3-27, is called a **point-contact transistor.**

Amplification

A current of several milliamperes will flow in the emitter circuit when the positive bias is only a fraction of a volt, so the impedance of the emitter circuit is quite low — of the order of a few hundred ohms. On the other hand, the output resistance of the collector circuit is of the order of tens of thousands of ohms. The **current gain** — the ratio of change in collector current to the change in emitter current — varies with the type of transistor and may range from somewhat less than 1 to a value as high as 3 or 4. However, the emitter current is flowing in a low impedance while the collector current is flowing in a high impedance, so there is a power gain in proportion to the impedances. This gain may be 20 db. or more.

The base circuit of a transistor also has considerable internal resistance, common to the emitter and collector circuits. Inherent feed-back occurs because the collector current flows through the base resistance and is thus introduced into the emitter circuit. If the current gain is greater than 1 this feed-back may cause self-oscillation.

Junction Transistors

Another form of transistor, called the junction type, is also shown in Fig. 3-27. This consists of a sandwich of germanium wafers having opposite conduction characteristics — that is, one type conducts because of a deficiency of electrons (p type) and the other because of an excess of electrons (n type). Junction transistors may be made either of an n-p-n or p-n-p sandwich. Biases of opposite polarity are used on the emitter and collector, just as in the case of point-contact transistors, but these biases are reversed when a p-n-p is substituted for an n-p-n.

Transistor Applications

Since transistors will both amplify and oscillate, they can perform many of the same functions as vacuum tubes. Their advantages are very small size and weight, no cathode power required, and operation at very low voltages and currents — of the order of 3 to 6 volts for the emitter and 10 to 25 volts for the collector. At present, their power-handling capacity is quite limited, confining their use to very low power applications. In many respects the characteristics of transistors are the opposite of those of vacuum tubes, so that the circuit techniques are quite different. Since transistors have only very recently been made available commercially, their application in amateur radio is largely a field for future exploration.

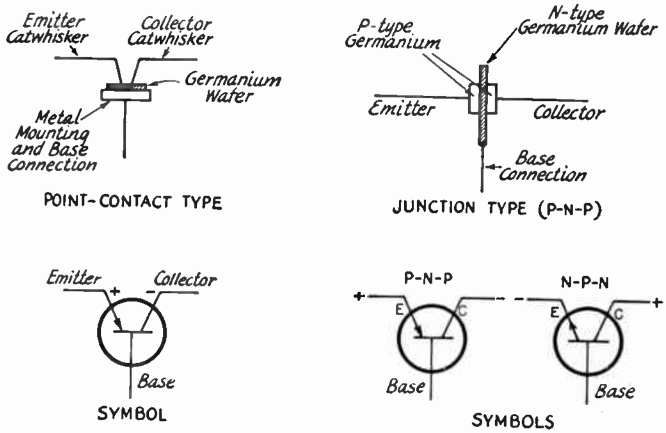


Fig. 3-27 — The point-contact transistor (left) and junction-type transistor (right). Plus and minus signs indicate polarities of bias voltages applied to the elements, with respect to the base.

High-Frequency Communication

Much of the appeal of amateur communication on the high frequencies lies in the fact that the results are not always predictable. Transmission conditions on the same frequency vary with the year and even with the time of day. Although these variations usually follow certain established cycles, many peculiar effects can be observed from time to time. Every radio amateur should have some understanding of the known facts about radio wave propagation so that he will stand some chance of interpreting the unusual conditions when they occur. The observant amateur is in an excellent position to make worthwhile contributions to the science, provided he has sufficient background to understand his results. He may discover new facts about propagation at the very-high frequencies or in the microwave region, as amateurs have in the past. In fact, it is through amateur efforts that most of the extended-range possibilities of various radio frequencies have been discovered, either through accident or long and careful investigation.

What To Expect on the Various Amateur Bands

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

The 3.5-Mc., or "80-meter," band is a more useful band during the night than during the daylight hours. In the daytime, one can seldom hear signals from a distance of greater than 200 miles or so, but during the darkness hours distances up to several thousand miles are not unusual, and transoceanic contacts are regularly made during the winter months. During the summer, the static level is high in some parts of the world.

The 7-Mc., or "40-meter," band has many of the same characteristics as 3.5, except that the distances that can be covered during the day and night hours are increased. During daylight, distances up to a thousand miles can be covered under good conditions, and during the dawn and dusk periods in winter it is possible to work stations as far as the other side of the world, the signals following the darkness path. The winter months are somewhat better than the summer ones. In general, summer static is much less of a problem than on 80 meters, although it can be serious in the semitropical zones.

The 14-Mc., or "20-meter," band is probably the best one for long-distance work. During portions of the sunspot cycle (discussed later in this chapter) it is open to some part of the world during practically all of the 24 hours, while at other times it is generally useful only during daylight hours and the dawn and dusk periods.

The 21-Mc., or "15-meter," band shows highly variable characteristics depending on the sunspot cycle. During sunspot maxima it is useful for long-distance work during a large part of the 24 hours, but in years of low sunspot activity it is almost wholly a daytime band, and sometimes unusable even in daytime. However, it is often possible to maintain communication over distances up to 1500 miles or more by sporadic-*E* ionization (described later), which may occur either day or night at any time in the sunspot cycle.

The 27-Mc. ("11-meter") and 28-Mc. ("10-meter") bands are generally considered to be DX bands during the daylight hours and good for local work during the hours of darkness, although at the peak of the sunspot cycle, they are "open" into the late evening hours for DX communication. At the sunspot minimum these bands are usually "dead" for long-distance communication in the northern latitudes. Nevertheless, sporadic-*E* propagation is likely to occur at any time, just as in the case of the 21-Mc. band. The v.h.f. and u.h.f. bands (50 Mc. and higher) are considered in detail in the chapter on v.h.f. propagation.

Characteristics of Radio Waves

Radio waves are basically of the same nature as light and heat, which also are forms of electromagnetic radiation. The principal difference is in the wavelength, which in the case of radio waves is much greater than the wavelengths of light or heat. However, all three types of radiation travel at the same speed (300,000,000 meters per second) in free space, and have similar properties in that they all can be reflected, refracted, and diffracted.

As described in the chapter on fundamentals, an electromagnetic wave is composed of moving fields of electric and magnetic force. The lines of force in the two fields are at right angles, and are mutually perpendicular to the direction of travel. A simple representation of a wave is shown in Fig. 4-1. In this drawing the electric lines are perpendicular to the earth and the magnetic lines are horizontal. They could, however, have any position with respect to earth so long as they remain perpendicular to each other.

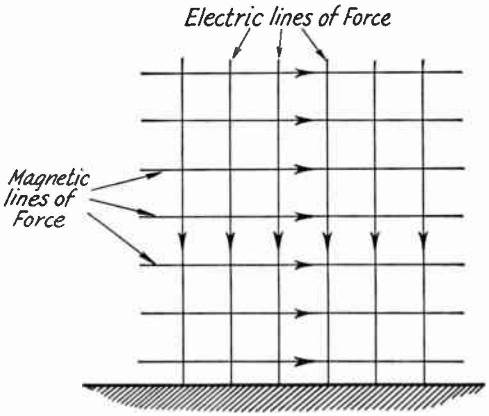


Fig. 4-1 — Representation of electrostatic and electromagnetic lines of force in a radio wave. Arrows indicate instantaneous directions of the fields for a wave traveling toward the reader. Reversing the direction of one set of lines would reverse the direction of travel.

The plane containing the continuous lines of electric and magnetic force shown by the grid- or mesh-like drawing in Fig. 4-1 is called the **wave front.**

Polarization

The **polarization** of a radio wave is taken as the direction of the lines of force in the electric field. If the electric lines are perpendicular to the earth, the wave is said to be **vertically polarized;** if parallel with the earth, the wave is **horizontally polarized.** The longer waves, when traveling along the ground, usually maintain their polarization in the same plane as was generated at the antenna. The polarization of shorter waves may be altered during travel, however, and sometimes will vary quite rapidly.

Medium of Propagation

The **medium** in which electromagnetic waves travel has a marked influence on the speed with which they move. When the medium is empty space the speed, as stated above, is 300,000,000 meters per second. It is almost, but not quite, that great in air, and is much less in some other substances. In dielectrics, for example, the speed is inversely proportional to the dielectric constant of the material.

When a wave meets a good conductor it cannot penetrate it to any extent (although it will travel through a dielectric with ease) because the electric lines of force are practically short-circuited.

Reflection

A light ray traveling through air of uniform characteristics goes in a straight line, but when it meets some object having different properties its path is shifted. If the "discontinuity" is sufficiently great in extent, as compared with the wavelength of light, and if the change in properties is abrupt, the ray may be reflected. The discontinuity may be either a change in the dielectric constant or the conductivity of the medium. Similarly, a radio wave will be reflected under comparable conditions. However, the discontinuity set up by the reflecting object must at least be comparable with the wavelength in size, to cause reflection of radio waves. Nevertheless, objects as small as an airplane, a tree, or even a man's body will reflect waves a few feet long and less.

Refraction

When a wave meets a discontinuity that it can penetrate, the change in speed causes its path to be deflected, if it enters at any angle other than the perpendicular to the surface of the new medium. That part of the wave front that enters the new medium first travels at the new speed before the trailing part of the wave front enters, and so the wave as a whole is swung around or **refracted.** The new direction depends on the difference in speed in the two media, and on the wavelength. Wave "bending" by refraction is the mechanism by which long-distance communication at high frequencies is possible. The medium in which the bending takes place is an ionized region, called the **ionosphere,** in the upper atmosphere. The composition and properties of the ionosphere are discussed later in this chapter.

Diffraction

When a wave grazes the edge of an object in passing, it tends to be bent around that edge. This effect, called **diffraction,** results in a diversion of part of the energy of those waves which normally follow a straight path, so they may be received at some distance below the summit of an obstruction or around its edges.

Spreading

The field intensity of a wave is inversely proportional to the distance from the source. Thus if one receiving point is twice as far from the transmitter as another, the field strength at the more distant point will be just half the field strength at the nearer point. This results from the fact that the energy in the wave front must be distributed over a greater area as the wave moves away from the source. This **inverse-distance law** is based on the assumption that there is nothing in the medium to absorb energy from the wave as it travels, which is true in free space but not in practical communication along the ground and through the atmosphere.

Types of Propagation

According to the altitude of the paths along which they are propagated, radio waves may be classified as **ionospheric waves, tropospheric waves** or **ground waves.**

The ionospheric wave or **sky wave** is that part of the total radiation that is directed toward the ionosphere. Depending upon variable conditions in that region, as well as upon transmitting wavelength, the ionospheric wave may or may not be returned to earth by the effects of refraction and reflection.

The tropospheric wave is that part of the total radiation that undergoes refraction and reflection in regions of abrupt change of dielectric constant in the troposphere, such as the boundaries between air masses of differing temperature and moisture content.

The ground wave is that part of the total radiation that is directly affected by the presence of the earth and its surface features. The ground

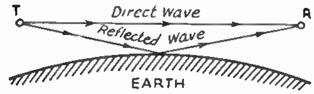


Fig. 4-2 — Showing how both direct and reflected waves may be received simultaneously.

wave has two components. One is the **surface wave,** which is an earth-guided wave, and the other is the **space wave** (not to be confused with the ionospheric or sky wave). The space wave is itself the resultant of two components — the **direct wave** and the **ground-reflected wave,** as shown in Fig. 4-2.

Ionospheric Propagation

PROPERTIES OF THE IONOSPHERE

Except for distances of a few miles, nearly all amateur communication on frequencies below 30 Mc. is by means of the sky wave. Upon leaving the transmitting antenna, this wave travels upward from the earth's surface at such an angle that it would continue out into space were its path not bent sufficiently to bring it back to earth. The medium that causes such bending is the **ionosphere,** a region in the upper atmosphere, above a height of about 60 miles, where free ions and electrons exist in sufficient quantity to have an appreciable effect on the speed at which the waves travel.

The ionization in the upper atmosphere is believed to be caused by ultraviolet radiation from the sun. The ionosphere is not a single region but is composed of a series of layers of varying densities of ionization occurring at different heights. Each layer consists of a central region of relatively dense ionization that tapers off in intensity both above and below.

Refraction and Reflection

The greater the intensity of ionization in a layer, the more the path of the wave is bent. The amount of bending also depends on the wavelength; the longer the wave, the more the path is bent for a given degree of ionization. Thus low-frequency waves are more readily bent than those of high frequency. For this reason the lower frequencies — 3.5 and 7 Mc. — are more "reliable" than the higher frequencies — 14 to 28 Mc.; there are times when the ionization is of such low value that waves of the latter frequency range are not bent enough to return to earth.

In addition to refraction, **reflection** may take place at the lower boundary of an ionized layer if the boundary is sharply defined; i.e., if there is an appreciable change in ionization within a relatively short interval of travel. For waves approaching the layer at or near the perpendicular, the change in ionization must take place within a difference in height comparable with the wavelength; hence, ionospheric reflection is more apt to occur at longer wavelengths (lower frequencies).

Absorption

In traveling through the ionosphere the wave gives up some of its energy by setting the ionized particles into motion. The energy **absorption** from this cause increases with the wavelength; that is, absorption is greater at lower frequencies. It also increases with the intensity of ionization, and with the density of the atmosphere in the ionized region.

Ionospheric absorption decreases the strength of the signal at the receiving point below the value that would be expected from the normal spreading of a wave traveling the same distance.

Virtual Height

Although an ionospheric layer is a region of considerable depth it is convenient to assign to it a definite height, called the **virtual height.** This is the height from which a simple reflection would give the same effect as the gradual refraction that actually takes place, as illustrated in Fig. 4-3. The wave traveling upward is bent back over a path having an appreciable radius of turning, and a measurable interval of time is consumed in the turning process. The virtual height is the height of a triangle having equal sides of a total length proportional to the time taken for the wave to travel from *T* to *R*.

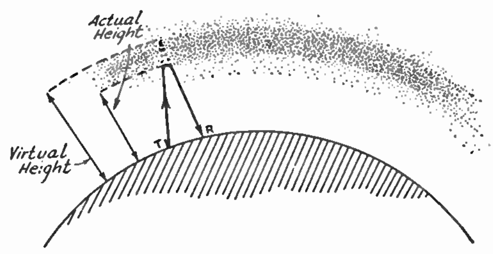


Fig. 4-3 — Bending in the ionosphere, and the echo or reflection method of determining virtual height.

Normal Structure of the Ionosphere

The lowest useful ionized layer is called the ***E* layer.** The average height of the region of maximum ionization is about 70 miles. The air at this height is sufficiently dense so that the ions and electrons set free by the sun's radiation do not travel far before they meet and recombine to form neutral particles, so the layer can maintain its normal intensity of ionization only in the presence of continuing radiation from the sun. Hence the ionization is greatest around local noon and practically disappears after sundown.

In the daytime there is a still lower ionized area, the ***D* region.** The *D*-region ionization is proportional to the height of the sun and is greatest at noon. Low-frequency waves (80 meters) are almost completely absorbed by this layer while it exists, and only the high-angle radiation is reflected by the *E* layer. (Lower-angle radiation travels farther through the *D* region and is absorbed.)

The second principal layer is the ***F* layer,** which has a height of about 175 miles at night. At this altitude the air is so thin that recombination of ions and electrons takes place very slowly, inasmuch as particles can travel relatively great distances before meeting. The ionization decreases after sundown, reaching a minimum just before sunrise. In the daytime the *F* layer splits into two parts, the **F_1 and F_2 layers,** with average virtual heights of, respectively, 140 miles and 200 miles. These layers are most highly ionized at about local noon, and merge again at sunset into the *F* layer.

SKY-WAVE PROPAGATION

Wave Angle

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

Skip Distance

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

The area between the end of the useful ground wave and the beginning of ionospheric-wave reception is called the **skip zone,** and the distance from the transmitter to the nearest point where the sky wave returns to earth is called the **skip distance.** The extent of skip zone depends upon the frequency and the state of the ionosphere, and also upon the height of the layer in which the refraction takes place. The higher layers give longer skip distances for the same wave angle. Wave angles at the transmitting and receiving points are usually, although not always, approximately the same for any given wave path.

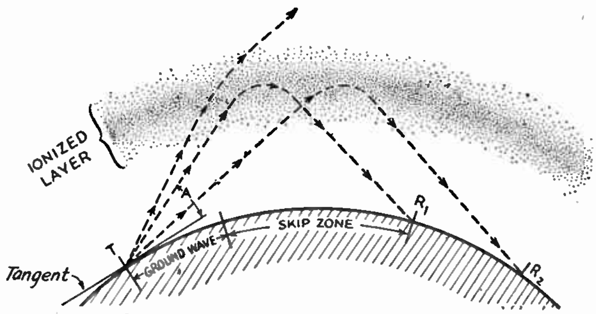


Fig. 4-4 — Refraction of sky waves, showing the critical wave angle and the skip zone. Waves leaving the transmitter at angles above the critical (greater than *A*) are not bent enough to be returned to earth. As the angle is decreased, the waves return to earth at increasingly greater distances.

Critical and Maximum Usable Frequencies

If the frequency is low enough, a wave sent vertically to the ionosphere will be reflected back down to the transmitting point. If the frequency is then gradually increased, eventually a frequency will be reached where this vertical reflection just fails to occur. This is the **critical frequency** for the layer under consideration. When the operating frequency is below the critical value there is no skip zone.

The critical frequency is a useful index to the highest frequency that can be used to transmit over a specified distance — the **maximum usable frequency (m.u.f.).** If the wave leaving the transmitting point at angle *A* in Fig. 4-4 is, for example, at a frequency of 14 Mc., and if a higher frequency would skip over the receiving point R_1, then 14 Mc. is the m.u.f. for the distance from *T* to R_1.

The greatest possible distance is covered when the wave leaves along the tangent to the earth; that is, at zero wave angle. Under average conditions this distance is about 4000 kilometers or 2500 miles for the F_2 layer, and 2000 km. for 1250 miles for the *E* layer. The distances vary with the layer height. Frequencies above these limiting m.u.f.'s will not be returned to earth at any distance. The 4000-km. m.u.f. for the F_2 layer is approximately 3 times the critical frequency for that layer, and for the *E* layer the 2000-km. m.u.f. is about 5 times the critical frequency.

Absorption in the ionosphere is least at the maximum usable frequency for the distance, and increases very rapidly as the frequency is lowered below the m.u.f. Consequently, best results with low power always are secured when the frequency is as close to the m.u.f. as possible.

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

Multihop Transmission

On returning to the earth the wave can be reflected upward and travel again to the ionosphere. There it may once more be refracted, and again bent back to earth. This process may be repeated several times. **Multihop** propagation of this nature is necessary for transmission over great distances because of the limited heights of the layers and the curvature of the earth, which restrict the maximum one-hop distance to the values mentioned in the preceding section. However, ground losses absorb some of the energy from the wave on each reflection (the amount of the loss varying with the type of ground and being least for reflection from sea water), and there is also absorption in the ionosphere at each reflection. Hence the smaller the number of hops the greater the signal strength at the receiver, other things being equal.

Fading

Two or more parts of the wave may follow slightly different paths in traveling to the receiving point, in which case the difference in path lengths will cause a phase difference to exist between the wave components at the receiving antenna. The total field strength will be the sum of the components and may be larger or smaller than one component alone, since the phases may be such as either to aid or oppose. Since the paths change from time to time, this causes a variation in signal strength called **fading.** Fading can also result from the combination of single-hop and multihop waves, or the combination of a ground wave with an ionospheric or tropospheric wave. The latter condition results in an area of severe fading in the region where the two waves have about the same intensity; better reception is obtained at either shorter or longer distances where one component is considerably stronger than the other.

Fading may be rapid or slow, the former type usually resulting from rapidly-changing conditions in the ionosphere, the latter occurring when transmission conditions are relatively stable.

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

Scatter

Even though the operating frequency is above the m.u.f. for a given distance, it is usually possible to hear signals from within the skip zone. This phenomenon, called **scatter,** is caused by random reflections from distances beyond the skip zone. Such reflections can occur when the transmitted energy strikes the earth at a distance and some of it is reflected back into the skip zone to the receiver. Other possible scatter sources are "patches" of ionization of different density than the average, or sporadic-E clouds (see later section). Scatter signals are weaker than those normally propagated, and also have a rapid fade or "flutter" that makes them easily recognizable.

It is probable that scatter also plays a considerable part in long-distance transmission (beyond the maximum one-hop distance) — particularly in cases where, with multihop propagation, the m.u.f. at some intermediate reflection point in the ionosphere is below the frequency actually being used.

● OTHER FEATURES OF IONOSPHERIC PROPAGATION

Cyclic Variations in the Ionosphere

Since ionization depends upon ultraviolet radiation, conditions in the ionosphere vary with changes in the sun's radiation. In addition to the daily variation, seasonal changes result in higher critical frequencies in the E layer in summer, averaging about 4 Mc. as against a winter average of 3 Mc. The F layer shows little variation, the critical frequency being of the order of 4 to 5 Mc. in the evening. The F_1 layer, which has a critical frequency near 5 Mc. in summer, usually disappears entirely in winter. The daytime maximum critical frequencies for the F_2 are highest in winter (10 to 12 Mc.) and lowest in summer (around 7 Mc.). The virtual height of the F_2 layer, which is about 185 miles in winter, averages 250 miles in summer. These values are representative of latitude 40 deg. North in the Western hemisphere, and are subject to considerable variation in other parts of the world.

Very marked changes in ionization also occur in step with the **11-year sunspot cycle.** Although there is no apparent direct correlation between sunspot activity and critical frequencies on a given day, there is a definite correlation between *average* sunspot activity and critical frequencies. The critical frequencies are highest during sunspot maxima and lowest during sunspot minima. During the period of minimum sunspot activity the lower frequencies — 7 and 3.5 Mc. — frequently are the only usable bands at night. At such times the 28-Mc. band is seldom useful for long-distance work, while the 14-Mc. band performs well in the daytime but is not ordinarily useful at night. The next sunspot minimum is forecast for the winter of 1954–55. The most recent maximum occurred in the winter of 1947–48.

Ionosphere Storms and Other Disturbances

Certain types of sunspot activity cause considerable disturbances in the ionosphere (**ionosphere storms**) and are accompanied by disturbances in the earth's magnetic field (**magnetic storms**). Ionosphere storms are characterized by a marked increase in absorption, so that radio conditions become poor. The critical frequencies also drop to relatively low values during a storm, so that only the lower frequencies are useful for communication. Ionosphere storms may last from a few hours to several days. Since the sun rotates on its axis once every 28 days, disturbances tend to recur at such intervals, if the sunspots responsible do not become inactive in the meantime. Absorption is usually low, and radio conditions therefore good, just preceding a storm.

Unusually high ionization in the region of the atmosphere below the normal ionosphere may increase absorption to such an extent that sky-wave transmission becomes difficult and sometimes even impossible. The length of such a disturbance may be several hours, with a gradual falling off of transmission conditions at the beginning and an equally gradual building up at the end of the period. **Fade-outs,** similar to the above in effect, are caused by sudden disturbances on the sun. They are characterized by very rapid ionization, with sky-wave transmission disappearing almost instantly, occur only in daylight, and do not last as long as the first type of absorption.

Magnetic storms frequently are accompanied by unusual auroral displays, creating an ionized "curtain" in the polar regions which can act as a reflector of radio waves. **Auroral reflection** may be observed on any frequency, depending upon the conditions, and it is always characterized by a flutter on all signals that makes voice work difficult. It is most noticeable in the northern latitudes and on signals traveling through the Auroral zone — that is, through the polar regions and over the North Atlantic.

Sporadic-E Ionization

Scattered patches or clouds of relatively dense ionization occasionally appear at heights approximately the same as that of the *E* layer. This **sporadic-*E*** ionization is most prevalent in the equatorial regions, where it is substantially continuous. In northern latitudes it is most frequent in the spring and early summer, but is present in some degree a fair percentage of the time the year 'round. It accounts for a good deal of the night-time short distance work on the lower frequencies (3.5 and 7 Mc.) and, when more intense, for similar work on 14 and 28 Mc. Exceptionally intense sporadic-*E* ionization is responsible for work over distances exceeding 400 or 500 miles on the 50-Mc. band.

There are indications of a relationship between sporadic-*E* ionization and average sunspot activity, but it does not appear to be directly related to daylight and darkness since it may occur at any time of the day. However, there is an apparent tendency for the ionization to peak at mid-morning and in the early evening.

Meteor Trails

A phenomenon that frequently occurs on signals from within the skip zone is a sudden increase in intensity, called a **burst.** Bursts are caused by meteors which, entering the earth's atmosphere at high speed, are followed by an ionized trail of rather high intensity. The ionization is caused by heating from the friction between the meteor and the air molecules in the ionosphere region. The ionization usually disappears in less than a second, but during that time it is often capable of reflecting signals up to 100 Mc. or so. The lower frequency limit depends on the length of the ionized trail. Bursts are frequently observed on the 14- and 28-Mc. bands, especially during those times of the year when "meteor showers" occur. When the meteor is moving in a direction somewhat parallel to the wave path, it can induce a rising or falling "whistle" on the signal, for a second or so.

Tropospheric Propagation

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

● PREDICTION CHARTS

The Central Radio Propagation Laboratory of National Bureau of Standards offers prediction charts three months in advance, by means of which it is possible to predict with considerable accuracy the maximum usable frequency that will hold over any path on the earth during a monthly period. The charts are based on ionosphere observations made at a number of stations throughout the world, coupled with considerable statistical data. They are conservative enough to enable the amateur to anticipate and plan his best operating times, particularly on the 14- and 28-Mc. bands. The charts can be obtained from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C. for 10 cents a copy or $1.00 per year on subscription. They are called "CRPL-D Basic Radio Propagation Predictions."

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 uncrowded 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 two major differences between receivers for 'phone reception and for c.w. reception. A 'phone signal has sidebands that make the signal take up about 6 or 8 kc. in the band, and the audio quality of the received signal is impaired if the passband of the receiver is less than half of this. On the other hand, a c.w. signal occupies only a few hundred cycles at the most, and consequently the passband of a c.w. receiver can be small. In either case, if the passband of the receiver is more than necessary, signals adjacent to the desired one can be heard, and the selectivity of the receiver is said to be poor. The detection process delivers directly the audio frequencies present as modulation on a 'phone signal. There is no modulation on a c.w. 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. If the source of the second radio frequency is a separate oscillator, the system is known as **heterodyne** reception; if the detector is made to oscillate and produce the second frequency, it is known as an **autodyne** detector. Modern superheterodyne receivers (described later) generally use a separate oscillator to generate the beat-note. Summing up the two differences, 'phone receivers can't use as much selectivity as c.w. receivers, and c.w. receivers require some kind of beating oscillator to give an audible signal. Broadcast receivers can receive only 'phone signals because no beat oscillator is included. **Communications receivers** include beat oscillators and often some means for varying the selectivity.

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 an audio output some stated amount (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 for comparing two receivers, because the passband width of the receiver plays a large part in the result.

The random motion of the molecules in the antenna and receiver circuits generates small voltages called **thermal-agitation noise** voltages. The frequency of this noise is random and the noise exists across the entire radio spectrum. Its amplitude increases with the temperature of the circuits. Only the noise in the antenna and first stage of a receiver is normally significant, since the noise developed in later stages is masked by the amplified noise from the first stage. The only noise that is amplified is that which falls within the receiver passband, so the noise appearing in the receiver output is less when the passband is reduced. Noise is also generated by the current flow within the first tube itself; this effect can be combined with the thermal noise and called **receiver noise.**

The limit of a receiver's ability to detect weak signals is the thermal noise generated in the input circuit. Even if a perfect noise-free tube were developed and used throughout the receiver, the limit to reception would be the thermal noise. (Atmospheric- and man-made noise is a *practical* limit below 20 Mc.) The degree to which a receiver approaches this ideal is called the **noise figure** of the receiver, and it is expressed as the ratio of noise power at the input of the receiver required to increase the noise output of the receiver 3 db. Since the noise power passed by the receiver is dependent on the passband, the figure shows how far the receiver departs from the ideal. The ratio is generally expressed in db., and runs around 6 to 12 db. for a good receiver, although figures of 2 to 4 db. have been obtained. Comparisons of noise figures can be made by the amateur with simple equipment. (See *QST*, August, 1949, page 20.)

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").

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 c.w. and 0.5 for 'phone. The minimum usable bandwidth at 6-db. down is about 150 cycles for c.w. reception and about 2000 cycles for 'phone.

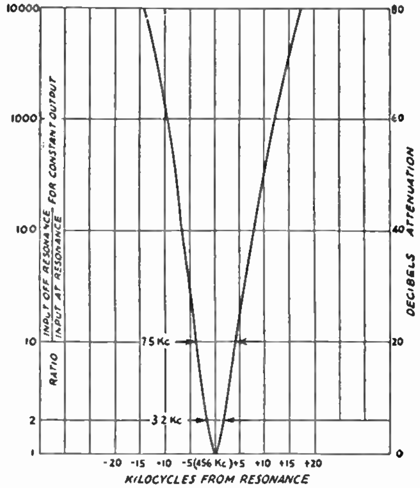


Fig. 5-1 — Typical selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios, the corresponding decibel steps are shown at the right.

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.

Fidelity

Fidelity is the relative ability of the receiver to reproduce in its output the modulation carried by the incoming signal. For perfect fidelity, the relative amplitudes of the various components must not be changed by passing through the receiver. However, in amateur communication the important requirement is to transmit intelligence and not "high-fidelity" signals.

Detection and Detectors

Detection is the process of recovering the modulation from a signal (see "Modulation, Heterodyning and Beats"). Any device that is "nonlinear" (i.e., whose output is not *exactly* proportional to its input) will act as a detector. It can be used as a detector if an impedance for the desired modulation frequency is connected in the output circuit.

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.

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.

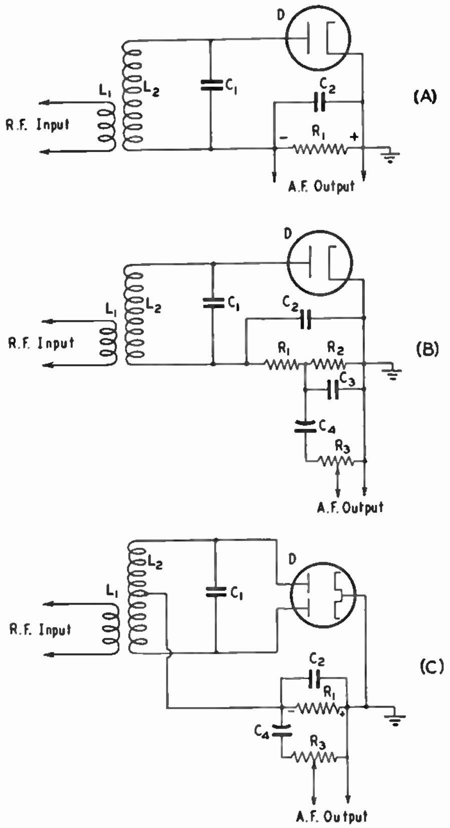


Fig. 5-2 — Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L_2C_1, is tuned to the signal frequency; typical values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

Circuits for both half-wave and full-wave diodes are given in Fig. 5-2. The simplified half-wave circuit at 5-2A includes the r.f. tuned circuit, L_2C_1, a coupling coil, L_1, from which the r.f. energy is fed to L_2C_1, and the diode, D, with its load resistance, R_1, and by-pass condenser, C_2. The flow of rectified r.f. current causes a d.c. voltage to develop across the terminals of R_1. The − and + signs show the polarity of the voltage. The variation in amplitude of the r.f. signal with modulation causes corresponding variations in the value of the d.c. voltage across R_1. In audio work the load resistor, R_1, is usually 0.1 megohm or higher, so that a fairly large voltage will develop from a small rectified-current flow.

The progress of the signal through the detector or rectifier is shown in Fig. 5-3. A typical modulated signal as it exists in the tuned circuit is shown at A. When this signal is applied to the rectifier tube, current will flow only during the part of the r.f. cycle when the plate is positive with respect to the cathode, so that the output of the rectifier consists of half-cycles of r.f. These current pulses flow in the load circuit comprised of R_1 and C_2, the resistance of R_1 and the capacity of C_2 being so proportioned that C_2 charges to the peak value of the rectified voltage on each pulse and retains enough charge between pulses so that the voltage across R_1 is smoothed out, as shown in C. C_2 thus acts as a filter for the radio-frequency component of the output of the rectifier, leaving a d.c. component that varies in the same way as the modulation on the original signal. When this varying d.c. voltage is applied to a following amplifier through a coupling condenser (C_4 in Fig. 5-2B), only the *variations* in voltage are transferred, so that the final output signal is a.c., as shown in D.

In the circuit at 5-2B, R_1 and C_2 have been divided for the purpose of providing a more effective filter for r.f. It is important to prevent the appearance of any r.f. voltage in the output of the detector, because it may cause overloading of a succeeding amplifier tube. The audio-frequency variations can be transferred to another circuit through a coupling condenser, C_4, 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 condenser 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 very little r.f. voltage appears across the load resistor, R_1, because the midpoint of L_2 is at the same potential as the cathode, or "ground" for r.f., and r.f. filtering is easier than in the half-wave circuit.

The reactance of C_2 must be small compared

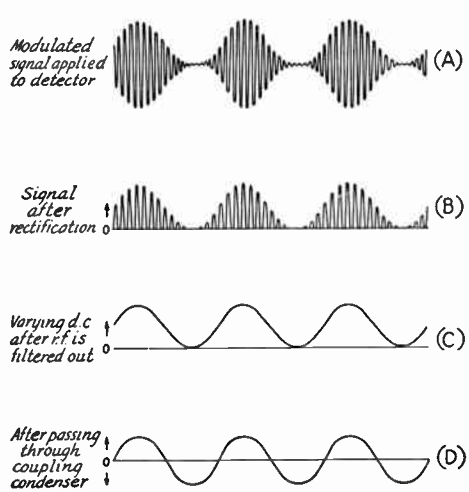


Fig. 5-3 — Diagrams showing the detection process.

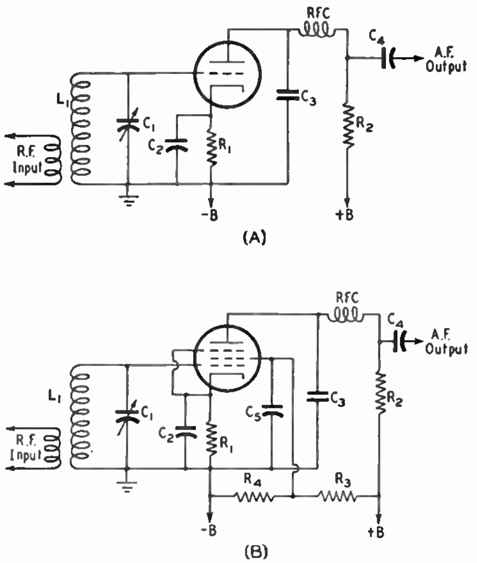


Fig. 5-4 — Circuits for plate detection. A, triode; B, pentode. The input circuit, L_1C_1, is tuned to the signal frequency. Typical values for the other components are:

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

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

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 amplitude in a fashion similar to the rectified current in a diode detector.

Circuits for triodes and pentodes are given in Fig. 5-4. C_3 is the plate by-pass condenser, and, with *RFC*, prevents r.f. from appearing in the output. The cathode resistor, R_1, provides the operating grid bias, and C_2 is a by-pass for both radio and audio frequencies. R_2 is the plate load resistance and C_4 is the output coupling condenser. In the pentode circuit at B, R_3 and R_4 form a voltage divider to supply the proper screen potential (about 30 volts), and C_5 is a by-pass condenser. 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 feed-back for the audio frequencies. The cathode resistor is by-passed for r.f. but not for audio, while the plate circuit is by-passed to ground for both audio and radio frequencies. R_2 forms, with C_3, an *RC* filter to isolate the plate from the "B" supply. An r.f. filter, consisting of a series r.f. choke and a shunt condenser, can be connected between the cathode and C_4 to eliminate any r.f. that might otherwise appear in the output.

The plate current is very low at no signal, increasing with signal as in the case of the plate detector. The voltage drop across R_1 consequently

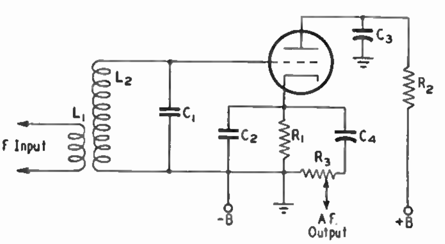


Fig. 5-5 — The infinite-impedance detector. The input circuit, L_2C_1, is tuned to the signal frequency. Typical values for the other components are:

C_2 — 250 μμfd.	R_1 — 0.15 megohm.
C_3 — 0.5 μfd.	R_2 — 25,000 ohms.
C_4 — 0.1 μfd.	R_3 — 0.25-megohm volume control.

A tube having a medium amplification factor (about 20) should be used. Plate voltage should be 250 volts.

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.

REGENERATIVE DETECTORS

By providing controllable r.f. feed-back (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 circuits of Fig. 5-6, the grid corresponds to the diode plate and the rectifying action is exactly the same as in a diode. The d.c. voltage from rectified-current flow through the grid leak, R_1, biases the grid negatively, 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, T_1, L_4 and L_3 are the plate load resistances, C_3 is a by-pass condenser and *RFC* an r.f. choke to eliminate r.f. in the output circuit.

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, as at 5-6 B and C. The operation is equivalent to that of the triode circuit. The screen by-pass condenser, C_5, should have low reactance for both radio and audio frequencies. R_2 and R_3 constitute a voltage divider on the plate supply to furnish the proper screen voltage. In both circuits, C_2 must have low r.f. reactance and high a.f. reactance compared to the resistance of R_1. 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.

The circuits in Fig. 5-6 are regenerative, the feed-back being obtained by feeding some signal to the grid back from the plate circuit. The amount of regeneration must be controllable, because maximum regenerative amplification is secured at the critical point where the circuit is just about to oscillate. The critical point in turn depends upon circuit conditions, which may vary with the frequency to which the detector is tuned. In the oscillating condition, a regenerative detector can be detuned slightly from an incoming c.w. signal to give *autodyne* reception.

The circuit of Fig. 5-6A uses a variable by-pass condenser, C_3, in the plate circuit to control regeneration. When the capacity is small the tube does not regenerate, but as it increases toward maximum its reactance becomes smaller until there is sufficient feed-back to cause oscillation. If L_2 and L_3 are wound end-to-end in the same direction, the plate connection is to the outside of the plate or "tickler" coil, L_3, when the grid connection is to the outside of L_2.

The circuit of 5-6B is for a pentode tube, regeneration being controlled by adjustment of the screen-grid voltage. The tickler, L_3, is in the plate circuit. The portion of the control resistor between the rotating contact and ground is by-passed by a large condenser (0.5

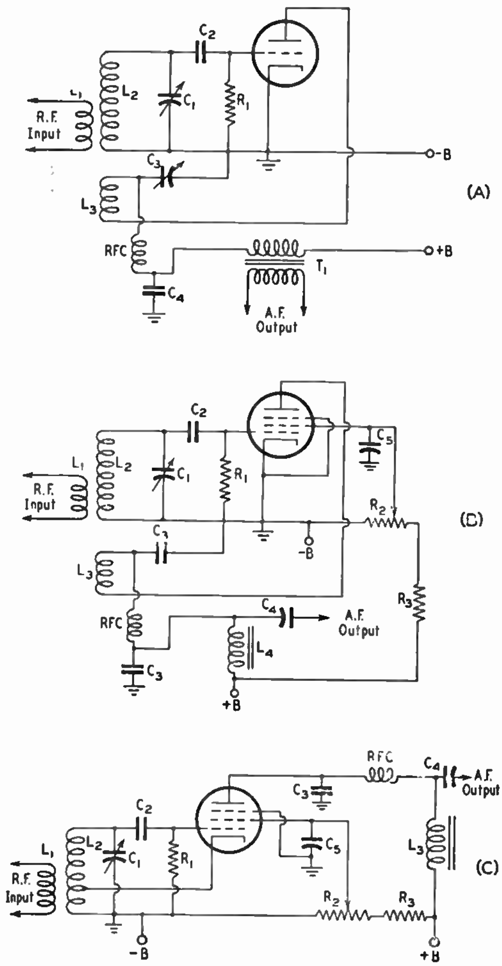


Fig. 5-6 — Triode and pentode regenerative detector circuits. The input circuit, L_2C_1, is tuned to the signal frequency. The grid condenser, C_2, should have a value of about 100 $\mu\mu$fd. in all circuits; the grid leak, R_1, may range in value from 1 to 5 megohms. The tickler coil, L_3, ordinarily will have from 10 to 25 per cent of the number of turns on L_2; in C, the cathode tap is about 10 per cent of the number of turns on L_2 above ground. Regeneration-control condenser C_3 in A should have a maximum capacity of 100 $\mu\mu$fd. or more; by-pass condensers C_3 in B and C are likewise 100 $\mu\mu$fd. C_5 is ordinarily 1 μfd. or more; R_2, a 50,000-ohm potentiometer; R_3, 50,000 to 100,000 ohms. L_4 in B (L_3 in C) is a 500-henry inductance, C_4 is 0.1 μfd. in both circuits. T_1 in A is a conventional audio transformer for coupling from the plate of a tube to a following grid. *RFC* is 2.5 mh. In A, the plate voltage should be about 50 volts for best sensitivity. Pentode circuits require about 30 volts on the screen; plate potential may be 100 to 250 volts.

μfd. or more) to filter out scratching noise when the arm is rotated. The feed-back is adjusted by varying the number of turns on L_3 or the coupling between L_2 and L_3, until the tube just goes into oscillation at a screen potential of approximately 30 volts.

Circuit C is identical with B in principle of operation. Since the screen and plate are in parallel for r.f. in this circuit, only a small amount of "tickler" — that is, relatively few turns between the cathode tap and ground — is required for oscillation.

Smooth Regeneration Control

The ideal regeneration control would permit the detector to go into and out of oscillation smoothly, would have no effect on the frequency of oscillation, and would give the same value of regeneration regardless of frequency and the loading on the circuit. In practice, the effects of loading, particularly the loading that occurs when the detector circuit is coupled to an antenna, are difficult to overcome. Likewise, the regeneration is usually affected by the frequency to which the grid circuit is tuned.

In all circuits it is best to wind the tickler at the ground or cathode end of the grid coil, and to use as few turns on the tickler as will allow the detector to oscillate easily over the whole tuning range at the plate (and screen, if a pentode) voltage that gives maximum sensitivity. Should the tube break into oscillation suddenly as the regeneration control is advanced, making a click, it usually indicates that the coupling to the antenna (or r.f. amplifier) is too tight. The wrong value of grid leak plus too-high plate and screen voltage are also frequent causes of lack of smoothness in going into oscillation.

Antenna Coupling

If the detector is coupled to an antenna, slight changes in the antenna (as when the wire swings in a breeze) affect the frequency of the oscillations generated, and thereby the beat frequency when c.w. signals are being received. The tighter the antenna coupling is made, the greater will be the feedback required or the higher will be the voltage necessary to make the detector oscillate. The antenna coupling should be the maximum that will allow the detector to go into oscillation smoothly with the correct voltages on the tube. If capacity coupling to the grid end of the coil is used, generally only a very small amount of capacity will be needed to couple to the antenna. Increasing the capacity increases the coupling.

At frequencies where the antenna system is resonant the absorption of energy from the oscillating detector circuit will be greater, with the consequence that more regeneration is needed. In extreme cases it may not be possible to make the detector oscillate with normal voltages. The remedy for these "dead spots" is to loosen the antenna coupling to a point that permits normal oscillation and smooth regeneration control.

Body Capacity

A regenerative detector occasionally shows a tendency to change frequency slightly as the hand is moved near the dial. This condition (**body capacity**) can be corrected by better shielding, and sometimes by r.f. filtering of the 'phone leads. A good, short ground connection and loosening the coupling to the antenna will help.

Hum

Hum at the power-supply frequency, even when using battery plate supply, may result from the use of a.c. on the tube heater. Effects of this type normally are troublesome only when the circuit of Fig. 5-6C is used, and then only at 14 Mc. and higher. Connecting one side of the heater supply to ground, or grounding the center-tap of the heater-transformer winding, will reduce the hum. The heater wiring should be kept as far as possible from the r.f. circuits.

House wiring, if of the "open" type, may cause hum if the detector tube, grid lead, and grid condenser and leak are not shielded. This type of hum is easily recognizable because of its rather high pitch.

Tuning

For c.w. reception, the regeneration control is advanced until the detector breaks into a "hiss," which indicates that the detector is oscillating. Further advancing the regeneration control after the detector starts oscillating will result in a slight decrease in the strength of the hiss, indicating that the sensitivity of the detector is decreasing.

The proper adjustment of the regeneration control for best reception of c.w. signals is where the detector just starts to oscillate. Then c.w. signals can be tuned in and will give a tone with each signal depending on the setting of the tuning control. As the receiver is tuned through a signal the tone first will be heard as a very high pitch, then will go down through "zero beat" and rise again on the other side, finally

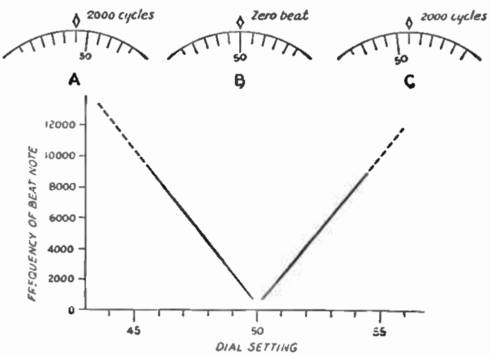


Fig. 5-7 — **As the tuning dial of a receiver is turned past a c.w. signal, the beat-note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past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

disappearing at a very high pitch. This behavior is shown in Fig. 5-7. 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. This phenomenon, is also called **"locking-in"**; the more stable of the two frequencies assumes control over the other. 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 c.w. reception. Further advancing the regeneration control makes the receiver less susceptible to blocking by strong signals, but also less sensitive to weak signals.

If the detector is in the oscillating condition and a 'phone signal is tuned in, a steady audible beat-note will result. While it is possible to listen to 'phone if the receiver can be tuned to exact zero beat, it is more satisfactory to reduce the regeneration to the point just before the receiver goes into oscillation. This is also the most sensitive operating point.

Tuning and Band-Changing Methods

Band-Changing

The resonant circuits that are tuned to the frequency of the incoming signal constitute a special problem in the design of amateur receivers, since the amateur frequency assignments consist of groups or bands of frequencies at widely-spaced intervals. The same coil and tuning condenser cannot be used for, say, 14 Mc. to 3.5 Mc., because of the impracticable maximum-to-minimum capacity ratio required, and also because the tuning would be excessively critical with such a large frequency range. It is necessary, therefore, to provide a means for changing the circuit constants for various frequency bands. As a matter of convenience the same tuning condenser usually is retained, but new coils are inserted in the circuit for each band.

One method of changing inductances is to use a switch having an appropriate number of contacts, which connects the desired coil and disconnects the others. The unused coils are sometimes short-circuited by the switch, to avoid the possibility of undesirable self resonances in the unused coils. 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 coils are advantageous when space in a multiband receiver is at a premium. They are also very useful when considerable experimental work is involved, because they are easier to work on than coils clustered around a switch.

Bandspreading

The tuning range of a given coil and variable condenser will depend upon the inductance of the coil and the change in tuning capacity. For ease of tuning, it is desirable to adjust the tuning range so that practically the whole dial scale is occupied by the band in use. This is called **bandspreading.** Because of the varying widths of the bands, special tuning methods must be devised to give the correct maximum-minimum capacity ratio on each band. Several of these methods are shown in Fig. 5-8.

In A, a small **bandspread condenser,** C_1 (15- to 25-μμfd. maximum capacity), is used in parallel with a condenser, C_2, which is usually large enough (100 to 140 μμfd.) to cover a 2-to-1 frequency range. The setting of C_2 will determine the minimum capacity of the circuit, and the maximum capacity for bandspread tuning will be the maximum capacity of C_1 plus the setting of C_2. The inductance of the coil can be adjusted so that the maximum-minimum ratio will give adequate bandspread. It is almost impossible, because of the non-harmonic relation of the various band limits, to get full bandspread on all bands with the same pair of condensers. C_2 is variously called the **band-setting** or **main-tuning** condenser. It must be reset each time the band is changed.

The method shown at B makes use of condensers in series. The tuning condenser, C_1, may have a maximum capacity of 100 μμfd. or more. The minimum capacity is determined principally by the setting of C_3, which usually has low capacity, and the maximum capacity by the setting of C_2, which is of the order of 25 to 50 μμfd. This method is capable of close adjustment to practically any desired degree of bandspread. Either C_2 and C_3 must be adjusted for each band or separate preadjusted condensers must be switched in.

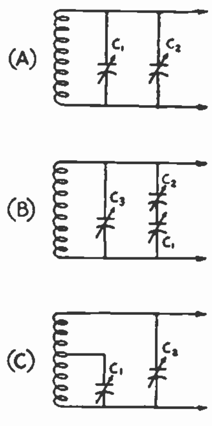


Fig. 5-8 — Essentials of the three basic band-spread tuning systems.

The circuit at C also gives complete spread on each band. C_1, the bandspread condenser, may have any convenient value of capacity; 50 μμfd. is satisfactory. C_2 may be used for continuous frequency coverage ("general coverage") and as a band-setting condenser. The effective maximum-minimum capacity ratio depends upon the capacity of C_2 and the point at which C_1 is tapped on the coil. The nearer the tap to the bottom of the coil, the greater the bandspread, and vice versa. For a given coil and tap, the bandspread will be greater if C_2 is set at higher capacity. C_2 may be mounted in the plug-in coil form and preset, if desired.

This requires a separate condenser for each band, but eliminates the necessity for resetting C_2 each time the band is changed.

Ganged Tuning

The tuning condensers of the several r.f. circuits may be coupled together mechanically and operated by a single control. However, this operating convenience involves more complicated construction, both electrically and mechanically. It becomes necessary to make the various circuits **track** — that is, tune to the same frequency at each setting of the tuning control.

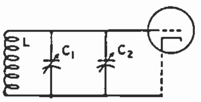


Fig. 5-9 — Showing the use of a trimmer condenser to set the minimum circuit capacity in order to obtain true tracking for gang-tuning.

True tracking can be obtained only when the inductance, tuning condensers, and circuit inductances and minimum and maximum capacities are identical in all "ganged" stages. A small **trimmer** or **padding** condenser may be connected across the coil, so that variations in minimum capacity can be compensated. The fundamental circuit is shown in Fig. 5-9, where C_1 is the trimmer and C_2 the tuning condenser. The use of the trimmer necessarily increases the minimum circuit capacity, but it is a necessity for satisfactory tracking. Midget condensers having maximum capacities of 15 to 30 μμfd. are commonly used.

The same methods are applied to bandspread circuits that must be tracked. The circuits are identical with those of Fig. 5-8. If both general-coverage and bandspread tuning are to be available, an additional trimmer condenser must be connected across the coil in each circuit shown. If only amateur-band tuning is desired, however, then C_3 in Fig. 5-8B, and C_2 in Fig. 5-8C, serve as trimmers.

The coil inductance can be adjusted by starting with a larger number of turns than necessary and removing a turn or fraction of a turn at a time until the circuits track satisfactorily. An alternative method, provided the inductance is reasonably close to the correct value initially, is to make the coil so that the last turn is variable with respect to the whole coil, or to use a single short-circuited turn the position of which can be varied with respect to the coil. The application of these methods is shown in Fig. 5-10.

Still another method for trimming the inductance is to use an adjustable brass (or copper) or powdered-iron core. The brass core acts like a single shorted turn, and the inductance of the coil is decreased as the brass core, or **"slug,"** is moved into the coil. The powdered-iron core has the opposite effect, and *increases* the inductance as it is moved into the coil. The Q of the coil is not affected materially by the use of the brass slug, provided the brass slug has a clean surface or is silverplated. The use of the powdered-iron core will 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.

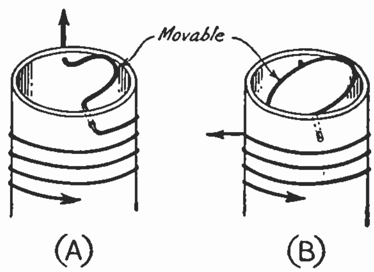


Fig. 5-10 — Methods of adjusting the inductance for ganging. The half-turn in A can be moved so that its magnetic field either aids or opposes the field of the coil. The shorted loop in B is not connected to the coil, but operates by induction. It will have no effect on the coil inductance when the axis of the loop is perpendicular to the axis of the coil, and will give maximum reduction of the coil inductance when rotated 90°. The loop can be a solid disk of metal and give exactly the same effect.

The Superheterodyne

For many years (up to about 1932) practically the only type of receiver to be found in amateur stations consisted of a regenerative detector and one or more stages of audio amplification. Receivers of this type can be made quite sensitive but 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.** C.w. signals are made audible by autodyne or heterodyne reception at the second detector.

As a numerical example, assume that an intermediate frequency of 455 kc. is chosen and that the incoming signal is 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 c.w. signal at the second detector of, say, 1000 cycles, the autodyning or heterodyning oscillator would be set to either 454 or 456 kc.

The frequency-conversion process permits r.f. amplification at a relatively low frequency, the i.f. High selectivity and gain can be obtained at this frequency, and this selectivity and gain are constant. The separate oscillators can be designed for 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 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 to 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.

The mixer should not require too much r.f. power from the h.f. oscillator, since it may be difficult to supply the power and yet maintain good oscillator stability. Also, the conversion efficiency should not depend too critically on the oscillator voltage (that is, a small change in oscillator output should not change the gain), since it is difficult to maintain constant output over a wide frequency range.

A change in oscillator frequency caused by tuning of the mixer grid circuit is called **pulling.** Pulling should be minimized, because the stability of the whole receiver depends critically upon the stability of the h.f. oscillator. Pulling decreases with separation of the signal and h.f.-oscillator frequencies, being less with high intermediate frequencies. Another type of pulling is caused by regulation in the power supply. Strong signals cause the supply voltage to change, and this in turn shifts the oscillator frequency.

Circuits

If the first detector and high-frequency oscillator are separate tubes, the first detector is called a "mixer." If the two are combined in one envelope (as is often done for reasons of economy or efficiency), the first detector is called a "converter." In either case the function is the same.

Typical mixer circuits are shown in Fig. 5-11. The variations are chiefly in the way in which the oscillator voltage is introduced. In 5-11A, a pentode functions as a plate detector; the oscillator voltage is capacity-coupled to the grid of the tube through C_2. Inductive coupling may be used instead. The conversion gain and input selectivity generally are good, so long as the sum of the two voltages (signal and oscillator) impressed on the mixer grid does not exceed the grid bias. It is desirable to make the oscillator voltage as high as possible without exceeding this limitation. The oscillator power required is negligible. If the signal frequency is only 5 or 10 times the i.f., it may be difficult to develop enough oscillator voltage at the grid (because of the selectivity of the tuned input circuit). However, the circuit is a sensitive one and makes a good mixer, particularly with high-G_m tubes like the 6AC7 and 6AK5. A good triode also works well in the circuit, and tubes like the 7F8 (one section), the 6J6 (one section), the 12AT7 (one section), and the 6J4 work well. When a triode is used, the signal frequency must be short-circuited in the plate circuit, and this is done by connecting the tuning capacitor of the i.f. transformer directly from plate to cathode.

It is difficult to avoid "pulling" in a triode or pentode mixer, and a pentagrid mixer tube provides much better isolation. A typical circuit is shown in Fig. 5-11B, and tubes like the 6SA7, 7Q7 or 6BE6 are commonly used. The oscillator voltage is introduced 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

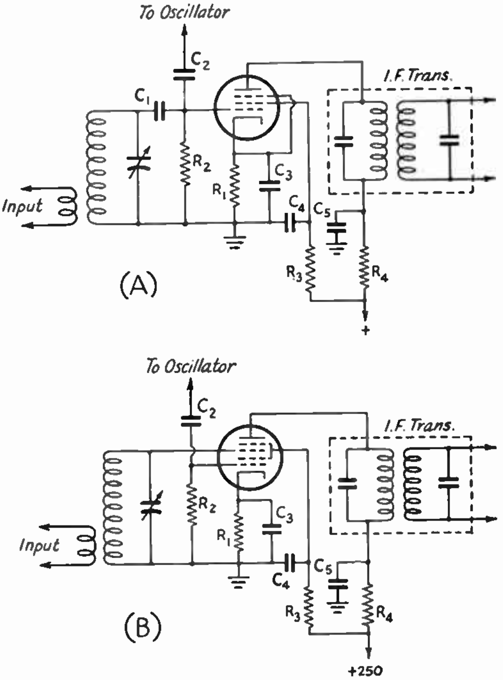


Fig. 5-11 — Typical circuits for separately-excited mixers. Grid injection of a pentode mixer is shown at A, and separate excitation of a pentagrid converter is given in B. Typical values for B will be found in Table 5-I — the values below are for the pentode mixer of A.

C_1 — 10 to 50 $\mu\mu$fd. R_2 — 1.0 megohm.
C_2 — 5 to 10 $\mu\mu$fd. R_3 — 0.47 megohm.
C_3, C_4, C_5 — 0.001 μfd. R_4 — 1500 ohms.
R_1 — 6800 ohms.

Positive supply voltage can be 250 volts with a 6AC7, 150 with a 6AK5.

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 feed-back of the oscillator circuit to give the proper oscillator r.f. voltage. This condition is checked by measuring the d.c. current flowing in grid resistor R_2.

A more stable receiver generally results, particularly at the higher frequencies, when separate tubes are used for the mixer and oscillator. Practically the same number of circuit com-

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](6AE8[1])	240	27,000	47,000	0.15–0.2	—	—	—	—
6SA7[2](7Q7[3])	0	18,000	22,000	0.5	150	18,000	22,000	0.5
6SB7Y[2]	0	15,000	22,000	0.35	68	15,000	22,000	0.35

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

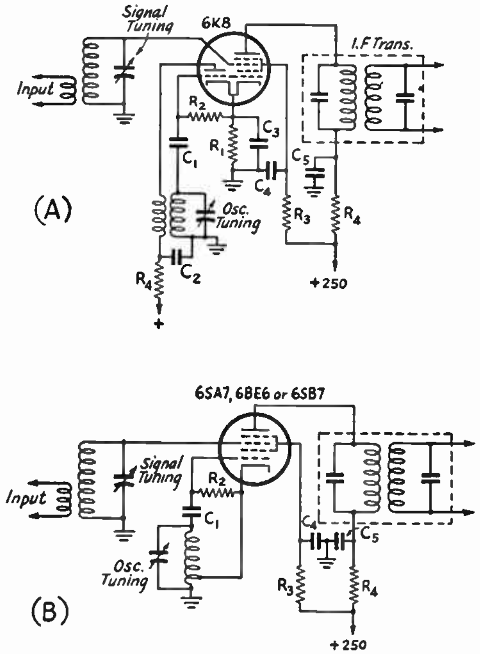


Fig. 5-12 — **Typical circuits for triode-hexode (A) and pentagrid (B) converters. Values for R_1, R_2 and R_3 can be found in Table 5-1; others are given below.**

C_1 — 47 μμfd.	C_3 — 0.01 μfd.
C_2, C_4, C_5 — 0.001 μfd.	R_4 — 1000 ohms.

ponents 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 5-1. The grid leak referred to is the oscillator grid leak or injection-grid return, R_2 of Figs. 5-11 and 5-12.

The effectiveness of converter tubes of the type just described becomes less as the signal frequency is increased. Some oscillator voltage will be coupled to the signal grid through "space-charge" coupling, an effect that increases with frequency. If there is relatively little frequency difference between oscillator and signal, as for example a 14- or 28-Mc. signal and an i.f. of 455 kc., this voltage can become considerable because the selectivity of the signal circuit will be unable to reject it. 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 percent of the signal frequency, for best results.

Audio Converters

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

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

THE HIGH-FREQUENCY OSCILLATOR

Stability of the receiver is dependent chiefly upon the stability of the h.f. oscillator, and particular care should be given this part of the receiver. The frequency of oscillation should be insensitive to mechanical shock and changes

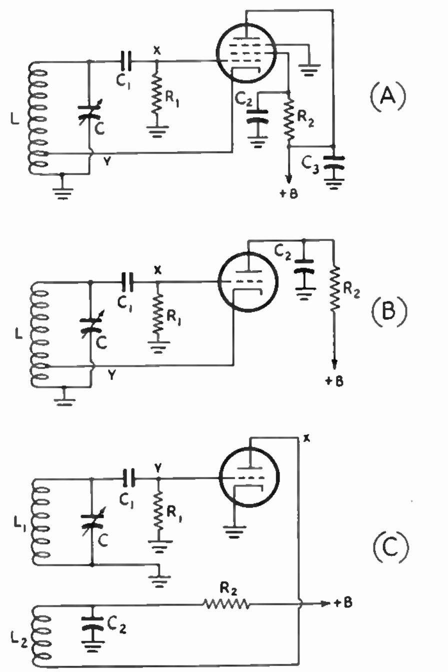


Fig. 5-13 — **High-frequency oscillator circuits. A, pentode grounded-plate oscillator; B, triode grounded-plate oscillator; C, triode oscillator with tickler circuit. Coupling to the mixer may be taken from points X and Y. In A and B, coupling from Y will reduce pulling effects, but gives less voltage than from X; this type is best adapted to mixer circuits with small oscillator-voltage requirements. Typical values for components are as follows:**

	Circuit A	*Circuit B*	*Circuit C*
C_1 —	100 μμfd.	100 μμfd.	100 μμfd.
C_2 —	0.1 μfd.	0.1 μfd.	0.1 μfd.
C_3 —	0.1 μfd.		
R_1 —	47,000 ohms.	47,000 ohms.	47,000 ohms.
R_2 —	47,000 ohms.	10,000 to 25,000 ohms.	10,000 to 25,000 ohms.

The plate-supply voltage should be 250 volts. In circuits B and C, R_2 is used to drop the supply voltage to 100–150 volts; it may be omitted if voltage is obtained from a voltage divider in the power supply.

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 condenser too close to a tube. Propping up the lid of a receiver will often reduce drift by lowering the terminal temperature of the unit.

Sensitivity to vibration and shock can be minimized by using good mechanical support for coils and tuning condensers, 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 condensers. They should have good alignment and no back-lash. If the condensers are mounted off the chassis on posts instead of brackets, it is almost impossible to avoid some back-lash unless the posts have extra-wide bases. The condensers should be selected with good wiping contacts to the rotor, since with age the rotor contacts can be a source of erratic tuning. All joints in the oscillator tuning circuit should be carefully soldered, 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, preferably 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. 5-13. The point at which output voltage is taken for the mixer is indicated in each case by *X* or *Y*. Circuits A and B will give about the same results, and require only one coil. However, in these two circuits the cathode is above ground potential for r.f., which often is a cause of hum modulation of the oscillator output at 14 Mc. and higher frequencies when a.c.-heated-cathode tubes are used. The circuit of Fig. 5-13C reduces hum because the cathode is grounded. It is 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 feed-back to obtain optimum results. Too much feed-back may cause the oscillator to "squeg" and generate several frequencies simultaneously; too little feed-back will cause the output to be low. In the tapped-coil circuits (A, B), the feed-back is increased by moving the tap toward the grid end of the coil. Using the oscillator shown at C, feed-back is obtained by increasing the number of turns on L_2 or by moving L_2 closer to L_1.

The Intermediate-Frequency Amplifier

One major advantage of the superhet is that high gain and selectivity can be obtained by using a good i.f. amplifier. This can be a one-stage affair in simple receivers, or two or three stages in the more 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 selectivity and gain are lowered. The difference in gain is least important.

An i.f. of the order of 455 kc. gives good selectivity and is satisfactory from the standpoint of image ratio and oscillator pulling at frequencies up to 7 Mc. The image ratio is poor at 14 Mc. when the mixer is connected to the antenna, but adequate when there is a tuned r.f. amplifier between antenna and mixer. At 28 Mc. and on the very-high frequencies, the image ratio is very poor unless several r.f. stages are used. Above 14 Mc., pulling is likely to be bad unless very loose coupling can be used between mixer and oscillator.

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

In choosing an i.f. it is wise to avoid frequencies on which there is considerable activity by the various radio services, since such signals may be picked up directly on the i.f. wiring. 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, this means that the i.f. amplifier should amplify equally well all frequencies within that band. In other words, the amplification must be uniform over a band 5 kc. wide, when the carrier is set at one edge. If the carrier is set in the center, a 10-kc. band 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 if the receiver is used in the lower-frequency bands.

Circuits

I.f. amplifiers usually consist of one or two stages. At 455 kc. two stages generally give all the gain usable, and also give suitable selectivity for 'phone reception.

A typical circuit arrangement is shown in Fig. 5-14. A second stage would simply duplicate the circuit of the first. The i.f. amplifier practically always uses a remote cut-off pentode-type tube operated as a Class A amplifier. For maximum selectivity, double-tuned transformers are used for interstage coupling, although single-tuned circuits or transformers with untuned primaries can be used for coupling, with a consequent loss in selectivity. All other things being equal, the selectivity of an i.f. amplifier is proportional to the number of tuned circuits in it.

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

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

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

When two stages are used the high gain will tend to cause instability and oscillation, so that good shielding, by-passing, and careful circuit arrangement to prevent stray coupling, with exposed r.f. leads well separated, are necessary.

I.F. Transformers

The tuned circuits of i.f. amplifiers are built up as transformer units consisting of a metal shield container in which the coils and tuning condensers are mounted. Both air-core and powdered iron-core universal-wound coils are used, the latter having somewhat higher *Q*s and hence greater selectivity and gain. 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

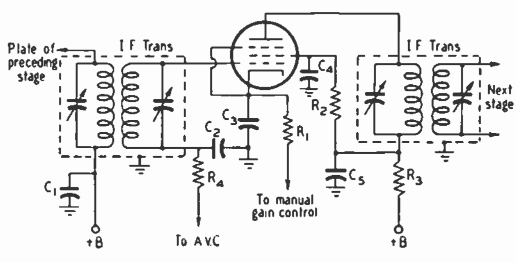


Fig. 5-14 — Typical intermediate-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

C_1 — 0.1 µfd. at 455 kc.: 0.01 µfd. at 1600 kc. and higher.
C_2 — 0.01 µfd.
C_3, C_4, C_5 — 0.1 µfd. at 455 kc.; 0.01 µfd. above 1600 kc.
R_1, R_2 — See Table 5-II. R_3 — 1800 ohms.
R_4 — 0.22 megohm.

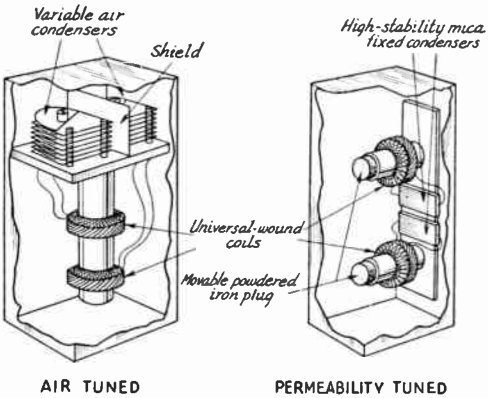


Fig. 5-15 — Representative i.f.-transformer construction. Coils are supported on insulating tubing or (in the air-tuned type) on wax-impregnated wooden dowels. The shield in the air-tuned transformer prevents capacity coupling between the tuning condensers. In the permeability-tuned transformer the cores consist of finely-divided iron particles supported in an insulating binder, formed into cylindrical "plugs." The tuning capacity is fixed, and the inductances of the coils are varied by moving the iron plugs in and out.

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

For tuning, air-dielectric tuning condensers 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-condenser tuning can be obtained by use of high-stability fixed mica condensers. Such stability is of great importance, since a circuit whose frequency "drifts" with time eventually will be tuned to a different frequency than the other circuits, thereby reducing the gain and selectivity of the amplifier. Typical i.f.-transformer construction is shown in Fig. 5-15.

Besides the type of i.f. transformer shown in Fig. 5-15, special units to give desired selectivity characteristics are available. For higher-than-ordinary adjacent-channel selectivity **triple-tuned** transformers, with a third tuned circuit inserted between the input and output windings, are 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. **Variable-selectivity** transformers also can be obtained. These usually are provided with a third (untuned) winding which can be connected to a resistor, thereby loading the tuned circuits and decreasing the Q to broaden the selectivity curve. The resistor is switched in and out of the circuit to vary the selectivity. Another method 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 applications.

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)	0.8	1.4	2.8
One stage, 455 kc. (air core)	8.7	17.8	32.3
One stage, 455 kc. (iron core)	4.3	10.3	20.4
Two stages, 455 kc. (iron core)	2.9	6.4	10.8
Two stages, 1600 kc.	11.0	16.6	27.4
Two stages, 5000 kc.	25.8	46.0	100.0

Tubes for I.F. Amplifiers

Variable-μ (remote cut-off) pentodes are almost invariably used in i.f. amplifier stages, since grid-bias gain control is practically always applied to the i.f. amplifier. Tubes with high plate resistance will have least effect on the selectivity of the amplifier, and those with high mutual conductance will give greatest gain. The choice of i.f. tubes has practically no effect on the signal-to-noise ratio, since this is determined by the preceding mixer and r.f. amplifier.

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 by-pass condenser directly on the bottom of the socket, crosswise between the plate and grid pins, to provide additional shielding. The outside foil of the condenser should be grounded.

● 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

TABLE 5-II
Cathode and Screen-Dropping Resistors for R.F. or I.F. Amplifiers

Tube	*Plate Volts*	*Screen Volts*	*Cathode Resistor*	*Screen Resistor*
6AB7[1]*	300		200 ohms	33,000 ohms
6AC7[1]	300		160	62,000
6AK5[2]	180	120	200	27,000
6AU6[2]	250	150	68	33,000
6BA6[2]*	250	100	68	33,000
6BH6[2]	250	150	100	33,000
6BJ6[2]*	250	100	82	47,000
6J7[1]	250	100	1200	270,000
6K7[1]*	250	125	240	47,000
6SG7[1]*	250	125	68	27,000
6SJ7[1]*	250	150	200	47,000
6SH7[1]	250	150	68	39,000
6SJ7[1]	250	100	820	180,000
6SK7[1]*	250	100	270	56,000
7G7/1232[3]	250	100	270	68,000
7H7[3]*	250	150	180	27,000

[1] Octal base, metal. [2] Miniature tube. [3] Lock-in base.
* Remote cut-off type.

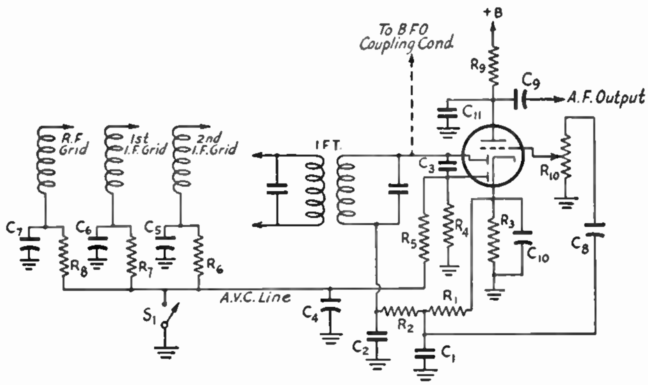


Fig. 5-16 — **Automatic volume-control circuit using a dual-diode-triode as a combined a.v.c. rectifier, second detector and first a.f. amplifier.**

R_1 — 0.27 megohm.
R_2 — 47,000 to 220,000 ohms.
R_3 — 1800 ohms.
R_4 — 2 to 5 megohms.
R_5 — 0.47 to 1 megohm.
R_6, R_7, R_8, R_9 — 0.22 megohm.
R_{10} — 0.5-megohm variable.
C_1, C_2, C_3 — 100 $\mu\mu$fd.
C_4 — 0.1 μfd.
C_5, C_6, C_7 — 0.01 μfd.
C_8, C_9 — 0.01 to 0.1 μfd.
C_{10} — 5- to 10-μfd. electrolytic.
C_{11} — 270 $\mu\mu$fd.

great amplification ahead of it. Therefore, the ability to handle large signals without distortion is preferable to high sensitivity. Plate detection is used to some extent, but the diode detector is most popular. It is especially adapted to furnishing automatic gain or volume control. The basic circuits have been described, although in many cases the diode elements are incorporated in a multipurpose tube that contains an amplifier section in addition to the diode.

The Beat Oscillator

Any standard oscillator circuit may be used for the beat oscillator required for heterodyne reception. Special beat-oscillator transformers are available, usually consisting of a tapped coil with adjustable tuning; these are most conveniently used with the circuits shown in Fig. 5-13A and B, with the output taken from *Y*. A variable condenser of about 25-$\mu\mu$fd. capacity may be connected between cathode and ground to provide fine adjustment of the frequency. The beat oscillator usually is coupled to the second-detector tuned circuit through a fixed condenser of a few $\mu\mu$fd. capacity.

The beat oscillator should be well shielded, to prevent coupling to any part of the 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 VOLUME 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 as the number of stages to which the a.v.c. bias is applied is increased. Control of at least two stages is advisable.

Circuits

A typical circuit using a diode-triode type tube as a combined a.v.c. rectifier, detector and first audio amplifier is shown in Fig. 5-16. One plate of the diode section of the tube is used for signal detection and the other for a.v.c. rectification. The a.v.c. diode plate is fed from the detector diode through the small coupling condenser, C_3. A negative bias voltage resulting from the flow of rectified carrier current is developed across R_4, the diode load resistor. This negative voltage is applied to the grids of the controlled stages through the filtering resistors, R_5, R_6, R_7 and R_8. When S_1 is closed the a.v.c. line is grounded, removing the a.v.c. bias from the amplifiers.

It does not matter which of the two diode plates is selected for audio and which for a.v.c. Frequently the two plates are connected together and used as a combined detector and a.v.c. rectifier. This could be done in Fig. 5-16. The a.v.c. filter and line would connect to the junction of R_2 and C_2, while C_3 and R_4 would be omitted from the circuit.

Delayed A.V.C.

In Fig. 5-16 the audio-diode return is made directly to the cathode and the a.v.c. diode is returned to ground. This places bias on the a.v.c. diode equal to the d.c. drop through the cathode resistor (a volt or two) and thus delays the application of a.v.c. voltage to the amplifier grids, since no rectification takes place in the a.v.c. diode circuit until the carrier amplitude is large enough to overcome the bias. Without this delay the a.v.c. would start working even with a very small signal. This is undesirable, because the full amplification of the receiver then could not be realized on weak signals. In the audio-diode circuit fixed bias would cause distortion, so the return there is directly to the cathode.

Time Constant

The time constant of the resistor-condenser combinations in the a.v.c. circuit is an important part of the system. It must be high enough so that the modulation on the signal is completely filtered from the d.c. output, leaving only an average d.c. component which follows the relatively slow carrier variations with fading. Audio-frequency variations in the a.v.c. voltage applied to the amplifier grids would reduce the percentage of modulation on the incoming signal. But the time constant must not be too great or the a.v.c. will be unable to follow rapid fading. The capacitance and resistance values indicated in Fig. 5-16 will give a time constant that is satisfactory for average reception.

C.W.

A.v.c. can be used for c.w. reception but the circuit is more complicated. The a.v.c. voltage must be derived from a rectifier that is isolated from the beat-frequency oscillator (otherwise the rectified b.f.o. voltage will reduce the receiver gain even with no signal coming through). This is generally done by using a separate a.v.c. channel connected to an i.f. amplifier stage ahead of the second detector (and b.f.o.). If the selectivity ahead of the a.v.c. rectifier isn't good, strong adjacent signals will develop a.v.c. voltages that will reduce the receiver gain while listening to weak signals. When clear channels are available, however, c.w. a.v.c. will hold the receiver output constant over a wide range of signal input. A.v.c. systems designed to work on c.w. signals must have fairly long time constants to work with slow-speed sending, and often a selection of time constants is made available.

Amplified A.V.C.

The a.v.c. system shown in Fig. 5-16 will not hold the audio output of the receiver exactly constant, although the variation becomes less as more stages are controlled by the a.v.c. voltage. The variation also becomes less as the delay voltage is increased, although there will, of course, be variation in output if the signal intensity is below the delay-voltage level at the a.v.c. rectifier. In the circuit of Fig. 5-16, the delay voltage is set by the proper operating bias for the triode portion of the tube. However, a separate diode may be used, as shown in Fig. 5-17A. Since such a system requires a large voltage at the diode, a separate i.f. stage is sometimes used to feed the delayed a.v.c. diode, as in Fig. 5-17B. A system like this, often called an "amplified a.v.c." system, gives superlative control action, since it maintains full receiver sensitivity for weak signals and substantially uniform audio output over a very wide range of signal strengths. To avoid a slight decrease in signal volume "on tune," the transformer coupling V_2 to V_3 should not be selective.

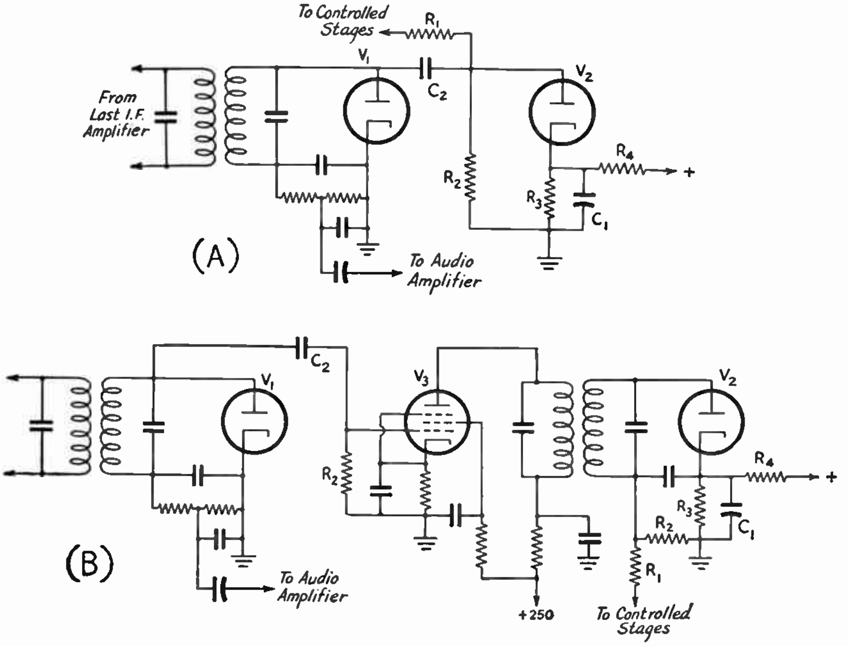


Fig. 5-17 — Delayed a.v.c. is shown at A, and amplified and delayed a.v.c. is shown in B. The circuit at B gives excellent a.v.c. action over a wide range, with no impairment of sensitivity for weak signals. For either circuit, typical values are:

C_1 — 0.001 μfd. R_1, R_2 — 1.0 megohm.
C_2 — 100 μμfd. R_3, R_4 — Voltage divider.

Resistors R_3 and R_4 are carefully proportioned to give the desired delay voltage at the cathode of diode V_2. Bleeder current of 1 or 2 ma. is ample, and hence the bleeder can be figured on 1000 or 500 ohms per volt. The delay voltage should be in the vicinity of 3 or 4 for a simple receiver and 10 or 20 in the case of a multitube high-gain affair.

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.

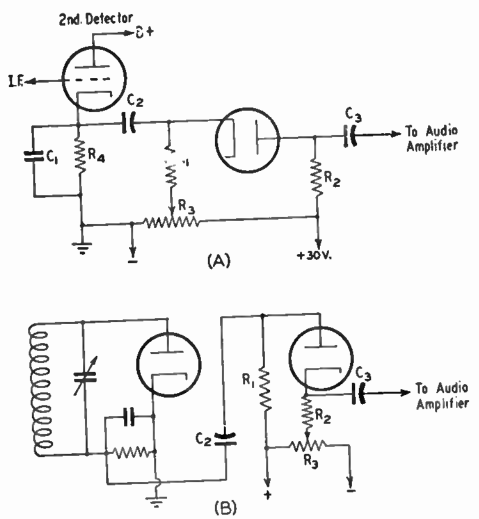


Fig. 5-18 — Series-valve noise-limiter circuits. A, as used with an infinite-impedance detector; B, with a diode detector. Typical values for components are as follows:

R_1 — 0.27 megohm.
R_2 — 47,000 ohms.
R_3 — 10,000 ohms.
R_4 — 20,000 to 47,000 ohms.
C_1 — 270 $\mu\mu$fd.
C_2, C_3 — 0.1 μfd.

All other diode-circuit constants in B are conventional.

Audio Limiting

A considerable degree of noise reduction in code reception can be accomplished by amplitude-limiting arrangements applied to the audio-output circuit of a receiver. Such limiters also maintain the signal output nearly constant during fading. These output-limiter systems are simple, and adaptable to most receivers. However, they cannot prevent noise peaks from overloading previous stages.

SECOND-DETECTOR NOISE LIMITER CIRCUITS

The circuit of Fig. 5-18 "chops" noise peaks at the second detector of a superhet receiver by means of a biased diode, which becomes nonconducting above a predetermined signal level. The audio output of the detector must pass through the diode to the grid of the amplifier tube. The diode normally would be nonconducting with the connections shown were it not for the fact that it is given positive bias from a 30-volt source through the adjustable potentiometer, R_3. Resistors R_1 and R_2 must be fairly large in value to prevent loss of audio signal.

The audio signal from the detector can be considered to modulate the steady diode current, and conduction will take place so long as the diode plate is positive with respect to the cathode. When the signal is sufficiently large to swing the cathode positive with respect to the plate, however, conduction ceases, and that portion of the signal is cut off from the audio amplifier. The point at which cut-off occurs can be selected by adjustment of R_3. By setting R_3 so that the signal just passes through the "valve," noise pulses higher in amplitude than the signal will be cut off. The circuit of Fig. 5-18A, using an infinite-impedance detector, gives a positive voltage on rectifi-

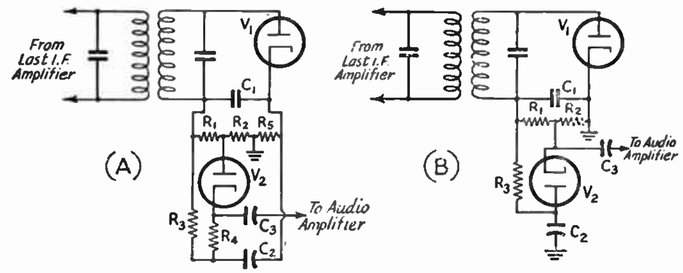


Fig. 5-19 — Self-adjusting series (A) and shunt (B) noise limiters. The functions of V_1 and V_2 can be combined in one tube like the 6H6 or 6AL5.

C_1 — 100 μμfd.
C_2, C_3 — 0.05 μfd.
R_1 — 0.27 meg. in A; 47,000 ohms in B.
R_2 — 0.27 meg. in A; 0.15 meg. in B.
R_3 — 1.0 megohm.
R_4 — 0.82 megohm.
R_5 — 6800 ohms.

cation. When the rectified voltage is negative, as it is from the usual diode detector, the circuit arrangement shown in Fig. 5-18B must be used.

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

Second-detector noise-limiting circuits that automatically adjust themselves to the receiver carrier level are shown in Fig. 5-19. In either circuit, V_1 is the usual diode second detector, R_1R_2 is the diode load resistor, and C_1 is an r.f. by-pass. A negative voltage proportional to the carrier level is developed across C_2, and this voltage cannot change rapidly because R_3 and C_2 are both large. In the circuit at A, diode V_2 acts as a conductor for the audio signal up to the point where its anode is negative with respect to the cathode. Noise peaks that exceed the maximum carrier-modulation level will drive the anode negative instantaneously, and during this time the diode does not conduct. The large time constant of C_2R_3 prevents any rapid change of 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. reception, but they are both quite effective for 'phone work. The series circuit (A) is slightly better than the shunt circuit.

I.F. Noise Silencer

In the circuit shown in Fig. 5-20, noise pulses are made to decrease the gain of an i.f. stage momentarily and thus silence the receiver for the duration of the pulse. Any noise voltage in excess of the desired signal's maximum i.f. voltage is taken off at the grid of the i.f. amplifier, amplified by the noise-amplifier stage, and rectified by the full-wave diode noise rectifier. The noise circuits are tuned to the i.f. The rectified noise voltage is applied as a pulse of negative bias to the No. 3 grid of the 6L7 i.f. amplifier, wholly or partially disabling this stage for the duration of the individual noise pulse, depending on the amplitude of the noise voltage. The noise-amplifier/rectifier circuit is biased by means of the "threshold control," R_2, so that rectification will not start until the noise voltage exceeds the desired signal amplitude. With automatic volume control the a.v.c. voltage can be applied to the grid of the noise amplifier, to augment this threshold bias. In a typical instance, this system improved the signal-to-noise ratio some 30 db. (power ratio of 1000) with heavy ignition interference, raising the signal-to-noise ratio from −10 db. without the silencer to +20 db. with the silencer.

● SIGNAL-STRENGTH AND TUNING INDICATORS

An indicator that will show relative signal strength is a useful receiver accessory. It is an aid in giving reports to transmitting stations, and it is helpful in aligning the receiver circuits, in conjunction with a test oscillator or other steady signal.

Two types of indicators are shown in Fig. 5-21. That at A uses an electron-ray tube, several types of which are available. The grid of the triode section usually is connected to the a.v.c. line. The particular type of tube used depends upon the voltage available for its grid; where the

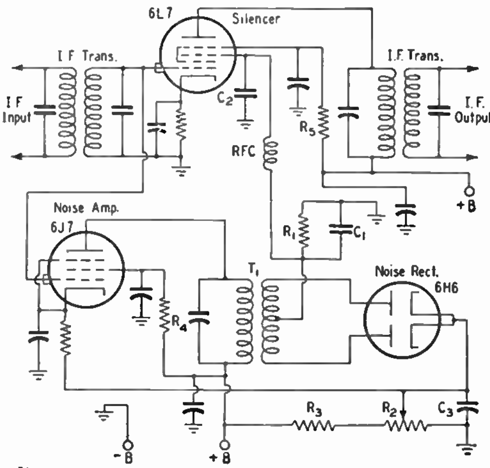


Fig. 5-20 — I.f. noise-silencing circuit. The plate supply should be 250 volts. Typical values for components are:

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

a.v.c. voltage is large, a remote cut-off type (6G5, 6N5 or 6AD6G) should be used in preference to the sharp cut-off type (6E5).

The system at B uses a milliammeter in a bridge circuit, arranged so that the meter readings increase with the signal strength. The voltage developed by the a.v.c. circuit is approximately a logarithmic function of the signal, so if the plate current of the tube is proportional to the grid voltage, the meter will read according to a linear decibel scale and will not be "crowded" at some point.

To adjust the system in Fig. 5-21B, 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.v.c. switch to "off" so the grid is shorted to ground, and adjust the 3000-ohm variable resistor for zero meter current. When the a.v.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. This will occur in the neighborhood of 15 volts with a 6J5 or 6SN7GT, and represents a rather high-amplitude signal.

The bridge circuit, while not exactly linear, is quite satisfactory from a practical standpoint. It will handle a signal range of well over 80 db. The meter cannot be "pinned" because the maximum reading occurs when the tube plate current is driven to zero, at which point further increases in a.v.c. bias cause no change.

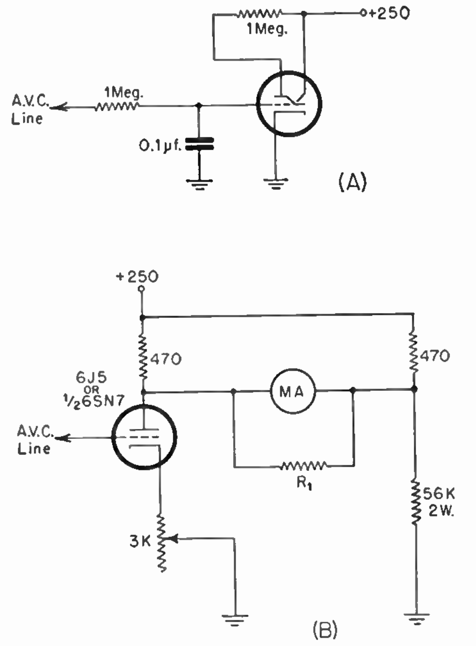


Fig. 5-21 — Tuning-indicator or S-meter circuits for superheterodyne receivers. A, electron-ray indicator; B, bridge circuit for a.v.c.-controlled tube.
MA — 0-1 or 0-2 milliammeter.
R_1 — See text.

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 feed-back may be controlled by the regular cathode-resistor gain control. When the i.f. is regenerative, it is preferable to operate the tube at reduced gain (high bias) and depend on regeneration to bring up the signal strength. This prevents overloading and increases selectivity.

The higher selectivity with regeneration reduces the over-all response to noise generated in the earlier stages of the receiver, just as does high selectivity produced by other means, and therefore improves the signal-to-noise ratio. However, the regenerative gain varies with signal strength, being less on strong signals, and the selectivity varies.

Crystal Filters

Probably the simplest means for obtaining high selectivity is by the use of a piezoelectric quartz crystal as a selective filter in the i.f. amplifier. Compared to a good tuned circuit, the Q of such a crystal is extremely high. The crystal is ground to be resonant at the desired intermediate frequency. It is then used as a selective coupler between i.f. stages.

Fig. 5-22 gives a typical crystal-filter resonance curve. For single-signal reception, the audio-frequency image can be reduced by a factor of 1000 or more. Besides practically eliminating the a.f. image, the high selectivity of the crystal filter provides good discrimination against signals very close to the desired signal and, by reducing the band-width, reduces the response of the receiver to noise.

Crystal-Filter Circuits; Phasing

Several crystal-filter circuits are shown in Fig. 5-23. Those at A and B are practically identical in performance, although differing in details. The crystal is connected in a bridge circuit, with the secondary side of T_1, the input transformer, balanced to ground either through a pair of condensers, C-C (A), or by a center-tap on the secondary, L_2 (B). The bridge is completed by the crystal and the *phasing condenser*, C_2, which has a maximum capacity somewhat higher than the capacity of the crystal in its holder. When C_2 is set to balance the crystal-holder capacity, the resonance curve of the crystal circuit is practically symmetrical; the crystal acts as a series-resonant circuit of very high Q and thus allows signals of the desired frequency to be fed through C_3 to L_3L_4, the output transformer. Without C_2, the holder capacity (with the crystal acting as a dielectric) would pass undesired signals.

In the circuit at C, the Q of the load circuit for the filter is adjusted by the setting of R, which in turn varies the bandwidth of the filter from "sharp" to a bandwidth suitable for 'phone reception. This circuit or a modification of it is found in practically all of the better communications receivers these days.

The "bandpass" crystal filter at D uses two crystals separated slightly in frequency to give a bandpass characteristic to the filter. If the frequencies are removed only a few hundred cycles from each, the characteristic is an excellent one for c.w. 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 adding additional selectivity. One popular method is to couple a BC-453 aircraft receiver (war surplus, tuning range 190 to 550 kc.) to the tail end of the 465-kc. i.f. amplifier in the communications receiver and use the resultant output of the BC-453. The aircraft receiver uses an 85-kc. i.f. amplifier that is quite sharp — 6.5 kc. wide at −60 db. — and it helps tremendously in separating 'phone signals and in backing up crystal filters for improved c.w. reception. (See *QST*, January, 1948, page 40.)

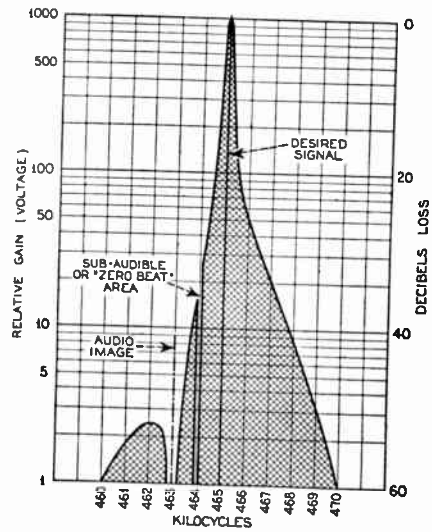


Fig. 5-22 — Graphical representation of single-signal selectivity. The shaded area indicates the over-all bandwidth, or region in which response is obtainable.

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

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

RADIO-FREQUENCY AMPLIFIERS

While selectivity to reduce audio-frequency images can be built into the i.f. amplifier, discrimination against radio-frequency images can only be obtained in circuits ahead of the first detector. These tuned circuits and their associated vacuum tubes are called **radio-frequency amplifiers.** For top performance of a communications receiver on frequencies above 7 Mc., it is mandatory that it have one or two stages of r.f. amplification, for image rejection and improved sensitivity.

Receivers with an i.f. of 455 kc. can be expected to have some r.f. image response at a signal frequency of 14 Mc. and higher if only one stage of r.f. amplification is used. (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. The better solution at 28 Mc. is to use a "triple-detection" superheterodyne, with one stage of r.f. amplification and a first i.f. of 1600 kc. or higher. A normal receiver with an i.f. of 455 kc. can be converted to a triple superhet by connecting a "converter" (to be described later) ahead of the receiver.

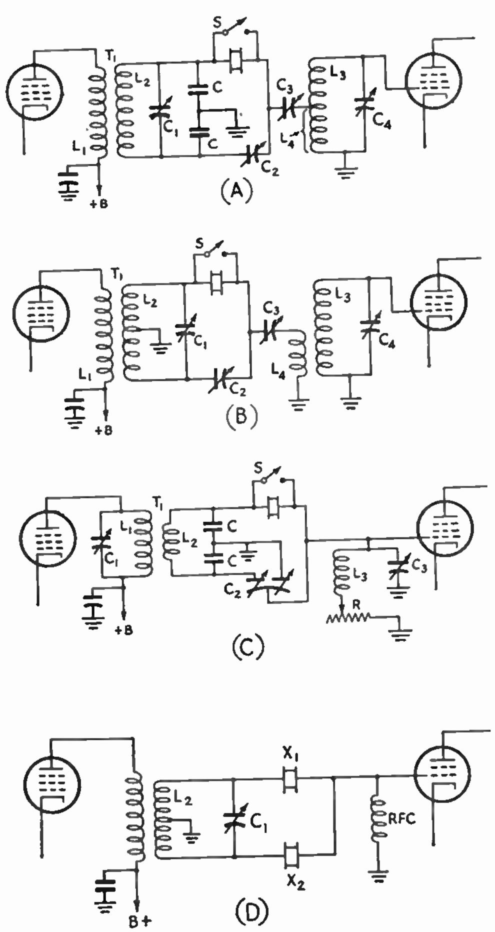


Fig. 5-23 — Crystal-filter circuits of four types. The first three give variable bandwidth, with (C) having the greatest range of selectivity.

For best selectivity, r.f. amplifiers should use high-*Q* circuits and tubes with high input and output resistance. Variable-μ pentodes are practically always used, although triodes (neutralized or otherwise connected so that they won't oscillate) are often used on the higher frequencies because they introduce less noise. Pentodes are better where maximum image rejection is desired, because they have less loading effect on the circuits.

FEED-BACK

Feed-back giving rise to regeneration and oscillation can occur in a single stage or it may appear as an over-all feed-back through several stages that are on the same frequency. To avoid feed-back in a single stage, the output must be isolated from the input in every way possible, with the vacuum tube furnishing the only coupling between the two circuits. 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 condensers in r.f. and i.f. circuits, the use of good by-pass condensers (mica or ceramic at r.f., paper or ceramic at i.f.), and returning all by-pass condensers (grid, cathode, plate and screen) with short leads to one spot on the chassis. If single-ended tubes are used, the screen or cathode by-pass condenser should be mounted across the socket, to serve as a shield between grid and plate pins. Less care is required as the frequency is lowered, but in high-impedance circuits, it is sometimes necessary to shield grid and plate leads and to be careful not to run them close together.

To avoid over-all feed-back in a multistage amplifier, attention must be paid to avoid running any part of the output circuit back near the input circuit without first filtering it carefully. Since the signal-carrying parts of the circuit (the "hot" grid and plate leads) can't be filtered, the best design for any multistage amplifier is a straight line, to keep the output as far away from the input as possible. For example, an r.f. amplifier might run along a chassis in a straight line, run into a mixer where the frequency is changed, and then the i.f. amplifier could be run back parallel to the r.f. amplifier, provided there was a very large frequency difference between the r.f. and the i.f. amplifiers. However, to avoid any possible coupling, it would be better to run the i.f. amplifier off at right angles to the r.f.-amplifier line, just to be on the safe side. Good shielding is important in preventing over-all oscillation in high-gain-per-stage amplifiers, but it becomes less important when the stage gain drops to a low value. In a high-gain amplifier, the power leads (including the heater circuit) are common to all stages, and they can provide the over-all coupling if they aren't properly filtered. Good by-passing and the use of series isolating resistors will generally eliminate any possibility of coupling through the power leads. R.f. chokes, instead of resistors, are used in the heater leads where necessary.

CROSS-MODULATION

Since a one- or two-stage r.f. amplifier will have a passband measured in hundreds of kc. at 14 Mc. or higher, strong signals will be amplified through the r.f. amplifier even though it is not tuned exactly to them. If these signals are strong enough, their amplified magnitude may be measurable in volts after passing through several r.f. stages. If an undesired signal is strong enough after amplification in the r.f. stages to shift the operating point of a tube (by driving the grid into the positive region), the undesired signal will modulate the desired signal. This effect is called **cross-modulation,** and is often encountered in receivers with several r.f. stages 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.

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.

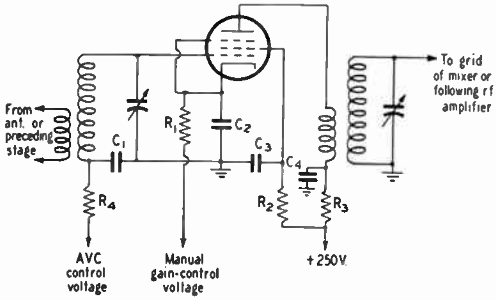


Fig. 5-24 — Typical radio-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:

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

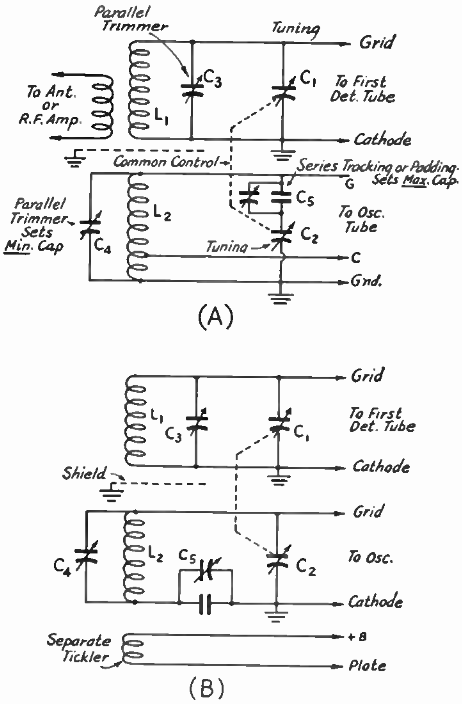


Fig. 5-25 — Converter-circuit tracking methods. Following are approximate circuit values for 450- to 465-kc. i.f.s, with tuning ranges of approximately 2.15-to-1 and C_2 having 140-$\mu\mu$fd. maximum, and the total minimum capacitance, including C_3 or C_4, being 30 to 36 $\mu\mu$fd.

Tuning Range	L_1	L_2	C_5
1.7–4 Mc.	50 μh.	40 μh.	0.0013 μfd.
3.7–7.5 Mc.	14 μh.	12.2 μh.	0.0022 μfd.
7–15 Mc.	3.5 μh.	3 μh.	0.0045 μfd.
14–30 Mc.	0.8 μh.	0.78 μh.	None used

Approximate values for 450- to 465-kc. i.f.s with a 2.5-to-1 tuning range, C_1 and C_2 being 350-$\mu\mu$fd. maximum, minimum including C_3 and C_4 being 40 to 50 $\mu\mu$fd.

Tuning Range	L_1	L_2	C_5
0.5–1.5 Mc.	240 μh.	130 μh.	425 $\mu\mu$fd.
1.5–4 Mc.	32 μh.	25 μh.	0.00115 μfd.
4–10 Mc.	4.5 μh.	1 μh.	0.0028 μfd.
10–25 Mc.	0.8 μh.	0.75 μh.	None used

Gain Control

To avoid cross-modulation and other overload effects in the first detector and r.f. stages, the gain of the r.f. stages is usually made adjustable. This is accomplished by using variable-μ tubes and varying the d.c. grid bias, either in the grid or cathode circuit. If the gain control is automatic, as in the case of a.v.c., the bias is controlled in the grid circuit Manual control of r.f. gain is generally done in the cathode circuit. A typical r.f. amplifier stage with the two types of gain control is shown in schematic form in Fig. 5-24.

Tracking

In a receiver with no r.f. stage, it is no inconvenience to adjust the high-frequency oscillator and the mixer circuit independently, because the mixer tuning is broad and requires little attention over an amateur band. However, when r.f. stages are added ahead of the mixer, the 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.**

Tracking methods for covering a wide frequency range, suitable for general-coverage receivers, are shown in Fig. 5-25. The **tracking capacity,** C_5, commonly consists of two condensers in parallel, a fixed one of somewhat less capacity than the value needed and a smaller variable in parallel to allow for adjustment to the exact proper value. The trimmer, C_4, is first set for the high-frequency end of the tuning range, and then the tracking condenser is set for the low-frequency end. The tracking capacity becomes larger as the percentage difference between the oscillator and signal frequencies becomes smaller (that is, as the signal frequency becomes higher). Typical circuit values are given in the tables under Fig. 5-25. The coils can be conveniently calculated with the *ARRL Lightning Calculator* and then trimmed in the circuit for best tracking.

In amateur-band receivers, tracking is simplified by choosing a bandspread circuit that gives practically straight-line-frequency tuning (equal frequency change for each dial division), and then adjusting the oscillator and mixer tuned circuits so that both cover the same total number of kilocycles. For example, if the i.f. is 455 kc. and the mixer circuit tunes from 7000 to 7300 kc. between two given points on the dial, then the oscillator must tune from 7455 to 7755 kc. between the same two dial readings. With the bandspread arrangement of Fig. 5-8A, the tuning will be practically straight-line-frequency if C_2 (bandset) is 4 times or more the maximum capacity of C_1 (bandspread), as is usually the case for strictly amateur-band coverage. C_1 should be of the straight-line-capacity type (semicircular plates).

Improving Receiver Sensitivity

The sensitivity (signal-to-noise ratio) of a receiver on the higher frequencies above 20 Mc. is dependent upon the bandwidth of the receiver and the noise contributed by the "front end" of the receiver. Neglecting the fact that 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, 7F8 and triode-connected 6AK5 are the best of the triodes. For best noise figure, the antenna circuit should be coupled a little heavier than optimum. This 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. Since the receiver manufacturer has no way to predict the type of antenna that will be used, he generally designs the input for some compromise value, usually around 300 or 400 ohms in the high-frequency ranges. If your antenna looks like something far different than this, the receiver effectiveness can be improved by proper matching. This can be accomplished by changing the antenna to the right value (as determined from the receiver instruction book) or by using a simple matching device as described later in this chapter. Overcoupling the input circuit will often improve sensitivity but it will, of course, always reduce the image-rejection contribution of the antenna circuit.

Commercial receivers can also be "hopped up" by substituting a high-g_m tube in the first r.f. stage if one isn't already there. The amateur must be prepared to take the consequences, however, since the stage may oscillate, or not track without some modification. A simpler solution is to add the "hot" r.f. stage ahead of the receiver.

Regeneration

Regeneration in the r.f. stage of a receiver (where only one stage exists) will often improve the sensitivity because the greater gain it provides serves to mask more completely the first-detector noise, and it also provides a measure of automatic matching to the antenna through tighter coupling. However, accurate ganging becomes a problem, because of the increased selectivity of the regenerative r.f. stage, and the receiver almost invariably becomes a two-handed-tuning device. Regeneration should not be overlooked as an expedient, however, and amateurs have used it with considerable success. High-g_m tubes are the best as regenerative amplifiers, and the feed-back should not be controlled by changing the operating voltages (which should be the same as for the tube used in a high-gain amplifier) but by changing the loading or the feed-back coupling. This is 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.

Extending the Tuning Range

As mentioned earlier, when a receiver doesn't cover a particular frequency range, either in fact or in satisfactory performance, a simple solution is to use a **converter.** A converter is another "front end" for the receiver, and it is made to tune the proper range or to give the necessary performance. It works into the receiver at some frequency between 1.6 and 10 Mc. and thus forms with the receiver a "triple-detection" superhet.

There are several different types of converters in vogue at the present time. The commonest type, since it is the oldest, uses a regular tunable oscillator, mixer, and r.f. stages as desired, and works into the receiver at a fixed frequency. A second type uses broad-banded r.f. stages in the r.f. and mixer stages of the converter, and only the oscillator is tuned. Since the frequency the converter works into is high (7 Mc. or more), little or no trouble with images is experienced, despite the broad-band r.f. stages. A third type of converter uses broad-banded r.f. and output stages and a fixed-frequency oscillator (self- or crystal-controlled). The tuning is done with the receiver the converter is connected to. This is an excellent system if the receiver itself is well shielded and has no external pick-up of its own. Many war-surplus receivers fall in this category. A fourth type of converter uses a fixed oscillator with ganged mixer and r.f. stages, and requires two-handed tuning, for the r.f. stages and for the receiver. The r.f. tuning is not critical, however, unless there are many stages

The broad-banded r.f. stages have the advantage that they can be built with short leads, since no tuning capacitors are required and the unit can be tuned initially by trimming the inductances. They are more prone to cross-modulation than the gang-tuned r.f. stages, however, because of the lack of selectivity. The fourth type of converter is probably the most satisfactory, particularly if a crystal-controlled high-frequency oscillator is used. It not only has the advantage of the best selectivity and protection against images and cross-modulation, but the crystal gives it a stability unobtainable with self-controlled oscillators. Amateurs who specialize in operation on 28 and 50 Mc. generally use good converters ahead of conventional communications receivers, and it pays off in better performance for the station.

While converters can extend the operating range of an existing receiver, their greatest advantage probably lies in the opportunity they give for getting the best performance on any one band. By selecting the best tubes and techniques for any particular band, the amateur is assured of top receiver performance. With separate converters for each of several bands, changes can be made in any one without disabling or impairing the receiver performance on another band. The use of converters ahead of the low-frequency receiver is rapidly becoming standard practice on the bands above 14 Mc.

Tuning a Receiver

C.W. Reception

For making code signals audible, the beat oscillator should be set to a frequency slightly different from the intermediate frequency. To adjust the beat-oscillator frequency, first tune in a moderately-weak but steady carrier with the beat oscillator turned off. Adjust the receiver tuning for maximum signal strength, as indicated by maximum hiss. Then turn on the beat oscillator and adjust its frequency (leaving the receiver tuning unchanged) to give a suitable beat note. The beat oscillator need not subsequently be touched, except for occasional checking to make certain the frequency has not drifted from the initial setting. The b.f.o. may be set on either the high- or low-frequency side of zero beat.

The best receiver condition for the reception of c.w. 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 gain of the first r.f. stage and the 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 at times when interference is present.

'Phone Reception

In reception of 'phone signals, the normal procedure is to set the r.f. and i.f. gain at maximum, switch on the a.v.c., and use the audio gain control for setting the volume. This insures maximum effectiveness of the a.v.c. system in compensating for fading and maintaining constant audio output on either strong or weak signals. On occasion a strong signal close to the frequency of a weaker desired station may take control of the a.v.c., in which case the weaker station may disappear because of the reduced gain. In this case better reception may result if the a.v.c. is switched off, using the manual r.f. gain control to set the gain at a point that prevents "blocking" by the stronger signal.

When receiving an AM signal on a frequency within 5 to 20 kc. from a single-sideband signal it may also be necessary to switch off the a.v.c. and resort to the use of manual gain control, unless the receiver has excellent skirt selectivity. No ordinary a.v.c. circuit can handle the syllabic bursts of energy from the SSB station.

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 c.w. reception, it is advisable to do all tuning with the filter in the circuit. Variable-selectivity filters permit a choice of selectivity to suit interference conditions.

An undesired carrier close in frequency to a desired carrier will heterodyne with it to produce a beat note equal to the frequency difference. Such a heterodyne can be reduced by adjustment of the phasing control in the crystal filter.

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, causes some reduction in naturalness.

Spurious Responses

Spurious responses can be recognized without a great deal of difficulty. Often it is possible to identify an image by the nature of the transmitting station, if the frequency assignments applying to the frequency to which the receiver is tuned are known. However, an image also can be recognized by its behavior with tuning. If the signal causes a heterodyne beat note with the desired signal and is actually on the same frequency, the beat note will not change as the receiver is tuned through the signal; but if the interfering signal is an image, the beat will vary in pitch as the receiver is tuned. The beat oscillator in the receiver must be turned off for this test. Using a crystal filter with the beat oscillator on, an image will peak on the side of zero beat opposite that on which 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.

Narrow-Band Frequency- and Phase-Modulation Reception

FM Reception

In the reception of NFM (narrow-band FM) by a normal AM receiver, the a.v.c. is switched off and the incoming signal is not tuned "on the nose," as indicated by maximum reading of the S-meter, but slightly off to one side or the other. This puts the carrier of the incoming signal on one side or the other of the i.f. selectivity characteristic (see Fig. 5-1). As the frequency of the signal changes back and forth over a small range with modulation, these variations in frequency are translated to variations in amplitude, and the consequent AM is detected in the normal manner. The signal is tuned in (on one side or the other of maximum carrier strength) until the audio quality appears to be best. If the audio is too weak, the transmitting operator should be advised to increase his swing slightly, and if the audio quality is bad ("splashy" and with serious distortion on volume peaks) he should be advised to reduce his swing. Coöperation between transmitting and receiving operators is a necessity for best audio quality. The transmitting station should always be advised immediately if at any time his bandwidth exceeds that of an AM signal, since this is a violation of FCC regulations, except in those portions of the bands where wide-band FM is permitted.

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

PM Reception

Phase-modulated signals can be received in the same way that NFM signals are, except that in this case the audio output will appear to be lacking in "lows," because of the differences in the deviation-*vs.*-audio characteristics of the two systems. This can be remedied to a considerable degree by advancing the tone control of the receiver to the point where more nearly normal speech output is obtained.

NPM signals can also be received on communications receivers by making use of the crystal filter, in which case there is no need for audio compensation. The crystal filter should be set to the sharpest position and the carrier should be tuned in on the crystal peak, *not* set off to one side. The phasing condenser should be set not for exact neutralization but to give a rejection notch at some convenient side frequency such as 1000 cycles off resonance. There is considerable attenuation of the side bands with such tuning, but it can readily be overcome by using additional audio gain. NFM signals received through the crystal filter in this fashion will have a "boomy" characteristic because the lower frequencies are accentuated.

Reception of Single-Sideband Signals

Single-sideband signals are generally transmitted with little or no carrier, and it is necessary to furnish the carrier at the receiver before proper reception can be obtained. Because little or no carrier is transmitted, the a.v.c. in the receiver has nothing that indicates the average signal level, and manual variation of the r.f. gain control is required.

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

Once the proper setting of the b.f.o. has been established by the procedure above, all further tuning should be done with the main tuning control. However, it is not unlikely that SSB stations will be encountered that are transmitting the other sideband, and to receive them will require shifting the b.f.o. setting to the other side of the receiver i.f. passband. The initial tuning procedure is exactly the same as outlined above, except that you will end up with a considerably different b.f.o. setting. The two b.f.o. settings should be noted for future reference, and all tuning of SSB signals can then be done with the main tuning dial. After a little experience, it becomes a simple matter to determine which way to tune the receiver if the receiver (or transmitter) drifts off to make the received signal sound low- or high-pitched.

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-μfd. blocking condenser to the receiver chassis. Lacking an a.c. voltmeter, the audio output can be judged by ear, although this method is not as accurate as the others. If the tuning meter is used as an indication, the a.v.c. of the receiver should be turned on, but any other indication requires that it be turned off. Lacking a test oscillator, a steady 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 spend a little time and haywire together a simple oscillator for test purposes.

Initial alignment of a new i.f. amplifier is as follows: The test oscillator is set to the correct frequency, and its output is coupled through a condenser to the grid of the last i.f. amplifier tube. The trimmer condensers of the transformer feeding the second detector are then adjusted for maximum output, as shown by the indicating device being used. The oscillator output lead is then clipped on to the grid of the next-to-the-last i.f. amplifier tube, and the second-from-the-last transformer trimmer adjustments are peaked for maximum output. This process is continued, working back from the second detector, until all of the i.f. transformers have been aligned. It will be necessary to reduce the output of the test oscillator as more of the i.f. amplifier is brought into use. 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-stage 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 condenser in the receiver to give maximum response on the test-oscillator signal, then reset the receiver dial to the low-frequency end of the range. Set the test-oscillator frequency near the frequency indicated by the receiver dial and tune the test oscillator until its signal is heard in the receiver. If the frequency of the signal as indicated by the test-oscillator calibration is higher than that indicated by the receiver dial, more inductance (or more capacity in the tracking condenser) is needed in the receiver oscillator circuit; if the frequency is lower, less inductance (less tracking capacity) is required in the receiver oscillator. Most commercial receivers provide some means for varying the inductance of the coils or the capacity of the tracking condenser, to permit aligning the receiver tuning with the dial calibration. Set the test oscillator to the frequency indicated by the receiver dial, and then adjust the tracking capacity or inductance of the receiver oscillator coil to obtain maximum response. After making this adjustment, recheck the high-frequency end of the scale as previously described. It may be necessary to go back and forth between the ends of the range several times before the proper combination of inductance and capacity is secured. In many cases, better over-all tracking will result if frequencies near but not actually at the ends of the tuning range are selected, instead of taking the extreme dial settings.

After the oscillator range is properly adjusted, set the receiver and test oscillator to the high-frequency end of the range. Adjust the mixer trimmer condenser for maximum hiss or signal, then the r.f. trimmers. Reset the tuning dial and test oscillator to the low-frequency end of the range, and repeat; if the circuits are properly designed, no change in trimmer settings should be necessary. If it is necessary to increase the trimmer capacity in any circuit, it indicates that more inductance is needed; 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 from this cause will be small, however, and it will suffice to bring the circuits into line at both ends of the range. If most reception is in a particular part of the range, such as an amateur band, the circuits may be aligned for maximum performance in that region, even though the ends of the frequency range as a whole may be slightly out of alignment.

Oscillation in R.F. or I.F. Amplifiers

Oscillation in high-frequency amplifier and mixer circuits 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.v.c. system to reduce the receiver gain drastically. Oscillation can be caused by poor connections in the common ground circuits. Inadequate or defective by-pass condensers in cathode, plate and screen-grid circuits also can cause such oscillation. A metal tube with an ungrounded shell may cause trouble. Improper screen-grid voltage, resulting from a shorted or too-low screen-grid series resistor, also may be responsible for such instability.

Oscillation in the i.f. circuits is independent of high-frequency tuning, and is indicated by a continuous squeal that appears when the gain is advanced with the c.w. beat oscillator on. It can result from defects in i.f.-amplifier circuits similar to those above. Inadequate screen or plate by-pass capacitance is a common cause of such oscillation. An additional by-pass condenser of 0.1- to 0.25-μfd. capacitance often will remedy the trouble.

Instability

"Birdies" or a mushy hiss occurring with tuning of the high-frequency oscillator may indicate that the oscillator is "squegging" or oscillating simultaneously at high and low frequencies. This may be caused by a defective tube, too-high oscillator plate or screen-grid voltage, excessive feed-back, or too-high grid-leak resistance.

A varying beat note in c.w. reception indicates instability in either the h.f. oscillator or beat oscillator, usually the former. The stability of the beat oscillator can be checked by introducing a signal of intermediate frequency (from a test oscillator) into the i.f. amplifier; if the beat note is unstable, the trouble is in the beat oscillator. Poor connections or defective parts are the likely cause. Instability in the high-frequency oscillator may be the result of poor circuit design, loose connections, defective tubes or circuit components, or poor voltage regulation in the oscillator plate- and/or screen-supply circuits. Mixer pulling of the oscillator circuit also will cause the beat note to "chirp" on strong c.w. signals because the oscillator load changes slightly.

In 'phone reception with a.v.c., a peculiar type of instability ("motorboating") may appear if the h.f.-oscillator frequency is sensitive to changes in plate voltage. As the a.v.c. voltage rises the currents of the controlled tubes decrease, decreasing the load on the power supply and causing its output voltage to rise. Since this increases the voltage applied to the oscillator, its frequency changes correspondingly, throwing the signal off the peak of the i.f. resonance curve and reducing the a.v.c. voltage, thus tending to restore the original conditions. The process then repeats itself, at a rate determined by the signal strength and the time constant of the power-supply circuits. This effect is most pronounced with high i.f. selectivity, as when a crystal filter is used, and can be cured by making the oscillator insensitive to voltage changes or by regulating the plate-voltage supply. The better receivers use VR-type tubes to stabilize the oscillator voltage — a defective VR tube will cause trouble with oscillator instability.

A One-Tube Regenerative Receiver

The receiver shown in Figs. 5-26, 5-27, 5-28 and 5-29 represents close to the minimum requirements of a useful short-wave receiver. Under suitable conditions, it is capable of receiving signals from many foreign countries. It is an excellent receiver for the beginner, because it is easy to build and the components are not expensive.

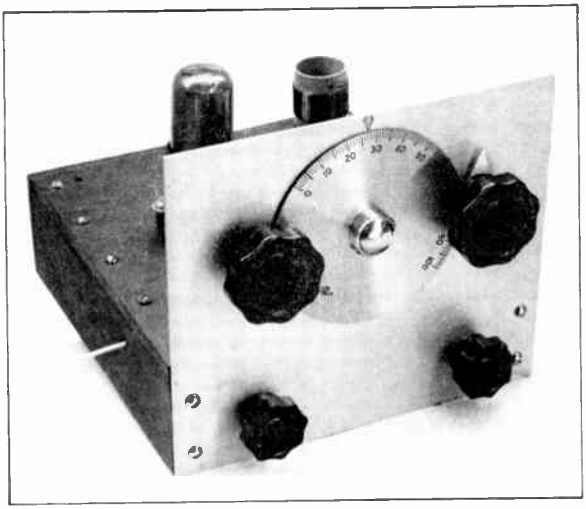


Fig. 5-26 — The simple one-tube regenerative receiver is built on a wood-and-Presdwood chassis, with an aluminum panel. The large left-hand knob drives the calibrated scale on the bandspread condenser. The large right-hand knob is for the band-set condenser.

From the circuit in Fig. 5-28, it can be seen that the only tube in the receiver is a 6SN7 twin triode. One section is used as a regenerative detector, the other triode section serving as an audio amplifier to the headphones. A variable antenna-coupling condenser, C_1, minimizes "dead spots" in the tuning range that might be caused by antenna-resonance effects. Two tuning condensers are used. The band-set condenser, C_4, tunes to the desired frequency band, and the bandspread condenser, C_2-C_3, allows the operator to tune slowly through the band. The bandspread condenser is a dual condenser made from a single midget variable, and on all of the amateur bands except 3.5 Mc. only the C_3 portion is connected in the circuit. The 3.5-Mc. coil includes a jumper that connects C_2 on that band. Regeneration is controlled by varying the plate voltage on the detector with R_4.

The mechanical design is made as simple as possible. Work on the chassis and the front panel can be done with only a No. 18 drill, a ½-inch drill, and a round file. There is no complicated metal work or bending. To reduce the panel size, the knob on the band-set condenser overlaps the friction-driven tuning dial.

The front panel is a 7 × 7-inch sheet of 1/16-inch aluminum. It carries the tuning controls, the regeneration adjustment and the antenna-coupling condenser shaft. The sides of the chassis are soft wood strips, 7 × 2 × 5/8 inches. The deck of the chassis is a 7 × 7-inch sheet of ¼-inch Presdwood (or Masonite). The 6SN7 socket is supported on 5/8-inch-long mounting pillars, and the 5-

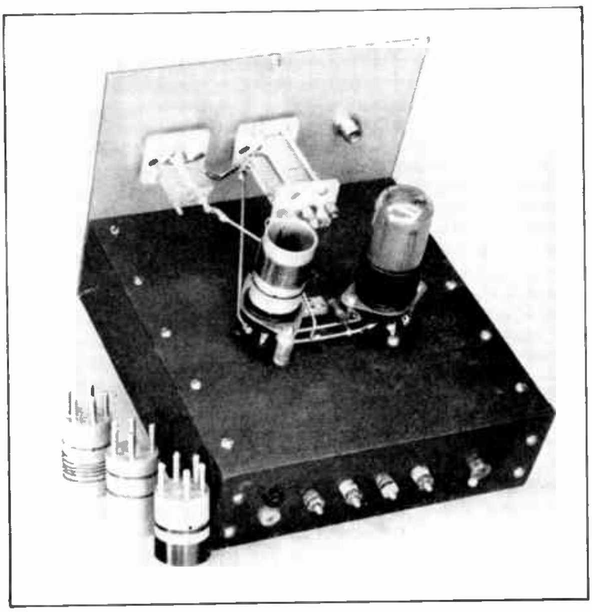


Fig. 5-27 — Another view of the one-tube regenerative receiver shows how the tube and coil sockets are mounted. The headphone tips plug into the two small tip jacks on the rear panel — the set of four machine screws and nuts is for connecting to the power supply.

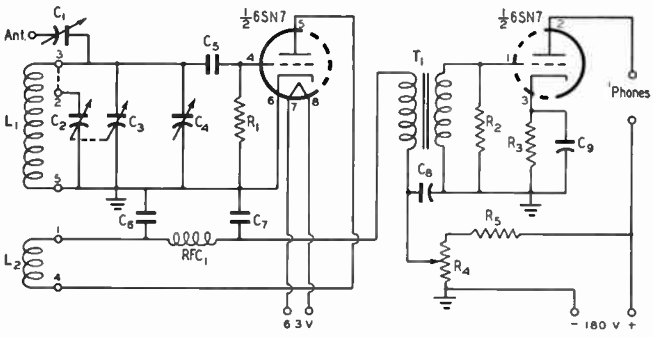


Fig. 5-28 — **Wiring diagram of the one-tube regenerative receiver.**

- C_1 — Homemade adjustable condenser. See text.
- C_2, C_3 — Reworked midget variable (Millen 21935). See text.
- C_4 — 100-μμfd. midget variable (Millen 20100).
- C_5 — 100-μμfd. mica.
- C_6, C_7 — 170-μμfd. mica.
- C_8 — 12-μfd. 150-volt electrolytic.
- C_9 — 10-μfd. 25-volt electrolytic.
- R_1 — 1.5 megohms, ½ watt.
- R_2 — 0.15 megohm, ½ watt.
- R_3 — 1500 ohms, ½ watt.
- R_4 — 50,000-ohm wire-wound potentiometer.
- R_5 — 33,000 ohms, 1 watt.
- RFC_1 — 2.5-mh. r.f. choke (National 100U).
- T_1 — Interstage audio transformer (Stancor A-4723).

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

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

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

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

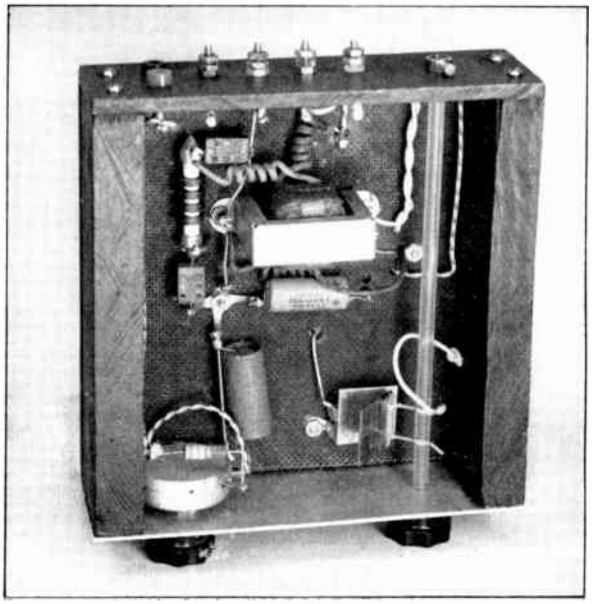


Fig. 5-29 — **This view underneath the one-tube regenerative receiver shows the arrangement of parts and the construction of the variable antenna-coupling condenser.**

COIL TABLE FOR THE ONE-TUBE REGENERATIVE RECEIVER

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

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

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

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

The power supply for the receiver, shown in Figs. 5-30 and 5-31, is simple to assemble because it is built on a wooden chassis. Two strips of 1½ × ¾-inch wood, 12 inches long, are nailed to two short end pieces. The separation between strips is just enough (1¼ inches) to clear the tube socket and electrolytic condensers, and the leads from the transformer and choke also pass through this opening. Binding posts are made in the same manner as on the receiver, with No. 6 machine screws and suitable nuts and washers.

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

Using the parts listed in Fig. 5-31 should result in a power supply that gives about 180 volts when connected to the receiver. However, if the 6SN7 in the receiver appears to run too hot (as tested by touching the tube after the receiver has been running for 5 or 10 minutes), the output voltage can be reduced by increasing the resistance at R_1 (Fig. 5-31). Adding

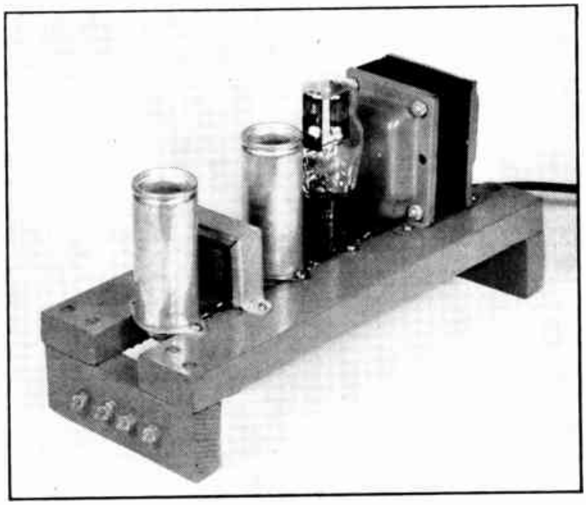


Fig. 5-30 — The power supply for the regenerative receiver is built on a simple wooden chassis.

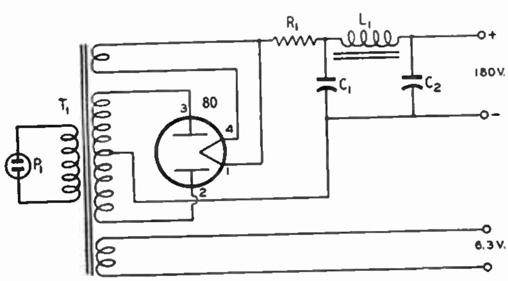


Fig. 5-31 — Circuit diagram of the power supply for the regenerative receiver.

C_1, C_2 — 16-μfd. 450-volt electrolytic (Mallory RS-217).
R_1 — 20,000-ohm 10-watt wire-wound.
L_1 — 7-henry 50-ma. filter choke (Stancor C-1707).
P_1 — 115-volt line plug.
T_1 — 275-0-275 volts at 50 ma., 6.3 v. at 2.5 amp., 5 v. at 2 amp. (Thordarson T22R30).

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

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

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

A Two-Band Five-Tube Superheterodyne

The five-tube superheterodyne shown in Figs. 5-32, 5-34 and 5-36 is a double-conversion receiver tuning the 3.5- and 7-Mc. amateur bands. It is not difficult to build, and it has stability and selectivity not surpassed by factory-built receivers costing much more.

As can be seen in Fig. 5-33, the circuit diagram, the receiver uses intermediate frequencies of 1700 and 100 kc. The 1700-kc. first i.f. permits using an oscillator that tunes only one range for the two bands. Tuning the oscillator from 5.2 to 5.7 Mc. gives an i.f. of 1700 kc. for the 3.5- to 4.0-Mc. range and the same i.f. for the 6.9- to 7.4-Mc. range. The oscillator components are soldered in place (no switching or plug-in coils) and the dial calibration is made once and can then be relied upon. To change bands, it is only necessary to swing the input condenser, C_1, to the 80- or 40-meter band. The 1700-kc. i.f. eliminates any pulling on the oscillator, in either range.

With no r.f. stage, the receiver's signal-to-noise ratio is determined by the mixer. The 6AC7 is the best tube available for the purpose. To minimize spurious responses, two tuned circuits are used in the input between antenna and converter grid. The stator plates of the dual condenser, C_1, are shielded from each other, as are the two coils L_2 and L_3, and the coupling between circuits is obtained by the 0.001-μfd. condenser.

The 1700-kc. signal from the first converter is converted in the 6K8 second converter to 100 kc. The use of a 1600-kc. crystal for the oscillator at this point permits using an r.f. gain control that has no effect on the frequency. No frequency change with gain-control setting is a desirable characteristic of any good receiver, so the 1600-kc. crystal at $2.75 is not a luxury. While the 1600-kc. oscillator could be made self-controlled, it would be almost certain to "pull" with gain-control changes.

The specified 1700-kc. transformer, T_1, is a relatively expensive item, but there can be no compromise at this point, because a poor transformer will not have enough rejection to avoid the secondary images (200 kc. away) that might otherwise ride through.

The 100-kc. output from the 6K8 is filtered through three tuned circuits and feeds a triode plate detector (½ 6SN7). This detector is regenerative, but the regeneration is fixed and doesn't have to be bothered with by the operator unless he changes tubes and the new tube has considerably different characteristics. The regeneration in the 100-kc. detector gives the receiver its single-signal c.w. reception characteristic, since there aren't enough tuned circuits to give it otherwise. The b.f.o. uses the other triode in the 6SN7 envelope, and stray coupling is used for the b.f.o. injection. No panel control of b.f.o. pitch is available, because the selectivity is not adjustable and the variable-pitch feature is not essential.

Up to this point the gain of the receiver is not too high, and two stages of audio amplification are used. Omitting the cathode by-pass condensers still leaves more than enough audio for any pair of high-impedance headphones.

By keeping the signal level low up to and through the selective stages, there is a minimum opportunity for overloading and cross-modulation, and the gain need be kept only high enough to prevent degrading the signal-to-noise ratio. Further, a regenerative stage has a tendency to "flatten out" with strong signals, so the regenerative detector is somewhat protected by holding the gain down. However, the receiver has quite adequate sensitivity — in any normal location and with a fair to good antenna, any signal that can be heard by a large receiver can be heard by this one, except in rare cases where the large receiver's superior selectivity makes the difference.

Construction

The construction of the receiver is unconventional in that two chassis are used, as shown in Figs. 5-32 and 5-34, and the panel is mounted away from the chassis. All of the electrical components are mounted on the aluminum 7 × 11 × 2-inch chassis, and this sits on an inverted 7 × 11 × 2-inch steel chassis that serves as a base and bottom cover. The bottom chassis has rubber feet (grommets) at its corners that prevent its slipping

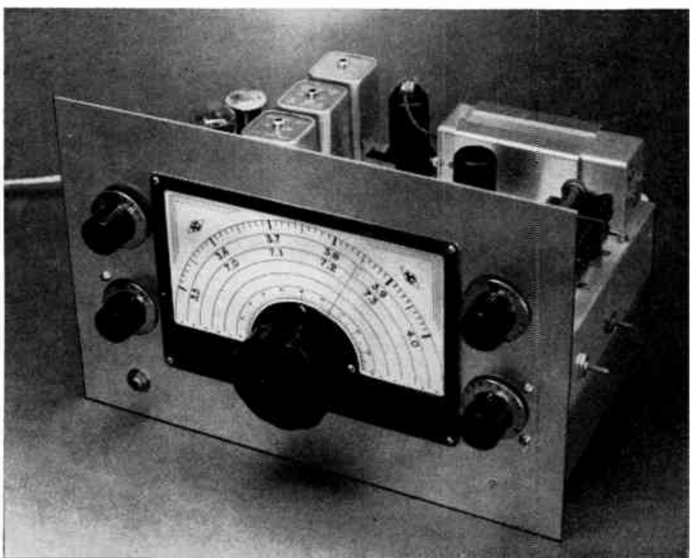


Fig. 5-32 — The five-tube double-conversion superheterodyne tunes the 3.5- and 7-Mc. bands without bandswitching. The controls on the left are audio volume (upper) and b.f.o. switch, and those on the right are antenna tuning (upper) and i.f. gain.

on the table. The 8 × 12-inch panel is supported away from the aluminum chassis on ½-inch-long brass collars, secured by suitable washers and 6-32 screws, as shown in Fig. 5-36. The panel is supported by two such collars at each end of the chassis and by two more that make up to two of the mounting screws of the National ACN dial at the center. The two center collars add to the strength of the assembly by furnishing additional support for the panel and dial, and they should not be omitted.

The aluminum chassis is bolted to the steel chassis by two 4¼-inch lengths of ⅛-inch diameter brass rod, threaded

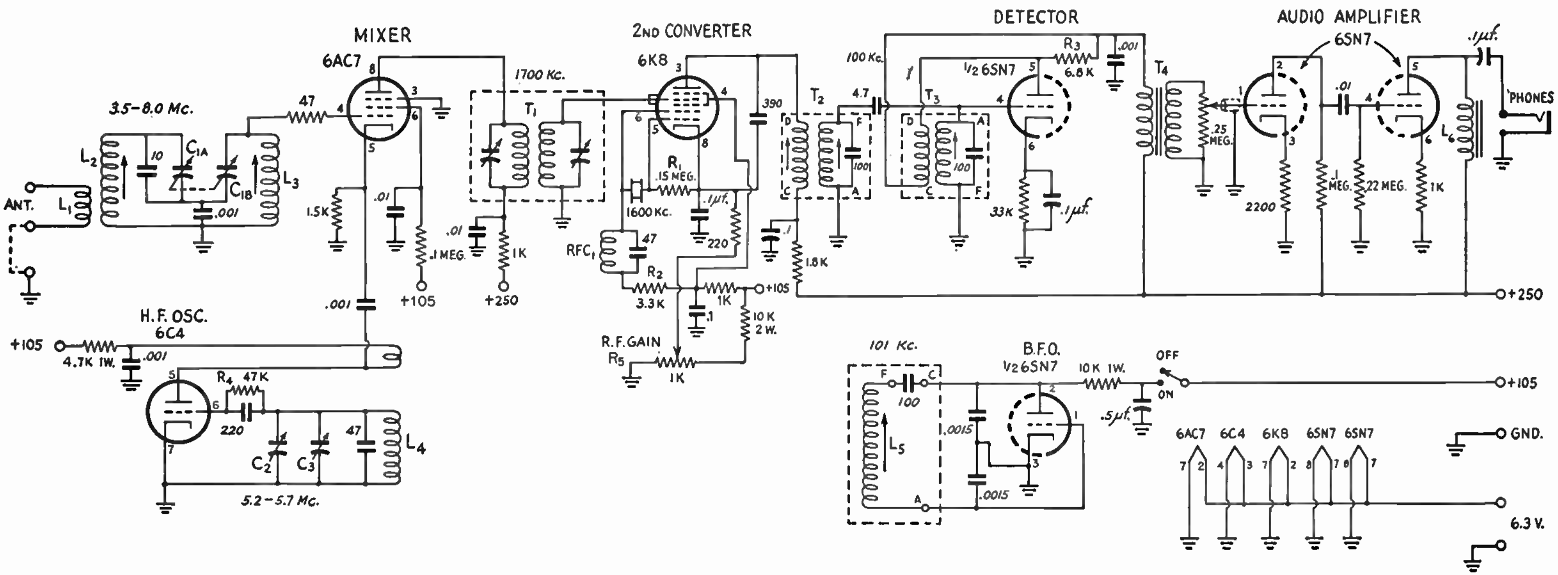


Fig. 5-33 — Wiring diagram of the five-tube receiver.

C_1 — 140-$\mu\mu$fd.-per-section dual variable (Hammarlund MCD-140-M).

C_2 — 35-$\mu\mu$fd. midget variable (Bud LC-1643 or Hammarlund HF-35).

C_3 — 100-$\mu\mu$fd. midget variable (National PSR-100).

R_5 — 1000-ohm wirewound potentiometer (Mallory A1MP).

All resistors ½-watt unless specified otherwise.

L_1 — 8 turns No. 30 d.c.c. close-wound over ground end of L_2.

L_2, L_3 — 35 turns No. 30 d.e.c. close-wound on National XR-50 slug-tuned form.

L_4 — 23 turns No. 24 bare space-wound 32 turns per inch, ⅝-inch diam. Tickler is 1¾ turns spaced 1 turn from L_4. See text. (Made from B & W 3008 Miniductor.)

L_5 — 20-mh. (approx.) slug-tuned coil (RCA 205R1).

T_1 — 1700-kc. i.f. transformer, modified (Millen 62161).

T_2, T_3 — 100-kc. transformers made from TV components (RCA 205R1). See text.

T_4 — Small 3:1 audio transformer (Stancor A-63-C).

RFC_1 — 750 μh. (National R-33).

The 1600-kc. crystal is a Peterson Radio type Z-2.

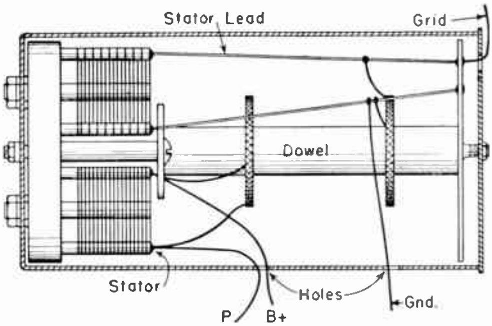


Fig. 5-35 — The 1700-kc. i.f. can is modified by drilling two holes in the side of the can.

On the transformer assembly proper, the old grid (green) and ground (black) wires are removed. On the tuning condenser connected to the coil nearest the tuning condensers, a new plate lead is connected to the stator and a new B+ lead to the rotor. The old plate lead (blue) becomes the new grid lead, and the old B+ lead (red) becomes the new ground lead by transferring it from the terminal to the rotor wire near the coil.

During reassembly, the new plate and B+ leads should be soldered to a length of wire that is passed through the shield-can hole before the entire assembly is completed. Otherwise it is difficult to snake out the new plate and B+ leads unless small flexible wire is used.

6-32 at each end. These rods pass through holes in the top and lip of each chassis. The only holes that are required in the steel chassis are those for the two tie rods, the four holes for the rubber feet, and a 1¼-inch diameter hole to clear the headphone jack.

In the oscillator circuit, the 35-μμfd. tuning condenser, C_2, is supported by a small aluminum bracket. The correct location of the condenser on the bracket can be found after the dial-and-chassis assembly has been completed. It is imperative to the smooth operation of the tuning condenser that the shaft of the condenser be correctly aligned with the coupling of the dial. The 100-μμfd. trimmer, C_3, is mounted under the chassis with its shaft extending through to the top, so that the capacitor is adjustable from above the chassis. Neither C_2 nor C_3 is grounded to the chassis through its mounting — leads from the rotors are grounded to the chassis at one point near the 6AC7 tube socket. The oscillator coil, L_4, is mounted by its leads on a small multiple tie point.

The shield between the input coils, L_2 and L_3, is made of thin aluminum. It has a notch in the edge that goes against the chassis side, to clear the antenna-coil leads, and it has a hole through it for the lead between the bottoms of L_2 and L_3. The dual condenser, C_1, is fastened to the chassis by a single 6-32 screw, and the head of this screw has a copper shield soldered to it for minimizing coupling between C_{1A} and C_{1B}. The shield is easily cut out from copper flashing and soldered to the screw head. The rotor assembly of C_1 must be removed to put the shield in place, but this is just a matter of loosening four screws. Don't touch the stator plates. The screw with the shield on it, which holds C_1 to the chassis, also holds the coil shield in place underneath the chassis.

The 1700-kc. i.f. transformer is mounted on its side because the chassis and panel sizes are such that the receiver can be mounted in a small cabinet, and mounting the transformer upright would prevent any such installation. To lay the transformer on its side, two ⅜-inch diameter holes are drilled in the side of the i.f. can, opposite the coils. The leads from the i.f. transformer are brought out these holes and through corresponding holes in the chassis. An end plate on the transformer has a clearance hole for the grid lead. Fig. 5-35 shows these modifications and how the leads are connected. The 1700-kc. transformer is fastened to the chassis with two clamps using spade bolts. An alternative method would be to make a bracket of the end plate and another bracket at the adjusting-screw end of the transformer.

The 100-kc. circuits use a TV component, the RCA 205R1 Horizontal Oscillator coil. As purchased, they have the soldering lugs and tuning screw out of the top of the can, but they are easily reversed by uncrimping the can and reversing the assembly. Before reassembly, however, there are

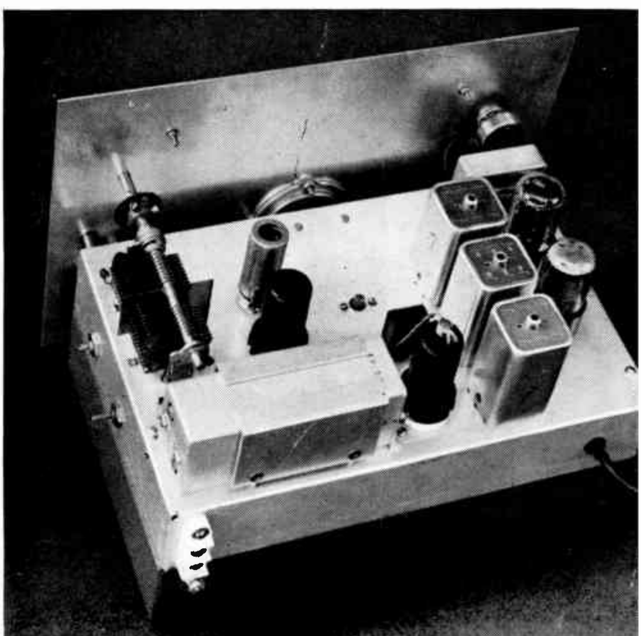


Fig. 5-34 — A top view of the five-tube superheterodyne shows how an aluminum and a steel chassis are combined for greater weight and strength. The 6C4 oscillator and 6AC7 mixer are at the left, and the two 6SN7s are at the extreme right. Note the shield between the stator sections of the condenser on the left.

a few things to be done. The large coil is used for the 100-kc. tuned circuit by connecting a 100-μμfd. mica condenser between Pins A and F and lifting the center-tap from Pin C. Don't break the center-tap — the easiest way is to scrape the two wires first to remove the insulation, flow a drop of solder on the scraped portion, and then cut the two wires away at the pin. The other winding is used as the primary in T_2 and the tickler in T_3. The primary in T_2 can be tuned from the top, because there is also an iron slug in this smaller coil.

In wiring the set, use tie points liberally so that no components will be floppy. The only shielded wires are the one running from the volume control to Pin 1 of the audio amplifier and the leads from T_3 to Pins 4 and 5 of the detector. The shields are grounded to the chassis at the ends and any other convenient points.

The oscillator coil, L_4, is made from B & W Miniductor. To separate the two coils of L_4, push the 3rd or 4th turn from one end of the piece of Miniductor through toward the center of the coil. Snip this wire with a pair of cutters and push the two ends back out. Each end is then peeled around for ½ turn. The two coils are adjusted to the right number of turns by working in from the outside ends.

The rotor of C_1 is connected underneath the chassis to the 0.001-μfd. coupling condenser by running a wire from the front support of the rotor through a ¼-inch clearance hole in the chassis. The 0.001-μfd. coupling condenser and L_2 and L_3 are grounded to the lug under L_2.

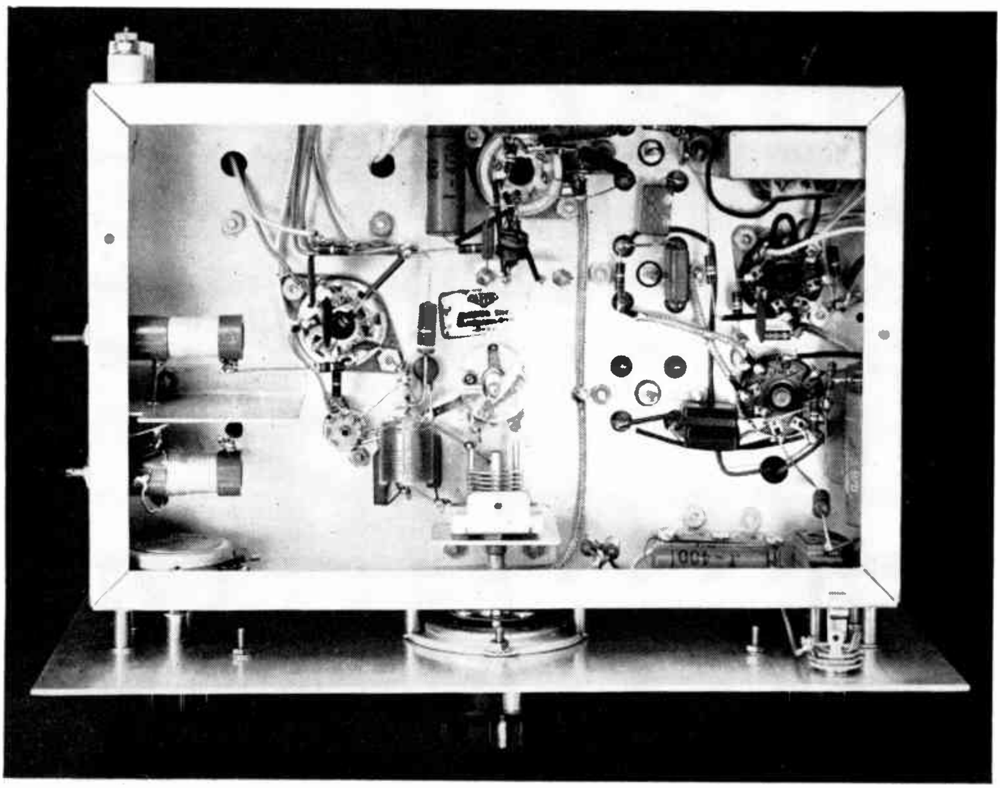
Adjustment

There are two types of adjustment that must be made to get the receiver working: adjusting the circuits to the proper frequencies and adjusting the oscillators and the regenerative detector to the proper amplitudes. To this latter end, leave the cathode end of R_1 disconnected in the original wiring, and lightly solder (so that it can be changed later) the lead from Pin 5 of the detector to Terminal C of T_3. Resistors that may require changing are R_2 and R_3, so don't solder them too well at first.

Connect a power supply to the receiver and see that the tubes light and that the power-supply voltages are approximately correct. The 250 volts can be anything 25 volts either side of 250, and the 105 volts, coming from a VR tube, will be nothing to worry about if the VR tube lights. A suggested power supply is shown in Fig. 5-37.

Next connect a low-range milliammeter between R_1 and cathode (+ lead to cathode) and apply power again. The grid current should read about 0.05 ma. (50 μa.). If it reads much more than this, try a slightly larger resistor at R_2, or a smaller one if the grid current is too low. Make these adjustments with the rotor arm of the r.f.

Fig. 5-36 — A bottom view of the five-tube superheterodyne. The audio choke, L_6, is in the upper right-hand corner, near where the power leads leave the chassis. The 6SN7 socket nearer the panel is the detector-b.f.o. section.



gain control at the grounded end.

Next check the oscillation of the 6C4 high-frequency oscillator. To do this, connect a 0–10 voltmeter across the 4700-ohm resistor in the plate circuit of the 6C4 (+ terminal to +105 side, − terminal to the 0.001-µfd. condenser). Observe the voltage reading and then touch your finger to the stator of C_2 or C_3. If the oscillator is working, the voltmeter reading will increase. If you get no change, it means the oscillator isn't working. With both coils of L_4 wound in the same direction (as they will be if Miniductor is used), the stator of the tuning condenser should be connected to the outer end of the larger coil, and Pin 5 of the 6C4 should be connected to the outside turn of the smaller coil.

If you can borrow a serviceman's test oscillator that will give a modulated signal at 1700 kc., this signal can be introduced at the grid of the 6K8 and the 100-kc. i.f. circuits can be peaked (b.f.o. turned off), listening in the headphones for maximum response. The 1700-kc. signal can then be transferred to the grid of the 6AC7 and the trimmers peaked on T_1. Lacking the signal generator, the alternative is to provide a modulated signal in the 80- or 40-meter band and couple it to the stator of C_{1B}. If the signal is from a crystal oscillator or VFO at 3750 kc. (for example), running from an unfiltered power supply to furnish the modulation, set the tuning dial vertical. If the signal is at 3500 kc., set the tuning condenser C_2 at almost full capacity. Rock C_3 slowly until the signal is heard. Then peak the 100-kc. transformers T_2 and T_3, reducing the signal input as necessary to avoid overloading. Next turn on the b.f.o. and adjust the slug in L_5 until a beat note is heard. Then peak the trimmers in T_1.

With the initial tuning of the 100-kc. channel done, the slugs of L_2 and L_3 can be adjusted for maximum signal, with no antenna connected. Set C_1 at almost full capacity, the signal near 3.5 Mc., and adjust the iron slugs for maximum in the headphones. If a VFO or crystal oscillator is furnishing the signal, there will probably be enough pick-up without any apparent coupling, but a short 6-inch wire connected to the antenna terminal may be required to pick up the output from a low-powered signal source.

It is not likely that the 100-kc. circuits will be tuned to the exact frequency that makes the calibrations coincide on 80 and 40 meters. While this isn't necessary, of course, it does make the dial look cleaner. To bring the calibrations into line, beg or borrow a frequency standard that will give signals at 100-kc. intervals. First locate the 4.0- and 7.0-Mc. points on the receiver dial, by referring the harmonics from the 100-kc. standard to the original signal you used for alignment. If, for example, the 80-meter signal you used was at 3650 kc., you know that the first 100-kc. harmonic you hear on the high-frequency side will be 3700 kc., and the first one on the low side will be 3600

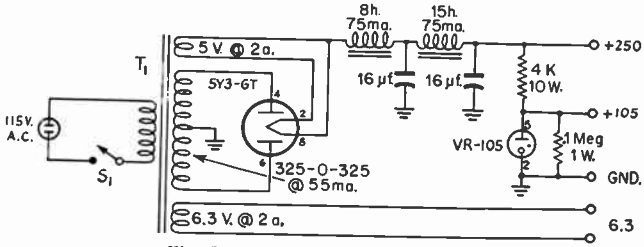


Fig. 5-37 — Suggested circuit diagram for the receiver power supply.

T_1 — Stancor PM-8407 or equivalent.
S_1 — S.p.s.t. toggle switch.

kc. The second harmonic of the 3650-kc. signal will furnish a check point at 7300 kc. (2 × 3650), so swinging C_1 to about ⅓ meshed (where it will peak the 7-Mc. signals) will allow you to locate the 7-Mc. points. Thus you will have 100-kc. intervals on the dial from 3.5 to 4.0 Mc. and from 6.9 to 7.4 Mc., but not necessarily coinciding. To make them coincide, some slight retuning of the 100-kc. transformers is required. If, for example, the 7.0-Mc. point occurs to the right of the 3.6-Mc. point, the 100-kc. amplifier is tuned low, and the slugs should be turned out slightly. A few trials will bring the circuits into place.

Now check the regeneration of the detector by connecting the lead from Pin 5 of the detector to D on T_3. If a steady beat is heard, indicating that the detector is oscillating, tune both circuits of T_2 and see if they will kill the oscillation. Their action is to load the regenerative detector to where it won't oscillate — if the action persists, try a 4700-ohm resistor at R_3 as a last resort. These circuits should be peaked on a modulated signal, with the b.f.o. turned off.

After the detector has been made regenerative, the calibration can again be checked as in a preceding paragraph, and any minor changes in tuning made as are found necessary. Once the 100-kc. circuits have been aligned they can be left alone, and if the 3.5- and 4.0-Mc. points don't come where you want them on the tuning dial, a slight adjustment of C_3 will correct it.

Connect a 140-µµfd. variable in series between antenna and the antenna post. On 80 meters, peak C_1 on a signal and rock the adjustment slug of L_2. If it tunes fairly sharp, the antenna coupling is not too tight on that band. Swing C_1 out until you are listening on 40 (to a signal) and again rock the slug on L_2. If it tunes broad, reduce the capacity of the 140-µµfd. antenna condenser until L_2 shows a definite peak. Note the settings of the condenser for the two bands.

The input condenser, C_1, will tune sharply on either band, and it should always be peaked when listening to a weak signal. Detuning it slightly will attenuate abnormally loud signals.

The power-supply requirements for the receiver are slight: about 15 ma. at 250 volts and 25 ma. at 105. A 60-ma. power supply will take care of this and the extra 10–12 ma. for a VR-105. A circuit diagram with suggested values is shown in Fig. 5-37.

A Clipper/Filter for C.W. or 'Phone

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

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

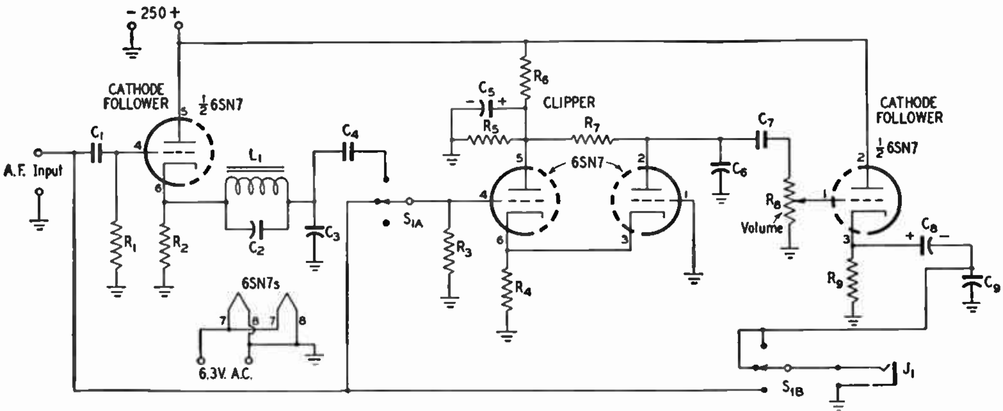


Fig. 5-38 — Circuit diagram of the audio clipper unit. Power requirements are 16 ma. at 250 v. d.c., 1.2 amp. at 6.3 v. a.c.

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

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

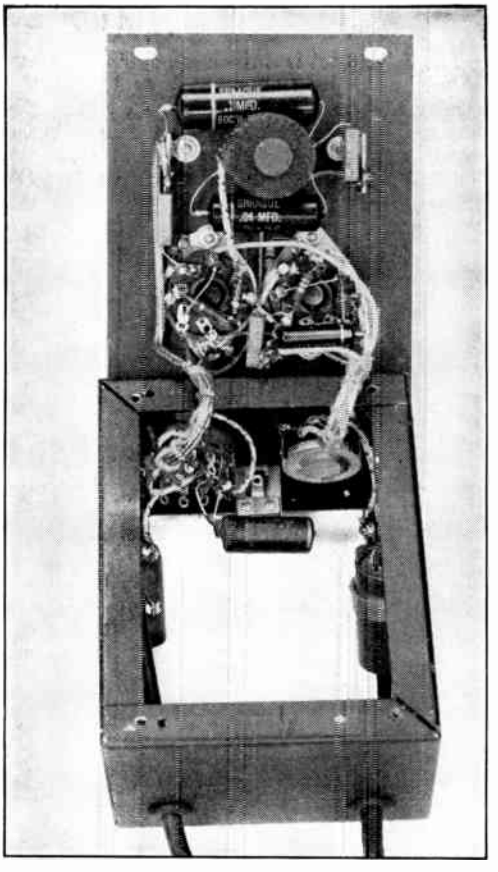


Fig. 5-40 — **Inside view of the clipper unit. The gain control, switch, headphone jack, and the larger fixed condensers are mounted on the walls of the box. The two tubes and the selective audio circuit are mounted on the removable panel. The selective circuit, consisting of the choke coil and two tubular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

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

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

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

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

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

The "Selectoject"

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

The wiring diagram of the Selectoject is shown in Fig. 5-41. Resistors R_2 and R_3, and R_4 and R_5, can be within 10 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. 5-41. The lead between the second detector and the receiver gain control should be broken and run in shielded leads to the two pins of the socket corresponding to those on the plug marked "A.F. Input" and "A.F. Output." If the receiver has a VR-150 included in it for voltage stabilization there will be no problem in getting the plate voltage — otherwise a suitable voltage divider should be incorporated in the receiver, with a 20- to 40-μfd. electrolytic condenser connected from the +150-volt tap to ground.

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

The Selectoject is useful as a means for obtaining much of the performance of a crystal filter from a receiver lacking a filter.

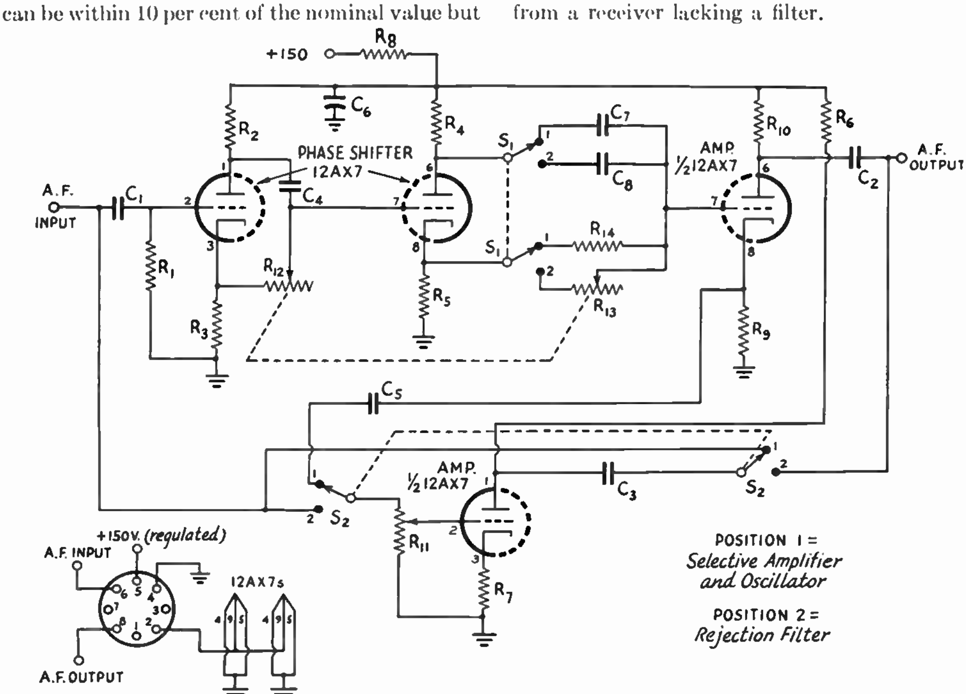


Fig. 5-41 — Complete schematic of Selectoject using 12AX7 tubes.

C_1 — 0.01-μfd. mica, 400 volts.
C_2, C_3 — 0.1-μfd. paper, 200 volts.
C_4, C_8 — 0.002-μfd. paper, 400 volts.
C_5 — 0.05-μfd. paper, 400 volts.
C_6 — 16-μfd. 150-volt electrolytic.
C_7 — 0.0002-μfd. 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}, R_{13} — Ganged 5-megohm potentiometers, standard audio taper (tuning control).
R_{14} — 0.12 megohm, ½ watt.
S_1, S_2 — D.p.d.t. toggle (can be ganged).

A Bandswitching Preselector for 14 to 30 Mc.

The performance of many receivers begins to drop off at 14 and 30 Mc. The signal-to-noise ratio is reduced, and trouble with r.f.-image signals becomes apparent. The preselector shown in Figs. 5-42 and 5-44 can be added ahead of any receiver without making any changes within the receiver, and a self-contained power supply eliminates the problem of furnishing heater and plate power.

As can be seen from the wiring diagram, Fig. 5-43, a 6AK5 r.f. pentode is used in the preselector. Both the grid and plate circuits are tuned, but the tuning condensers are ganged and only one control is required. The gain through the amplifier is controlled by changing the cathode voltage, through R_3. A selenium rectifier is used to supply plate power, and the heater power comes from a step-down transformer. The chassis is at r.f. ground but the d.c. circuit is isolated, to prevent short-circuiting the a.c. line through external connections to the preselector.

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

As shown in Fig. 5-42, the ganged condensers are controlled from the front panel by a National MCN dial, and a small knob to the right of this dial is connected to the antenna trimmer, C_4, for peaking the tuning with various antennas. The a.c. line is controlled by S_2, a toggle switch mounted on the panel.

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

The shield partition between the two switch sections (Fig. 5-44) straddles the tube socket and shields the grid from the plate circuit. The switched ends of all coils are supported by their respective switch points, and the other ends are soldered to tie points mounted on the chassis. The mica trimmers, C_9 and C_{10}, are supported on short lengths of stiff wire, and a hole in the side of the chassis is required to reach C_{10} with an aligning tool.

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

COIL TABLE FOR THE PRESELECTOR

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

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

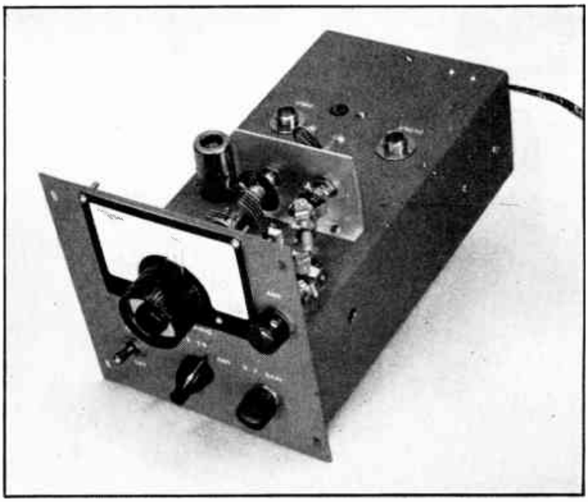


Fig. 5-42 — A bandswitching preselector for 14 and 28 Mc. A single 6AK5 amplifier is used, and the power supply is included in the unit. The antenna-trimming condenser is mounted on the small aluminum partition.

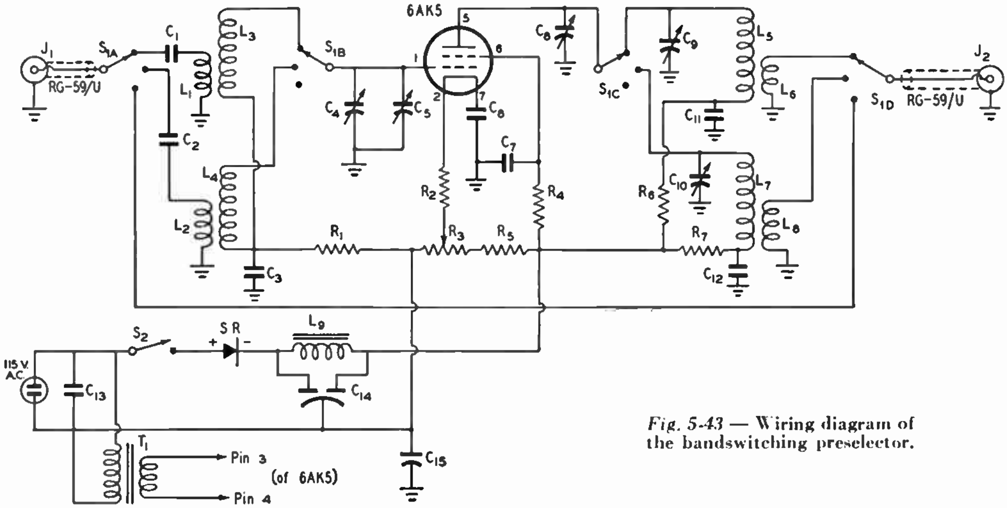


Fig. 5-43 — Wiring diagram of the bandswitching preselector.

C_1, C_2 — 10-μμfd. mica.
C_3, C_6, C_7, C_{11}, C_{12} — 680-μμfd. mica.
C_4 — 15-μμfd. midget variable (Millen 20015).
C_5, C_8 — 50-μμfd. midget variable (Millen 19050).
C_9, C_{10} — 3- to 30-μμfd. mica trimmer.
C_{13}, C_{15} — 0.01-μfd. paper, 400 volts.
C_{14} — Dual 40-μfd. 150-volt electrolytic.
R_1 — 27,000 ohms.
R_2 — 330 ohms.
R_3 — 5000-ohm wire-wound potentiometer.
R_4 — 4700 ohms.
R_5 — 18,000 ohms, 2 watts.
R_6, R_7 — 470 ohms.
L_1–L_8 — See coil table.
L_9 — 20-henry 30-ma. filter choke.
J_1, J_2 — Coaxial-cable jack (Jones S-101).
S_1 — 2-gang 2-circuit 5-position ceramic (Mallory 177C).
S_2 — S.p.s.t. toggle.
SR — 50 ma. selenium rectifier.
T_1 — 6.3-volt transformer.

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

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

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

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

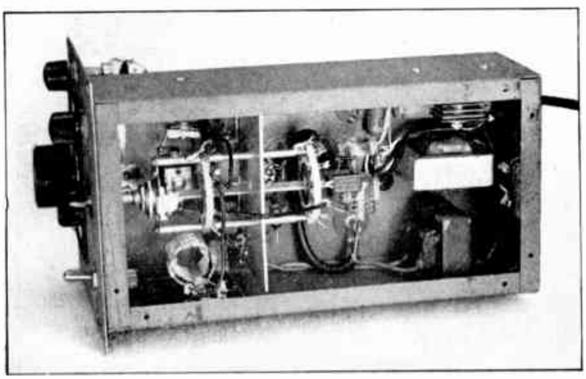


Fig. 5-44 — A view underneath the chassis of the bandswitching preselector, showing the shield partition between switch sections and the selenium rectifier and associated filter.

An Antenna-Coupling Unit for Receiving

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

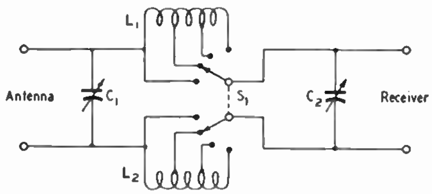


Fig. 5-45 — Circuit diagram of the coupling unit.

C_1 — 140-μμfd. midget variable (Millen 22140).
C_2 — 100-μμfd. midget variable (Millen 22100).
L_1, L_2 — 25 turns No. 26 d.c.c. space-wound to occupy 1 inch on 1-inch diameter form (Millen 45000), tapped at 2, 5, 8, 12 and 18 turns.
S_1 — 2-circuit 5-position single-section ceramic wafer switch (Mallory 173C).

One convenient type of antenna-coupling unit for receivers uses the familiar pi-section filter circuit, and can be used to match a wide range of antenna impedances. The diagram of a compact unit of this type is shown in Fig. 5-45. Through proper selection of condensers and inductances, a match can be obtained over a wide range of values. The device can be placed close to the receiver and left connected all of the time, since it will have little or no effect on the lower frequencies. A short length of 300-ohm Twin-Lead is convenient for connecting the antenna coupler to the receiver.

The antenna coupler is built in a 5 × 7 × 2-inch metal chassis. All of the components except the two coils are mounted on the front and rear faces. The condensers are mounted off the panel by the spacers furnished with the condensers, and a clearance hole for the shaft prevents any short-circuit to the panel. The coils, wound on Millen 45000 phenolic forms, are fastened to the chassis with brass screws, and the coils should be wound on the forms as far away as possible from the mounting end. The switch should be wired so that the switching sequence puts in, in each coil, 2 turns, 5 turns, 8 turns, 12 turns, 18 and 25 turns.

The unit is adjusted for maximum signal by switching to different coil positions and adjusting C_1 and C_2. It will not be necessary to retrim the condensers except when going from one end of a band to the other, and when the unit is not in use, as on 7 and 3.5 Mc., the coils should be set at the minimum number of turns and the condensers set at minimum. The small reactances remaining have a negligible effect. The coil in the grounded side should be shorted if coaxial-line feed is used.

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Fig. 5-46 — A compact coupling network for matching a balanced line to the receiver on 14 and 28 Mc.

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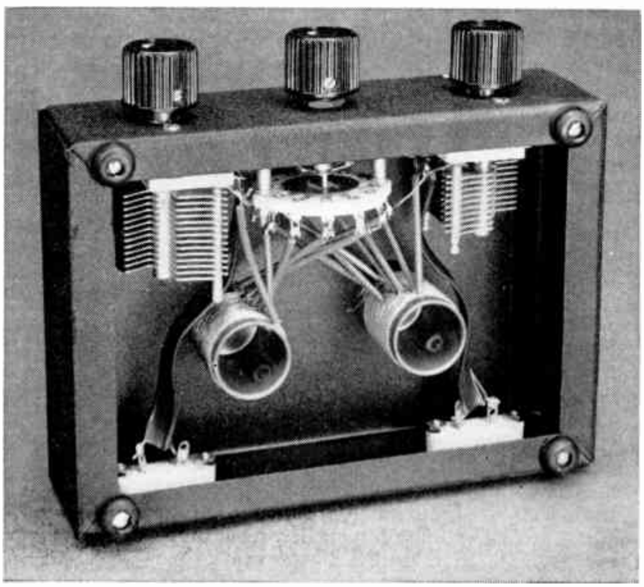


Receiver Matching to Tuned Lines

The pi-section coupler shown in Figs. 5-45 and 5-46 can be used in many instances for matching a balanced open-wire line to the receiver, and it can be used with an unbalanced line by short-circuiting the inductance in the grounded side of the unbalanced line. However, there are many applications where another type of coupler is slightly more advantageous, as when an all-band antenna system with tuned feeders is used, or where a wide range of line impedances may be encountered. This other type of coupler, shown in Figs. 5-47, 5-48 and 5-49, is simply a scaled-down transmitter coupler, with provision for either series or parallel tuning. The change from series to parallel tuning is made simply by the manner in which the antenna connection plate is plugged into the unit.

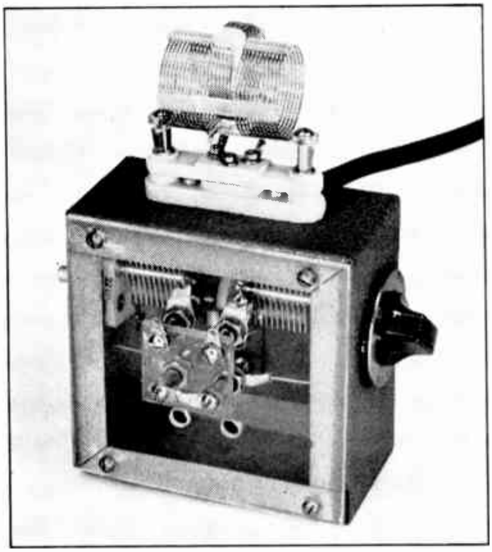


Fig. 5-47 — A small tuned coupler for matching the receiver to a tuned line. The unit is made either series- or parallel-tuned by the position of the antenna connection block.

As can be seen in the wiring diagram, Fig. 5-48, when the antenna connection plate is plugged in so that all four contacts are engaged, the two condensers are connected across the coil in series, to give parallel tuning. When the plate is dropped down, so that only the antenna plugs engage at *A* and *B*, the unit is connected for series tuning. Small low-power transmitting coils with swinging links are used.

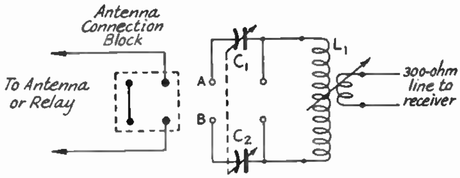


Fig. 5-48 — Circuit of the tuned antenna coupler. C_1, C_2 — 100-μμfd. midget variable (Millen 22100). L_1 — Coil to tune to band in use, with swinging link (National AR-16).

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

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

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

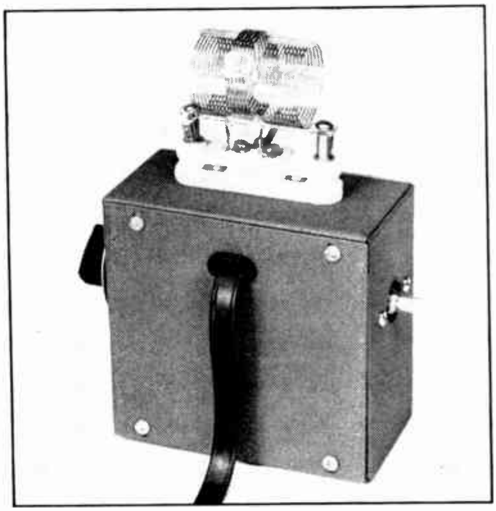


Fig. 5-49 — Another view of the tuned antenna coupler.

A One-Tube Converter for 10 and 11 Meters

The 10- and 11-meter converter shown in Figs. 5-50 and 5-52 is a simple unit that can be built in a few hours, for a cost of less than fourteen dollars. The converter uses a fixed-tune i.f. and tunable input and oscillator circuits, in preference to a fixed-frequency oscillator and a tunable output circuit. With a one-tube converter of the latter type, it is almost impossible to avoid picking up at least a few signals in the tuning range of the receiver. Using a tunable oscillator and a fixed-frequency output circuit permits one to select an i.f. free from interference. The plate-current demand is only 5 ma., and it is usually possible to operate the converter from the receiver power supply.

As can be seen in Fig. 5-51, the Hartley circuit is used in the oscillator portion of the 6BA7 pentagrid converter. A padding condenser, C_2, is switched in through S_1 to change the range for 11-meter operation. Condenser C_4 is used for tuning, and the input circuit is tuned to either range with C_1. The screen grid of the 6BA7 is operated at about 65 volts, since higher voltages will increase the total tube current without any marked improvement in performance. However, since the available supply voltage will vary with different receivers, the value of the screen dropping resistor, R_2, cannot be specified, and it must be calculated, as described later.

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

Construction

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

It can be seen in Fig. 5-50 that the oscillator tuning condenser, C_4, is mounted on ¼-inch metal pillars. A National type GS-10 stand-off insulator is located at the front-right-hand side of C_4, and a soldering lug at the top end of this insulator is soldered to the stator terminal lug of the condenser. This added support for the tuning condenser improves oscillator stability, by preventing rocking of C_4 as the control shaft is turned. A feed-through bushing at the other front terminal of the condenser is used to support and insulate the lead passing through the chassis to the coil below. The padder condensers for the oscillator circuit, C_3 and C_5, are mounted on the rear terminal lugs of the tuning condenser.

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

The screen and decoupling resistors, R_2 and R_3, respectively, are supported at the power-supply ends by a tie-point strip which is held in place by the same screw that anchors the soldering lug for L_3. If the receiver supply voltage is known at this time, it is possible to calculate the correct value for the screen-dropping resistor, and the resistor can be mounted on the tie-point strip. The resistor value is obtained from the equation

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

Example: Supply voltage 260; the resistor value is

$$\frac{260 - 65}{0.0046} = 42{,}391 \text{ ohms.}$$ Anything within 20% of this figure would be satisfactory.

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

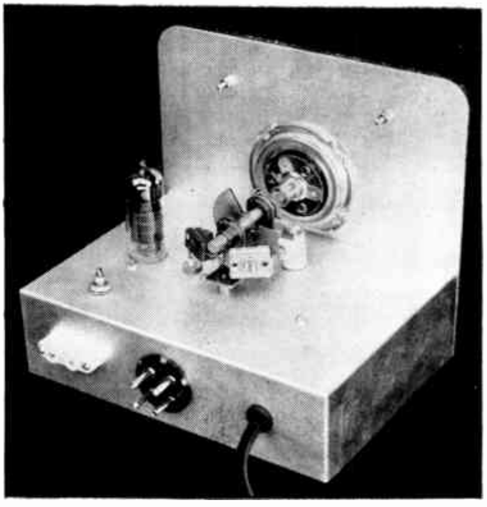


Fig. 5-50 — A one-tube converter for extending the tuning range of a receiver to 10 and 11 meters. The crystal socket on the back of the chassis receives the antenna plug (Millen 37412).

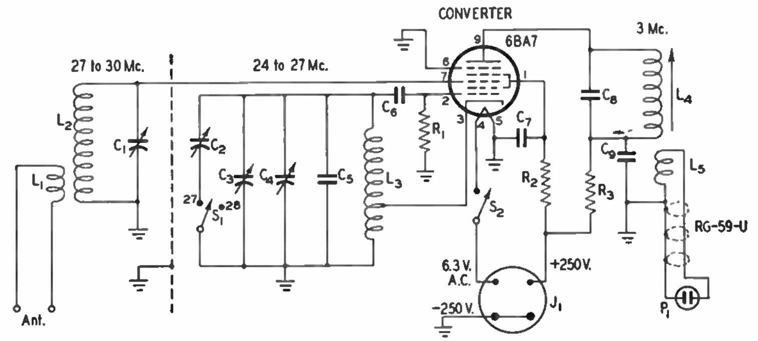


Fig. 5-51 — Circuit diagram of the low-cost 10- and 11-meter converter.

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

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

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

Testing

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

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

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

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

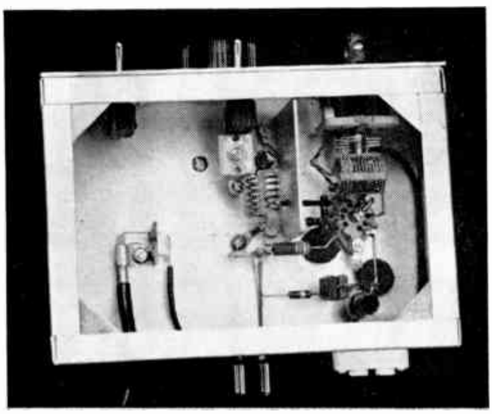


Fig. 5-52 — A bottom view of the one-tube converter. The toggle switches are for band-changing and opening the heater circuit.

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

The principle of using a fixed high-frequency oscillator in a converter and tuning the receiver the converter works into can be elaborated upon by using a stage of r.f. amplification ahead of the mixer and by using a crystal-controlled oscillator for maximum stability. Since such a converter is generally used on a high frequency where fundamental crystals are not available, it is necessary to use a harmonic of a lower-frequency crystal. A crystal-controlled converter of this type is shown in Figs. 5-53 and 5-55. A separate converter is required for the 14-, 21- and 27-/28-Mc. bands, since by using separate converters it is possible to simplify their construction and to maximize their performance.

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

The wiring diagram is shown in Fig. 5-54. A neutralized triode-connected 6AK5 is used for the r.f. amplifier. There is some question as to its necessity on 14 and 21 Mc., where the atmospheric noise is generally high enough to limit the maximum usable sensitivity. A pentode-connected 6AK5 could probably be used with no detectable difference in performance on 14 and 21, but the triode is easy to handle and you don't lose anything by using it. Using high-impedance circuits with the pentode might give trouble from regeneration, unless the stage were neutralized. Adjustable antenna coupling and a Faraday screen are included to accommodate various antenna systems and to eliminate capacity coupling to the antenna line. The r.f. stage runs at 105 volts on the plate, since this gives the best noise figure. The separate plate lead also offers an opportunity to kill the converter by opening this circuit. The 6AK5 pentode mixer is easy to handle and quiet enough so that its noise doesn't impair the over-all performance. A triode mixer might be used, but the pentode runs with low current and is quiet.

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

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

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

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

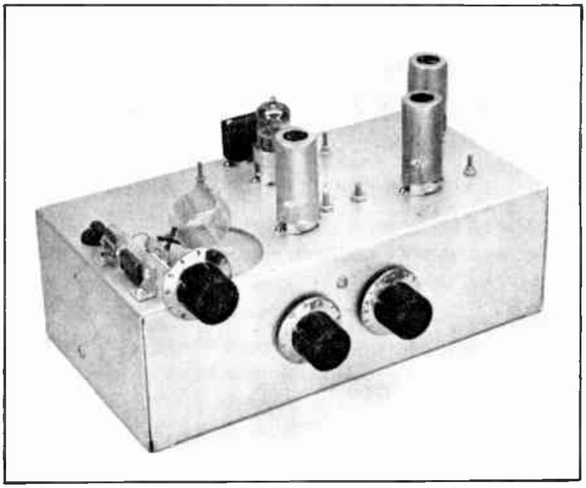


Fig. 5-53 — A 28-Mc. crystal-controlled converter. The adjustable antenna coupling can be seen at the left front. The tube shields, from left to right, cover the triode-connected 6AK5 r.f. amplifier, the 6AK5 mixer and the 6C4 cathode follower. The unshielded tube is the 6J6 oscillator-multiplier.

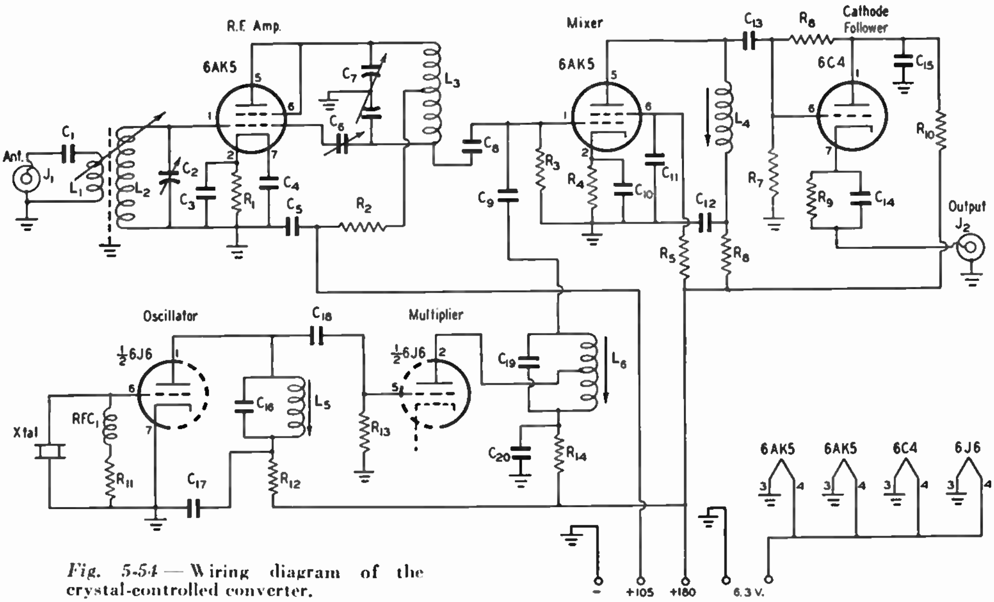


Fig. 5-54 — Wiring diagram of the crystal-controlled converter.

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

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

A 2¼-inch diameter hole is punched in the chassis, so that the externally-mounted antenna coil, L_1, can be coupled to the grid coil, L_2. The Faraday screen is then mounted across this hole on the underside of the chassis. To construct the Faraday shield, first cut a piece of ⅛-inch-thick polystyrene (Millen Quartz-Q) to measure 2½ by 3¼ inches, and drill a pair of holes at one end to clear No. 6 screws, for mounting the finished shield. (These are the same screws that hold the mounting strip for the antenna condenser, C_1, visible in Fig. 5-53.) At the opposite end of the poly sheet, drill a small hole in each corner, for securing the wire used in making the shield. Then wind No. 20 tinned wire tightly around the poly sheet in the long direction, spacing it with string or more No. 20 wire. When the winding is finished and secured at both ends, unwind the spacing string (or wire) and remove it. If you have done the job carefully, you will have neat parallel lines of wire across the polystyrene, all equally spaced and all lying fairly flat. Then apply two or three heavy coats of Duco cement to *one side only*, allowing sufficient time between coats for the cement to harden thoroughly. When this has been done, it will be found an easy job to cut each wire on the uncemented side. Straighten out the wires so that you now have a flat sheet of parallel wires, and trim off the wires at the mounting holes end of the sheet along a line inside the mounting holes. Figs. 5-55 and 5-56 show what this looks like. When trimming these wires, be careful to see that no wire is left touching an adjacent one. Trim the wire ends at the other end to about ½ inch from the polystyrene. Clamp the shield in a vise, between two pieces of wood, and wrap each wire end around a piece of No. 12 tinned copper, as shown in Fig. 5-56. With a good hot iron, run a bead of solder along the bus, and your shield is finished. Work fast, and no heat will reach the poly. The shield is mounted with the smooth side exposed through the hole, and one end of the No. 12 bus is grounded at the r.f. tube socket.

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

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

COIL TABLE FOR THE CRYSTAL-CONTROLLED CONVERTER

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

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

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

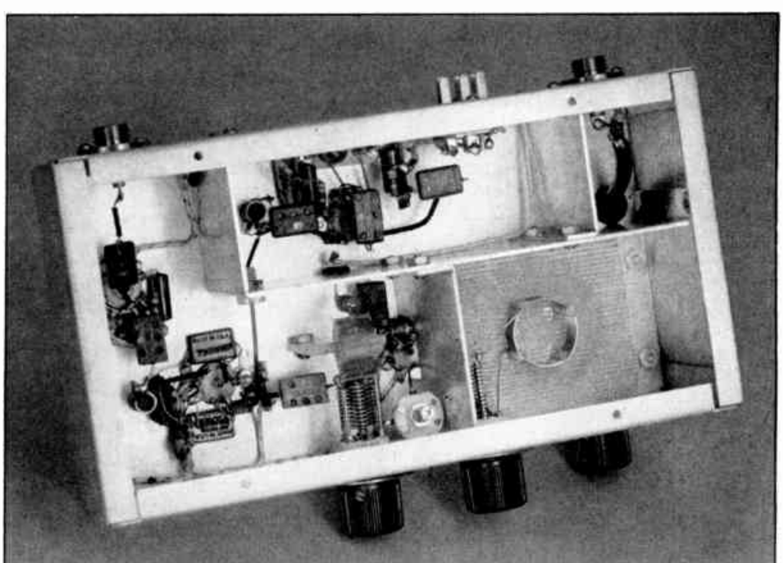


Fig. 5-55 — This view of the underside of the converter with the bottom cover removed shows the Faraday shield at the lower right, the shield straddling the r.f. amplifier socket (lower center) and the shielded oscillator section (top center). The neutralizing condenser for the r.f. stage is adjusted through a hole in the bottom cover.

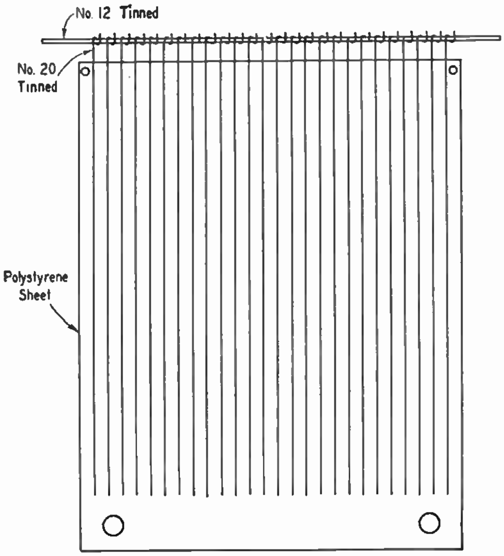


Fig. 5-56 — Constructional details of the Faraday shield, before soldering the ends of the No. 20 wires to the No. 12 wire bus.

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

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

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

Power Supply

The circuit diagram of a suitable power supply for use with the converters is shown in Fig. 5-57, although any source of 6.3 volts a.c. and 105 and 180 volts d.c. will do. One set of connections runs to the converter in use, and the other goes to a small control box located on the operating table. If desired, the a.c. switch can be incorporated in the power supply, but the plate switch, in the 105-volt lead to the r.f. stage, should be handy to the operator. A switch can be provided for shifting the power from one converter to another. Since separate receiving antennas are generally used at these frequencies, the antennas do not require switching.

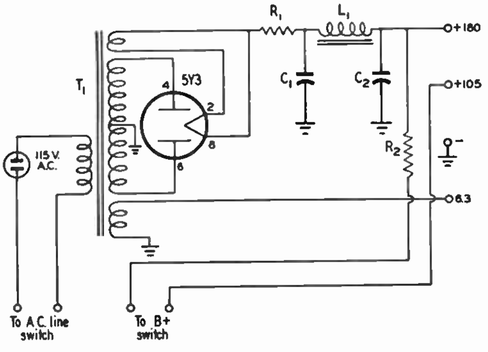


Fig. 5-57 — A power supply for the crystal-controlled converter.

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

An All-Purpose Super-Selective I.F. Amplifier

The amplifier shown in Figs. 5-58 and 5-60 is designed to connect to any receiver at the grid of the first i.f. tube, to give superior selectivity for either 'phone or c.w. reception. The signals at 455 kc. are heterodyned to 50 kc. and filtered through either or both of two selective amplifiers. One of the amplifiers uses 11 high-Q tuned circuits to give a selectivity characteristic that is about 350 cycles wide at 6 db. down and 1300 cycles wide at 60 db. down. The other amplifier uses 9 "stagger-tuned" circuits that give a 2300-cycle bandwidth at 6 db. down and 5 kc. at 60 db. down. The broader amplifier has its tuning adjusted so that it is centered about 1700 cycles higher in frequency than the sharp one. Thus, when a 'phone carrier is tuned to fall in the center of the sharp amplifier, one sideband falls in the broader amplifier. The outputs of the amplifiers are fed to a common detector, and the relative amplitude of carrier and sideband at the detector can be changed by controlling the gains through the two amplifiers. By emphasizing the carrier at the detector, "exalted-carrier" reception is obtained, which has the advantage that fewer distortion products are generated on a signal in the presence of QRM. For c.w. reception, only the sharp amplifier is used, while the reception of SSB signals requires only the broad amplifier.

The complete circuit of the amplifier is shown in Fig. 5-59. Receiver output at 455 kc., at as low a level as possible (to avoid overloading), is fed into the 6BE6 converter stage, where a crystal-controlled oscillator is selected either 50 kc. higher or lower, to use the selectable-sideband principle.[1] A third position of the switch, S_1, permits running both crystals at once, for alignment purposes, as described later.

The two i.f. amplifiers follow the converter, and two 6BJ6 variable-μ pentodes are used in each channel. There are isolation resistors and condensers in each power lead to prevent any over-all feed-back.

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

The resistor, R_{50}, between gain control, R_{17}, and ground, is used to bring the relative maximum gains of the two channels to approximate equality. The gain of the broad channel will vary with the degree of stagger-tuning, so R_{50} should be inserted only after the alignment procedure has been completed. Its value, of course, may work out differently than that shown.

The detector uses two 12AU7 dual triodes in in the "product detector" circuit. The advantage of the circuit is that it minimizes intermodulation at the detector and doesn't require a big b.f.o. signal for exalted-carrier reception. A signal-level indicator circuit connected to the sharp amplifier doesn't indicate b.f.o. voltage, so the signal-level meter reads the same with b.f.o. on or off.

The signal-level circuit, labeled "A.V.C.-Rect." in Fig. 5-59, consists of a cathode follower driving a diode. In three positions of S_2, the rectified current simply works the meter, but an a.v.c. voltage is applied throughout the amplifier in the fourth position.

The tuning meter is important. It permits the operator to center the carrier in the sharp amplifier, and also warns him when the amplifier is in danger of overloading. Overloading will tend to nullify the advantages of high selectivity, so it is important that the unit always be operated below this point. The manual gain controls will take care of about 60-db. range.

The series trap, RFC_5C_{48}, is tuned to 50 kc. to by-pass the r.f. and prevent its getting on the audio grids. A choice of two low-impedance outputs is provided, for 'phones and loudspeaker.

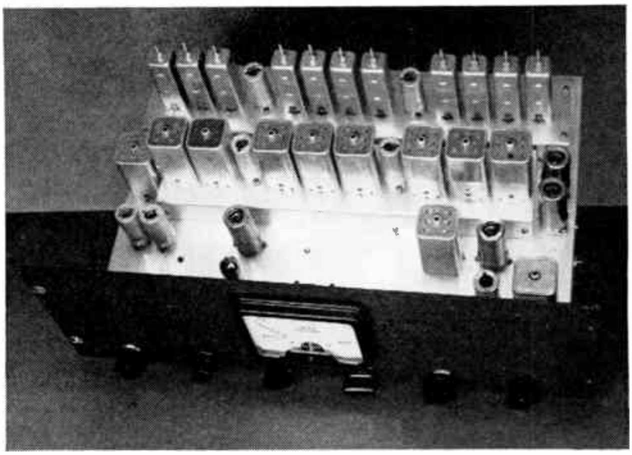
Construction

There are only a few departures from conventional construction technique in this amplifier. Miniature tubes were used only to provide room for the tuned circuits — on a larger chassis or with a different layout, metal tubes should be perfectly satisfactory. However, no attempt should be made to save space by mounting the

Fig. 5-58 — The super-selective i.f. amplifier uses two channels in parallel — a sharp one for c.w. or for 'phone carrier, and a broad one for a 'phone sideband.

The sharp i.f. is the strip at the rear of the chassis, and the broad one is just in front of it. The two tubes at the right-hand end of the broad amplifier are the "product detector." The b.f.o. can is at the front right, next to the tube, and the near-by tube and can are in the signal-metering circuit.

The controls, from left to right, are sideband selector switch, audio volume, broad i.f. gain, sharp i.f. gain, function switch, and b.f.o. pitch control.



tuned circuits in anything but a straight line. The shield cans do not provide complete magnetic shielding at 50 kc., and it is possible to couple right through the thin aluminum.

The i.f. strips proper are built on aluminum channels. All power leads are brought out through shielded wires, to minimize coupling via the common power circuits. Using the shielded wire is also an aid to construction, because the shields are soldered to lugs at points near the tube sockets, and the isolating resistors are then mounted between tube socket (or coil terminal) and the exposed ends of the shielded wires. The Hallicrafters coils leave no room for the associated shunt condensers, so they are connected directly across the terminals.

The RCA coils, used in the broad amplifier, must be reworked slightly before using. As supplied, the terminals come out the top of the can, so the coil must be removed by untwisting four small tabs. The coil to be used is connected to Terminals A and F, and another coil connected to Terminals C and D should have its leads snipped. The 390-$\mu\mu$fd. silver-mica condenser can then be soldered to Terminals A and F before the assembly is replaced in the shield can.

The b.f.o. coil, L_1, uses both coils of the RCA 205R1 connected in series. This is done by lifting the single wire from Terminal C and connecting it to Terminal F. Externally, Terminals A and D are used.

The main chassis is aluminum, 12 by 17 by 2 inches, and the front panel is a standard relay-rack affair 7 inches high. The shielded leads from the i.f. strips proper are brought out through holes to tie points conveniently located away from signal circuits. Two short pieces of RG-59/U coaxial cable are used — one from the input jack at the rear of the chassis up to the 6BE6 grids, and the other from the output of the sharp

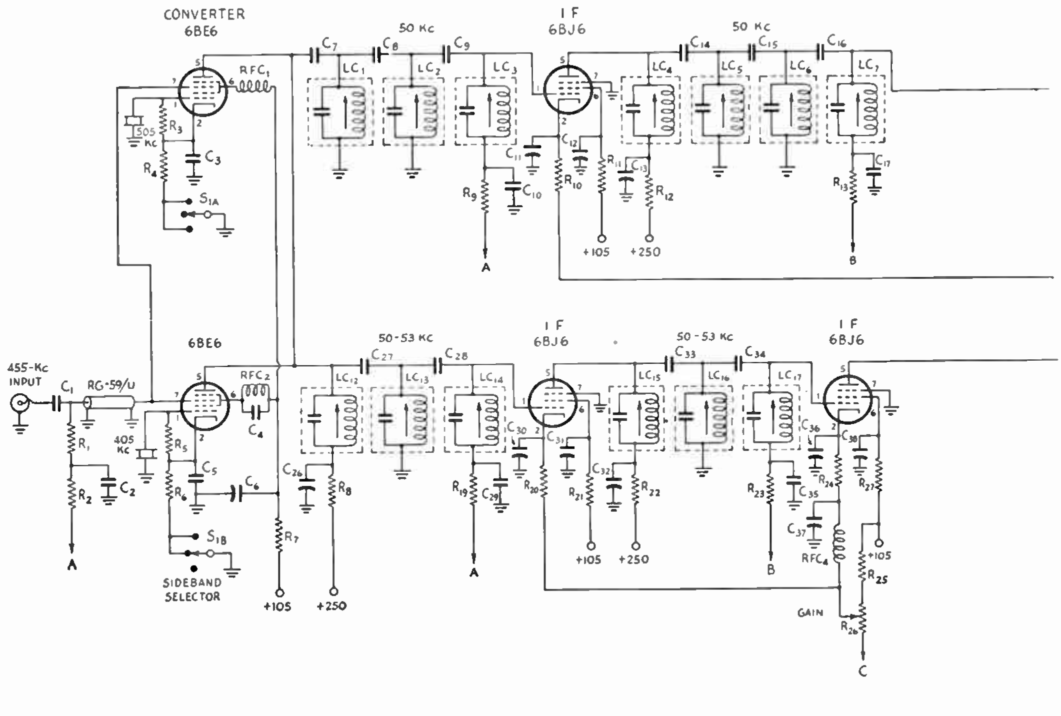


Fig. 5-59 — **Wiring diagram of the 50-kc. selective amplifier.**

C_1 — 0.005-μfd. ceramic.
C_2, C_6, C_{11}, C_{12}, C_{13}, C_{18}, C_{19}, C_{20}, C_{21}, C_{26}, C_{30}, C_{31}, C_{32}, C_{36}, C_{37}, C_{38}, C_{39}, C_{42}, C_{44}, C_{45}, C_{59} — 0.1-μfd. 400-volt.
C_3, C_5, C_{10}, C_{17}, C_{29}, C_{35}, C_{43}, C_{52} — 0.01-μfd. ceramic.
C_4 — 47-$\mu\mu$fd. ceramic.
C_7, C_8, C_9, C_{14}, C_{15}, C_{16}, C_{22}, C_{23}, C_{24} — 2.4-$\mu\mu$fd. mica (two 4.7-$\mu\mu$fd. in series if lower value not available).
C_{25} — 100-$\mu\mu$fd. ceramic.
C_{27}, C_{28}, C_{33}, C_{34}, C_{40}, C_{41} — 4.7-$\mu\mu$fd. mica.
C_{46}, C_{51} — 16-μfd. 450-volt electrolytic.
C_{47} — 0.002-μfd. ceramic.
C_{48} — 250–970-$\mu\mu$fd. adjustable mica (El Menco 306).
C_{49} — 0.001-μfd. ceramic.
C_{50}, C_{53} — 10-μfd. 50-volt electrolytic.
C_{54} — 470-$\mu\mu$fd. ceramic.
C_{55} — 35-$\mu\mu$fd. midget variable.
C_{56} — 220-$\mu\mu$fd. silver mica.
C_{57}, C_{58} — 3300-$\mu\mu$fd. silver mica.
C_{60}, C_{61} — 20-μfd. 50-volt electrolytic.
C_{62} — 10-$\mu\mu$fd. ceramic.
R_1 — 0.15 megohm.
R_2, R_9, R_{13}, R_{19}, R_{23}, R_{32}, R_{40}, — 0.1 megohm.
R_3, R_5 — 0.12 megohm.
R_4, R_6 — 330 ohms.
R_7, R_8 — 2700 ohms.
R_{10}, R_{14}, R_{20}, R_{24}, R_{48} — 100 ohms.
R_{11}, R_{12}, R_{15}, R_{16}, R_{21}, R_{22}, R_{27}, R_{28} — 10,000 ohms.
R_{17}, R_{26} — 2000-ohm wire-wound potentiometer.
R_{18}, R_{25} — 27,000 ohms, 1 watt.
R_{29} — 1500 ohms.

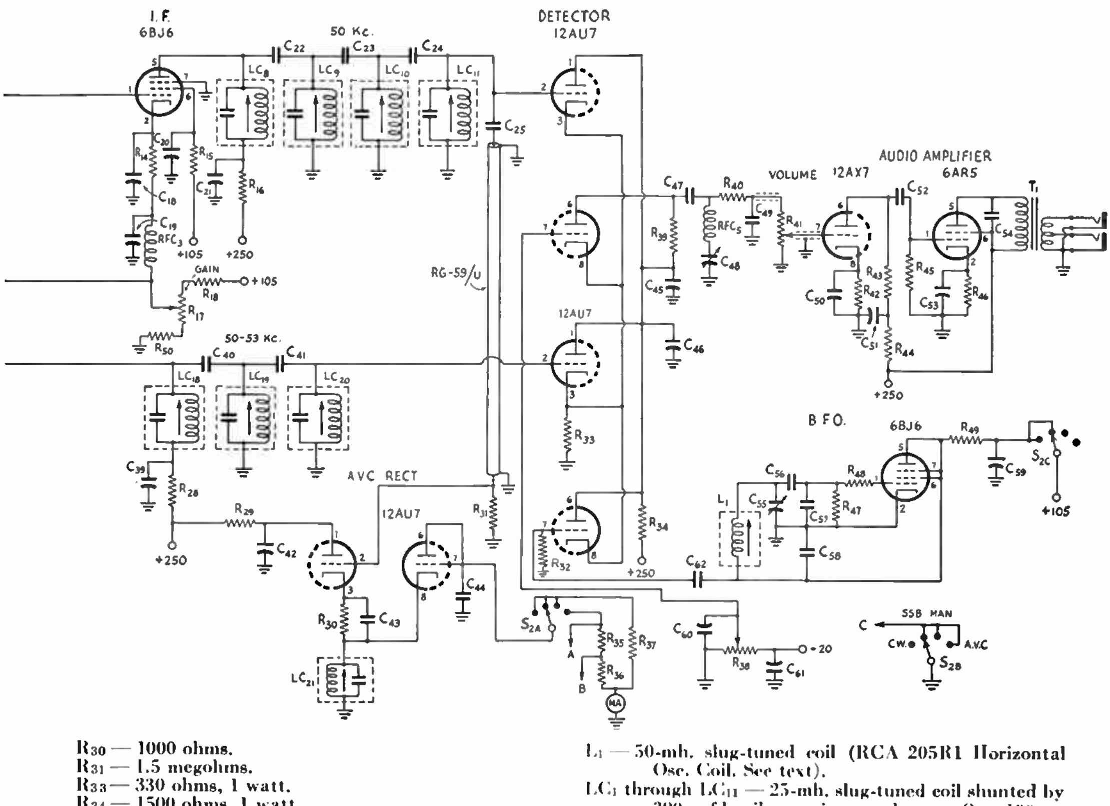
i.f. amplifier to the grid of the 12AU7 a.v.c.-rectifier. The input and output signal leads from the i.f. amplifiers are fed through Millen 32150 ceramic bushings, where the projecting wire serves as a tie point. The detector bias control, R_{38}, is mounted at the rear of the chassis, since it need not be touched after the original adjustment for minimum detection in a single channel, except when a 12AU7 detector tube is replaced.

Alignment

The best point in a receiver to take off the signal for this i.f. amplifier is at the grid of the first i.f. stage in the receiver. If the receiver has a crystal filter between mixer and i.f. stage, it won't be used normally. The crystal filter can be used, but it requires getting two oscillator crystals for the sharp i.f. amplifier of just the right frequency.

The frequency to which the selective amplifier is aligned is determined by the frequencies of the two crystals in the 6BE6 converters. Assume that the nominal i.f. frequency of the receiver is 455 kc., and that the available crystals are 408 and 505 kc. The sharp i.f. will then be aligned to half the difference, or 48.5 kc. (408 + 48.5), but the fact that this is 1.5 kc. higher than the nominal 455 is nothing to worry about.

Set a signal generator or test oscillator to half the crystal-oscillator difference (e.g., 48.5 kc.) and align the sharp channel by working back from the detector, introducing the signal first at the grid of the second 6BJ6, and aligning the following circuits, and then introducing the signal at the first 6BJ6 and then the 6BE6 mixer. The final touching up of the sharp amplifier is done by switching S_1 to the point where both 6BE6s are operative and tuning a signal at 455 kc. until it "zero beats" with itself, as heard in the output. The sharp circuits are then given a fi-



R_{30} — 1000 ohms.
R_{31} — 1.5 megohms.
R_{33} — 330 ohms, 1 watt.
R_{34} — 1500 ohms, 1 watt.
R_{35} — 4700 ohms.
R_{36} — 6800 ohms.
R_{37} — 12,000 ohms.
R_{38} — 5000-ohm wire-wound potentiometer.
R_{39}, R_{44} — 47,000 ohms.
R_{41} — 0.5-megohm volume control.
R_{42} — 2200 ohms.
R_{43}, R_{45} — 0.22 megohm.
R_{46} — 450 ohms, 1 watt.
R_{47} — 47,000 ohms, 1 watt.
R_{49} — 68,000 ohms, 1 watt.
R_{50} — 270 ohms; adjust to balance gains.
All resistors ½ watt unless specified otherwise.

L_1 — 50-mh. slug-tuned coil (RCA 205R1 Horizontal Osc. Coil. See text).
LC_1 through LC_{11} — 25-mh. slug-tuned coil shunted by 390-μμfd. silver mica condenser. $Q = 100$ at 50 kc. (Hallicrafters 50B489).
LC_{12} through LC_{21} — 25-mh. slug-tuned coil shunted by 390-μμfd. silver mica condenser. $Q = 60$ at 50 kc. (RCA 205R1 Horizontal Osc. Coil modified. See text).
RFC_1, RFC_2 — 750-μh. r.f. choke (National R-33).
RFC_2, RFC_4 — 10-mh. r.f. choke (National R-50-1).
RFC_5 — 25-mh. r.f. choke (Millen 34225).
MA — 0.2-ma. milliammeter.
S_1 — Two-circuit 3-position wafer switch.
S_2 — Three-circuit 4-position wafer switch.
T_1 — 8-watt output transformer (Merit A-2901).

nal peaking, as indicated by the tuning meter. During alignment procedures, always work with a minimum signal and with the gain control, R_{17}, advanced to maximum gain.

The b.f.o. is aligned by switching it on, setting C_{55} to the center of its range, and adjusting the slug in L_1 to zero beat on a signal peaked through the sharp amplifier.

The broad i.f. amplifier is "stagger-tuned," which means that alternate circuits are tuned to the same frequency. First, peak circuits LC_{12} through LC_{20} to a slightly higher (1.5 kc.) frequency than the sharp channel. While doing this, the lead from the meter circuit can be transferred from LC_{11} to LC_{20}, and the signal introduced to the grid of a 6BE6. Then set the signal source to a frequency 750 cycles higher than the frequency at which the sharp channel was peaked, and peak circuits LC_{12}, LC_{14}, LC_{16}, LC_{18} and LC_{20}, as indicated by the meter. Then set the signal source to a frequency 2750 cycles higher than the sharp-channel frequency, and peak circuits LC_{13}, LC_{15}, LC_{17} and LC_{19}. Now, varying the frequency of the signal source, the response indicated by the meter will show a response that has two unequal peaks. The peaks can be equalized, or nearly so, by readjustment of LC_{12}. The lead from the meter circuit can now be returned to LC_{11}.

If an audio output meter is available, get a final check on the response of the broad amplifier by setting the b.f.o. to the midfrequency of the sharp amplifier and, with the sharp amplifier turned down, swing the input signal across the range and watch the audio response. It should be fairly flat from about 500 to 2700 cycles or so, dropping off rapidly beyond that.

Without access to a signal generator, it may be necessary to rig up a 50- or a 450-kc. oscillator with good stability and a slow tuning rate.

Operation

The operator has his choice of several types of operation with this amplifier. For highly-selective c.w. reception, use switch S_2 in the "C.W." position, with the b.f.o. offset to give the favorite beat-note frequency. Signals will drop in and out rapidly as one tunes across a band, and a slow tuning rate is highly desirable. For less critical reception of c.w., or for net operation, switch to "SSB" and use the broad i.f. characteristic, reducing the gain in the sharp channel to a minimum. The same settings maintain for the reception of SSB 'phone signals — the b.f.o. is set to the midfrequency of the sharp channel and all tuning is done with the main tuning dial of the receiver.

Regular AM 'phone signals are received with S_2 set either to "MAN." or "A.V.C.," depending upon the QRM conditions. In either case, the carrier is peaked on the meter for accurate tuning, and the two gain controls are set for best listening. In "MAN." operation this will usually mean riding gain on the sharp channel so that the meter never goes beyond half-scale, and with the broad-amplifier gain control backed off proportionately. In "A.V.C.," both controls can be run wide open, but as one tunes across some signals the set may overload until the tuning is centered on the desired carrier. A heterodyne on one sideband will be eliminated by switching S_1. "Practice" is the only advice one can give on handling the i.f. amplifier to its greatest capabilities, always remembering that you have the choice of two sidebands to listen to plus the ability to vary the relative amplitudes of carrier and sidebands.

As in all selective amplifiers, overload is the big enemy, and it is generally best to run the audio volume at or near maximum and the i.f. gain at the lowest usable value.

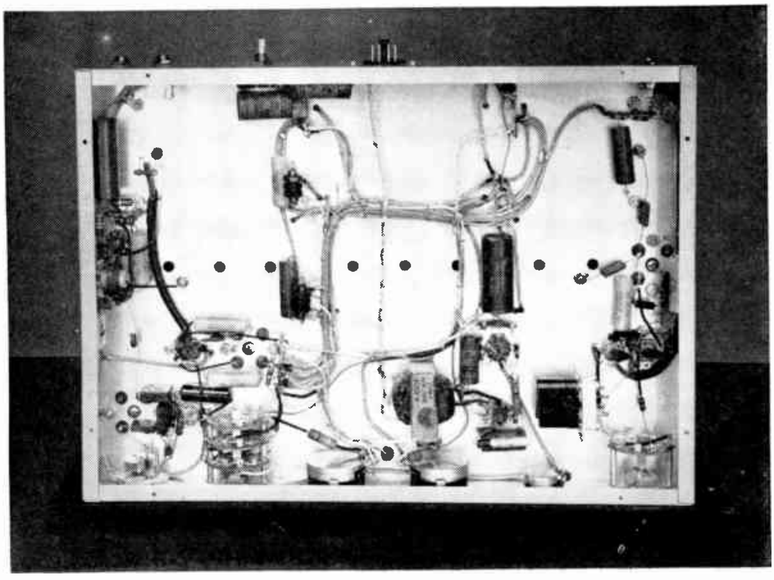


Fig. 5-60 — This view underneath the chassis shows the two oscillator crystals at the lower right. Most of the shielded leads are power leads to the i.f. strips, although some of the low-level audio leads are also run in shielded wire. The eight holes across the center are for access to the tuning slugs of the broad i. f. strip.

High-Frequency Transmitters

The principle 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.** If such a straight amplifier is placed in an intermediate position between two other transmitter stages it is sometimes called a **buffer amplifier.**

Because it becomes increasingly difficult to maintain oscillator frequency stability as the frequency is increased, it is most usual practice in working at the higher frequencies to operate the oscillator at a low frequency and follow it with one or more **frequency multipliers** as required to arrive at the desired output frequency. A frequency multiplier is an amplifier that delivers output at a multiple of the exciting frequency. A **doubler** is a multiplier that gives output at twice the exciting frequency; a **tripler** multiplies the exciting frequency by three, etc. 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 **VFO** (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. The most satisfactory oscillator circuits require the use of a screen-grid tube.

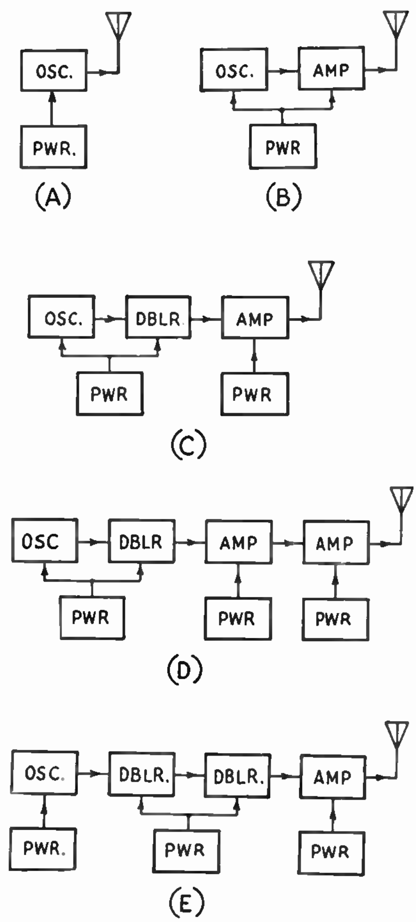


Fig. 6-1 — Block diagrams showing typical combinations of oscillator and amplifiers and power-supply arrangements for transmitters. A wide selection is possible, depending upon the number of bands in which operation is desired and the power output.

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 feed-back required to provide proper excitation. Crystal heating short of the danger point results in frequency drift to an extent depending upon the way the crystal is cut. Excitation should always be adjusted to the minimum necessary for proper operation.

Crystal-Oscillator Circuits

Fig. 6-2 shows three commonly-used crystal-oscillator circuits. All are of the **electron-coupled** type in which the screen of the tube serves as the plate of a triode oscillator. A separate output tank circuit is used in the actual plate circuit. Because of the shielding effect of the screen and suppressor grids, the coupling between the two circuits is comparatively small and exists principally through the common electron stream within the tube. Thus when the load is coupled to the output circuit, its effect will be much less than if it were coupled directly to the frequency-generating circuit.

In the **Tri-tet** circuit of A, the screen is the grounded "plate" of a t.g.t.p. triode oscillator, the crystal taking the place of the coil-and-condenser grid tank. 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 condenser of equal value.

In the **grid-plate** circuit of Fig. 6-2B, the oscillating circuit is the equivalent of a grounded-plate Colpitts. Excitation is adjusted by changing the ratio of the two capacitances, C_6 and C_7. The oscillating circuit of the **modified Pierce** oscillator in C is also basically a Colpitts, this time with a grounded cathode. The grid-cathode and screen-cathode capacitances serve the same purpose as the two condensers connected across the circuit in B. To obtain proper adjustment of excitation, the screen-cathode capacitance is augmented by C_9 which may be adjusted for optimum excitation.

In these circuits, output at multiples of the crystal frequency may be obtained by tuning the plate tank circuit to the desired harmonic, the output obtainable dropping off, of course, at the higher harmonics.

If the behavior of these circuits is to be predicted with any degree of accuracy, the tube used must be one having good screening. From all considerations, the 6AG7 is recommended. With a well-screened tube and proper excitation adjustment, the output plate tuning characteristic

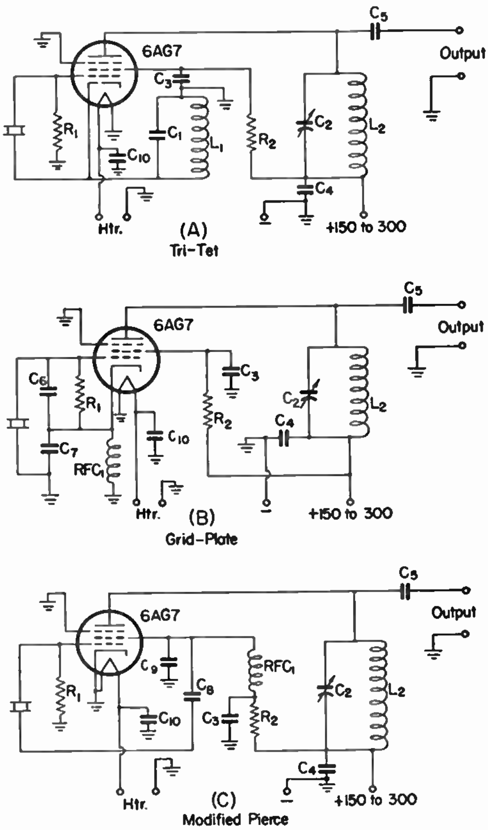


Fig. 6-2 — Commonly-used crystal-controlled oscillator circuits. Values are those recommended for a 6AG7 tube. (See reference in text for other tubes.)

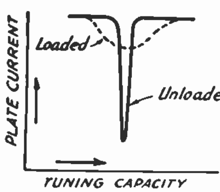
C_1 — Feed-back-control condenser — 3.5-Mc. crystals — approx. 220-μμfd. mica. — 7-Mc. crystals — approx. 150-μμfd. mica.
C_2 — Output tank condenser — 100-μμfd. variable for single-band tank; 250-μμfd. variable for two-band tank (see text).
C_3 — Screen by-pass — 0.001-μfd. disk ceramic.
C_4 — Plate by-pass — 0.001-μfd. disk ceramic.
C_5 — Output coupling condenser — 50 to 100-μμfd. mica.
C_6 — Excitation-control condenser — approx. 10-μμfd. mica.
C_7 — Excitation-control condenser — 220-μμfd. mica.
C_8 — D.c. blocking condenser — 0.001-μfd. mica.
C_9 — Excitation-control condenser — 220-μμfd. mica.
C_{10} — Heater by-pass — 0.001-μfd. disk ceramic.
R_1 — Grid leak — 0.1 megohm, ½ watt.
R_2 — Screen resistor — 47,000 ohms, 1 watt (see text if oscillator is to be keyed).
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. (See text.)
RFC_1 — 2.5-mh. 50-ma. r.f. choke.

at the crystal fundamental, as well as at harmonics, will be similar to that shown in Fig. 6-3 and tuning will cause less than 25 cycles change in frequency. Crystal current, under these conditions, should not be excessive. If the oscillator is to be keyed, best characteristics will be obtained by omitting the screen resistor, R_2, and connecting the screen lead to a regulated source of 75 to 150 volts.

If a tube with poorer screening is used, the effect of tuning the output circuit will not be greatly different at harmonics of the crystal frequency, but the operation at the crystal fundamental may be altered drastically. When the output circuit is tuned near resonance, oscillation may stop entirely, necessitating a critical adjustment to one side of resonance for good keying characteristics and to prevent a marked rise in crystal current. Under these conditions, the frequency may vary as much as 200 cycles.

Crystal current may be estimated by observing the relative brilliance of a 60-ma. dial lamp connected in series with the crystal. For stable operation, crystal current should be limited as much as possible and satisfactory output should be obtained with a current of 40 ma. or less. If the oscillator is to be keyed, the lamp should be removed to prevent chirps.

Fig. 6-3 — Plate tuning characteristic of electron-coupled circuits with a well-screened tube. The plate-current dip at resonance broadens and is less pronounced when the circuit is loaded.



For best harmonic output a tube with high mutual conductance should be used. This is especially important in the circuit of Fig. 6-2C. The 6AG7 also meets this requirement. A low-*C* output tank circuit is desirable, especially for harmonic output. However, if a tank condenser large enough to cover two adjacent bands with the same coil is used, the output at the crystal fundamental and at the harmonic will be approximately the same, since the *L*/*C* ratio will be high when the circuit is tuned to the harmonic, where low *C* is of the greater importance.

For best performance with a 6AG7 tube, the values given under Fig. 6-2 should be followed closely. (For a discussion of values for other tubes, see *QST* for March, 1950, page 28.)

Quartz-Crystal Characteristics

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

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

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

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

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

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

AT-cut — 66.2
BT-cut — 100.78

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

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

Regrinding Crystals

Because crystals near any desired frequency can be purchased reasonably these days, it is not profitable for the amateur to cut and grind his own blanks. However, frequently it may be desirable to make a limited increase in the frequency of a crystal at hand. Indispensable requirements are a piece of plate glass, a good micrometer, supplies of Size 800 aluminum oxide for light grinding, and Size 400 silicon carbide for coarse grinding, and a test oscillator. A test oscillator of the regenerative type, such as the one shown in Fig. 6-2B, is preferred. The oscillator should be equipped with a grid-current milliammeter, preferably one with a 0.5-ma. scale. The grid current should be checked first with the crystal to be reground, and preferably with several others known to have satisfactory activity, to obtain an average of the grid current to be expected for normal crystal activity.

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

The grinding compound should be sprinkled on the glass plate and moistened with water to make a very thin paste. One side of the crystal should be marked at a corner with a pencil and *all* of the grinding should be done on the *opposite* side. The crystal should be swirled around in figure-eight paths. The path should be changed frequently to another part of the glass plate so that the plate will be worn evenly. Light pressure with the finger on a corner of the crystal should be used. Make three or four "8's" to each of the corners in succession and then repeat. Use lighter pressure and make fewer "8's" as the desired frequency is approached.

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

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

Lowering Frequency

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

VARIABLE-FREQUENCY OSCILLATORS

The frequency of a VFO 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 undesirable modulation.

VFO Circuits

Fig. 6-4 shows the most commonly used circuits. They are designed to minimize the effects mentioned above. All are of the **electron-coupled** type discussed in connection with crystal oscillators.

The oscillating circuits in Figs. 6-4A and B are the Hartley type; those in C and D are Colpitts circuits. There is little choice between the circuits of A and C. In both, all of the effects mentioned, 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. 6-4D (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 condensers across the coil. In addition, the tube capacitances are shunted by large condensers, so the effects of the tube — changes in electrode voltages and loading — are still further reduced. In contrast 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 $C_{11} + C_{12}$ to C_{13} or C_{14} (which are usually equal) 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, although there will be some sacrifice in output.

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 VFO 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 one or more fixed-tuned amplifiers or doublers, as shown in Fig. 6-5, to give the desired isolation between the oscillator and the first tunable amplifier stage. The fixed-tuned circuits may consist of nonresonant chokes or, for greater output or frequency multiplying, slug-tuned coils adjusted to the desired frequency. A voltage-regulated supply is recommended.

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

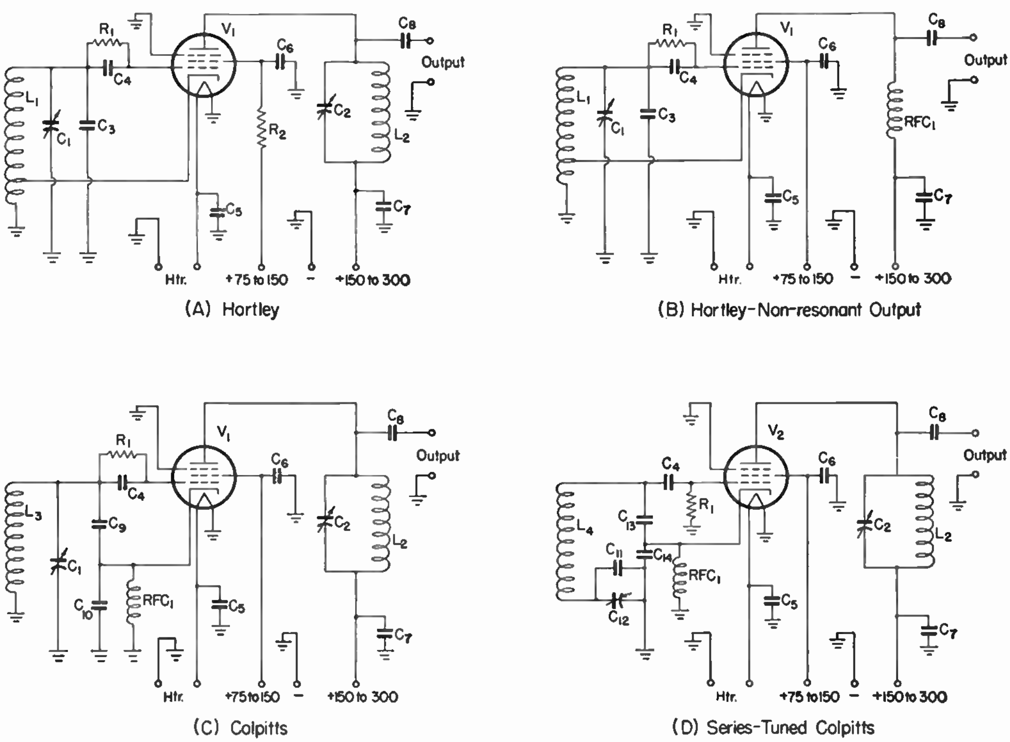


Fig. 6-4 — VFO circuits. Approximate values for 3.5 Mc. are given below. For 1.75 Mc., all tank-circuit values of capacitance and inductance, all tuning capacitances and C_{13} and C_{14} should be doubled; for 7 Mc., they should be cut in half.

C_1 — Oscillator bandspread tuning condenser — 150-μμfd. variable.
C_2 — Output-circuit tank condenser — 100-μμfd. variable.
C_3 — Oscillator tank condenser — 500-μμfd. zero-temp. mica.
C_4 — Grid coupling condenser — 100-μμfd. zero-temp. mica.
C_5 — Heater by-pass — 0.001-μfd. disk ceramic.
C_6 — Screen by-pass — 0.001-μfd. disk ceramic.
C_7 — Plate by-pass — 0.001-μfd. disk ceramic.
C_8 — Output coupling condenser — 50 to 100-μμfd. mica.
C_9 — Oscillator tank condenser — 680-μμfd. zero-temp. mica.
C_{10} — Oscillator tank condenser — 0.0022-μfd. zero-temp. mica.
C_{11} — Oscillator bandspread padder — 50-μμfd. variable air.
C_{12} — Oscillator bandspread tuning condenser — 25-μμfd. variable.
C_{13}, C_{14} — Tube-coupling condenser — 0.001-μfd. zero-temp. 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 preferred; other well-screened types usable.
V_2 — 6AG7 required.

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.

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 VFO. The circuit of Fig. 6-4D 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.

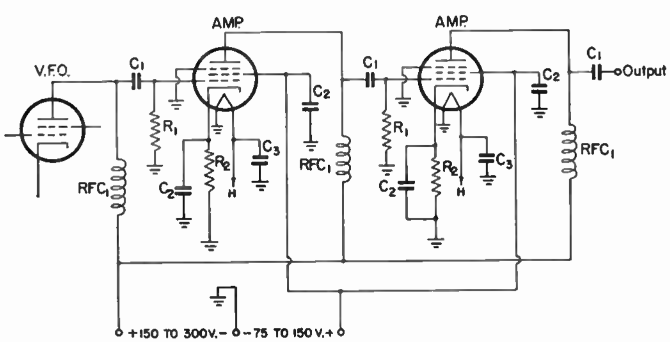


Fig. 6-5 — Diagram showing two isolating amplifier stages following a VFO. Well-screened tubes, such as the 6SK7 or similar types are recommended.

C_1 — Coupling condenser — 100-μμfd. mica.
C_2 — By-pass condenser — 0.001-μfd. disk ceramic.
C_3 — Heater by-pass — 0.001-μfd. disk ceramic.
R_1 — Grid leak — 50,000 ohms.
R_2 — Cathode biasing resistor — 200 to 500 ohms, 1 watt.
RFC_1 — 2.5-mh. 50-ma. r.f. choke or slug-tuned coil.

Variable condensers should have ceramic insulation, good bearing contacts and should preferably be of the double-bearing type, and fixed condensers should have zero temperature coefficient. The tube socket also should have ceramic insulation and special attention should be paid to the selection of a tank 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. 6-4D, the condenser 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. 6-3. 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 VFO Stability

A VFO should be checked thoroughly before it is placed in regular operation on the air. Since succeeding amplifier stages may affect the signal characteristics, final tests should be made with the complete transmitter in operation. Almost any VFO will show signals of good quality and stability when it is running free and not connected to a load. A well-isolated monitor is a necessity. Perhaps the most convenient, as well as one of the most satisfactory, well-shielded monitoring arrangements is a receiver combined with a crystal oscillator, as shown in Fig. 6-6. (See "Crystal Oscillators," this chapter.) The crystal frequency should lie in the band of the lowest frequency to be checked and in the frequency range where its harmonics will fall in the higher-frequency bands. The receiver b.f.o. is turned off and the VFO signal is tuned to beat with the signal from the crystal oscillator instead. In this way any receiver instability caused by overloading of the input circuits, which may result in "pulling" of the h.f. oscillator in the receiver, or by a change in line voltage to the receiver when the transmitter is keyed, will not affect the reliability of the check. Most present-day crystals have a sufficiently-low temperature coefficient to give a satisfactory check on drift as well as on chirp and signal quality if they are not overloaded.

Harmonics of the crystal may be used to beat with the transmitter signal when monitoring at the higher frequencies. Since any chirp at the lower frequencies will be magnified at the higher frequencies, accurate checking can best be done by monitoring at the latter.

The distance between the crystal oscillator and receiver should be adjusted to give a good beat between the crystal oscillator and the transmitter signal. When using harmonics of the crystal oscillator, it may be necessary to attach a piece of wire to the oscillator as an antenna to give sufficient signal in the receiver.

Checks may show that the stability is sufficiently good to permit oscillator keying at the lower frequencies, where break-in operation is of greater value, but that chirp becomes objectionable at the higher frequencies. If further improvement does not seem possible, it would be logical in this case to use oscillator keying at the lower frequencies and amplifier keying at the higher frequencies.

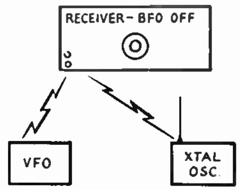


Fig. 6-6—Set-up for checking VFO stability. The receiver should be tuned preferably to a harmonic of the VFO frequency. The crystal oscillator may operate somewhere in the band in which the VFO is operating. The receiver b.f.o. should be turned off.

R. F. Power Amplifiers

R.f. power amplifiers used in amateur transmitters usually are operated under Class C conditions (see chapter on vacuum-tube fundamentals). Fig. 6-7 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. The main objective, of course, is to deliver as much fundamental power as possible (or as desired) 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.

PLATE TANK Q

The Q is determined (see chapter on electrical laws and circuits) by the L/C ratio and the load resistance of the tube (not the resistance of the load circuit). The tube load resistance is related, in approximation, to 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 12 for various ratios is shown in Fig. 6-9. A Q of 12 is a value chosen as an average that will satisfy most of the requirements to be discussed. Certain specific considerations may make a higher or lower value desirable. For a given plate-voltage/plate-current ratio, the Q will vary directly as the tank capacitance, twice the capacitance doubles the Q etc.

Effect of Q on Tube Plate Efficiency

For good tube plate efficiency, the voltage drop across the tank (which determines the instantaneous plate voltage) should approach a sine wave characteristic. Although the plate *current* flowing through the tank is in the highly-distorted form of short pulses containing considerable harmonic energy, a resonant circuit discriminates against harmonic *voltages* across the circuit according to the Q of the circuit. If the Q is sufficiently high, the wave shape of the voltage drop across the tank circuit will be essentially sinusoidal. So far as tube plate efficiency is concerned, requirements will be met satisfactorily if the tank Q is 5 or greater. However, as the Q is increased, the current circulating in the tank circuit becomes greater, increasing the tank-circuit loss. If the Q is greater than about 20, the losses in the tank circuit caused by the increasingly greater tank current will offset any further improvement in plate efficiency.

Harmonic Output Reduction

Strictly speaking, a high-Q tank circuit does not "attenuate" harmonics. The plate *current* pulses remain unchanged with Q. However, it has been explained above that the harmonic *voltage* drop across the tank circuit (a pure sine wave has no harmonic content) decreases with an increase in Q and therefore when the load circuit is coupled across the tank circuit capacitively, as shown in Fig. 6-7B, the harmonic voltage across the load will be reduced as the Q of the tank circuit is increased.

When inductive coupling is used, as in Fig. 6-7A, harmonic reduction in the load comes about for a different reason. At resonance, as explained in the chapter on electrical laws and circuits, there is a build-up of fundamental current in the tank circuit, and this current becomes greater as the Q is increased. As the current through the tank coil increases, the same power in the load will be obtained with looser inductive coupling (a smaller coupling coefficient). Since the harmonic current through the coil remains fixed irrespective of Q, the amount of harmonic energy coupled out becomes less as the coupling is decreased.

As stated above, tank-circuit loss increases with Q, so that the choice of Q must be a compromise depending upon whether efficiency or harmonic reduction is considered the more important.

Q vs. Coupling

Also, as explained above, it is seen that the Q has an influence on coupling to a load when the coupling is inductive. The higher the Q, the larger the tank current and the smaller the coefficient of coupling to the load can be for a given value of current in the load. Conversely, the lower the Q, the greater the coefficient of coupling must be.

Q and Broadbanding

Amateur frequencies are in bands — not spot frequencies — and it becomes desirable to design the circuits of the transmitter so that it may be operated within a band with a minimum of retuning. It is therefore desirable to use the minimum Q that will satisfy the previously discussed requirements.

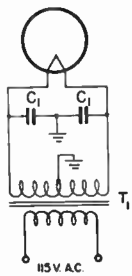


Fig. 6-8 — Filament center-tap connections to be substituted in place of cathode connections shown in diagrams when filament-type tubes are substituted. T_1 is the filament transformer. Filament by-passes C_1, should be 0.001-μfd. disk ceramic condensers.

OUTPUT COUPLING SYSTEMS

Coupling to Flat Coaxial Lines

When the load R in Fig. 6-7A 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 characteristic impedance of the cable. This reduces losses in the cable to a minimum 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. 6-10C, if

1) The plate tank circuit has reasonably high value of Q. A value of 10 or more is usually sufficient.

2) The inductance of the pick-up or link coil is close to the optimum value for the frequency

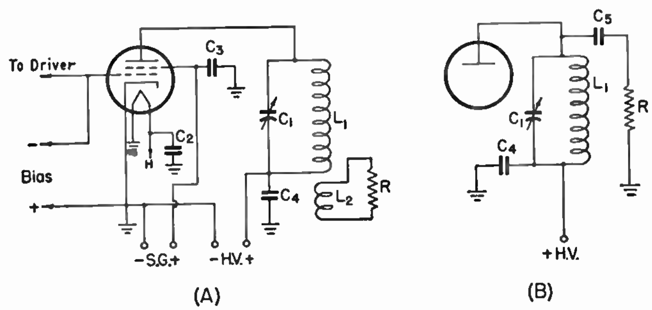


Fig. 6-7 — Output coupling circuits. A — Inductive link coupling. B — Capacitive coupling.

C_1 — Plate tank condenser — see text and Fig. 6-9 for capacitance, Fig. 6-29 for voltage rating.
C_2 — Heater by-pass — 0.001-μfd. disk ceramic.
C_3 — Screen by-pass — voltage rating depends on method of screen supply. See section on screen considerations. Voltage rating same as plate voltage will be safe under any condition.
C_4 — Plate by-pass — 0.001-μfd. disk ceramic or mica. Voltage rating same as C_1, plus safety factor.
C_5 — Coupling condenser — see Fig. 6-18.
L_1 — To resonate at operating frequency with C_1. See LC chart in miscellaneous-data chapter and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator*.
L_2 — Reactance equal to line impedance. See reactance chart in miscellaneous-data chapter and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator*.
R — Representing load.

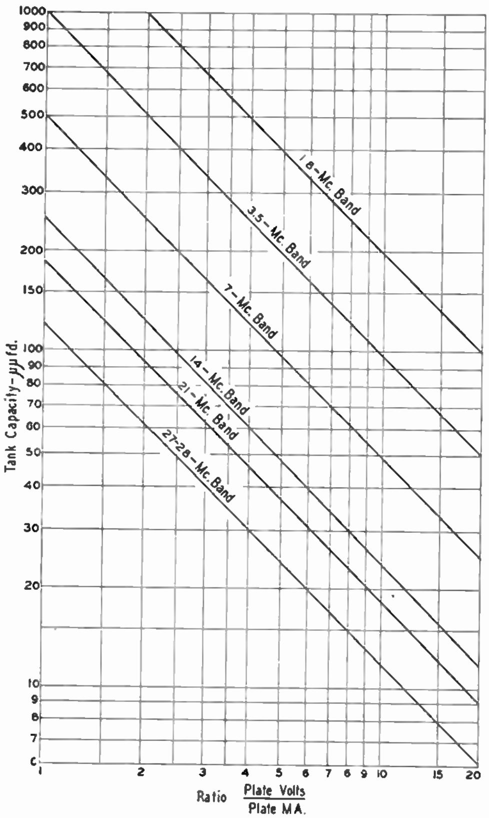


Fig. 6-9 — Chart showing plate tank capacitance required for a Q of 12. To use the chart, divide the tube plate voltage by the plate current in milliamperes. Select the vertical line corresponding to the answer obtained. Follow this vertical line to the diagonal line for the band in question, and thence 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 condenser is used in a balanced circuit, the capacitance of *each section* may be one half of the value given by the chart.

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 characteristic 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

Capacitance in μμfd. 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.

system shown in Fig. 6-10C 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.

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, which has a Z_0 of 75 ohms or less, 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. 6-10 at A and B. The values of inductance and capacitance in the coupling circuits are not highly critical, but the L/C ratio must not be too small. 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.

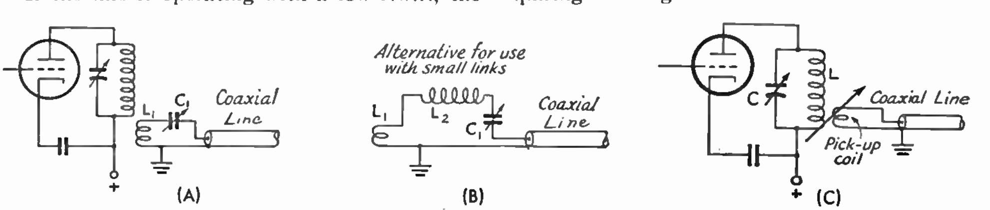


Fig. 6-10 — With flat transmission lines power transfer is obtained with looser coupling if the line input is tuned to resonance. C_1 and L_1 should resonate at the operating frequency. See table for maximum usable value of C_1. If circuit does not resonate with maximum C_1 or less, inductance of L_1 must be increased, or added in series at L_2.

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. 6-10B.

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 condenser, C_1, is tuned through the value of capacitance given by the table. The coupling between the two coils should then be increased until the amplifier loads normally, 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 as the frequency is changed. If the input impedance variations are not large, C_1 may be used as a loading control, no changes in the coupling between L_1 and the tank coil being necessary.

The degree of coupling between L_1 and the amplifier tank coil will depend on the coupling-circuit Q. With a Q of 2, the coupling should be tight — comparable with the coupling that is typical of "fixed-link" manufactured coils. With a swinging link it may be necessary to increase the Q of the coupling circuit in order to get sufficient power transfer. 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 a low-impedance transmission line, as shown in Fig. 6-11. The output condenser, C_2,

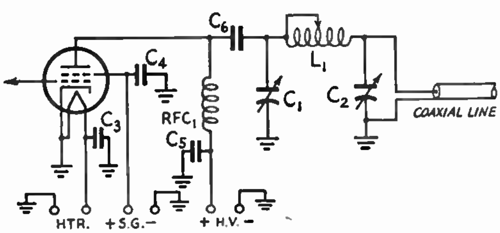


Fig. 6-11 — Pi-section output tank circuit.

C_1 — Input condenser — see text and Fig. 6-9 for capacitance. For voltage rating see C_1, Fig. 6-7.
C_2 — Output condenser — adjustable to half reactance of line impedance — see text and reactance chart in chapter of miscellaneous data. Voltage rating — receiving spacing good for 1 kw. at 50 or 75 ohms if line is terminated in link, otherwise plate voltage plus 25%.
C_3 — Heater by-pass — 0.001-μfd. disk ceramic.
C_4 — Screen by-pass — see Fig. 6-7.
C_5 — Plate by-pass — see Fig. 6-7.
C_6 — Plate blocking condenser — 0.001-μfd. disk ceramic or mica. Voltage rating same as C_1.
L_1 — Inductance approx. same as L_1, Fig. 6-7.

should be adjustable to a reactance of about half of the characteristic impedance of the line. C_1, the input condenser, and L_1 should have values approximately the same as used in a conventional tank circuit for a Q of 12 (see Fig. 6-9).

A decrease in the capacitance of C_2, or the inductance of L_1, will increase the coupling and vice versa. Each time L_1 or C_2 is changed, C_1 must be readjusted for resonance.

R.F. AMPLIFIER-TUBE OPERATION

Driving Power, Efficiency, Dissipation and Power Input

One of the most significant tube ratings is the maximum plate-dissipation rating. This is the power that can be safely dissipated in the tube as heat. It is the difference between r.f. power output and the d.c. power input to the plate. For a given dissipation rating, the theoretical power output from a tube depends on the efficiency with which it can be made to operate. The P_o/P_d curve of Fig. 6-12 shows the theoretical power output obtainable at various efficiencies in terms of the plate-dissipation rating. For instance, at an efficiency of 60 per cent, the curve shows that the output will be 1.5 times the dissipation rating, while at an efficiency of 90 per cent a power of 9 times the dissipation rating might be obtained. However, the P_i/P_d curve shows that the power input at 90 per cent would have to be 10 times the dissipation rating. An input of this magnitude would exceed the power-input rating (plate voltage × plate current) of the tube, which is based on cathode emission and electrode insulation. Also, referring to Fig. 6-13, it is seen that the higher efficiencies are obtainable only by the use of an inordinate amount of driving power. In other words, the *power amplification* decreases rapidly. The typical operating conditions given in the tube tables represent a compromise of these factors. Fig. 6-12 shows the usual practical efficiencies attainable for various classes of tube operation. For instance, at an efficiency of 75 per cent, a Class C amplifier could normally be operated at a power input of 4 times its plate dissipation. A doubler, however, normally operating at about 35 per cent efficiency, could handle an input of only about 1.5 times its dissipation rating. The efficiencies shown for Class B amplifiers are for full excitation and full input.

The figures for driving power listed in the tube tables do not include coupling-circuit losses and to assure adequate excitation, the driver tube should be capable of an output power three or four times the rated driving power of the amplifier. For normal operation, proper excitation is indicated when rated d.c. grid current is obtained at rated bias (see tube tables).

Depending on the material from which the plate is made, the plate will show no color, or varying degrees of redness, when operating at rated dissipation. This can be checked by operating the tube without excitation, but with plate

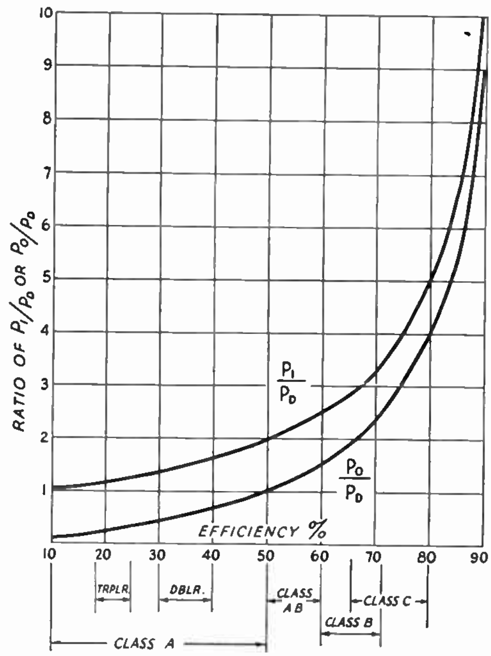


Fig. 6-12 — **Curves showing the relationship of power output (P_o), power input (P_i), plate dissipation (P_d) and efficiency according to class of amplifier tube operation.**

and screen voltages applied, for a period approximating normal operation. Fixed bias should be applied to bring the plate current to some low value at the start. The bias should be gradually reduced until the input to the tube (plate voltage × plate current in decimal parts of an ampere) equals the rated dissipation. The color of the plate at this input should be noted so that it can be compared with the color showing in normal operation. A brighter color in operation would indicate that the dissipation rating is being exceeded. However, most tubes of recent design do not show color at rated dissipation.

Maximum Grid Current

Maximum grid dissipation usually is expressed in terms of the maximum grid current at which the tube should be operated to prevent damage to the tube. A common result of excessive grid heating is a condition where the grid current gradually falls off. If the bias is supplied largely by grid-leak action, the bias drops and the tube draws excessive plate current. The total effect is one in which the temperature of the tube rapidly rises to the danger point. Sometimes, but not always, the tube will restore itself to normal if all power, except filament, is turned off for several minutes. If the overload has been serious or prolonged, with a thoriated-filament tube, it may be possible to reactivate the filament, as described below, but sometimes the tube will be permanently damaged.

Filament Voltage

The filament voltage for the indirectly-heated cathode-type tubes found in low-power classifications may vary 10 per cent above or below rating without seriously reducing the life of the tube. But the voltage of the higher-power filament-type tubes should be held closely between the rated voltage as a minimum and 5 per cent above rating as a maximum. Make sure that the plate power drawn from the power line does not cause a drop in filament voltage below the proper value when plate power is applied.

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

Bias and Tube Protection

The portion of the excitation cycle over which the amplifier draws plate grid current (operating angle) is governed by applying a negative biasing voltage between grid and cathode. Recommended values will be found in the tube tables. Several methods of obtaining bias are shown in Fig. 6-14. 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. 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. This protection can be supplied by obtaining all bias from

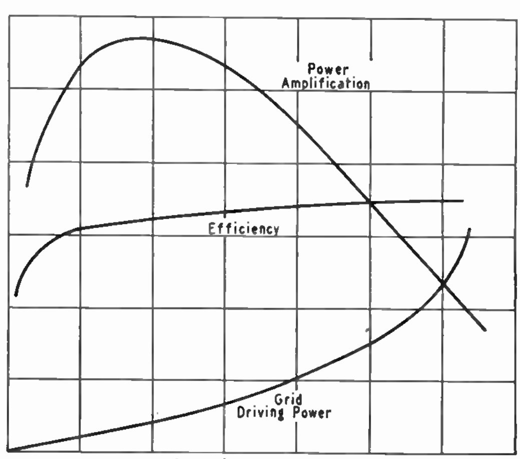


Fig. 6-13 — **Curves showing relationship of driving power, power amplification and plate-circuit efficiency of an r.f. power-amplifier stage.**

a source of fixed voltage, as shown in Fig. 6-14B. 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 indicated in C. The grid-leak resistance in this case is calculated as above, except that the fixed voltage used 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 or above-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.

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 dissipation is not exceeded. In this case plate-modulated 'phone ratings should be used for c.w. operation, however.

In Fig. 6-14F, bias is obtained from the voltage drop across a resistor in the cathode (or filament center-tap) lead. Protective bias is obtained by the voltage drop across R_5 as a result of plate (and screen) current flow. Since plate current must flow to obtain a voltage drop across the resistor, it is obvious that cut-off protective bias cannot be obtained by this system. When excitation is applied, plate (and screen) current increases and the grid current also contributes to the drop across R_5, thereby increasing the bias to the operating value. Since the voltage between plate and cathode is reduced by the amount of the voltage drop across R_5, the over-all supply voltage must be the sum of the plate and operating-bias voltages. 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 by-pass condenser. 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.

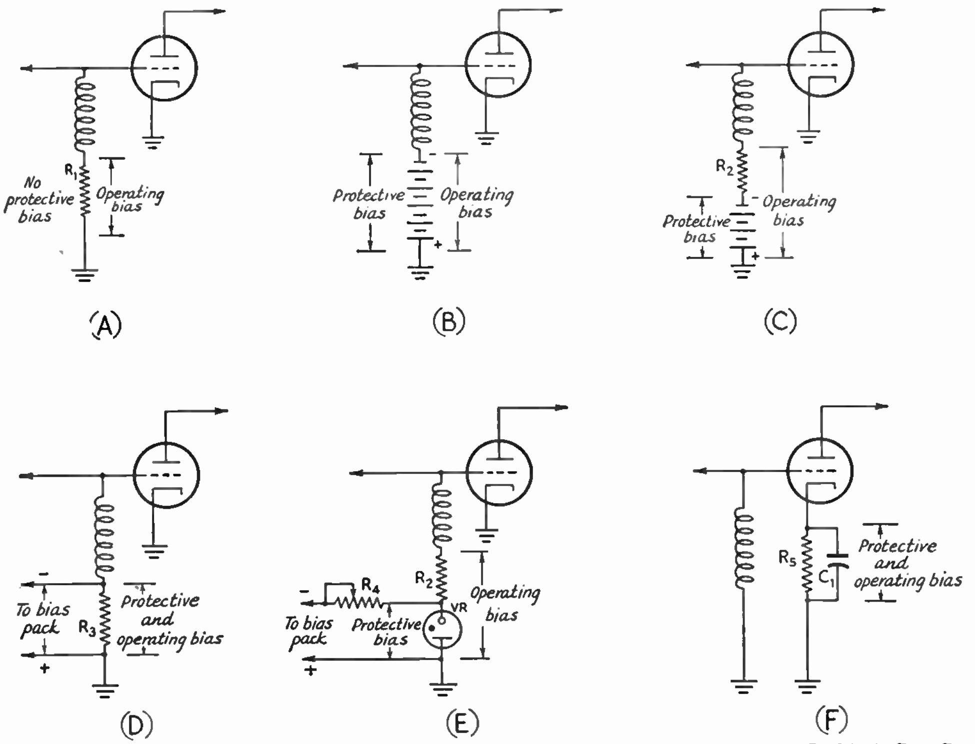


Fig. 6-14 — Various systems for obtaining protective and operating bias for r.f. amplifiers. A — Grid-leak. B — Battery. C — Combination battery and grid leak. D — Grid leak and adjusted-voltage bias pack. E — Combination grid leak and voltage-regulated pack. F — Cathode bias.

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 screen-clamper tube, as shown in Fig. 6-15. The grid-leak bias of the amplifier tube with excitation is applied 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 to the amplifier is removed. One VR tube should be used for each 40 ma. of screen current, other tubes being added in parallel if needed.

Screen Considerations

Since the power taken by the screen does not contribute to the r.f. output, it is dissipated entirely in heating the screen, so the dissipation can be calculated simply by multiplying the screen voltage by the screen current.

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 of resistance 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.

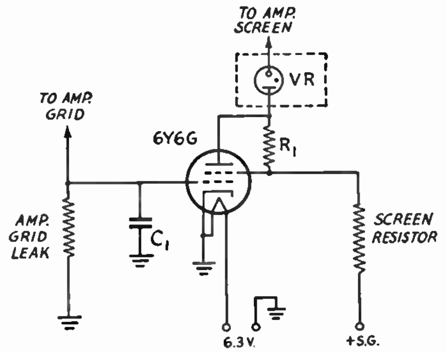


Fig. 6-15 — Screen clamper circuit for protecting screen-grid power tubes. The VR tube is needed only for complete cut-off.

C_1 — 0.001-μfd. disk ceramic. R_1 — 100 ohms.

Some further adjustment may be necessary, as mentioned above, so an adjustable resistor with a total resistance above that calculated should be provided.

FEEDING EXCITATION TO THE GRID

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.

As explained earlier, the grid of a Class C amplifier must be driven positive in respect to cathode over a portion of the excitation cycle, and rectified grid current flows in the grid-cathode circuit. This represents an average resistance across which the exciting voltage must be developed by the driver stage. 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:

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

For normal operation, the values of 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. 6-16. 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 wavelength, 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. 6-17) 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 physical coupling between the two coils, 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.

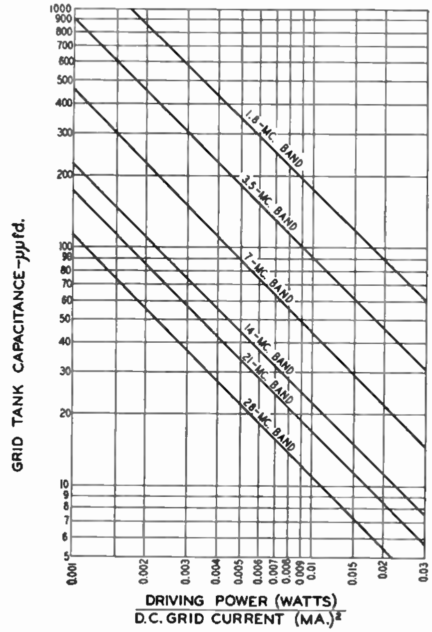


Fig. 6-17 — Chart showing required grid tank capacitance for a Q of 12. To use, divide the driving power in watts by the square of the d.c. grid current in milliamperes and proceed as described under Fig. 6-9. Driving power and grid current may be taken from the tube tables. When a split-stator condenser is used in a balanced grid circuit, the capacitance of *each section* may be half that shown by the chart.

Once the s.w.r. has been brought down to 1 to 1, the frequency should be shifted over the band so that the variation in s.w.r. can be observed, without changing C_1 or the coupling between L_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

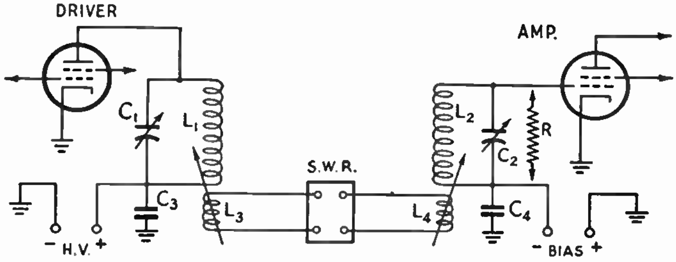


Fig. 6-16 — Coupling excitation to the grid of an r.f. power amplifier by means of a low-impedance coaxial line

C_1, C_3, L_1, L_3 — See corresponding components in Fig. 6-7.

C_2 — Amplifier grid tank condenser — see text and Fig. 6-17 for capacitance, Fig. 6-30 for voltage rating.

C_4 — 0.001-μfd. disk ceramic.

L_2 — To resonate at operating frequency with C_2. See LC chart in miscellaneous-data chapter and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator*.

L_4 — Reactance equal to line impedance — see reactance chart in miscellaneous-data chapter 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 in line only while line is made flat.

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 chapter) 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. 6-16. 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. Unless the constants happen to tune the link near resonance, any appreciable reactance, inductive or capacitive, 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 for this reason 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. The disadvantages of such a resonant link are obvious. 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 break-down 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 depends so much on the dimensions of the link line used that it 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, maintaining as close as possible equal inductances in each coil, 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 condenser of 300 μμfd. 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 condenser should be connected in series with the inner conductor. If the line is long enough to have appreciable reactance, the variable condenser 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 condenser, will affect the resonant frequency and it may take an adjustment of all three before the condenser will show a pronounced effect on the coupling. When the system has been made resonant, coupling may be adjusted by varying the link condenser.

Simple Capacitive Interstage Coupling

The capacitive system of Fig. 6-18A 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 feed-back 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 condenser, C_2, but no impedance transforming is possible. The driver load impedance is the sum of the amplifier grid resistance and the reactance of the coupling condenser in series, the coupling condenser serving simply as a series reactor. Driver load resistance increases with a decrease in the capacitance of the coupling condenser.

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-Section Tank as Interstage Coupler

A pi-section tank circuit, as shown in Fig. 6-18B, 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 condenser, 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 condenser 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 $\mu\mu$fd. 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. 6-18B, parallel driver plate feed and amplifier grid feed are necessary.

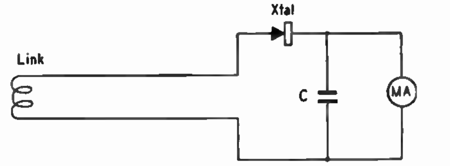


Fig. 6-19 — Circuit of sensitive neutralizing indicator. *Xtal* is a 1N34 crystal detector, *MA* a 0–1 direct-current milliammeter and *C* a 0.001-μfd. mica by-pass condenser.

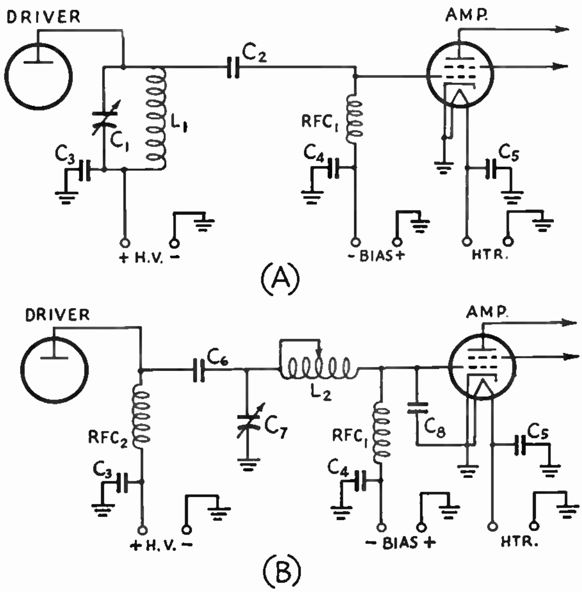


Fig. 6-18 — Capacitive-coupled amplifiers. A — Simple capacitive coupling. B — Pi-section coupling.

C_1 — Driver plate tank condenser — see text and Fig. 6-7 for capacitance, Fig. 6-29 for voltage rating.

C_2 — Coupling condenser — 50 to 150 $\mu\mu$fd. mica, as necessary for desired coupling. Voltage rating sum of driver plate and amplifier biasing voltages, plus safety factor.

C_3 — Driver plate by-pass condenser — 0.001-μfd. disk ceramic or mica. Voltage rating same as plate voltage, plus safety factor.

C_4 — Grid by-pass — 0.001-μfd. disk ceramic.

C_5 — Heater by-pass — 0.001-μfd. disk ceramic.

C_6 — Driver plate blocking condenser — 0.001-μfd. disk ceramic or mica. Voltage rating same as C_2.

C_7 — Pi-section input condenser — see text and Fig. 6-9 for capacitance. Voltage rating same as C_1.

C_8 — Pi-section output condenser — 100-$\mu\mu$fd. mica. Voltage rating same as driver plate voltage plus safety factor.

L_1 — To resonate at operating frequency with C_1. See *LC* chart in miscellaneous-data chapter and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator*.

L_2 — Pi-section inductance — See text. Approximately same as L_1.

RFC_1 — Grid r.f. choke — 2.5-mh. Current rating minimum of grid-current to be expected.

RFC_2 — Driver plate r.f. choke — 2.5 mh. Current rating minimum of plate current expected.

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 condenser, or by-pass 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. 6-19. 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 condenser tuned for any indication of r.f. feed-through. Experiment with shielding and rearrangement of parts will show whether the isolation can be improved.

Neutralizing Circuits

The plate-grid capacitance of screen-grid tubes is reduced to a fraction of a micro-microfarad by the interposed grounded screen. Nevertheless, the power sensitivity of these tubes is so great that only a very small amount of feed-back 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. 6-20A 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, once correct polarization has been determined. A wrong connection will cause the amplifier to oscillate still more strongly. In the case of capacitive coupling, one of the link coils will be coupled to the plate tank coil of the driver stage.

A capacitive neutralizing system for screen-grid tubes is shown in Fig. 6-20B. C_2 is the neutralizing condenser. The capacitance should be chosen so that at some adjustment of C_2, the ratio of C_2 to C_1 equals the ratio of the tube grid-plate capacitance to the grid-cathode capacitance. If C_1 is 0.001 μfd., then

$$C_2 = \frac{1000\ C_{gp}}{C_{gk}}$$

The grid-cathode capacitance must include all strays directly across the tube capacitance, including the capacitance of the tuning-condenser stator to ground. This may amount to 5 to 20 $\mu\mu$fd. In the case of capacitance coupling, as shown in Fig. 6-20C, 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 condensers 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. There should be no plate voltage applied to the amplifier.

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

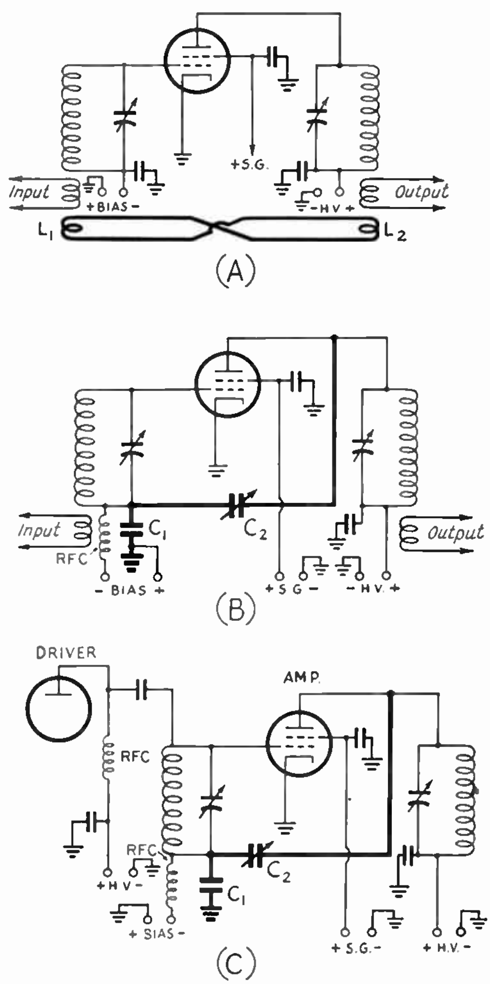


Fig. 6-20 — Screen-grid neutralizing circuits. A — Inductive neutralizing. B-C — Capacitive neutralizing.

C_1 — Grid by-pass condenser — approx. 0.001-μfd. mica. Voltage rating same as biasing voltage in B, same as driver plate voltage in C.

C_2 — Neutralizing condenser — approx. 2 to 10 $\mu\mu$fd. — 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.

The device shown in Fig. 6-19 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 condenser should be readjusted for maximum reading after each change in neutralizing.

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

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

When neutralizing an amplifier of medium or high power, it may not be possible to bring the reading of the rectifier indicator down to zero, but a minimum point in the adjustment of the neutralizing control should be found where higher readings are obtained on either side.

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 will increase v.h.f. harmonics. The best loading system is the use of a pi-section filter, as shown in Fig. 6-18B. 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-μμfd. mica condenser for C_5, wired directly between tube terminals will provide sufficient loading for most screen-grid tubes.

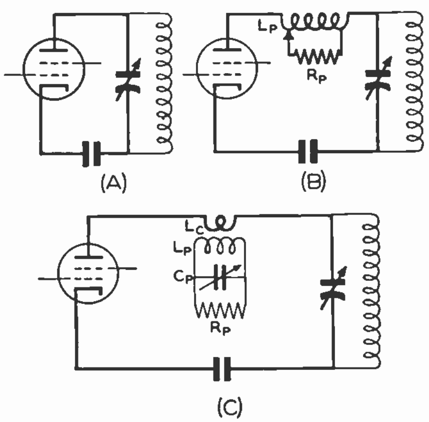


Fig. 6-21 — A — Usual parasitic circuit. B — Resistive loading of parasitic circuit. C — Inductive coupling of loading resistance into parasitic circuit.

V.H.F. Parasitic Oscillation

Unless steps are taken to prevent it, parasitic oscillation in the v.h.f. range will take place in almost every r.f. power amplifier. The heavy lines of Fig. 6-21A 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 above the parasitic frequency, in which the tube will be self-neutralized. Therefore, a v.h.f. parasitic oscillation may be suppressed by adding sufficient inductance, L_p, to tune the circuit into this region. However, to avoid TVI, the self-neutralizing frequency must not be above 100 Mc., preferably 120 Mc. When it is lower, the circuit must be limited to 100 or 120 Mc. and the parasitic suppressed by loading the circuit with resistance, R_p. A coil of 4 or 5 turns, ¼ inch in diameter, is a good starting size. With the tank condenser 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 by-pass condensers 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. 6-21C. 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-μμfd. mica trimmer should serve as the tuning condenser, 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 section 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. 6-17 should be used for the same Q. The plate load resistance is halved so that the plate tank condenser capacitance for a single tube (Fig. 6-9) also should be doubled. The total grid current will be doubled, so to maintain the same grid bias, the grid-leak resist-

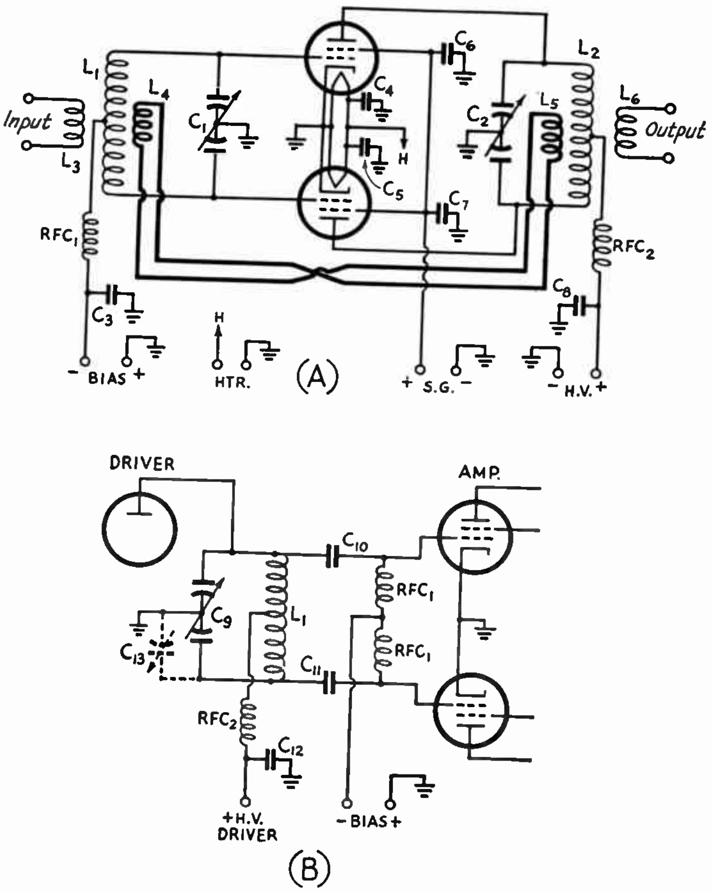


Fig. 6-22 — Push-pull screen-grid amplifier circuits.

A — Inductive-link coupling. B — Capacitive coupling.

C_1 — Split-stator grid tank condenser — see text and Fig. 6-17 for capacitance, Fig. 6-30 for voltage rating.

C_2 — Split-stator plate tank condenser — see text and Fig. 6-9 for capacitance Fig. 6-29 for voltage rating.

C_3 — Grid by-pass condenser — 0.001-μfd. disk ceramic.

C_4, C_5 — Filament by-pass — 0.001-μfd. disk ceramic.

C_6, C_7 — Screen by-pass — 0.001-μfd. disk ceramic or mica. Voltage rating depends on maximum voltage to which screen may soar, depending on how it is supplied. Voltage rating equal to plate voltage will be safe in any case.

C_8 — Plate by-pass — 0.001-μfd. disk ceramic or mica. Voltage rating same as plate voltage for c.w.; twice this value for plate modulation, plus safety factor.

C_9 — Driver plate tank condenser — see section on simple capacitive coupling with single tube. For same Q, each section should have half the capacitance shown in Fig. 6-9. Voltage rating of each section should be twice d.c. plate voltage of driver.

C_{10}, C_{11} — Coupling condenser — 50- to 150-μμfd. mica. Voltage rating twice driver plate voltage.

C_{12} — 0.001-μfd. disk ceramic or mica. Voltage rating same as plate voltage plus safety factor.

C_{13} — See text.

L_1, L_2 — To resonate at operating frequency. See *LC* chart in miscellaneous-data chapter and inductance formula in electrical-laws chapter, or use ARRL *Lightning Calculator*.

L_3, L_4 — Coupling links — reactance equal to feed-line impedance. See reactance chart in miscellaneous-data chapter and inductance formula in electrical-laws chapter.

L_4, L_5 — Neutralizing links — usually a turn or two will be sufficient.

RFC_1 — 2.5-mh. r.f. choke, to carry grid current.

RFC_2 — 2.5-mh. r.f. choke to carry plate current.

ance should be half that used for a single tube. The required driving power is doubled. The capacitance of a neutralizing condenser, 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 individual chokes in each plate and grid lead, rather than one in the common leads. Input and output capacitances are doubled, which may be a factor in efficient operation at higher frequencies.

PUSH-PULL AMPLIFIERS

Circuits for push-pull amplifiers are shown in Fig. 6-22. With this arrangement both grid-input impedance and optimum plate load resistance are doubled. For the same *Q*, *each section* of the split-stator tank condensers should have half the capacitance for a single tube drawing the same total plate current and having the same grid impedance shown by Figs. 6-9 and 6-17. This means that the total tank-circuit capacitance is one-quarter that for a single tube and that the inductances of the tank coils must be quadrupled to resonate at the same frequency. Other values remain the same, except that the total grid, screen and plate currents will be twice the values for a single tube and the stage will require twice the driving power.

In Fig. 6-22A, inductive link coupling is shown. The neutralizing circuit is shown in heavy lines and may not be necessary. Fig. 6-22B shows capacitive coupling to the grids. The driver in this case must be provided with a balanced output circuit. To maintain balanced excitation, it may be necessary to place C_{13}, shown in dashed

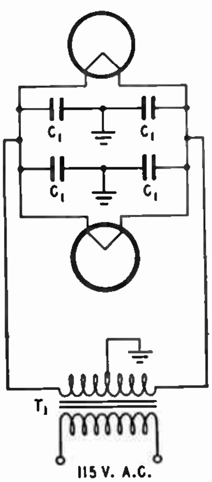


Fig. 6-23 — Connections for tubes in push-pull when filament-types are used. The by-pass condensers, C_1, should be 0.001-μfd. disk ceramic, one placed close to each filament terminal. T_1 is the filament transformer.

lines, across the lower portion of the circuit to balance the driver-tube output capacitance across the upper half. The remainder of circuit B is the same as A. If a neutralizing link is needed, it should be coupled at the center of the driver plate tank coil.

It is advisable to use separate screen and heater by-pass condensers, especially when TVI is a factor. Fig. 6-23 shows equivalent "cathode" connections to be substituted when filament-type tubes are used. Also, individual v.h.f. parasitic chokes will be necessary.

Balance in Push-Pull Amplifiers

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

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

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 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 the introduction of neutralization. The link neutralizing system of Fig. 6-20A is convenient in such a contingency.

Push-Pull Multiplier

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

Push-Push Multipliers

A two-tube circuit which works well at even harmonics, but not at the fundamental or odd harmonics, is shown in Fig. 6-24. It is known as the **push-push** circuit. The grids are connected in push-pull while the plates are connected in parallel. The efficiency of a doubler using this circuit may approach that of a straight amplifier under similar operating conditions, because there is a plate-current pulse for each cycle of the output frequency.

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

The grid tank circuit is tuned to the frequency of the driving stage and should have the same constants as the grid tank circuit of a push-pull amplifier (see Fig. 6-22). The plate tank circuit is tuned to an even multiple of the exciting frequency, usually the second harmonic, and should have the same values as a straight amplifier for the harmonic frequency (see Fig. 6-9), bearing in mind that the total plate current of both tubes determines the *C* to be used.

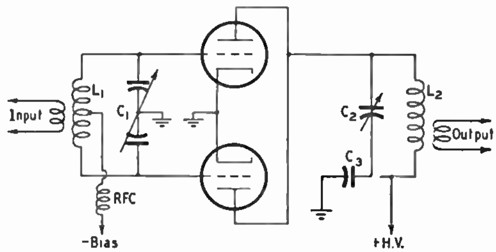


Fig. 6-24 — Circuit of a push-push frequency multiplier for even harmonics.

C_1L_1 and C_2L_2 — See text.

C_3 — Plate by-pass — 0.001-µfd. disk ceramic or mica. Voltage rating equal to plate voltage plus safety factor.

RFC — 2.5-mh. r.f. choke.

● TRIODE AMPLIFIERS

Circuits for triode amplifiers are shown in Fig. 6-25. Neglecting references to the screen, all of the foregoing information applies equally well to triodes. All triode straight amplifiers must be neutralized, as Fig. 6-25 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 feed-back 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 condensers 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. 6-25B, the amplifier grid 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. 6-25C 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.

● TUNING A TRANSMITTER

Fig. 6-26 shows where milliammeters and voltmeters may be connected to obtain desired readings. Metering of all stages is usually not necessary except for initial adjustments. After preceding stages have been adjusted for proper operating conditions, a transmitter can often be tuned up using only grid- and plate-current milliammeters in the final-amplifier circuit.

While cathode metering often is used for reasons of safety to the operator and meter insulation, it is frequently difficult to interpret readings that are the resultant of three currents, one of which may be falling while the other two are increasing. Fig. 6-27 shows a commonly-used system for switching a single meter to read current in any of several different circuits. The resistors, *R*, are connected in the various circuits in place of the milliammeters shown in Fig. 6-26. Since the resistance of *R* is several times the internal resistance of the milliammeter, it will have no practical effect upon the reading of the meter itself.

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

The first step in adjusting each stage is to check for parasitic oscillation as discussed earlier. The second step is to adjust neutralizing if neutralization is required.

While it is usually possible to make all initial tuning adjustments of low-power stages with plate voltage applied, it is preferable to disconnect the plate voltage until adjustments of excitation have been made. Starting with the oscillator, its output tank circuit should be resonated as indicated by a dip in the plate-current reading (see Fig. 6-3), or by a maximum reading of grid current to the following stage if it is coupled capacitively. Both readings should occur simultaneously. At this point, the frequency of the oscillator output should be checked with an absorption wavemeter to make sure that it is tuned to the desired band. If transmission-line coupling is used, the coupling to the grid of the amplifier should first be adjusted for minimum standing-wave ratio as described earlier. After this adjustment, the coupling at the oscillator end of the line only should be altered. If the amplifier grid current is much above rated value, the coupling to the oscillator should be reduced. Conversely, if the amplifier grid current is low, coupling should be increased. As the coupling is increased, the oscillator should draw more plate current and the dip at resonance should become less pronounced, as indicated in Fig. 6-3. If it is possible to increase the coupling to the point where the oscillator plate current is up to the rated value and yet the required grid current is not up to rated value, the biasing voltage should

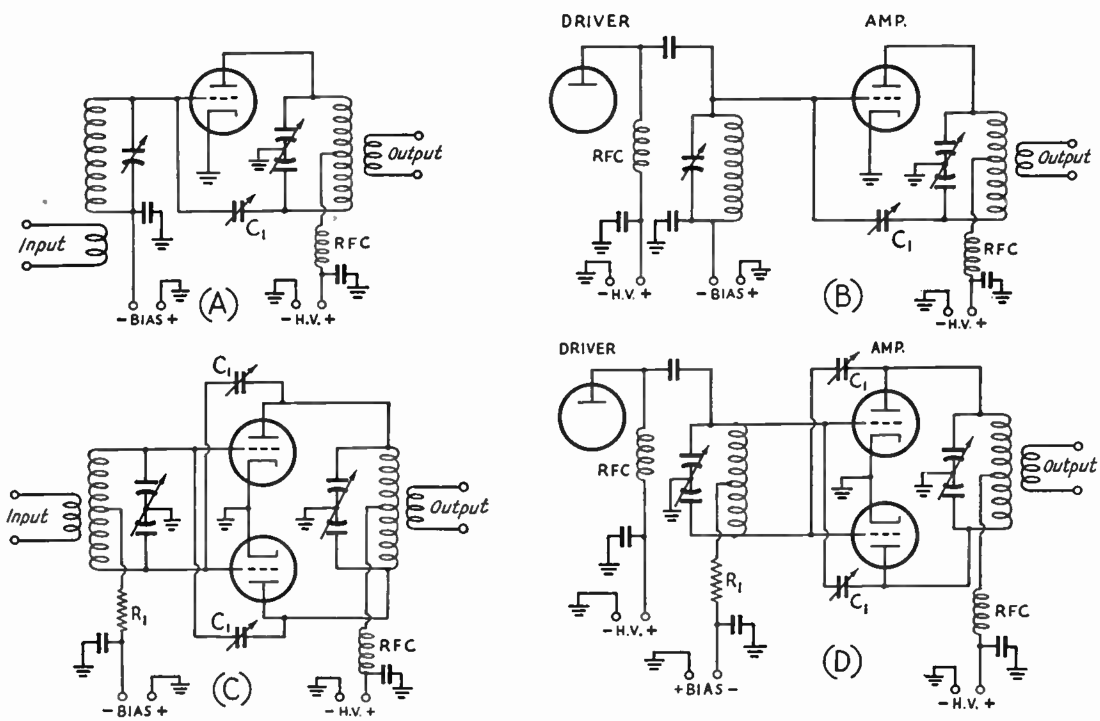


Fig. 6-25 — Triode amplifier circuits. A — Link coupling, single tube. B — Capacitive coupling, single tube. C — Link coupling, push-pull. D — Capacitive coupling, push-pull. Aside from the neutralizing circuits, which are mandatory with triodes, the circuits are the same as for screen-grid tubes, and should have the same values throughout. The neutralizing condenser, 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.

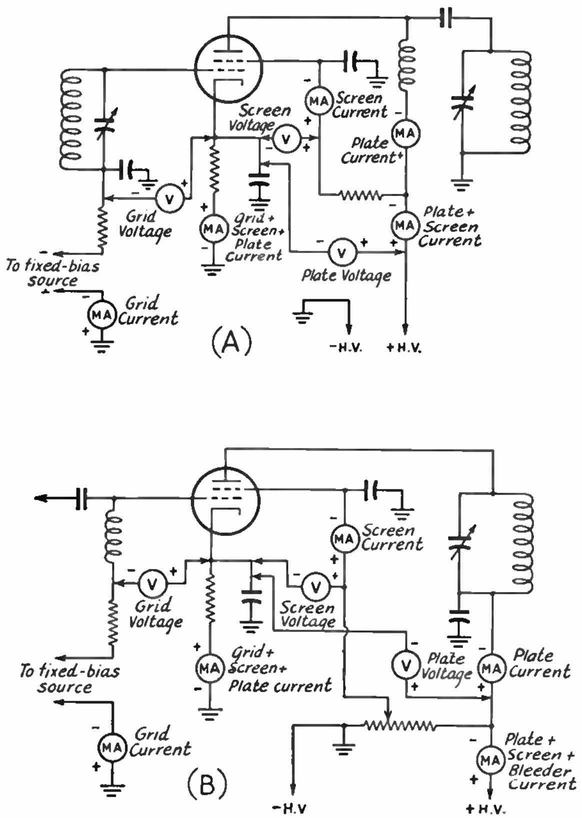


Fig. 6-26 — **Diagrams showing placement of voltmeter and milliammeter to obtain desired measurements. A — Series grid feed, parallel plate feed and series screen voltage-dropping resistor. B — Parallel grid feed, series plate feed and screen voltage divider.**

be measured with a high-resistance (20,000 ohms per volt) voltmeter. If the stage has a simple biasing resistor from grid to ground, connect a 2.5-mh. r. f. choke in series with the voltmeter prod going to the grid. The bias should be measured with the stage operating under excitation. If the biasing voltage measures too high, any fixed bias should be reduced and then, if necessary, the grid-leak resistance. If the driver is operating up to rated plate current and rated grid current cannot be obtained with the required bias, the indication is that the screen and/or plate voltage of the oscillator must be raised if this can be done with safety to the oscillator tube. However, it should be borne in mind that even if an intermediate stage is underdriven, it still may furnish the required driving power for the following stage. Therefore, it is, of course, advisable to check this before making any drastic changes in the oscillator.

The same process is followed in tuning up following amplifier stages, step by step. If there is any difficulty in obtaining the desired excitation to any particular stage, be sure that the screen voltage of the driver stage is up to normal as discussed earlier in the section on screen-grid considerations. If the excitation is adjusted first without plate and screen voltages it may be found that the grid current will change when these voltages are applied and the stage is loaded. It is normal for grid current to drop somewhat when these voltages are applied and still further when the load is coupled, especially with triodes. When this occurs, excitation should be increased, to bring the grid current back to rated value.

If it is found that grid current increases when the plate tank circuit is tuned slightly to the high-frequency side of resonance, this indicates regeneration. This may be of little consequence in exciter stages so long as oscillation does not result under any normal tuning condition. But in the final amplifier, especially if it is to be modulated, it is a condition to be avoided by better shielding or more accurate neutralization.

The main objective in the end, of course, is to obtain adequate excitation to the final amplifier and, in general, any adjustment of earlier stages that will produce this result without overloading anywhere along the line will be satisfactory. In conservative design, the full power capability of the exciter stages may not be needed. In the interests of v.h.f. harmonic reduction, it is desirable to provide an excitation control so that the excitation to the final amplifier can be limited to that necessary for satisfactory operation. This can be in the form of a potentiometer control of the screen voltage of the first

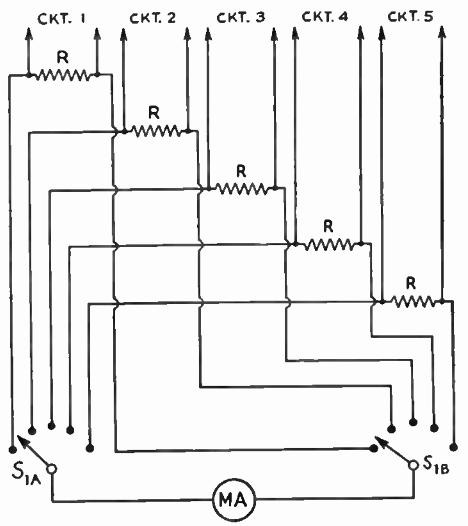
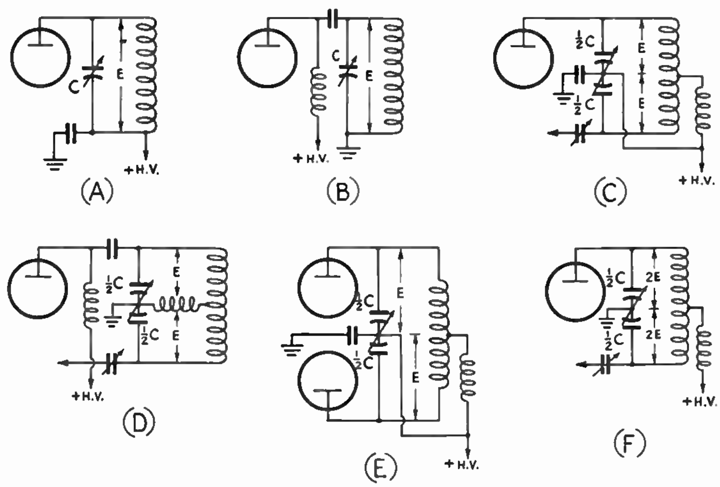


Fig. 6-27 — **Method of switching a single milliammeter. The resistors, *R*, should be 10 to 20 times the internal resistance of the meter; 47 ohms will usually be satisfactory. S_1 is a 2-section rotary switch. Its insulation should be ceramic for high voltages, and an insulating coupling should always be used between shaft and control knob.**

stage after the oscillator. Then reduction in screen voltage of this stage will reduce excitation all along the line, which is desirable.

MEASURING POWER OUTPUT

The power output of any transmitter stage can be checked with reasonable accuracy by simply coupling an ordinary lamp to the output tank circuit and comparing its brilliance with that of another lamp of the same size operating from a.c. Since it is difficult to judge power accurately when the lamp is over or under normal brilliance, the lamp selected should have a wattage rating as close as possible to that expected from the amplifier. Flashlight bulbs can be used for low power. At frequencies above 7 Mc. sufficient coupling usually is obtained by connecting the lamp in series with a few turns of wire that can be slipped over or inside the tank coil, as shown in Fig. 6-28A. But at 3.5 and 7 Mc., it is usually necessary to tap the bulb directly across a portion of the tank coil, as shown at B. WARNING! *Turn off the high voltage when tapping a series-fed tank circuit.* The coupling should be adjusted until the plate current at resonance is the rated loaded value for the tube. A more accurate dummy load is described in *QST* for March, 1951, page 32.



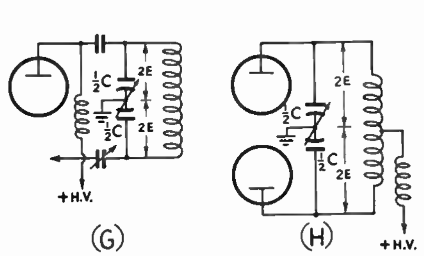


Fig. 6-29 — Diagrams showing the peak voltage for which the plate tank condenser should be rated for c.w. operation with various circuit arrangements. *E* is equal to the d.c. plate voltage. The values should be doubled for plate modulation. The circuit is assumed to be fully loaded. Circuits A, C and E require that the tank condenser be insulated from chassis or ground, and from the control.

COMPONENT RATINGS AND INSTALLATION

Plate Tank-Condenser Voltage

In selecting a tank condenser with a spacing between plates sufficient to prevent voltage breakdown, the peak r.f. voltage across a tank circuit under load, but without modulation, may be taken conservatively as equal to the d.c. plate voltage. If the d.c. plate voltage also appears across the tank condenser, this must be added to the peak r.f. voltage, making the total peak voltage twice the d.c. plate voltage. If the amplifier is to be plate-modulated, this last value must be doubled to make it four times the d.c. plate voltage, because both d.c. and r.f. voltages double with 100-per-cent plate modulation. At the higher plate voltages, it is desirable to choose a tank circuit in which the d.c. and modulation voltages do not appear across the tank condenser, to permit the use of a smaller condenser with less plate spacing. Fig. 6-29 shows the peak voltage, in terms of d.c. plate voltage, to be expected across the tank condenser in various circuit arrangements. These peak-voltage values are given assuming that the amplifier is loaded to rated plate current. Without load, the peak r.f. voltage will run much higher. Since a c.w. transmitter may be operated without load while adjustments are being made, although a modulated amplifier never should be operated without load, it is sometimes considered logical to select a condenser for a c.w. transmitter with a peak-voltage rating equal to that required for a 'phone transmitter of the same power. However, if minimum cost and space are considerations, a condenser with half the spacing required for 'phone operation can be used in a c.w. transmitter for the same carrier

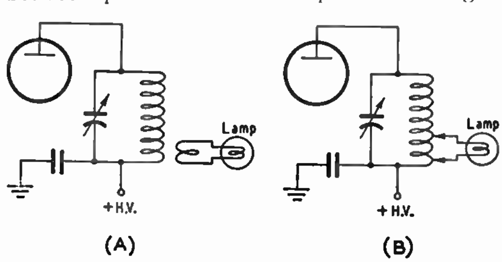


Fig. 6-28 — Using a lamp bulb for an approximate check on the output of an oscillator or amplifier. The coupling should be adjusted to make the stage draw rated plate current when tuned to resonance. Special caution should be used in tapping the lamp directly on the coil when series plate feed is used. *Always turn off the power before making a change in the tap.*

output, as indicated under Fig. 6-29, if power is reduced temporarily while tuning up without load.

In the circuits of Fig. 6-29C, D and E the rotors are deliberately connected to the positive side of the high-voltage supply, eliminating any difference in d.c. potential between the rotors and stators.

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

Plate tank condensers 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 condenser 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 condenser 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 condenser 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 Condensers

In the circuit of Fig. 6-30, the grid tank condenser 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 condenser should be this same value.

The grid tank condenser 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 terminal. The rotor ground lead or by-pass lead should be run directly to the nearest point on the chassis or other shielding. In the circuit of Fig. 6-30A, the same insulating precautions mentioned in connection with the plate tank condenser should be used.

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 6-17 will be greater than that for which the coil is designed and turns must be removed if a Q of 12 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 condenser. Leads up to 6 or 8 inches are permissible. It is more important to keep the tank condenser 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 condenser shaft, either alongside the condenser or above it.

Plate-Blocking and By-Pass Condensers

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

Wherever their voltage rating will permit (500 volts), 0.001-μfd. disk ceramic condensers should be used as by-passes, since, when applied correctly (see TVI chapter), they are series resonant in the TV range and therefore are an important measure in filtering power-supply leads. For higher voltages, use 0.001-μfd. mica by-passes.

R.F. Chokes

The r.f. choke in parallel plate feed must have high impedance at the operating frequency to avoid loss. In multiband transmitters, if it is found that the choke heats excessively on one or more bands, the only solution is to use a different choke for these bands.

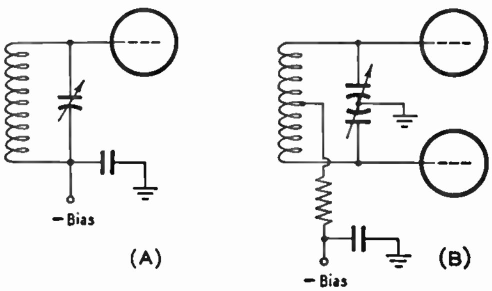


Fig. 6-30 — The voltage rating of the grid tank condenser in A should be equal to the biasing voltage plus about 20 per cent of the plate voltage. This same rating should be applied to each section of the split-stator condenser in B.

A One-Tube Two-Band Transmitter for the Novice

Figs. 6-31, 6-32, and 6-33 show the details of a low-power crystal-oscillator transmitter covering the 3.5- and 7-Mc. bands. It is complete with power supply, and an output circuit that will feed directly into a simple antenna without the need for an antenna tuner. The circuit diagram appears in Fig. 6-32. A 6AG7 pentode is used in an oscillator of the grid-plate type. The output circuit, consisting of C_{10}, C_{11} and L_1, is in the form of a pi-section network that will couple into a wire of random length. The circuit is keyed in the cathode circuit.

J_1 is an octal tube socket that is used as a combination crystal socket and key jack. R_1 is the grid leak. C_1 and C_2 are excitation-control condensers. RFC_1 is necessary to prevent short-circuiting C_2 for r.f. when the key is closed. R_2 is the screen voltage-dropping resistor that reduces the voltage to the screen. RFC_2 is the plate feed choke. Plate current is measured by the milliammeter, MA_1. C_7 is the plate blocking condenser, and C_3, C_5 and C_6 are by-pass condensers.

The power supply is a simple one delivering about 350 volts. The smoothing filter, consisting of C_8, C_9 and L_2, is of the condenser-input type. R_3 is the bleeder resistor. S_1 turns the power supply on and off.

Construction

The parts are assembled on a 7 × 12 × 3-inch aluminum chassis. In the placement of parts in the transmitter, the power-supply section is kept in a line at the back of the chassis. The r.f. components are mounted toward the front of the chassis. As can be seen in the photographs, there are three octal sockets — one for the 5Y3 rectifier, one for the 6AG7 oscillator, and the third which is used as a crystal socket and key jack.

With the exception of the three sockets and the meter, all the mounting holes can be made with an ordinary hand drill. For the socket holes, one can purchase, or borrow, a socket punch. The meter hole can be started with the socket punch and then enlarged with a half-round or rattail file. The variable condensers are mounted directly against the under side of the chassis. In placing them, be sure that their shafts extend far enough out from the front of the chassis to accommodate the tuning knobs. These condensers are of the broadcast-receiver replacement type, and can be purchased locally, or from one of the large mail-order houses. They are usually listed as single-gang midget t.r.f. condensers and have a maximum capacitance of more than 300 μμf.

The power transformer is mounted in such a manner that the high-voltage leads and the 5-volt rectifier leads are brought out at a point closest to the 5Y3 rectifier socket. A three-terminal tie point is mounted close to the transformer 115-volt leads to furnish terminals for the power switch and transformer leads. After the sockets, a.c. switch, meter, and feed-through bushings for holding L_1 are all mounted in place, the wiring can be started.

Wiring

Connect the two 115-volt transformer primary leads (black), each to one of the tie points. Then also connect one of the power-cord wires to one of these tie points, and one terminal of the power switch, S_1, to the other. Connect the remaining side of S_1, and the remaining power-cord wire to the third tie point. Fasten one of the 6.3-volt transformer leads (green) to a soldering lug under the tie-point mounting screw. The remaining 6.3-volt transformer wire (green) is connected to Pin 7 on the 6AG7 socket.

For the high-voltage wiring, the center-tap

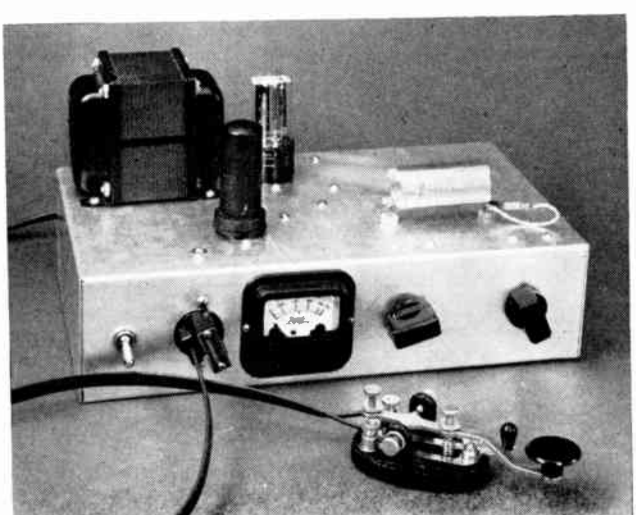


Fig. 6-31 — Top view of the Novice 2-band transmitter. L_1 at the top right-hand side is shown in the 80-meter position. The shorting clip is clipped to the feed-through bushing. The lead to the key is a short piece of 300-ohm Twin-lead which is terminated in a Millen 300-ohm plug. This type of plug is the correct size for octal socket Pins 2 and 4.

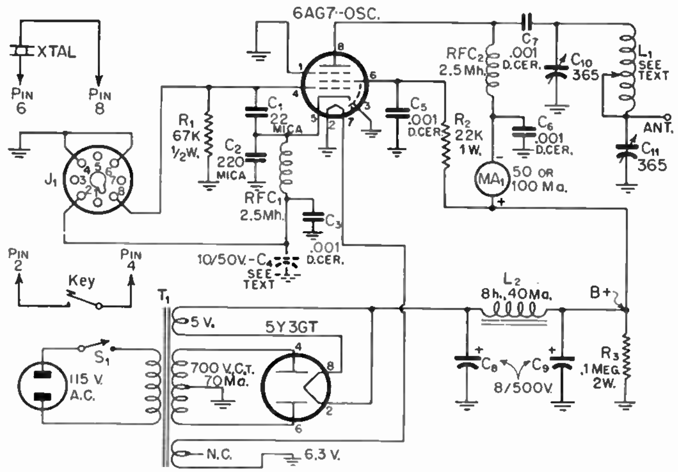


Fig. 6-32 — Circuit diagram of the beginner's transmitter.

wire of the high-voltage secondary (red and yellow) is connected to ground, one of the high-voltage leads (red) is connected to Pin 4 of the 5Y3 socket, while the other red lead goes to Pin 6. One of the 5-volt rectifier-filament leads (yellow) is connected to Pin 8 of the 5Y3 socket, and the other yellow lead is run to Pin 2. Also connected to Pin 2 of the 5Y3 socket is a lead from the choke, L_2, and the lead marked + from C_8. The other side of C_8, or the *negative* side, is grounded. The remaining lead of L_2, the *plus* side of C_9, and a lead from R_3, are all run to a terminal on a tie point. The *negative* side of C_9 and the other lead from R_3 are grounded. This completes the power-supply wiring.

Pins 1, 2, and 3 of the 6AG7 socket are connected together with a bare wire and the wire run to ground. Also, one side of C_2 must be grounded, so it can be connected to one of these pins. The other side of C_2 is run to Pin 5. A lead to RFC_1 is also connected to Pin 5. One side of C_1, one side of R_1, and a lead to Pin 8 of J_1 are all soldered to Pin 4 of the 6AG7 socket. The other side of R_1 is grounded, while the remaining side of C_1 goes to Pin 5. Pins 4 and 6 of the crystal socket are also grounded. The remaining side of RFC_1 is connected to Pin 2 of J_1. Also connected to Pin 2 is one side of C_3. The other side of C_3 is grounded.

The screen resistor, R_2, is connected between the B+ (+ terminal of C_9) terminal and Pin 6 of the 6AG7 socket. Also connected to Pin 6 is one side of C_5. The other side of C_5 is grounded. A lead is connected between the B+ terminal and the + side of the meter. The other terminal of the meter is connected to one side of RFC_2. Also connected to this point on RFC_2 is one side of C_6, the other side of C_6 being grounded. The remaining side of RFC_2 is connected to Pin 8 of the 6AG7 socket and C_7 is connected between this side of RFC_2 and the stator section of C_{10} is also connected to the nearest of the two feed-through bushings holding L_1. The stator of C_{11} is connected to the other feed-through bushing, and a lead is run from this bushing to the transmitter output terminal mounted on the back side of the chassis. This should complete all wiring below the chassis.

Coil

As shown in the parts list, L_1 is a Barker & Williamson stock No. 3016 coil with 13 turns removed from each end. For 40-meter operation, it is necessary to short out a large part of the coil. This is accomplished by use of a short clip lead. One end of the lead is connected along with one end of L_1 to the output bushing (the one connected to C_{11}). The other end of L_1 is soldered to the input bushing. To operate on 40 meters, it is necessary to attach the clip to the 30th turn of L_1, from the input side. In order not to short out the 29th and 31st turns, they can be bent in toward the axis of the coil.

Testing

An 80-meter crystal between 3700 and 3750 kc. will be needed for 80-meter operation. For 40-meter work, one between 3588 and 3598 kc. will be required. (The crystal frequency is doubled for 7-Mc. operation.)

In tuning up on 80 meters, insert the crystal in Pins 6 and 8 of the octal socket. The key leads are inserted in Pins 2 and 4. A 115-volt 10- or 15-watt light bulb will serve as an artificial load for testing purposes. Connect the bulb to the output of the rig by soldering a piece of wire to the center terminal in the base of the bulb, and one to the screw shell portion. One of the wires is then connected to the output terminal of the transmitter and the other to the chassis. The 115-volt a.c. switch is turned on and the tubes allowed a minute or so to warm up. After the rig has been on for a minute, close the key. Tune the station receiver to the crystal frequency and the transmitter's signal should be heard. The input condenser, C_{10}, is slowly tuned through its range. Two things should happen — the dummy load lamp should light and the meter should show a dip, or lower reading, at the point where the bulb lights. Also, the signal should be louder at this point. Now tune the output condenser, C_{11}, across its range and the bulb should brighten at one point, and the signal get louder in the receiver. Also, the meter should show a greater reading than before. Switching back and forth between the two condensers, always tune for maximum brilliance in the bulb.

Antenna

An antenna may now be substituted for the lamp. The type of output circuit used in the rig will load with almost any length of wire. However, it will load with a 30-foot length of wire on both 80 and 40 meters a great deal easier than with some lengths. One end of the wire should be connected to the output terminal and the other end suspended on an insulator attached to a cord or rope slung from the highest available support. (See the antenna chapter for methods of bringing the wire in to the transmitter.)

Output Indicator

The transmitter *can* be tuned up by the meter, but sometimes a beginner may become confused trying to interpret the readings he gets. A simple device to show that the antenna is taking power consists of two pieces of wire, about two feet long, and a 2-volt 0.06-ampere flashlight bulb, either No. 48 or 49. The bulb is connected between the two pieces of wire, one lead to the tip of the bulb base and the other lead to the shell of the base, making a four-foot length of wire with the bulb in the center. One end of this wire is connected to the output terminal, while the other end is clipped on the antenna, three or four feet up. Scrape the wire at this point if it is insulated. When the transmitter is turned on and the condensers are tuned, a point will be reached in the tuning where the bulb will glow, or light up. Tune the condensers for maximum brilliance in the bulb; this is an indication that maximum power is going into the antenna.

Forty-meter tune-up procedure is the same as for 80 with the exception of using the correct crystal, and shorting out the section of L_1. Remember to listen on the receiver when tuning up the transmitter on 40 or 80. When tuning up on 40, the signal should be definitely louder on 40 than on 80 meters, and vice versa for 80-meter tune-up.

When the oscillator is fully loaded and tuned to resonance, the plate current should run between 20 and 30 ma., representing a power input of 7 to 10 watts.

(This unit originally described in the November, 1953, issue of *QST*.)

Shopping List for Novice Transmitter

22-μμf. mica condenser.
220-μμf. mica condenser.
4 0.001-μf. disk ceramic condensers.
2 8-μf. 500-volt midget electrolytic condensers.
67,000-ohm resistor, ½ watt.
22,000-ohm resistor, 1 watt.
0.1-megohm resistor, 2 watts.
2 2½-mh. r.f. chokes (National R100S or Millen 34102).
2 variable condensers (midget type t.r.f. one-gang broadcast receiver replacement).
70 turns of No. 24 wire, 1-inch diam., 2¼ inches long (B & W 3016 with 13 turns removed from each end).
8-hy. 40-ma. filter choke (Thordarson T20C52).
Power transformer: 350–0–350 volts r.m.s., 70 ma.; 5 v., 2 amp.; 6.3 v., 2½ amp. (Thordarson TS-24R02).
3 octal sockets.
Single-pole single-throw toggle switch.
2 feed-through insulators (National TPB).
Tip jack (Amphenol type 7818).
2 three-point terminal strips.
0–50 or 0–100 d.c. milliammeter (Shurite).
Aluminum chassis 3 by 7 by 12 inches.
6 feet of hook-up wire.
6AG7 tube.
5Y3 tube.
6 solder lugs.
18 6-32 × ½-inch nuts, bolts, and washers.
Two tuning knobs to fit ¼-inch shaft
Crystal.

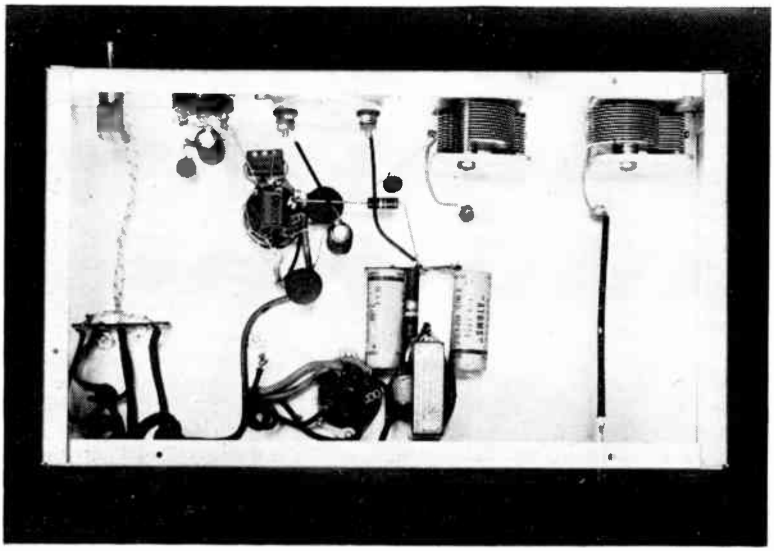


Fig. 6-33 — Bottom view of the Novice one-tube transmitter showing the wiring of parts. The power supply components are mounted along the back side while the r.f. section runs along the front. The output lead from the feed-through bushing is clearly visible on the right-hand side. The only openings at the back are the output terminal and the 115-volt a.c. leads.

A Sweep-Tube Transmitter for 3.5 and 7 Mc.

Figs. 6-34 through 6-37 show a low-power transmitter using a single TV-receiver sweep-tube triode. It will deliver an output of about 10 watts on 80 or 40 meters. Power supply and antenna tuner are included.

As shown by Fig. 6-36, the oscillator utilizes one section of a 6BL7. J_1 is the keying jack, and also serves as the oscillator metering jack. The plate tank, C_2L_1, covers the frequency range of 3.75 to 9.2 Mc.

Plate voltage for the oscillator is held to approximately 200 volts by a series-dropping resistor, R_2, and output from the stage is capacity-coupled to the final through C_6.

The amplifier employs grid-leak bias, has a split-stator plate circuit, and is neutralized by means of capacitor C_7. J_2 is the metering jack and S_1 is the plate-voltage on-off switch. With excitation available and with S_1 open, a meter plugged into J_2 will register amplifier grid current. When the switch is closed, the meter will indicate the combined plate and grid currents.

Output from the amplifier is link-coupled to the antenna tuner, $C_{13}L_4$. The tuner components have been wired to feed-through bushings and the antenna feeder terminals in a manner which permits adjustment of the LC ratio for either series or parallel tuning. An accompanying chart lists the jumper connections which should be used for setting up the tuner circuit.

The power supply employs a condenser-input filter and delivers approximately 330 volts when

Fig. 6-34 — The sweep-tube transmitter is housed in a hinged cover metal cabinet. The knobs across the bottom of the 7 × 10-inch panel, from left to right, control the oscillator, amplifier and the antenna coupler. S_1 is located directly above J_1 and to the left of the panel indicator. S_2 is mounted above the amplifier metering jack, J_2.

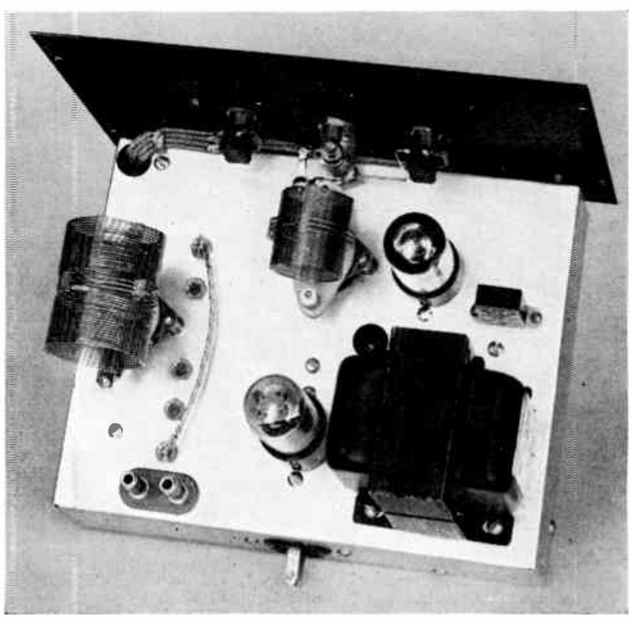


Fig. 6-35 — This interior view shows the antenna coil centered at the left edge of the 2 × 7 × 9-inch aluminum chassis. Five feed-through bushings for the antenna circuit are located to the right of the coil and the feeder terminals are at the rear of the base. L_2, the oscillator tube, and the crystal are at the front right-hand section of the chassis and the 5Y3GT is on the center line just to the left of the power transformer. A ⅝-inch hole, equipped with a rubber grommet, to the front of T_1, provides through-chassis clearance for a neutralizing tool. The a.c. input connector is located on the rear wall of the chassis.

loaded by the transmitter. S_2 is the on-off switch for the supply.

Construction

Three photographs of the transmitter show how the components are laid out on the chassis and the panel. The jacks, switches, and the panel indicator are the only parts actually mounted on the panel of the Bud type C-993 cabinet. Tuning capacitors for the oscillator and the amplifier are mounted on the front wall of the chassis and C_{13} of the coupler is mounted on small pillars at the right side (rear view) of the base. C_{13} must be insulated from ground. An insulated shaft coupling between the capacitor and a panel bearing assembly are provided. Quarter-inch metal pillars space the panel and base at either end of the unit. Three-eighths-inch holes are drilled in the panel for the tuning shafts of the three capacitors, and 1⅛-inch openings are punched in the front wall of the chassis to provide clearance for the panel-mounted jacks.

No. 16 tinned is used for the r.f. wiring, and Belden shielded wire No. 8885 is used for the leads running to the switches and the pilot lamp. The strip of flashing copper that supports the neutralizing condenser, C_7, is ½-inch wide at one end and tapers down to ⅛ inch at the tube socket end. C_7 is mounted in a ¼-inch hole, drilled at the wide end of the strip.

The three jumpers for the antenna circuit are made with ordinary hook-up wire and Millen type 36021 grid connectors. The holes in the connectors must be enlarged by reaming so that they will fit over the small National type TPB polystyrene bushings that serve as Terminals 1 through 5 of Fig. 6-36.

Testing

A 15-watt lamp bulb equipped with short wire leads, a 0–100-ma. meter, a key and a voltmeter should be available for testing the transmitter. The first test is made with the key plugged into J_1, with S_1 set at the open position and with the

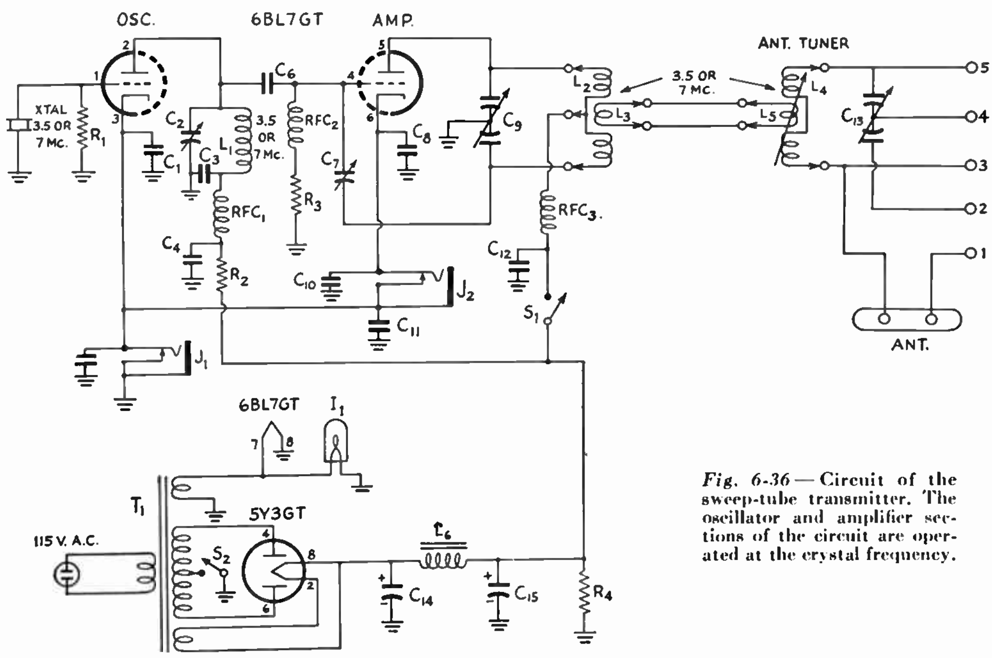


Fig. 6-36 — Circuit of the sweep-tube transmitter. The oscillator and amplifier sections of the circuit are operated at the crystal frequency.

C_1, C_3, C_4, C_8, C_{12} — 0.005-μfd. disk ceramic.
C_2 — 140-μμfd. variable (Hammarlund HF-140).
C_5, C_{10}, C_{11} — 0.001-μfd. disk ceramic.
C_6 — 15-μμfd. mica or ceramic.
C_7 — 1–8-μμfd. tubular trimmer (Erie 532–10).
C_9, C_{13} — 100-μμfd.-per-section variable (Bud LC-1663)
C_{14}, C_{15} — 8-μfd. 450-volt electrolytic (Sprague TVA-1704).
R_1 — 68,000 ohms, ½ watt.
R_2 — 10,000 ohms, 5 watts.
R_3 — 10,000 ohms, ½ watt.
R_4 — 50,000 ohms, 10 watts.
L_1 — 33 turns No. 24, ¾-inch diam., $1\frac{1}{32}$ inches long (B & W Miniductor No. 3012).
L_2 — 3.5 Mc. — 40 μh. — 46 turns No. 24, 1¼-inch diam., 1½ inches long, center-tapped (B & W 80MCL).
7 Mc. — 14 μh. — 26 turns No. 22, 1¼-inch diam., 1½ inches long, center-tapped (B & 40MCL).
L_3 — 3.5 and 7 Mc. — Each 3 turns No. 18, wound with turns spaced wire diam., over center of L_2.
L_4 — 3.5 Mc. — 37 μh. — 38 turns No. 16, 1¾-inch diam., $2\frac{7}{16}$ inches long. Wound in 2 sections with $\frac{7}{16}$-inch space at center for L_5 (B & W 80JVL).
7 Mc. — 12.8 μh. — 22 turns No. 16, 1¾-inch diam., $2\frac{5}{16}$ inches long. 2 sections with $\frac{7}{16}$-inch space at center for L_5 (B & W 40JVL).
L_5 — 3.5 and 7 Mc. — Each 3 turns No. 16, 1¾-inch diam., turns spaced wire diam.
L_6 — 8-henry 75-ma. filter choke (Stancor C1355).
I_1 — 6.3-volt panel-indicator assembly.
J_1, J_2 — Closed-circuit jacks.
RFC_1, RFC_2 — 1-mh. r.f. choke (National R-50).
RFC_3 — 2.5-mh. r.f. choke (National R-100S).
S_1, S_2 — S.p.s.t. toggle switch.
T_1 — Power transformer: 340 volts r.m.s. each side of center tap, 70 ma.; 5 volts, 2 amp.; 6.3 volts, 2.5 amp. (Stancor PC8408).

Antenna-Coupler Connection Chart

	Jumper Connections		
Tuning	*Low-C*	*Med.-C*	*High-C*
Parallel	1–5 2–3	1–5 3–4	1–5 2–5 3–4
Series	1–2	1–4	1–4 2–5

voltmeter connected across R_4. The supply output should exceed 400 volts when S_2 is closed.

Next, turn off the supply and insert a 3.5-Mc. crystal in the holder and a 3.5-Mc. coil in the amplifier. The meter should be plugged into J_2 and S_1 must be open for the time being. Now, turn on the power, close the key and tune the oscillator plate capacitor, C_2, for an amplifier grid current of approximately 10 ma. If the crystal kicks out as the maximum capacitance of C_2 is reached, the plate tank is tuned too close to the crystal frequency and it is necessary to retune to the high frequency side of resonance. Make certain that the oscillator is not tuned for *maximum* output inasmuch as this results in excessive crystal current. If the meter is transferred to J_1, it should show a cathode current of 30 ma.

The next step is that of neutralizing the amplifier. Start with C_7 set for minimum capacitance (slug all the way out) and then increase the capacitance until the amplifier plate condenser, C_9, can be swung through resonance without affecting the amplifier grid current. S_1 must be open during this adjustment.

If the lamp is to be used as the test load, connect it to the antenna terminals and insert the 7-Mc. coil in the coupler. Start the loading adjustments with very loose coupling between L_4 and L_5 and with the oscillator adjusted for an amplifier grid current of 5 or 6 ma. Now, close S_1 and tune C_9 for resonance. The amplifier cathode current should be approximately 25 ma. with the stage lightly loaded and may be increased to 55 or 60 ma. by increasing the coupling between L_4 and L_5 and by adjustment of C_{13}. As the loading is increased, make certain that the amplifier and the tuner are kept at resonance by retuning both C_9 and C_{13}.

With the amplifier fully loaded, the power supply output voltage will drop to approximately 325 volts and, as a result, the cathode current for the oscillator section of the 6BL7 will be lower than that recorded earlier. About 15 ma. is correct for the oscillator and this current may be checked by inserting the meter plug into J_1. Of course, with the amplifier in operation, it is necessary to subtract the amplifier cathode current from the reading registered at J_1 in order to determine the true oscillator drain.

The set-up for testing the transmitter at 7 Mc. is identical to that used at the lower frequency except for the antenna coupler connections. At 7 Mc., the bulb loads best with the coupler circuit adjusted for low-C operation. One precaution must be observed with the 7-Mc. crystal in use. *Always* start the oscillator adjustment with the tank capacitor, C_2, set for *minimum* capacitance and then tune for an amplifier grid current of not more than 5 or 6 ma.

For adjustment of the coupler for a particular antenna, see the transmission-line chapter.

(Original description, *QST*, April, 1953.)

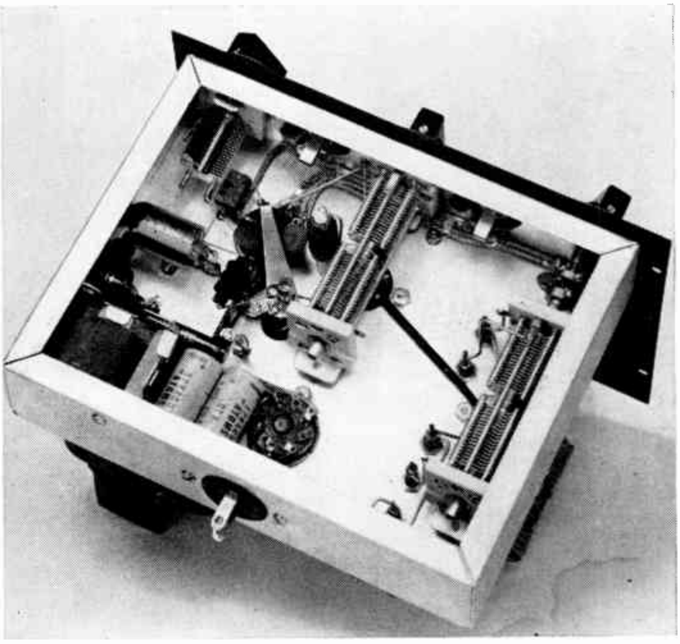


Fig. 6-37 — Bottom view showing L_1 and RFC_2 mounted on tie-point strips to the left and the rear of the 6BL7 tube socket, respectively. RFC_1 is parallel with the left wall of the chassis and RFC_3 stands up to the left of C_9. R_2 and R_4 are in front of L_6 and the filter capacitors at the rear of the chassis. The neutralizing capacitor, C_7, is supported by the rear stator terminal of C_9 and by a strip of flashing copper which also serves as the capacitor-to-grid lead. Holes, 1⅛ inches in diameter, punched in the chassis just below the centers of C_9 and C_{13}, provide clearance for the coil-socket wiring.

A Beginner's 35-Watt Transmitter

Figs. 6-38 through 6-40 illustrate a 35-watt two-stage transmitter for the 40- and 80-meter bands. The necessary power supply is included. The circuit is shown in Fig. 6-38. A 6AG7 Pierce crystal oscillator operating at 3.5 Mc. drives a 6L6, either as a straight amplifier on 80, or as a doubler to 40 meters. RFC_1 is resonant at about 5 Mc. — sufficiently close to either band to provide the required drive to the amplifier, yet far enough removed to prevent oscillation in the 6L6 stage. The output tank circuit, C_9L_1, has sufficient tuning range to include both bands without changing coils; the socket and plug-in form are merely a convenient means of mounting the coil. The output link is designed to feed an antenna tuner through a coax line. Both stages have parallel plate feed, and are keyed simultaneously in the cathode circuit. I_1 is a dial lamp, used here as a tuning indicator. If desired, it may be replaced with a 150-ma. d.c. milliammeter, either mounted on a bracket on top of the chassis, or set in the front edge.

With the components specified, the power supply should deliver a voltage of 350 or more under load. A condenser-input filter is used. (Although a metal-can dual filter condenser, mounted on top of the chassis, is shown, cardboard tubular condensers, mounted under the chassis may be substituted if desired.)

Wiring

Details of construction are covered in the photographs and their captions.

The power supply is wired first, using insulated tie points as junctions wherever a transformer of filter-condenser will not conveniently reach a desired terminal. (All power wiring should be kept close against the chassis, while r.f. wiring should be spaced well away from the chassis.) The heaters of the 6AG7 and 6L6 are wired next.

Pin 8 of the 6L6 and Pin 5 of the 6AG7 are wired together and C_2 and C_6 are installed. A lead is then run from Pin 5 of the 6AG7 to the key jack and C_{14} is installed across the key jack, keeping the leads of C_{14} as short as possible. This completes the cathode keying circuit.

The square condenser appearing over the 6AG7 socket is C_3 and is connected between Pin 6 and ground. R_2, the screen dropping resistor, is connected from Pin 6 to the tie point between the tubes. The B+ lead is run to this tie point, and both R_2 and R_4 are tied to it. RFC_1 goes from Pin 8 of the 6AG7 to the tie point of the B+ lead. The condenser below RFC_2 is C_8 — it is con-

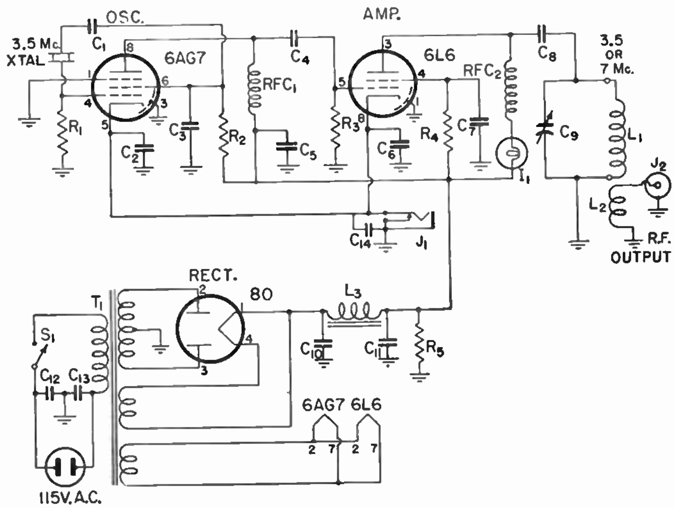
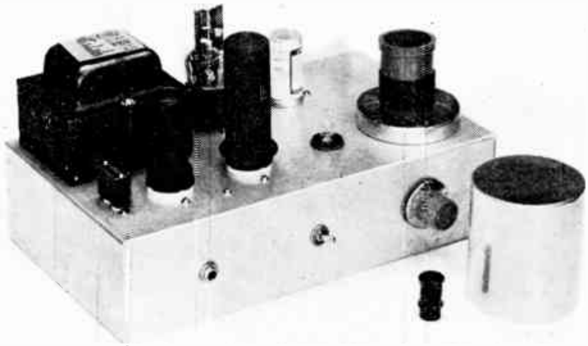


Fig. 6-38—Circuit diagram of the Novice 35-watt transmitter

C_1, C_2, C_5, C_6, C_7, C_8 — 0.005-μfd. 500-volt disk-type ceramic (Sprague).
C_3, C_4 — 100-μμfd. mica.
C_9 — 235-μμfd. variable (Bud MC-1859).
C_{10}, C_{11} — 10-μfd. 450-volt electrolytic (see text).
C_{12}, C_{13}, C_{14} — 0.001-μfd. 500-volt disk-type ceramic (Sprague) (see text).
R_1 — 56,000 ohms, ½ watt.
R_2 — 22,000 ohms, 1 watt.
R_3 — 18,000 ohms, 1 watt.
R_4 — 18,000 ohms, 1 watt.
R_5 — 50,000 ohms, 10 watts.
L_1 — 3.5–7.0 Mc. — 15 turns No. 18 enamel, 1½-inch diam., close-wound (National XR-4 coil form).
L_2 — 5-turn link No. 18 enamel, close-wound below tank coil L_1.
L_3 — Filter choke, 10.5 henrys, 110 ma., 220 ohms (Merit C-2993).
I_1 — No. 46 pilot-lamp bulb, 6–8 volts, 250 ma., blue bead.
J_1 — Closed-circuit jack.
J_2 — Coax connector, chassis-mounting type.
RFC_1 — 100-μh. r.f. choke (Millen 34300).
RFC_2 — 2.5-mh. r.f. choke (National R100-S).
S_1 — S.p.s.t. toggle switch.
T_1 — Power transformer, 350 volts r.m.s. each side of center, 120 ma.; 6.3 volts, 4.7 amp.; 5 volts, 3 amp. (Merit P-2953).

Fig. 6-39 — The aluminum chassis is 7 × 12 × 3 inches. Power-supply components are along the rear edge, while the crystal socket, 6AG7, 6L6, I_1 and the shielded coil are in line at the front. Centered along the front edge are the key jack, power switch and the single tuning control. All sockets are submounted. The rectifier and the coil take 4-prong sockets; the two tubes take octal sockets. The coil shield is ICA type 1549. The substitution of an upright transformer will avoid cutting a large hole in the chassis.



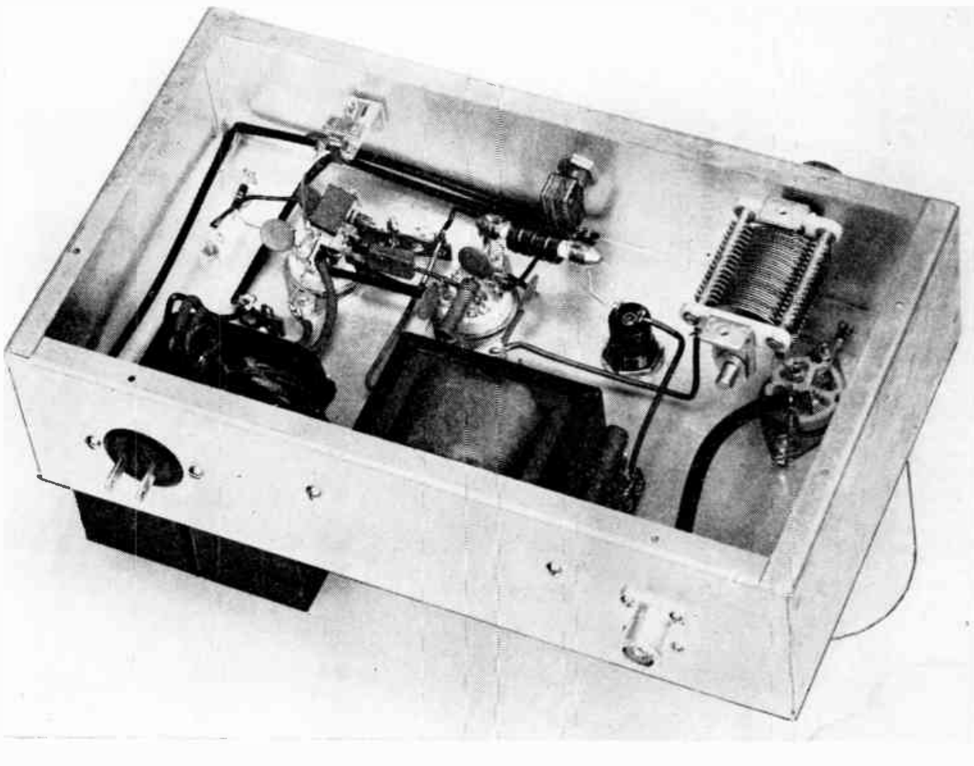
nected from Pin 3 of the 6L6 to a tie point and then to the stator of C_9. The link output terminals on the coil socket are connected to the coax connecter with a short length of coax cable. The v.h.f. filter condensers, C_{12} and C_{13}, are at the power connector with leads as short as possible.

Testing

The transmitter may be tested by connecting a 25-watt electric bulb between the center contact of the coax connector and chassis. When the power is turned on, and the key closed, the indicator lamp, I_1, should light up brightly. Then, starting at maximum capacitance, slowly adjust the tuning condenser, toward minimum capacitance until the indicator lamp dims. This is resonance at 80 meters, and the 25-watt lamp should light up as the indicator lamp dims. Further readjustment of the tuning condenser toward minimum capacitance should show a second resonance point, this time at 40 meters, and the 25-watt lamp should light again.

Information on the construction and adjustment of antenna couplers will be found in the chapter on transmission lines. The 6L6 may be loaded up to a maximum of 100 ma. plate current.

Fig. 6-40 — The a.c. power connector and coax output connector are at the rear. The filter choke, L_3, is fastened against the rear of the chassis. The choke to the rear of the power switch is RFC_2. The tuning condenser is in the upper right.



A Compact 75-Watt 6-Band Transmitter

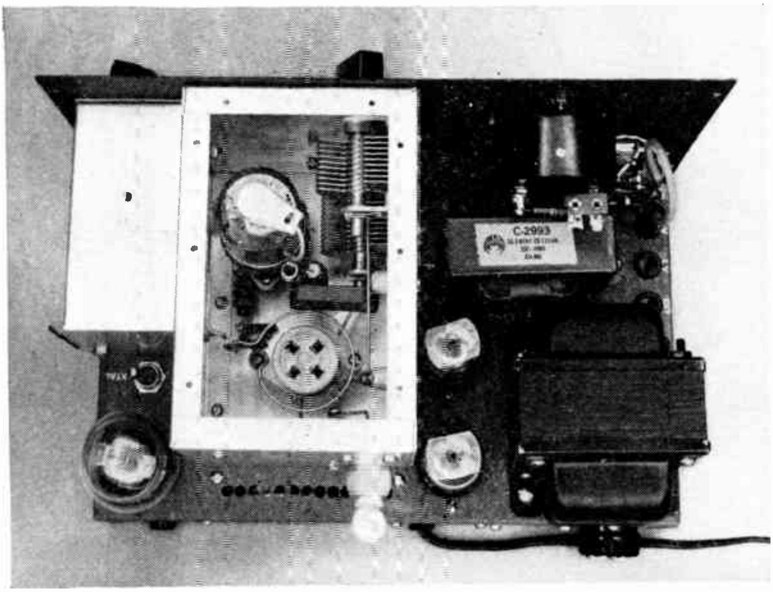
Fig. 6-41 — The complete 75-watt 6-band transmitter fits into an 8 × 14 × 8-inch cabinet. Along the bottom, from left to right, are the two power switches (S_5 and S_6), the key jack (J_7), the "operate-test" switch (S_4) and the crystal socket. Across the center are the meter switch (S_3), the amplifier tank control (C_9) and the oscillator tuning condenser (C_8). To the right of the meter at the top are the loading condenser (C_{10}) and the oscillator bandswitch (S_2).

Figs. 6-41 through 6-47 show the circuit and photographs of a two-stage transmitter delivering an r.f. output of 50 watts on all bands from 3.5 to 28 Mc., inclusive. It is complete with power supply and a versatile metering system on a 11 × 7 × 2-inch chassis. Provision is made for connection of a VFO, a plate-and-screen modulator and also an external emergency power supply.

As the circuit diagram of Fig. 6-44 shows, a 5763 is used in a grid-plate oscillator circuit. C_2 is a mica trimmer that permits adjustment of oscillator excitation for proper keying and drive to the amplifier. S_1 grounds the cathode through C_3 so that the 5763 can be driven from a VFO through the crystal socket. L_1 is tapped to cover 3.5 through 28 Mc. with a switch, S_2. The oscillator output with either 3.5- or 7-Mc. crystals, at either fundamental or second harmonic, is more than adequate for proper drive to the 6146 amplifier. Sufficient drive is also obtained quadrupling from 3.5-Mc. crystals to 14 Mc., or tripling to 21 Mc. from 7-Mc. crystals. Quadrupling from 7-Mc. crystals, however, does not supply adequate excitation, so frequency is doubled in the output stage for 27- or 28-Mc. operation, unless 9-Mc. crystals for tripling, or 28-Mc. crystals, are available.

Plug-in coils are used in the output tank circuit. Since both stages are parallel-fed in the plate circuit, the power supply need not be turned off while changing coils. The amplifier is

Fig. 6-42 — The oscillator is in the 2 × 4 × 4-inch box to the left, with the crystal-VFO switch and 5Y4G immediately behind. The amplifier is in the 4 × 5 × 6-inch box. C_9 (bottom) and C_{10} (top) are mounted against the right-hand side of the box. The coil socket is to the rear surrounded by the 1-turn neutralizing link. C_8, RFC_4 and L_2 are immediately in front of the coil socket. To the right are the two 6X5GTs, the power transformer and L_8. The pin jacks toward the front are metering jacks. The holes at the rear are for ventilation.



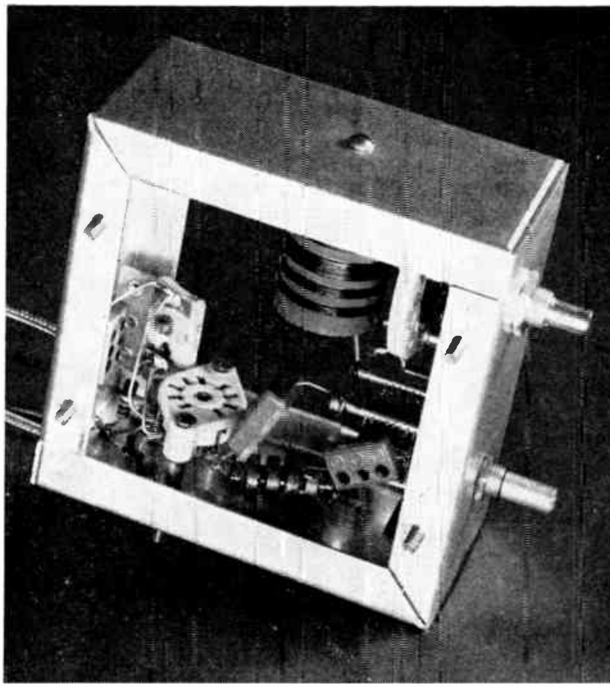


Fig. 6-43 — Inside of the oscillator box from the amplifier side. RFC_1 and C_7 are in the foreground in this view. Leads from C_7 and L_5 are precut to pass through to the amplifier compartment.

neutralized by means of a simple inductive link system (L_5 and L_6). L_2 is a v.h.f. parasitic suppressor.

Both stages are keyed simultaneously in the cathode circuit for break-in operation, the key being plugged in at J_7.

Power Supply

An economical power supply delivering voltages for both stages is included on the chassis. A voltage of 600 (under load) for the final amplifier is obtained from an inexpensive broadcast replacement transformer through the use of a bridge rectifier circuit. The center tap of this system provides a voltage of 230 for operating the oscillator and the screen of the amplifier, the latter through the dropping resistor, R_9. The choke, L_8, in the high-voltage filter, it should be noted, is connected in the negative side of the supply. When using the built-in supply, a plug with the pins shorted, as indicated by the dotted lines, should be inserted in J_8. When using an emergency supply, appropriate voltages can be introduced through J_8 after the shorting plug has been removed.

COIL DATA

Oscillator Coil, L_1: Wound with No. 26 enameled wire on 1-inch diameter form (Millen 45000) in four sections.

1st section: 20 turns close-wound
2nd section: 10 turns close-wound
3rd section: 5 turns close-wound
4th section: 4 turns spaced wire diameter

Taps taken off between sections. Spacing between sections approximately ⅛ inch. Fourth section (21–28 Mc.) turn spacing should be adjusted to cover 30 Mc. with oscillator condenser, C_6, near minimum capacitance.

Amplifier coils, L_3L_4:

Band		Wire Size	Turns	Turns/inch	Space Between Coils
3.5 Mc.	L_3	22 enam.	15	20	⅛ in.
	L_4	22 enam.	20	close-wound	
7 Mc.	L_3	18 enam.	10	10	3/16 in.
	L_4	18 enam.	8	close-wound	
14 Mc.	L_3	18 enam.	5	10	0.2 in.
	L_4	18 enam.	5	10	
21-28 Mc.	L_3	18 enam.	3	10	0.2 in.
	L_4	18 enam.	3	10	

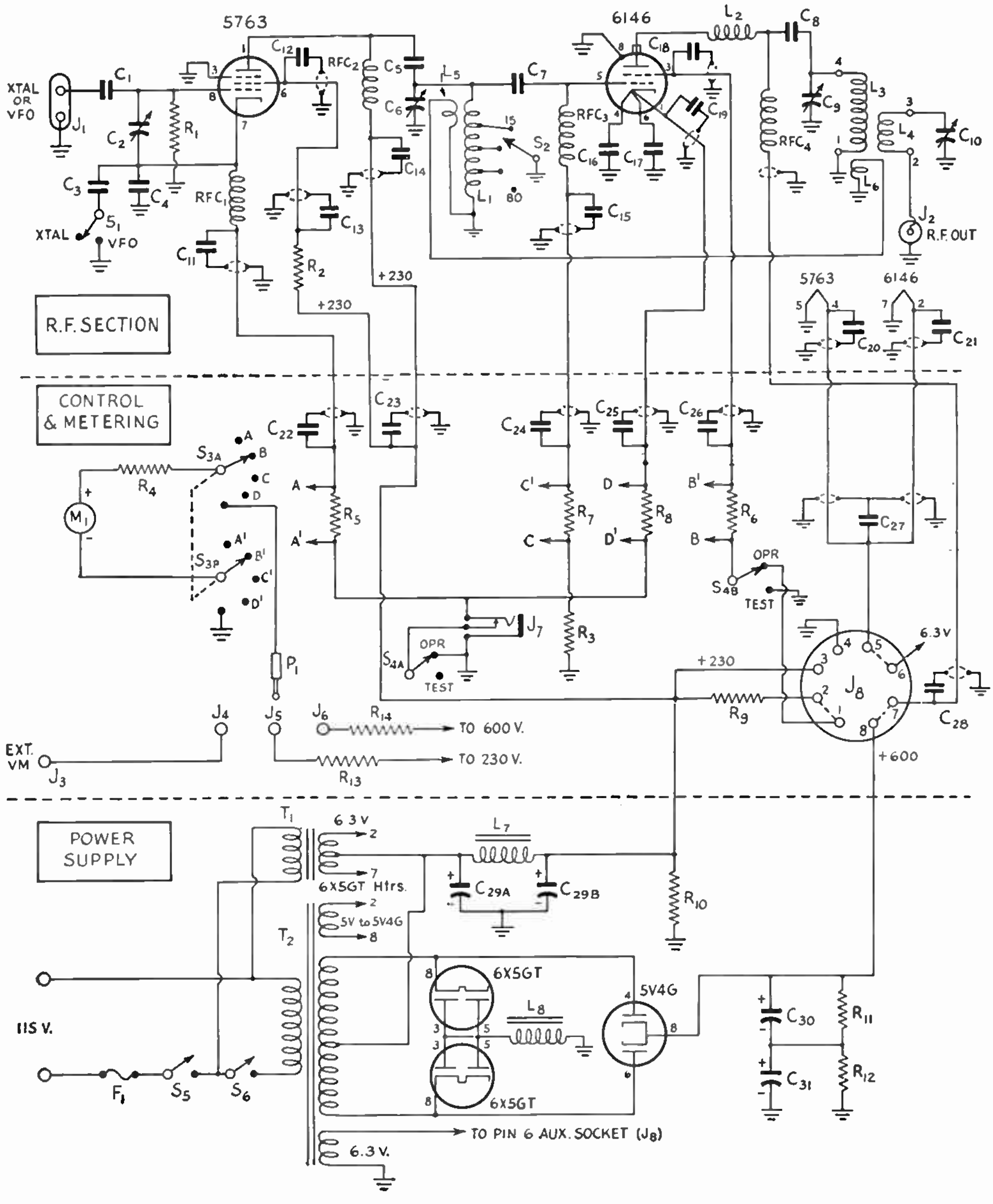
Coils wound on 1½-inch diameter forms (National XR-4) with L_3 at bottom and plate terminal down. See Fig. 6-44 for connections in coil form and socket.

Metering Circuits

A 1-ma. milliammeter, M_1, is used for measuring the essential currents and voltages. It is connected as a voltmeter having a full-scale range of 5 volts by adding R_4 in series. Current is determined by measuring the voltage drop across resistors of proper value inserted in series with the circuits in which current is to be measured. This permits the use of standard resistors as current shunts. The ranges selected here are as follows: oscillator cathode current, 50 ma.; amplifier grid current, 10 ma.; amplifier screen current, 20 ma.; amplifier cathode current, 200 ma. In addition, three tip jacks mounted on the chassis can be selected by a test prod connected to one position on the meter switch. One, J_5, is connected to the power-supply low-voltage terminal through R_{13} which is a multiplier giving a full-scale meter reading of 300 volts. A second tip jack, J_6, is similarly connected to the high-voltage terminal through a 1000-volt multiplier, R_{14}. The third tip jack, J_4, connects to another similar jack, J_3, at the rear of the chassis so that the meter can be used for external measurements, such as an indicator for an s.w.r. bridge or in an r.f. voltmeter for checking power output.

Test-Operate Switch

A useful adjunct is the "test-operate" switch, S_4. In the "operate" position, the amplifier screen is connected to its normal supply. In the "test" position, the screen is grounded. This limits the plate current to about 15 or 20 ma. which results in just about the right amount of power to operate an s.w.r. bridge. If the 6146 is to be plate-screen modulated, the screen voltage must be obtained from the high-voltage tap through a dropping resistor, rather than from the low-voltage tap. In this case, the cathode should never be opened while the power supply is on, because the voltage ratings of both the tube screen and the ceramic by-pass condensers will be greatly exceeded. S_{4A} guards against this by grounding the cathode through an auxiliary contact of J_7 when the key is removed. Then S_4 becomes the on-off switch, opening both cathode circuits (through S_{4A}) and grounding the amplifier screen (through S_{4B}) when the switch is in the "test" position. To turn the oscillator on and close the amplifier cathode circuit for "test" use, a closed key, or shorted plug, must be inserted in the closed-circuit jack, J_7.



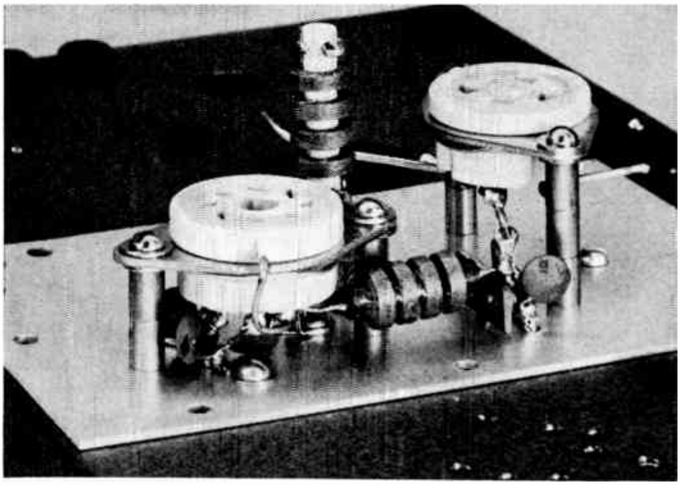


Fig. 6-45 — The bottom plate of the amplifier box is fastened permanently to the chassis and the amplifier partially assembled before fastening the box in place. RFC_3 is in the foreground, RFC_4 standing at the rear. The coil socket at the right is spaced up 1⅜ inches, the tube socket ¾ inch. Notice the "zero-length" leads to the disk ceramic condensers.

Neutralizing Coils

The neutralizing coil, L_5, is made simply by drilling two small holes diametrically opposite close to the outer (15-meter) end of the form. A piece of rather stiff wire is threaded through the holes and then the wire inside the form is pressed into a half-turn shape with the finger. Connections are made to each end outside the form and the half-turn may be rotated in the holes to adjust neutralization. L_6 is a single turn of No. 12 wire, approximately 1⅞ inches in diameter, supported at one end under the screw holding the socket, and at the other end by a tie-point mounted on the same screw.

Adjustment

With the key open, the supply voltage at the high tap should measure about 800 volts and 300 at the low tap. If the 5V4G is removed from its socket, the voltage at the low tap will be about 400.

With the switch set in the "test" position, the oscillator tuning should be adjusted for maximum amplifier grid current. A reading of 4 ma. indicates adequate drive, although on some bands it may run as high as 10 ma. If the minimum reading of 4 ma. is not obtained, adjust C_2. Up to a certain point, increasing this capacitance will increase the oscillator output, but too much feedback may result in chirpy keying. C_2 should be adjusted for the best compromise between adequate drive and good keying characteristics. The oscillator cathode current should run 25 to 30 ma. on all bands.

Neutralization is adjusted by moving the half turn L_5 closer to or farther away from the oscillator tank coil. With S_4 in the "test" position, the oscillator should be adjusted for maximum amplifier grid current on 21 Mc., and the amplifier plate tank circuit tuned to resonance. If the amplifier is not neutralized, there will be a noticeable kick in grid current as the plate tank condenser is swung through resonance. The neutralizing half turn should be adjusted carefully for minimum change in grid current. The same procedure should be followed for 14 Mc. If the neutralizing must be readjusted, the half turn should be set for the best average result for the two bands. The amplifier should then be checked for oscillation with S_4 in the "operate" position. The amplifier plate current at resonance should swing the meter off scale when the key is closed.

Fig. 6-44 — Circuit diagram of the complete transmitter. Dotted lines in J_8 indicate jumpers in plug used for normal operation.

C_1, C_7 — 220-μμfd. mica.
C_2 — 3-30-μμfd. ceramic trimmer, compression type.
C_3 — 0.002-μfd. mica.
C_4 — 100-μμfd. mica.
C_5 — 0.002-μfd. mica.
C_6 — 50-μμfd. midget variable (Bud LC-1644).
C_8 — 0.001-μfd. mica, 1200 volts, case type CM-45.
C_9 — 235-μμfd. variable, 0.024-inch spacing (Bud type MC-1859).
C_{10} — 140-μμfd. variable, 0.024-inch spacing (Bud type MC-1856).
C_{11} to C_{27}, inclusive — 0.001-μfd. disk ceramic, ⅜-inch diam., 600 volts.
C_{28} — 470-μμfd. mica, 1200 volts, case type CM-45.
C_{29} — Dual 8-μfd. electrolytic, 450 volts.
C_{30}, C_{31} — 40-μfd. electrolytic, 450 volts.
R_1 — 0.1 megohm, ½ watt.
R_2, R_3 — 27,000 ohms, 1 watt.
R_4 — 5000 ohms, ½ watt.
R_5 — 100 ohms, ½ watt.
R_6 — 263 ohms (270), ½ watt.
R_7 — 555 ohms (560), ½ watt.
R_8 — 25 ohms (27), ½ watt.
R_9 — 4700 ohms, 1 watt.
R_{10} — 0.1 megohm, 1 watt.
R_{11}, R_{12} — 20,000 ohms, 10 watts
R_{13} — 0.5 megohm, ½ watt.
R_{14} — 1 megohm, 1 watt.
L_1 — See coil data.
L_2 — 4 turns No. 16, 5⁄16-inch diam., ⅜ inch long.
L_3, L_4 — See coil data.
L_7 — Filter choke, 40 ma., 300 ohms, approximately.
L_8 — 10.5 henrys, 110 ma., 250 ohms.
F_1 — Fuse, 2 amp.
J_1 — Crystal socket.
J_2 — Coax connector, chassis-mounting type.
J_3, J_4, J_5, J_6 — Tip jacks, insulated type (Ampheno 78-1P).
J_7 — Closed-circuit 'phone jack.
J_8 — Octal socket.
M_1 — 0-1 d.c. milliammeter.
P_1 — 'Phone tip test plug.
S_1, S_5, S_6 — S.p.s.t. toggle.
S_2 — Single-pole 5-position ceramic wafer (Centralab 2500 or 2501).
S_3 — 2-pole 5-position bakelite wafer, non-shorting type (Centralab type 1405).
S_4 — D.p.d.t. toggle.
RFC_1, RFC_2, RFC_3 — 2.5 mh., 75 ma. (Millen 34300-2500).
RFC_4 — 2.5 mh., 250 ma. (Millen 34102).
T_1 — Filament transformer, 6.3 v., 1.2 amp.
T_2 — Power transformer, 360 v. each side c.t., 120 ma.; 5 v., 3 amp.; 6.3 v., 3 amp. or more.

NOTE: Manufacturer's part numbers given above are to indicate size and style. Similar components are generally available from a number of different suppliers.

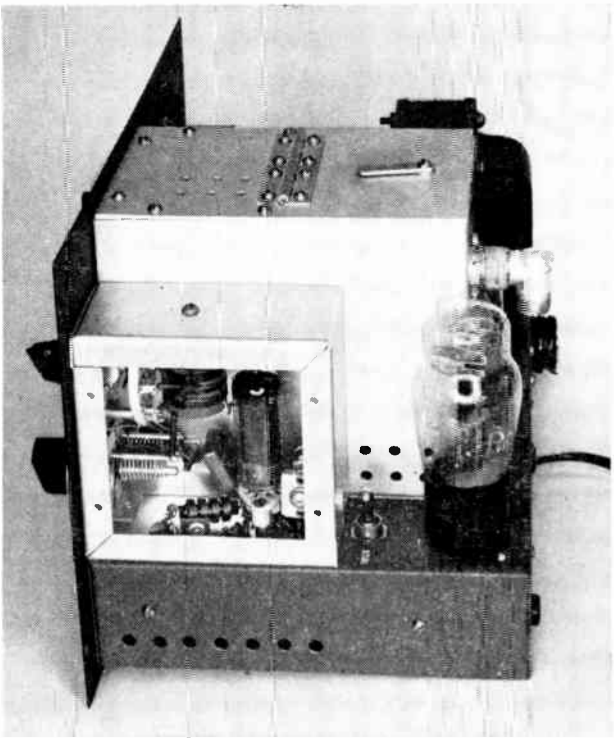


Fig. 6-46 — Looking into the oscillator compartment. L_1 and S_2 are at the top with C_6 below. RFC_2 and C_{14} are supported on a tie point in the foreground. R_1, C_2 and C_4 are to the rear of the tube. C_5 is soldered between C_6 and the tube socket. RFC_1 and C_7 are hidden by the tube and tuning condenser. The cover of the amplifier is hinged at the center for changing coils. The latch at the rear engages the rear lip of the box so that the lid is drawn down tight. Notice the numerous ventilating holes.

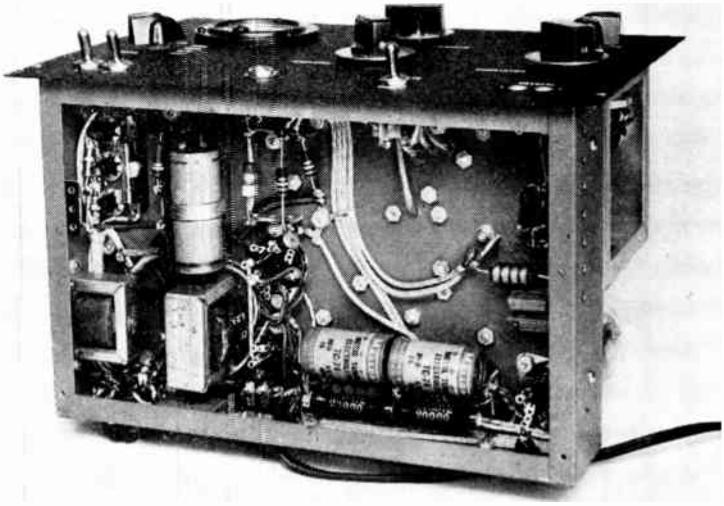
Do not close the key more than momentarily for this check.

The output coupling system is designed to work into a flat 50- or 75-ohm line, either to an antenna or to an antenna tuner. The amplifier may be loaded to a cathode current of 140 Ma. on all bands except 28 Mc. Under load, the amplifier grid current should be adjusted to 2 to 2.5 ma. by detuning the oscillator tank circuit. The screen current under these conditions should run between 10 and 12 ma. At 28 Mc., with the final amplifier doubling, the grid current should be adjusted to the maximum possible (5 to 6 ma. under load) and the cathode current limited to about 120 ma. Loading can be adjusted by C_{10} which tunes the link circuit.

In fringe areas, a low-pass filter may be required for 21- and 28-Mc. operation. On lower frequencies, or in the presence of good TV signals, the use of a conventional antenna tuner will usually be adequate to suppress TVI.

(Originally described in *QST*, December 1952.)

Fig. 6-47 — Bottom view of the 6-band transmitter. The high-voltage filter condensers, C_{30} and C_{31}, and their equalizing resistors, R_{11} and R_{12}, are at the rear of the chassis. T_1 and L_7 are to the left, with C_{29} and R_{10} above. J_5 is in the extreme rear left-hand corner. At top center, supported on insulated tie points, l. to r., are R_9, R_2 and R_3. In the upper left-hand corner are R_4, R_{13} and R_{14} and F_1. RFC_1, C_1 and C_3 are to the right. Shielded wiring and disk-ceramic condensers are applied according to method described in the chapter on TVI.



A Completely-Shielded 90-Watt Transmitter or Exciter

The transmitter shown in Figs. 6-48 through 6-52 is designed for the reduction of v.h.f. harmonic radiation without requiring special construction for shielding purposes. It uses a standard 3 by 4 by 17 inch chassis as the main enclosure. The plug-in coils are provided with individual shields using 3-inch diameter removable shield cans that also are standard items.

The final amplifier is a 6146, driven by a 6AG7 frequency multiplier that is driven in turn by a 6AG7 crystal oscillator-multiplier. Provision is made for driving the latter tube from an external VFO. The power output is approximately 60 watts on all bands from 3.5 through 28 Mc. at the 90-watt input c.w. rating of the 6146. With plate modulation the 67-watt input rating gives a carrier output of close to 50 watts.

Oscillator Circuit

The crystal oscillator uses the grid-plate circuit and is intended for use with either 3.5- or 7-Mc. crystals. Its plate circuit, L_1C_4 in Fig. 6-49, covers the range from 7 to 14.5 Mc. and L_1 is wired permanently in the circuit. When using 7-Mc. crystals C_4 is tuned toward its high-capacity end when 7-Mc. output is required for the following stage, and near the low-capacity end when the buffer is driven on 14 Mc. With 3.5-Mc. crystals C_4 is set near maximum capacity for 7-Mc. excitation of the buffer, and at or below midscale for 3.5-Mc. excitation. The tuning in the latter case corresponds to the setting that gives minimum harmonic output from the oscillator; at 3.5 Mc. enough fundamental voltage gets through to the buffer grid to give it adequate drive. Coil changing in the oscillator circuit is avoided by this method.

For VFO input the feed-back condenser, C_2, is shorted to ground for r.f. by S_1. The crystal should be removed from its socket when using the VFO. A coaxial connector is used for the VFO circuit, and the VFO should be of the type that includes the length of coax as part of its tuned output circuit. The VFO output can be on either 3.5 or 7 Mc., depending on the final output frequency and the choice of method of operation, as described later.

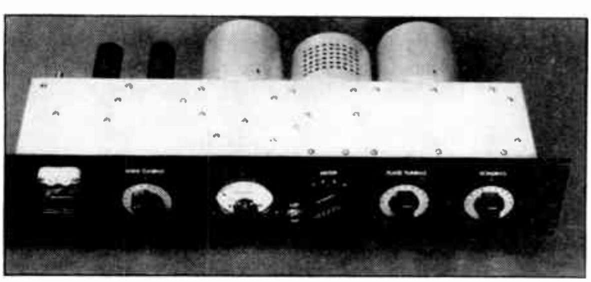


Fig. 6-48 — **A compact and completely shielded low-power transmitter using a 6146 as the final amplifier. It can be used at an input of 90 watts on c.w. or 67 watts for plate-modulated 'phone. The unit is mounted on a 3½-inch rack panel.**

Frequency Multiplier

The frequency multiplier or buffer stage is coupled to the final amplifier grid by a pi network. This type of circuit permits using a relatively large fixed capacitance, C_9, directly from grid to ground in the amplifier circuit and is highly advantageous in preventing v.h.f. harmonics generated in the grid circuit from developing an appreciable voltage between grid and ground. This not only prevents amplification of such harmonics in the plate circuit but also helps keep harmonic currents from flowing in the d.c. grid return lead.

C_9 is also useful in stabilizing the final amplifier to prevent self-oscillation at the operating frequency. The larger the capacitance of C_9 in comparison with the capacitance in use at C_7, the greater the impedance step-down between the buffer plate and the amplifier grid, thus the buffer plate resistance is reflected as a comparatively low resistance at the grid of the amplifier. This, together with the fact that any energy fed back from the amplifier plate circuit through the tube's grid-plate capacitance cannot develop much feed-back voltage across the large fixed capacitance between grid and cathode, effectively prevents self-oscillation and avoids the necessity for neutralization of the amplifier. The optimum circuit values for this purpose are given in Fig. 6-49 and the buffer coil table.

On 3.5 Mc. additional capacitance, C_8, is connected in parallel with C_9 to provide proper circuit operation. On all frequencies the buffer tuning condenser, C_7, is near minimum capacity at the proper operating setting. A 50 $\mu\mu$fd. condenser can be used instead of the one specified in Fig. 6-49, if desired.

L_2 and L_3 are small coils in the buffer grid and plate circuits to prevent v.h.f. parasitic oscillations in the buffer stage.

Amplifier Output Circuit

The amplifier output circuit also is a pi network, designed specifically for working into essentially resistive loads between 50 and 75 ohms. It is therefore suitable for working into properly terminated coaxial cable of the usual impedance values. In cases where the antenna is fed by types of line other than coax, an antenna matching network or antenna tuner of the coax-coupled type described in the chapter on transmission lines should be used. This permits operating the coax link at a low standing-wave ratio and provides the proper load for the 6146 amplifier circuit.

The amplifier tank condenser, C_{12}, is a split-stator type connected to the coil socket in such a way that only one section is used on all bands except 3.5 Mc., where the second section is connected in by means of a jumper in the coil form.

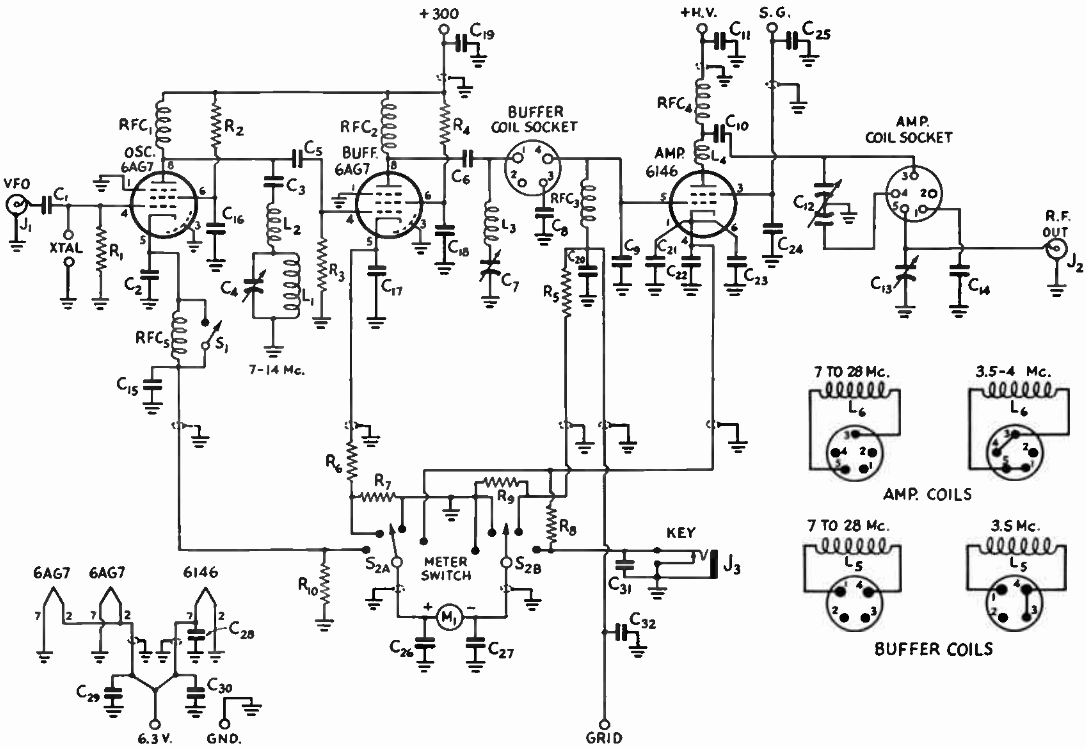


Fig. 6-49 — Circuit diagram of the transmitter.

C_1, C_3, C_5, C_6 — 470-μμfd. mica.
C_2 — 150-μμfd. mica.
C_4, C_7 — 140-μμfd. variable (Millen 19140)
C_8, C_9 — 100-μμfd. silver mica.
C_{10} — 0.001-μfd. mica, 1200-volt working.
C_{11} — 470-μμfd. mica, 1200-volt working.
C_{12} — 100-μμfd. per section variable, 1000-volt spacing (National TMS-100D).
C_{13} — 325-μμfd. variable (Millen 19325).
C_{14} — 470-μμfd. silver mica.
C_{15} to C_{32}, inc. — 0.001-μfd. ceramic, midget size.
R_1, R_3 — 47,000 ohms, ½-watt.
R_2 — 47,000 ohms, 1 watt.
R_4 — 15,000 ohms, 1 watt.
R_5 — 27,000 ohms, 1 watt.
R_6 — 150 ohms, ½ watt.
R_7 — 2.2 ohms (2X shunt for 0–25 milliammeter).
R_8 — 0.24 ohms (10 × shunt for 0-25 milliammeter).
R_9, R_{10} — 100 ohms, ½ watt.
J_1, J_2 — Coax connectors, chassis type.
J_3 — Closed-circuit jack.
RFC_1, RFC_3, RFC_4 — 2.5 mh. r.f. choke (National R-100S).
RFC_2 — 1-mh. r.f. choke (National R-300S).
RFC_5 — 2.5-mh. r.f. choke (Millen 34300–2500).
L_1 — 13 turns No. 22, diameter 1 inch, length 1 inch.
L_2 — 16 turns No. 30 d.c.c. on ½-watt resistor.
L_3 — 6 turns No. 14, diameter 5/16 inch, length 1 inch.
L_4 — 8 turns No. 18, diameter ¼ inch, length ⅝ inch.
L_5, L_6 — See coil table.
M_1 — 0–25 d.c. milliammeter (Simpson Model 125).
S_1 — S.p.s.t. toggle.
S_2 — 2-pole, 4-position wafer switch, non-shorting (Centralab 2505).

L_4 in the amplifier plate lead is for the purpose of preventing v.h.f. parasitic oscillation in the amplifier.

Other Circuit Details

Cathode currents of all three tubes can be measured by means of the meter switching arrangement shown in Fig. 6-49. The amplifier grid current also can be measured. The 0–25 milliampere scale is used directly for measuring the oscillator cathode current and amplifier grid current, the meter being shunted by 100-ohm resistances in each of these two positions to preserve circuit continuity when the switch is in other positions. In the switch position for measuring buffer cathode current the meter is shunted by a low resistance that multiplies the scale by 2, and when the final amplifier cathode current is measured the meter is similarly shunted by a resistance that multiplies the range by 10 so that the full-scale reading is 250 milliamperes. The values of multiplier resistance required in these two cases will depend on the type of instrument used and should be adjusted to the proper value experimentally. The method is described in the chapter on measuring equipment.

Loading is controlled by the output condenser, C_{13}. Although it has the highest capacitance available in condensers of this construction, it is not large enough for proper operation of the pi network on 3.5–4 Mc., so an additional capacitance, C_{14}, is connected in on this band by means of a jumper in the coil form. This large fixed capacitance restricts the adjustment range possible with C_{13}, so two coils are needed for proper loading in this band. The one covering the 3500–3750-kc. range is adjusted for proper loading to maximum permissible tube input at c.w. ratings,

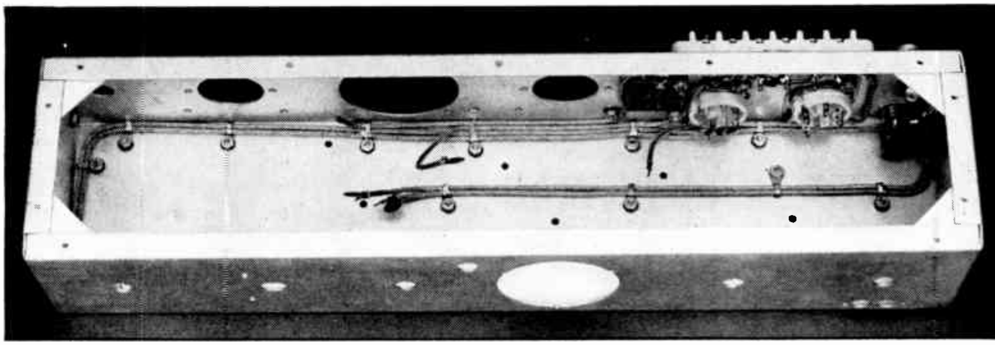


Fig. 6-50 — The shielded power wiring should be installed before the r.f. components are permanently mounted, including the ceramic by-passes across the ends of the shielded wires. The wires running along the center of the chassis go to the heater and grid choke of the final amplifier. The two that follow the chassis corner at the left are from the oscillator and buffer cathodes to the meter switch.

and the 3750–4000-kc. coil is similarly adjusted for sufficient range to give maximum tube input at 'phone ratings.

Amplifier cathode keying is shown in Fig. 6-49, but any method may be used with appropriate changes in the diagram. A lead is brought out from the "hot" end of the amplifier grid leak, R_5, so that the d.c. voltage developed by excitation may be used to control a screen protective tube if an earlier stage is keyed. The circuit constants in the oscillator and buffer stages in Fig. 6-49 are such that both these tubes can run without excitation, with a 300-volt plate supply, without exceeding the plate dissipation rating of either 6AG7. This permits keying the VFO when separate VFO input is used.

Shielded wiring for preventing harmonics from flowing on supply leads is indicated in the circuit diagram. These leads should be by-passed by midget ceramic condensers at the points indicated, using the technique described in the TVI chapter. The corresponding technique for high-voltage mica by-passes is used for the amplifier high-voltage plate lead.

All three tubes have parallel plate feed. This permits grounding the tank condensers directly to the chassis, which is advantageous both mechanically and electrically. In the buffer and amplifier stages parallel feed is a necessity because the pi networks cannot be series-fed.

Construction

All of the circuits with the exception of the buffer and amplifier coils are inside the chassis. The metal 6AG7s provide their own shielding. The 6146 mounts through the rear chassis wall and is covered by the same type of shield can (ICA No. 1549) as is used to cover the tank coils except that it is trimmed down a bit in length and is drilled with ⅛-inch holes above and below the tube to give ventilation. The location of the principal components is shown in the bottom view.

Since the space underneath the chassis is limited, some care must be used to fit the parts in. The best plan is first to lay out the complete transmitter and drill all holes in the chassis, making sure that everything is provided for before anything is permanently mounted. Make the partitions and amplifier tube mounting bracket and fit them in place before drilling any mounting holes for them in the chassis. Mounting holes in these pieces may then be used to locate the corresponding chassis holes. The tube socket bracket and final tank condenser together form a separate subassembly on which most of its wiring may be done, including the shielded cathode lead to the meter switch, after the mechanical fit has been checked. The bracket is drilled to clear the rear shaft extension of the condenser and uses holes already present in the condenser back plate for mounting. The plate blocking condenser, C_{10}, is mounted on the screw which is part of the stator plate assembly; this condenser must be as close as possible to the condenser so that it will clear the coil socket mounted on the rear chassis wall. A short stand-off insulator is mounted just to the left of the tube socket, at the left in the bottom view, to mount the plate lead and one end of the parasitic choke, L_4.

The center partition should have a ½-inch hole at the point where the amplifier grid lead comes through from the buffer stage, and should be cut out about ⅛ inch at the bottom where it must fit over the shielded wiring laid on the

Buffer and Amplifier Coil Table

Coils wound on 1½ inch diameter forms (National XR-4 and XR-5)

	Wire Size	*No. of Turns*	*Turns per Inch*	*L, uh.**
Buffer coil, L_5				
3.5 — 4 Mc.	26	42	28	48
7 Mc.	22	25	20	18.4
14 Mc.	18	10	10	3.5
21 Mc.	18	5	10	1.34
27 — 30 Mc.	18	3½	10	0.86
Amplifier coil, L_6				
3.5 — 3.75 Mc.	18	23⅓	16	14.5
3.75 — 4 Mc.	22	25⅓	20	18.7
7 Mc.	18	17⅓	12	8.3
14 Mc.	18	10⅓	8	3.25
21 Mc.	16	6⅓	5	1.36
27 — 30 Mc.	16	4⅓	5	0.84

* Measured values with coil unshielded.

chassis. These parts and the meter shield should be the last things mounted, after all other assembly and wiring has been completed.

The shielded wiring should be laid in first, as shown in Fig. 6-50. Soldering lugs may be used as hold-downs, the wire shield being spot soldered to each such lug. Start the leads, fitted with ceramic by-passes, at the output terminal strip or tube socket, as the case may be, and run them to their final locations, temporarily mounting the part at which they terminate to get the exact lead length. Then trim the wire and install the ceramic by-pass when called for in the diagram.

After the shielded wiring is in place, install the amplifier coil socket and wiring, leaving enough lead length to reach the tank condenser to be mounted later. This coil socket must be mounted with the ring *outside* the chassis in order to provide sufficient clearance for the amplifier-tube subassembly. Then complete the oscillator and buffer assembly and wiring, except that the buffer coil socket should not be mounted because it interferes with installing the amplifier subassembly. Also mount and wire the key jack and meter switch, including mounting and finishing shielded leads for the meter.

When this has been done the amplifier tube subassembly may be permanently installed and the connections to it completed. After installation the amplifier plate choke should be mounted, using the chassis hole for the 6146 for access. The buffer coil socket and amplifier output condenser, C_{13}, may then be installed and the wiring completed. The last operation is to mount the meter shield.

Since the size of some parts is critical, in view of the limited space, the specific components used in the unit shown are designated in the circuit caption.

Operation

The final amplifier is operated straight through on all bands and the buffer amplifier preferably, although not necessarily, is operated as a frequency multiplier. On bands where the buffer is used as a straight amplifier care must be taken to choose tuning conditions that do not permit self-oscillation in the buffer stage. On 3.5 Mc. with either crystal or VFO control there is no tendency for the buffer to self-oscillate because its grid circuit is not resonant at the operating frequency. On this frequency the principal precaution to be observed is that C_4 should be tuned so that the drive at *harmonics* of the input frequency is not excessive. The proper setting for C_4 is the one that results in maximum amplifier grid current when the buffer plate circuit is properly resonated.

When operating on 7 Mc., C_4 should be toward minimum capacitance, but not far enough to resonate at 14 Mc. Adjust for maximum amplifier grid current, with the buffer plate circuit resonated, by varying C_4 toward minimum capacity. When the amplifier grid current is maximum, pull out the crystal or shut off the VFO and the grid current should drop to zero. If it does not, decrease C_4 until it does. The grid current should be ample with C_4 set so there is no danger of buffer oscillation.

For 14-Mc. operation, set C_4 near maximum capacitance so that the buffer is driven on 7 Mc. and operates as a doubler. Adjust for maximum amplifier grid current. On 21 Mc., operate the buffer as a tripler, driving it on 7 Mc. and adjusting C_4 in the same way as for 14 Mc.

The preferable method of operation on 27–30 Mc. is to use a 7-Mc. crystal or VFO, adjust C_4 to resonate at 14 Mc., and then double in the buffer stage. In this case C_4 will be near minimum capacity. Alternatively, a 3.5-Mc. crystal or

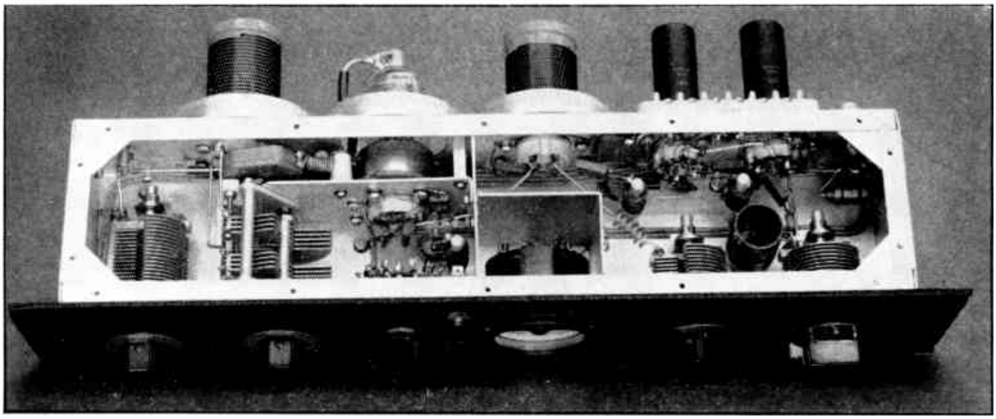


Fig. 6-51 — Bottom view of the transmitter completely wired. The oscillator plate coil, L_1, is between the two variable condensers at the right. The amplifier circuit occupies the left-hand portion of the chassis in this photograph. The chassis is 3 by 4 by 17 inch aluminum and is covered by a 4 × 17 aluminum bottom plate (not shown). The bracket on which the amplifier socket is mounted is supported at one end by the plate tank condenser and at the other by a partition that shields the amplifier section from the oscillator-buffer section. The amplifier plate choke is mounted on the chassis between the tube-socket bracket and the chassis wall, just below the plate-lead terminal. The meter is enclosed by a right-angle shield to prevent stray harmonic pick-up that might cause radiation through the meter hole in the panel.

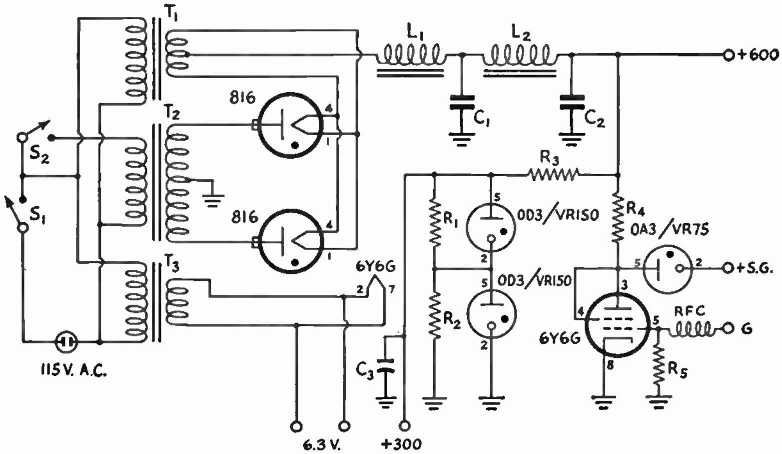


Fig. 6-52 — Power supply diagram for the 6146 exciter-transmitter. Note: For plate modulation, disregard the screen-supply circuit shown, and supply d.c. screen voltage through a 35,000-ohm 10-watt resistor connected from the "hot" end of the modulation transformer secondary to screen.

C_1, C_2 — 4-μfd. 1000-volt paper.
C_3 — 8-μfd. 450-volt electrolytic.
R_1, R_2 — 0.1 megohm, 1 watt.
R_3 — 4000 ohms, 25 watts.
R_4 — 25,000 ohms, 10 watts.
R_5 — 0.5 megohm, ½ watt.
L_1 — 5/25 henrys, 225 ma.
L_2 — 4.5 henrys, 200 ma.
T_1 — Filament transformer: 2.5 v., 4 amp., 1500-volt insulation.
T_2 — Plate transformer: 800 v. each side c.t., 225 ma.
T_3 — Filament transformer: 6.3 v., 6 amp.
S_1, S_2 — S.p.s.t. toggle.
RFC — 2.5 mh. r.f. choke.

VFO may be used, in which case the optimum method is to double in the oscillator plate circuit, the setting of C_4 being near maximum capacity, and use the buffer as a quadrupler. This results in higher amplifier grid current, in the average case, than can be obtained by quadrupling in the oscillator stage and doubling in the buffer. The grid drive for the final amplifier is less than when using 7-Mc. crystals or VFO, but is sufficient for operating the 6146 at maximum ratings on either c.w. or 'phone. Care must be used to select the right harmonic when quadrupling in the buffer, since the tuning range is sufficient to reach both 21 and 28 Mc. on the 28-Mc. coil. In all the preliminary tuning, it is excellent practice to check the actual frequency of each circuit, particularly the buffer plate circuit, with an absorption wavemeter.

With any of the types of operation described above, the maximum grid current through the 27,000-ohm amplifier grid resistor should be from 3 ma. to about 4.5 ma., with the amplifier fully loaded. These values are in excess of the normal operating figures, the optimum current being 2.5 to 3 ma. for c.w. operation and 1.8 to 2 ma. for plate-modulated 'phone. This is for a plate-supply voltage of 600, with a plate current of 150 ma. for c.w. operation and 113 ma. for 'phone.

The method of tuning the amplifier is the same on all bands. Assuming that the load has been adjusted to represent a pure resistance, or nearly so, of 50 to 75 ohms, set C_{13} to maximum capacitance, apply plate and screen voltage, and adjust C_{12} for minimum plate current. Then decrease the capacity of C_{13} by a small amount and reresonate C_{12}. Continue until the plate current at the minimum of the dip is the desired value. Since the off-resonance plate current of the 6146 may run as high as 250 ma. it is advisable to do preliminary testing at reduced plate and screen voltage, until the proper operating conditions have been once established.

If the load is not the type that is represented by a properly-terminated coax line it may or may not be possible to control the loading adequately by means of C_{13}. The pi network constants are fairly critical as to loading, and if proper loading cannot be secured it is an indication that the coax line is not flat.

Power Supply

The oscillator and buffer require a total current of approximately 50 ma. at 300 volts. In order to avoid the excessive plate dissipation that might occur with a supply that gives more than 300 volts, the plate voltage should be regulated by means of VR tubes. The plate currents taken by the oscillator and buffer do not vary greatly from band to band, the oscillator current being about 20 ma. on all bands and the buffer taking about 25 ma. on all except 7 Mc. where it is about 12.

The amplifier requires a 600-volt plate supply capable of an output current of 150 ma., approximately. The screen current averages about 12 ma. through a dropping resistor of 35,000 ohms, the optimum value. A suggested power supply circuit is given in Fig. 6-52. This utilizes a single plate transformer designed to deliver 600 volts at 225 ma. through a choke-input filter.

Compared with other beam tetrodes, the 6146 operates with quite low screen voltage and the ordinary screen protective tube circuit does not reduce the screen voltage to a low-enough value to prevent excessive plate dissipation when there is no r.f. excitation. The circuit shown here consequently includes a VR-75 to cut off the screen voltage under such conditions. To compensate for the voltage drop through the VR tube the screen resistor is reduced to 25,000 ohms.

(Originally described in *QST*, Feb., 1952.)

A 500-Watt Multiband VFO Transmitter

Figs. 6-53 through 6-59 show the circuit and other details of a 500-watt transmitter with VFO frequency control, capable of operation in any band from 3.5 to 28 Mc. It is completely shielded and all tuning adjustments, including band changing, may be done with the panel controls.

As the circuit of Fig. 6-56 shows, the VFO uses a 5763 in a Clapp circuit operating over a range of 3370 to 4000 kc., split into three bandspread ranges, tuned by C_1, which is fitted with a calibrated dial. These ranges, selected by proper setting of C_2, are 3500 to 3750 kc., 3370 to 3405 kc. (for 11-meter operation) and 3750 to 4000 kc. for 75-meter 'phone work.

The oscillator circuit is followed by two isolating stages. The first is a 6C4 connected as a cathode follower, which is very effective in reducing reaction on the oscillator by subsequent stages. The result is a keyed VFO with good characteristics, even on 10 meters. Since the output of the cathode follower is quite small, it is followed by a 5763 in an amplifier fixed tuned in the 3.5-Mc. region.

Frequency multiplying to reach the higher-frequency bands is done in the next two stages, the first using a 5763, while the second employs the larger 6146 to drive the final amplifier. These two stages are tuned with multiband tuners — circuits which have a tuning range that includes all necessary bands. Thus no switching or plug-in coils are needed. Neither of these two stages is operated as a straight amplifier, except on 80 meters. Frequency is doubled in the 6146 stage for output on 40, 20 and 10 meters, and tripled for output on 15 meters. The 5763 stage is operated at 3.5 Mc. for 80- and 40-meter output, doubles to 7 Mc. for 20- and 15-meter output, and quadruples to 14 Mc. for 10-meter output. Excitation to the final is adjusted by the potentiometer in the screen circuit of this stage.

The 813 in the final amplifier also uses a multiband tuner to cover all bands. This stage is always operated as a straight amplifier, and should be entirely stable without neutralization. The only switching necessary is in the output link circuit in changing between high- and low-frequency bands. Loading is adjusted by C_{10}.

A 50-ma. meter may be switched to read plate current in the exciter stages, grid current in the driver and final-amplifier stages, or screen current to the 813. The ½-ohm resistor in the 6146 high-voltage lead multiples the meter-sale reading by three. A separate 500-ma. meter is used to check plate current to the 813.

The two-circuit rotary switch, S_1, is used to bias the screens of the 6146 and 813 negative while tuning up the preceding stages and setting the VFO to frequency. In the first position, both screens are biased; in the second position, only the 813 screen is biased, while positive voltage from a voltage divider is applied to the screen of the 6146 so that this stage may be tuned up. In the third and fourth positions, positive voltage is applied to both screens, but in the last position, it is applied to the 813 screen through an audio choke so that the stage may be screen-plate modulated.

Two bias rectifiers are included in the unit, to supply fixed bias to the 6146 and 813, so that the plate currents will be cut off during keying intervals. Both rectifier systems operate from a single 6.3-volt filament transformer connected in reverse. The bias transformer, T_2, is operated from the 6.3-volt winding of the filament transformer, T_1.

Two a.c. outlets are provided for connecting the primaries of external high- and low-voltage supplies into the control circuit consisting of three toggle switches. B_1 is the ventilating blower that starts operating as soon as the filament switch is closed. The blower is essential where so much power is confined in a small space. The jack, J_3, provides a means of keying the final amplifier, rather than the oscillator, if desired. It also permits plugging in a simple cathode modulator of the type described in the chapter on speech amplifiers and modulators.

It is highly important that the VFO box make good contact with the chassis; otherwise the VFO may be adversely affected by feed-back from the

Fig. 6-53 — The standard-rack panel is 12¼ inches high. Controls (National HRS) along the bottom, centers spaced at intervals of 2⅛ inches either side of center, are, left to right, for C_4, S_3, C_5, C_2, S_1 (Centralab 1405), S_2 and C_{10}. Power toggles are below at the center, spaced 1 inch apart. The calibrated VFO dial (National SCN) for C_1 is at the center, with the excitation control (National P dial) to the left, and the dial (National AM) for C_9 to the right. National CFA chart frames outline the rectangular openings for the recessed meters, 50-ma. to the left, 500-ma. to the right. The shielding enclosure is built up using aluminum angle, perforated sheet (also used for the bottom plate), and self-tapping screws.

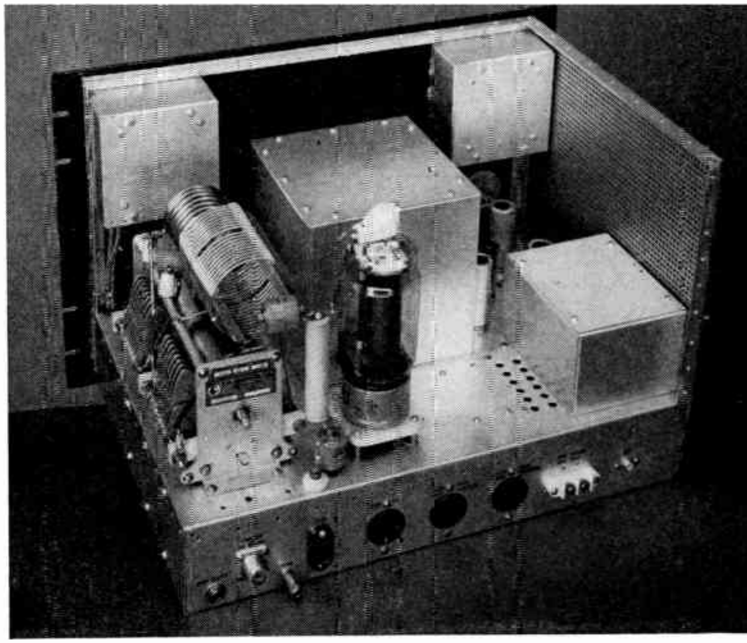


Fig. 6-54 — The components are assembled on a 17 × 12 × 3-inch aluminum chassis. The meters are housed in 4 × 4 × 2-inch boxes, the VFO enclosure is 6 × 6 × 6, while the box enclosing L_3 and L_4, to the right, measures 3 × 4 × 5 inches. The special plate choke, RFC_1, to the left of the 813, is close-wound with 129 turns No. 26 d.c.c. wire, on a Millen 31004 1 1/16-inch ceramic pillar. C_8 is fastened to the top of the choke, while C_7 is mounted below near the h.v. feed-through. (Both C_7 and C_8 are Sprague 20DK-T5.) The small cones, fastened to the condenser frame by drilling holes in the assembly rods, support L_9. A screw, tapped into the same rod, anchors the grounded end of L_7, whose outer end connects to the rear stator terminal below. The 813 socket is mounted on ½-inch pillars, over a 2¼-inch hole in the chassis. Along the rear apron are J_3, J_2, + h.v. (Millen 37001) and ground terminals, a.c. power-input connector, two a.c. outlets, low-voltage input terminals, and key connector.

◆

adjacent final tank when working on 80 meters. Mounting screws spaced an inch around the bottom lip of the box, and correspondingly in the top cover, should eliminate this completely.

Coils

L_1 (35 μh.) is a B & W 80-BCL coil with the link and base removed. L_2 is given under Fig. 6-58. L_3 (2.6 μh.) is 31 turns of B & W 3003 miniductor, while L_4 (5.3 μh.) is 30 turns of Type 3011. L_5 (1.5 μh.) consists of 11 turns of No. 16, ¾-inch diameter, 1 3/16 inch long. L_6 (8.9 μh.) has 29½ turns of B & W 3015 miniductor. L_9 (1.6 μh.) has 7 turns of ¼-inch copper tubing, 2 inches diameter, 2⅛ inches long.

L_7 (5.1 μh.) and L_8 (4.2 μh.) are made as follows from B&W 3905-1 strip coil: Count off 10¼ turns, clip the wire without breaking the support bars. Bend the last quarter turn out. This portion is L_7. Remove the next ¾ turn to make a ¼-inch space between L_7 and L_8. Count off 10 turns more, cut the remainder of the coil stock off. Unwind the last turn on L_8 to make the necessary lead to the stator of C_9. Tap L_8 at the 8th turn from L_7.

Adjustment

A 400-volt 250-ma. supply is required for the exciter and the screen of the final amplifier. For full rated output from the 813, a supply delivering 2000 to 2200 volts at 300 ma. (including bleeder current) is needed. The amplifier may, of course, be operated at lower plate voltage with less power input. The diagram of a suitable power unit is shown in Fig. 6-59.

The VFO tuning ranges should first be adjusted. Set S_1 to the first position, biasing the screen of the 6146. Adjust the screen potentiometer in the 5763 multiplier stage to zero, and turn on the filaments and the low-voltage supply. Set C_1 at 95 degrees on the dial (near minimum capacitance). Set C_2 accurately at midscale. Then, listening on a calibrated receiver, adjust C_3 until the VFO signal is heard at 3750 kc.

Now, tune the receiver to 3500 kc., and turn C_1 toward maximum capacitance until the VFO signal is heard. This should be close to the lower end of the dial. By carefully bending the rearmost rotor plate of C_1 toward the rear, it should be

◆

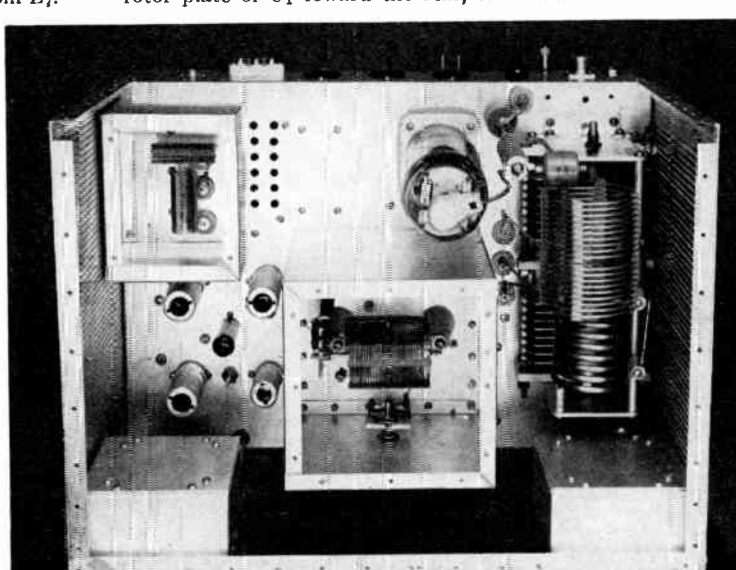


Fig. 6-55 — The VFO box is placed with its front wall 1 3/16 inches back of the panel, central on the chassis. L_1 is mounted on 2-inch cones to center it in the box. The shaft of C_1 (Cardwell PL-6001 minus last stator plate) is central on the box front, at a height to match that of C_9. C_2 (Cardwell PL-6002) is mounted, between C_1 and the coil, shaft downward, to engage the right-angle drive below. C_3 (Cardwell PL-6009) is similarly mounted, to the left of C_2. Grouped to the left are V_4, L_2, and V_3 in front, with V_5 and V_1 to the rear, and V_2 in the center. Feed-throughs in the bottom of the coil box to the rear connect L_3 and L_4 to C_4 below. The ventilating holes are over the 6146. C_9 (Johnson 200DD35) is placed with its shaft 2¼ inches from the end of the chassis, and its rear end plate 1⅝ inches in from the back edge. The three feed-throughs to the left connect L_8 to S_2.

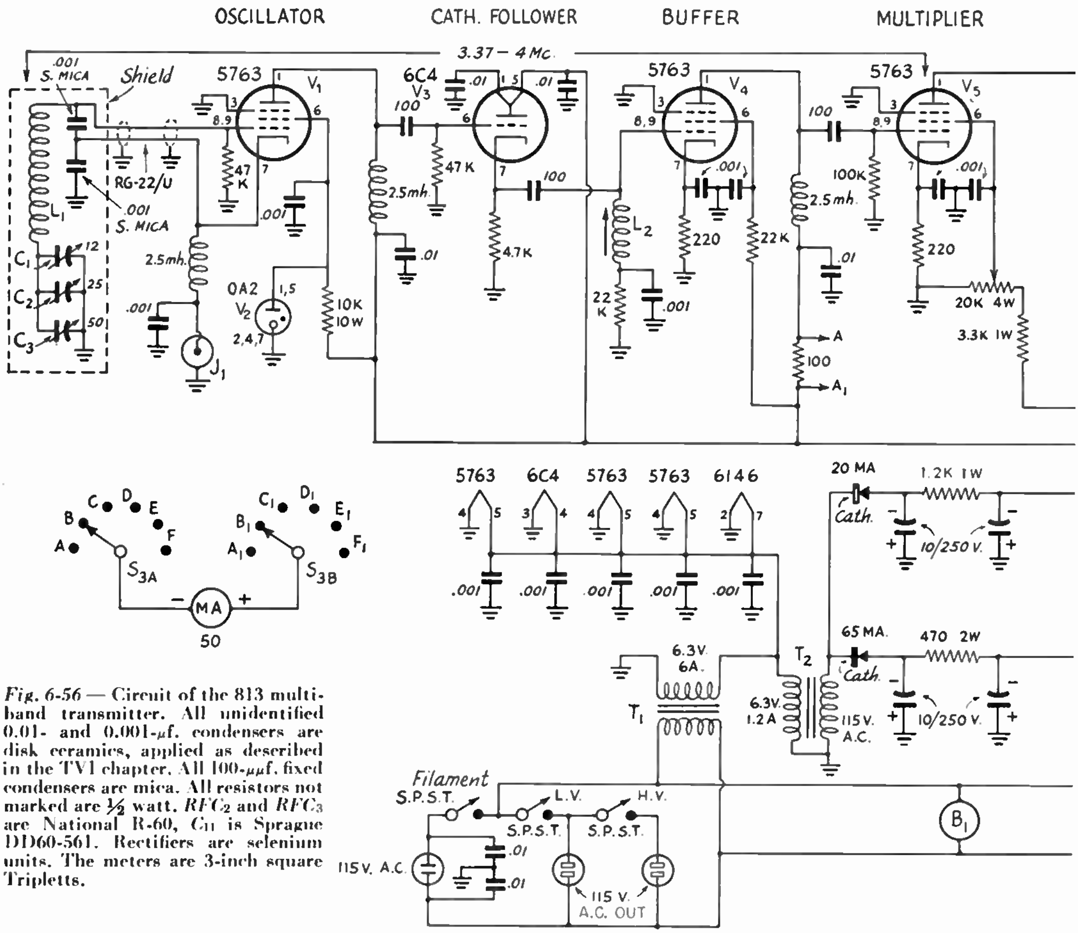


Fig. 6-56 — Circuit of the 813 multi-band transmitter. All unidentified 0.01- and 0.001-μf. condensers are disk ceramics, applied as described in the TVI chapter. All 100-$\mu\mu$f. fixed condensers are mica. All resistors not marked are ½ watt. RFC_2 and RFC_3 are National R-60, C_{11} is Sprague DD60-561. Rectifiers are selenium units. The meters are 3-inch square Tripletts.

possible to adjust the range of 3500 to 3750 kc. so that it covers from 5 to 95 degrees on the dial. Some slight readjustment of C_3 may be necessary during the plate-bending process to keep the band centered on the dial.

Now, set C_1 at about 15 degrees. Set the receiver at 3750 kc. and reduce the capacitance of C_2 until the VFO signal is heard. Then, tuning the receiver to 4000 kc., the VFO signal should be heard when its dial is set at about 85 degrees. Mark this setting of C_2 accurately.

If it is desired to center the 11-meter band on the dial, set C_1 at midscale. Increase the capacitance of C_2 until the VFO signal is heard at 3387 kc. Mark this setting of C_2 also accurately.

Tuning Chart for the 813 Transmitter

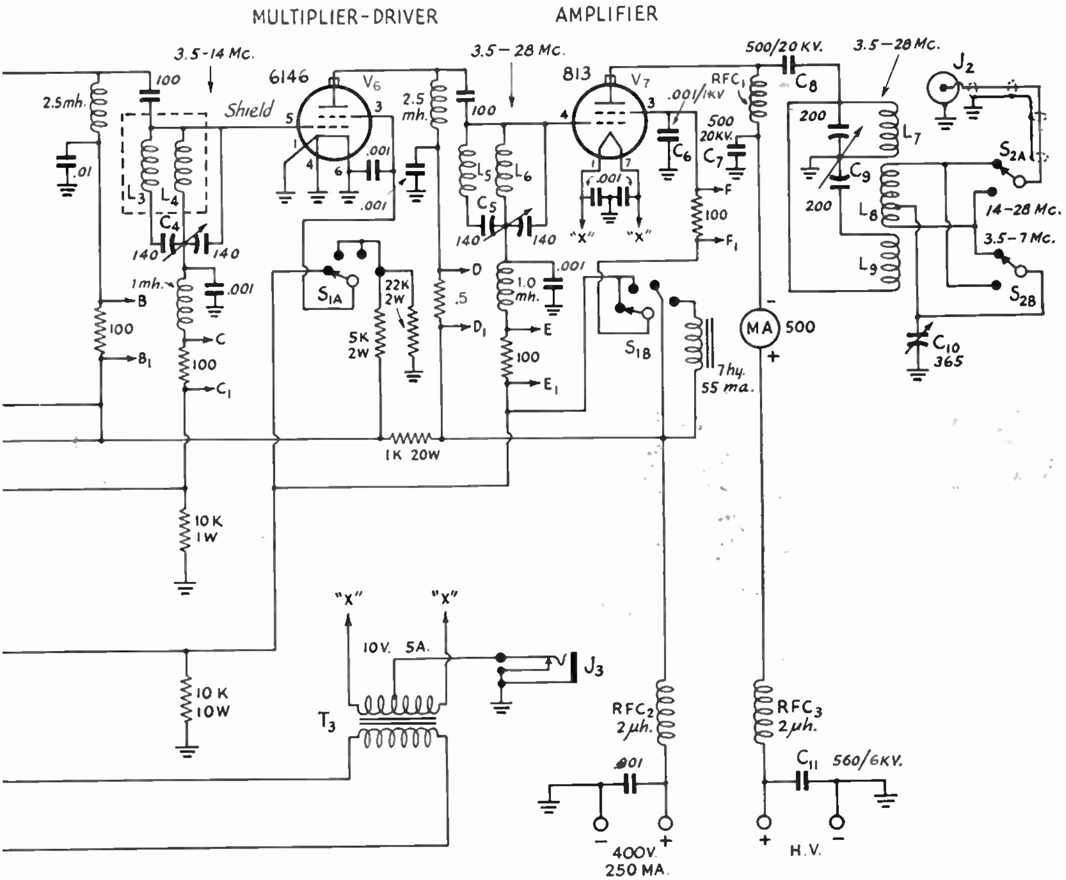
Output Band (Mc.)	*C₄ Dial*[1]	*Band (Mc.)*	*C₆ Dial*[1]	*Band (Mc.)*	*C₉ Dial*[2]
3.5	8.8	3.5	6.1	3.5	77
7	8.8	3.5	0.5	7	9
14	1.5	7	9.5	14	82
21	1.5	7	3.7	21	26
27–28	4.7	14	1.8	28	7

[1] 10-division dial — 10 max. capacitance.
[2] 100-division dial — 100 max. capacitance.

The next step can be done most easily with a high-resistance voltmeter connected across the grid leak of the 5763 buffer amplifier. Set C_1 and C_2 at minimum capacitance, and adjust the slug in L_2 for maximum grid voltage. Then watch the grid voltage as C_2 is swung through its range. If there is appreciable increase in grid voltage as C_2 is turned toward maximum capacitance, tune L_2 to a higher frequency by moving the slug out more. By correct adjustment of the slug, the grid voltage should remain essentially constant over the entire usable frequency range.

Now readjust C_2 to midscale and turn the meter switch to read 6146 grid current, and turn the excitation control to give a reading of 2 or 3 ma. Resonate the output tank circuit of the 5763 frequency multiplier at 80 meters (near maximum capacitance) as indicated by maximum 6146 grid current.

Next, turn S_1 to the second position, so that screen voltage is applied to the 6146, but not to the 813. Turn the meter switch to read 6146 plate current, and resonate the 6146 output tank circuit as indicated by the plate-current dip (near maximum capacitance). Turning the meter



switch to read 813 grid current adjust the excitation control to give a final-amplifier grid-current reading of about 25 ma.

The 813 should be tested initially at reduced plate voltage. Plate voltage can be reduced by inserting a 150-watt lamp in series with the high-voltage transformer primary. A 300-watt lamp bulb connected across the output connector can be used as a dummy load for testing. Make sure that S_2 is turned to the low-frequency position. This position is used for 3.5- and 7-Mc. operation. The other position is used for 14, 21 and 28 Mc. Turn S_1 to the third position to apply screen voltage to the 813, apply plate voltage and resonate the output tank circuit (near maximum capacitance) as indicated by a dip in plate current. Full plate voltage may now be applied and C_{10} adjusted to give proper loading (220 ma. maximum). Adjust the excitation control to give a final-amplifier grid current of 15 to 20 ma.

Tuning up on the other bands is done in a similar manner, by adjusting the tuners in each circuit to the correct band to obtain the desired multiplication. The table shows the approximate dial setting for each band, but each should be checked with an absorption wavemeter and the setting logged for future reference.

A suitable antenna tuner should be used between the transmitter output and the antenna. Antenna tuners are described in the chapter on transmission lines.

(Originally described in *QST* for January, 1954.)

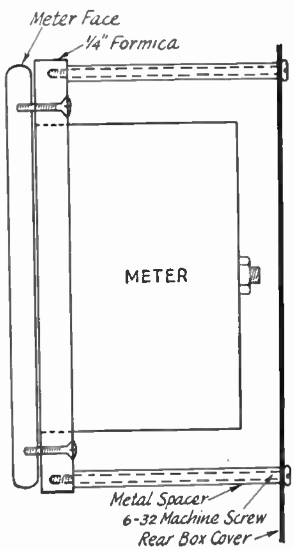


Fig. 6-57 — Sketch showing method of mounting meters in shielding boxes. The meters are suspended from the rear covers.

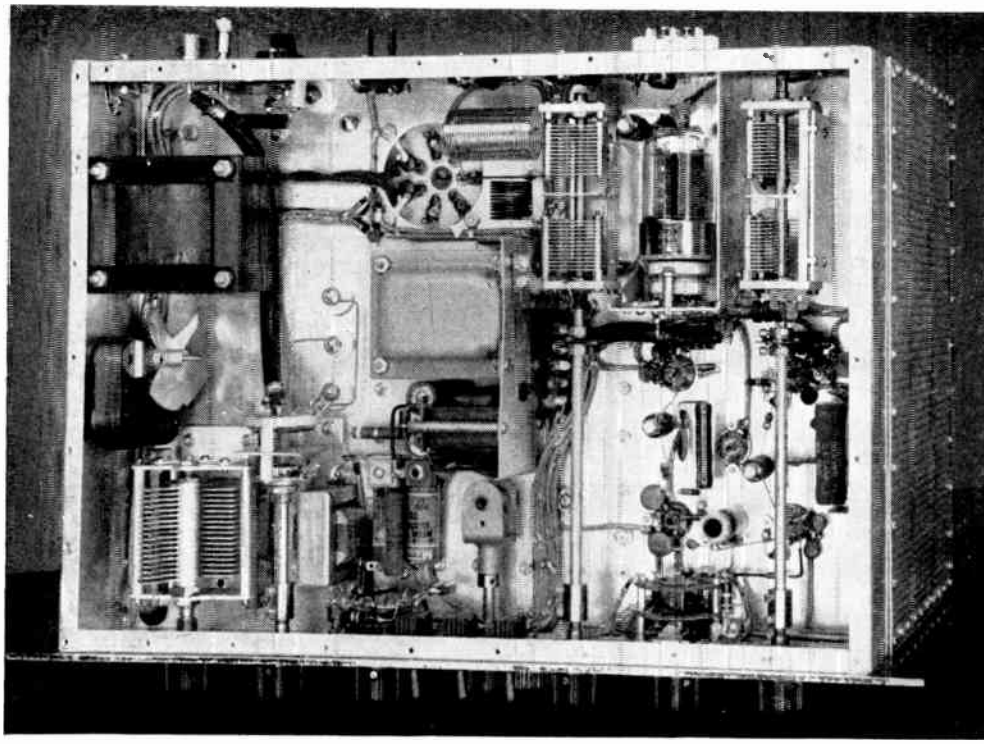
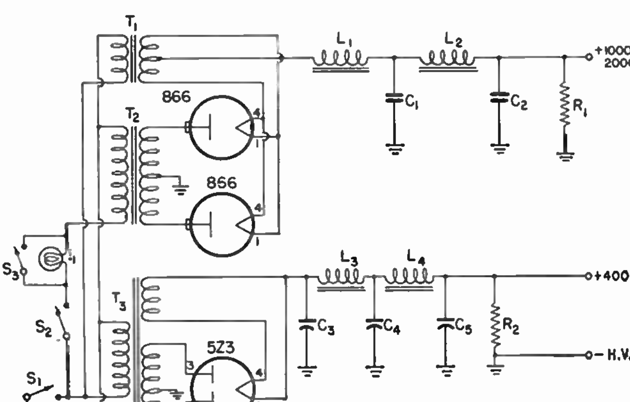


Fig. 6-58 — The panel drops 3/16 inch below the bottom edge of the chassis. The National ACD-1 right-angle drive for C_2 is at the center. The other controls along the bottom are placed 1½ inches up from the bottom edge of the chassis, and the corresponding components mounted so that their shafts line up with the controls. Panel bushings should be provided for the shafts of C_{10} (Cardwell PL-7006), and the right-angle drive; panel-bearing shaft units for C_4 and C_5 (Cardwell PL-6043), and S_2 (Centralab 2505). The 6146 is mounted on a 5 × 2¼-inch bracket between C_4 and C_5, whose shafts are fitted with insulating couplings. C_5 is mounted on spacers, while C_4 is mounted on its side on a bracket. T_1 (Triad F-18A) and T_2 (Triad F-14X) are mounted on another bracket at the center. L_5 and L_6, at right angles, are soldered between the terminals of C_5 and Pin 4 of the 813 socket, seen through the 2¼-inch hole in the chassis. C_{10} and S_2 are mounted on small brackets. T_3 (Triad F-23U) and the blower (available from Allied Radio, Chicago, No. 72-702 motor and 72-703 fan) are to the left. The screwdriver-slotted shaft of C_3 may be seen between the shaft of C_5 and the shielded power wires to the left. All power wiring is done with shielded wire (Belden 8656, Birnbach 1820, or shielded ignition wire for the 2000-volt line; Belden 8885 for the rest). L_2, behind S_3 (Centralab 1411), is a National XR-50 slug-tuned form close-wound with 93 turns No. 36 enameled wire.

Fig. 6-59 — Circuit of a suitable power supply for the 813 transmitters.



C_1, C_2 — 4-μfd. 2000-volt oil-filled.
C_3, C_4, C_5 — 4-μfd. 600-volt electrolytic.
R_1 — 25,000 ohms. 200 watts.
R_2 — 15,000 ohms, 25 watts.
L_1 — 5/25-h. 300-ma. swinging.
L_2 — 20-h. 300-ma. smoothing.
L_3, L_4 — 7-h. 250-ma. filter choke.
I_1 — 150-watt lamp (Tune up)
S_1, S_2 — 10-amp. switch.
S_3 — 3-amp. switch.
T_1 — 2.5 volts, 10 amp.
T_2 — 2000 volts d.c., 300 ma.
T_3 — 400-0-400 r.m.s., 250 ma.; 5 volts, 3 amp. (UTC S-40)

A 200-Watt Transmitter for 160 and 80 Meters

Figs. 6-60 through 6-64 show circuits and constructional details of a 200-watt transmitter designed primarily for the 160-meter band. However, it will also work well doubling frequency to the 80-meter band in the output stage.

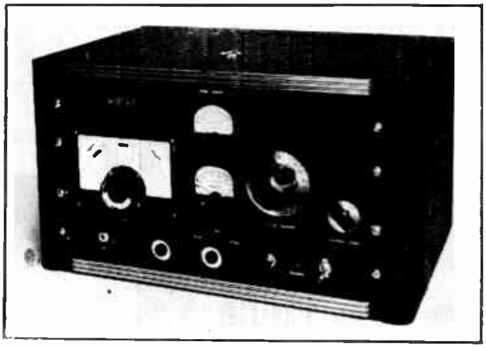


Fig. 6-60 — A front view of the 160-meter transmitter designed by W1TRF, showing the panel layout. The VFO is directly calibrated for 160 and 80 meters on the National SCN dial. The lower row of controls are, left to right, keying jack, buffer plate tuning, meter switch and the filament switch. To the right of the two meters are the final plate-tuning and the swinging-link controls.

Circuit

The circuit is shown in Fig. 6-62. A 6AG7 is used in the series-tuned VFO which works on 160. The oscillator plate circuit, which is untuned, is capacity coupled to another 6AG7 in the buffer stage. Cathode bias is supplied to the buffer stage by R_3. The buffer screen voltage is taken from the regulated source that supplies the VFO section. The buffer operates straight through and is coupled to the final-amplifier grid by C_{14}. An 813 was chosen because of its low drive requirements and its adaptability to a wide range of plate voltages — it is possible to run an input of 200 watts with a plate voltage as low as 1200. The stage is neutralized by means of a simple home-made condenser, C_{17}. The conventional neutralizing connection, shown in dotted lines, was not used in this instance. Stray wiring capacitances are such that the circuit is "over-neutralized," requiring the introduction of positive, instead of negative, feed-back for neutralization. Therefore, the neutralizing capacitance is directly from grid to plate. However, the use of different components, or a slightly different layout, may require the conventional connection shown in dotted lines, rather than the one used.

Fixed bias is supplied to the final amplifier by a 50-ma. selenium rectifier and a small filament transformer, T_2, working in reverse from the 6.3-volt filament supply. A VR-150 is used to stabilize the biasing voltage. Screen voltage is supplied from the high-voltage source through R_8 and R_9 to provide a simple means of modulating both plate and screen.

Construction

The transmitter is constructed entirely on a standard 10 × 17-inch chassis with a 10½-inch panel. The VFO portion is built on the left-hand side of the chassis. The 6AG7 socket is inverted so that the tube extends below the chassis. This method allows all of the wiring on the socket to be enclosed within the shield. C_3, C_4, C_5 and the grid resistor, R_1, are all soldered directly to the socket, and the filament by-pass condensers, C_{23} and C_{24}, as well as the screen by-pass condenser, C_7, are soldered directly to ground from their respective pins. Shielded power wires are brought into the compartment through rubber grommets. The r.f. plate lead to the coupling condenser, C_8, is made of a short piece of RG/59-U coaxial cable and this also is brought up through the chassis along with the power leads. L_1, the VFO coil, is close-wound on a 1-inch Millen form and is mounted on a half-inch cone insulator. The ends of the winding are soldered directly to their connections. Two half-inch spacers are used to hold the VFO tuning condenser, C_2, above the chassis so as to line the shaft up with the drive mechanism of the National SCN dial. The oscillator padder, C_1, and its mounting bracket are bolted firmly to the chassis. A 3 × 4 × 5-inch aluminum utility box is used to cover the VFO circuit. A small opening cut in the front cover allows the tuning dial to turn freely.

The oscillator plate choke, RFC_2, and the buffer grid choke, RFC_3, are mounted vertically. The choke terminals are used as tie points for the coupling condenser, C_8, and the buffer grid re-

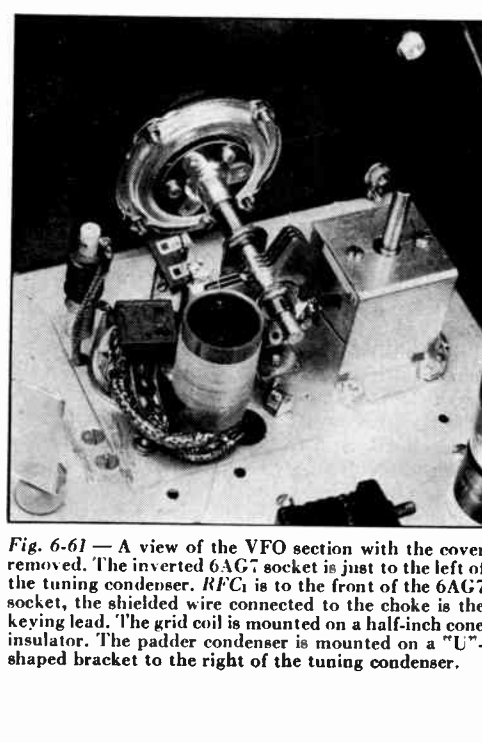


Fig. 6-61 — A view of the VFO section with the cover removed. The inverted 6AG7 socket is just to the left of the tuning condenser. RFC_1 is to the front of the 6AG7 socket, the shielded wire connected to the choke is the keying lead. The grid coil is mounted on a half-inch cone insulator. The padder condenser is mounted on a "U"-shaped bracket to the right of the tuning condenser.

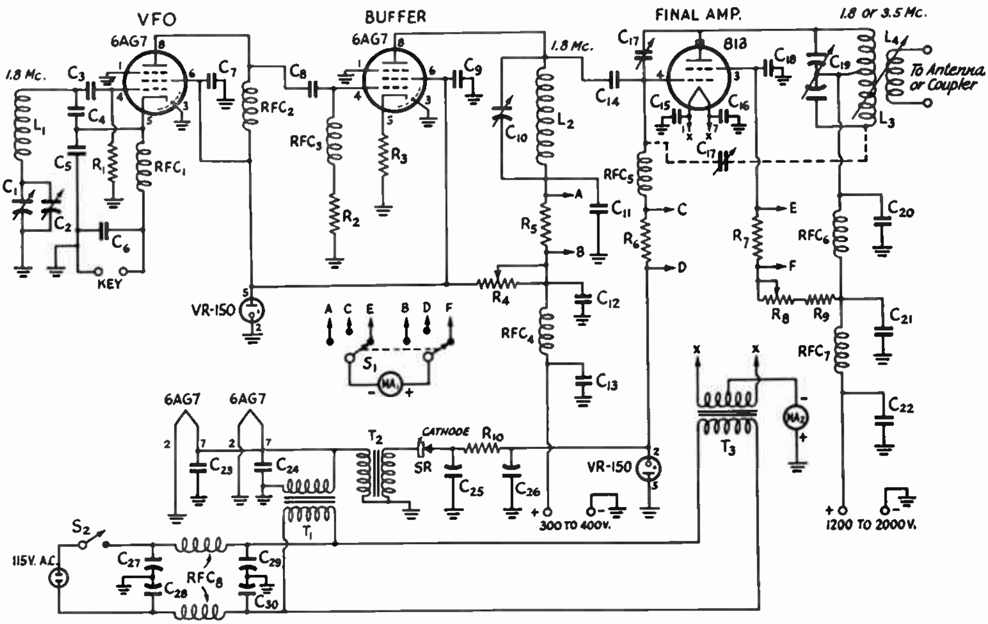


Fig. 6-62 — Circuit diagram of the 200-watt 160-meter transmitter.

C_1 — 100-μμfd. variable (Millen 22100).
C_2 — 50-μμfd. variable (Millen 19050).
C_3, C_8, C_{14} — 100-μμfd. mica.
C_4, C_5 — 680-μμfd. silvered mica.
C_6, C_7, C_9, C_{11}, C_{12}, C_{13}, C_{15}, C_{16}, C_{23}, C_{24} — 0.01-μfd. disc ceramic.
C_{10} — 140-μμfd. variable (Millen 19140).
C_{17} — Neutralizing capacitance: see text.
C_{18}, C_{20}, C_{21}, C_{22} — 0.001-μfd. 5000-volt mica.
C_{19} — Dual-section variable, 200-μμfd.-per-section (National TMC-200-D).
C_{25}, C_{26} — 8-μfd. 250-volt electrolytic.
C_{27}, C_{28}, C_{29}, C_{30} — 0.1-μfd. molded.
R_1, R_2 — 22,000 ohms, ½ watt.
R_3 — 220 ohms, 1 watt.
R_4 — 10,000 ohms, 10 watts, adjustable.
R_5, R_6, R_7 — 100 ohms, 2 watts.
R_8 — 25,000 ohms, 50 watts, adjustable.
R_9 — 25,000 ohms, 50 watts.
R_{10} — 500 ohms, 2 watts.
L_1 — 100 μh. — 68 turns No. 30 s.s.c. close-wound on 1-inch form.
L_2 — 70-μh. — 96 turns No. 24, 1-inch diam., 3 inches long (B & W 3016 Miniductor).
L_3 — 1.8 Mc. — 90 μh. — 56 turns No. 16, 3-inch diam., 6 inches long over-all, ¾-inch space at center for L_4 (B & W 160TVH or TVL with mounting for plug-in link).
3.5 Mc. — 40 μh. — 38 turns No. 14, 3-inch diam., 6 inches long over-all, ¾-inch space at center for L_4 (B & W 80TVH or TVL).
L_4 — 5-turn variable link (B & W 3555).
MA_1 — D.c. milliammeter, 50-ma. scale.
MA_2 — D.c. milliammeter, 500-ma. scale.
RFC_1, RFC_2, RFC_3, RFC_5 — 2.5-mh. r.f. choke (National R-100-S).
RFC_4, RFC_7 — 7-μh. r.f. choke (Ohmite Z-50).
RFC_6 — 1-mh. r.f. choke (National R-152).
RFC_8 — Line-filter choke (Ohmite Z-21).
S_1 — Single-wafer double-pole 3-position ceramic rotary.
S_2 — S.p.s.t. toggle.
SR — 50-ma. selenium rectifier.
T_1 — 6.3-volt 3-amp. filament transformer (Stancor P-5014 or equiv.).
T_2 — 6.3-volt 1.2-amp. filament transformer (Stancor P-6134 or equiv.).
T_3 — 10-volt 5-amp. filament transformer (Stancor P-6139 or equiv.).

sistor, R_2. The buffer tuning condenser, C_{10}, is mounted directly in front of the tube socket on the vertical bracket supplied with the condenser. A B & W 3016 Miniductor has just about the right inductance for L_2.

The 813 socket is mounted directly on the chassis to the right of the buffer-tube socket, with the coupling condenser, C_{14}, placed so that the leads are as short as possible. RFC_5, the 813 grid choke, is in front of the tube socket, near the grid-meter shunt. The meter shunting resistors for the buffer plate and the 813 grid circuits are fastened to a pair of two-terminal lug strips. The 813 screen-current shunt is mounted on two small cone insulators and is connected with high-voltage insulated wire, since the screen voltage rises to the supply value when the tube is not being driven. All external power leads have v.h.f. filters. The components are placed in the enclosure formed by the aluminum barrier shield running the length of the chassis.

The neutralizing "condenser," C_{17}, consists of a strip of aluminum about a half inch wide and 2 or 3 inches long, bent at right angles and mounted on a feed-through insulator near the socket grid terminal. The feed-through is connected to the grid terminal and neutralizing is adjusted by altering the length of the strip or by bending it closer to, or farther from, the tube.

The output tank condenser, C_{19}, is mounted above the chassis on half-inch cone insulators. The shaft is connected to the tuning dial through a ceramic-insulated shaft coupling. The jack bar for L_3 is supported on National GS-1 pillar insulators and mounted alongside the tank condenser. Another insulated shaft coupling is used to extend the shaft of the swinging link to the panel. A length of coaxial cable is run from the link assem-

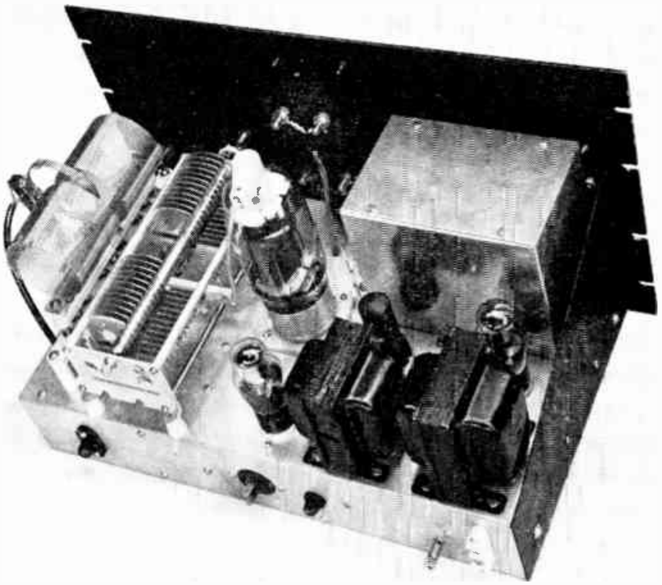


Fig. 6-63 — Top view of the 160-meter chassis removed from the cabinet. On the rear edge of the chassis are the two filament transformers and the VR tube for the bias supply. T_2 is underneath. In front of the transformers are the 6AG7 buffer tube, the VR-150 regulator for the VFO and the aluminum box shielding the oscillator section. To the left of the 813 are the final tank condenser and the swinging-link assembly. Along the rear of the chassis are the high-voltage connector, the 115-volt input connector, the grounding post and the exciter low-voltage connector.

bly to the antenna terminal along the left drop of the chassis.

The shielding barrier is spaced 3 inches from the rear. This enclosure contains all of the a.c. wiring, the line chokes and the bias supply. The high voltage to the final is routed through a feed-through in the shield. L_2 is cemented between two ceramic cone stand-off insulators on the other side of the barrier.

The circuit of a suitable power supply for this unit is shown in Fig. 6-59. A power transformer having a rating of 700 volts, c.t., 70 ma. may be substituted for the one specified under T_3. S_1 turns on the low-voltage supply and the filaments of the high-voltage rectifiers. S_2 turns on the high-voltage transformer. When S_3 is open, a 115-volt lamp, I_1, is connected in series with the primary of the high-voltage transformer to reduce voltage during adjustment.

Adjustment

After turning on the low-voltage supply, the slider on R_4 should be adjusted to the point where the VR tube just stays ignited with the key closed. At resonance, the buffer plate current should be about 22 ma. and screen current approximately 8 ma. This should produce an 813 grid current of 18 or 20 ma. When the key is opened, the buffer plate current should drop to about 12 ma. while the screen current is reduced almost to zero. If there is any variation in buffer plate current as the tank circuit is turned through its range with the key open, a check should be made for parasitic oscillation, as discussed earlier in this chapter.

In tuning up the final amplifier, the screen resistor, R_8, should be adjusted to leave about 20,000 ohms in the circuit and quarter or half maximum plate voltage applied. A dummy load should be connected and the output tank tuned to resonance. As the load is adjusted to take current, the plate and screen voltages can be increased slowly while checking the stability. For normal operation at maximum legal input, the screen voltage is raised to 350 and the plate voltage to 1200 or 1250. The coupling to the antenna or load can then be adjusted, by means of the variable link, to bring the power input up to 200 watts.

In the case of 80-meter operation, it may be of some advantage to raise the screen voltage to 400.

(Originally described in *QST* for July 1952.)

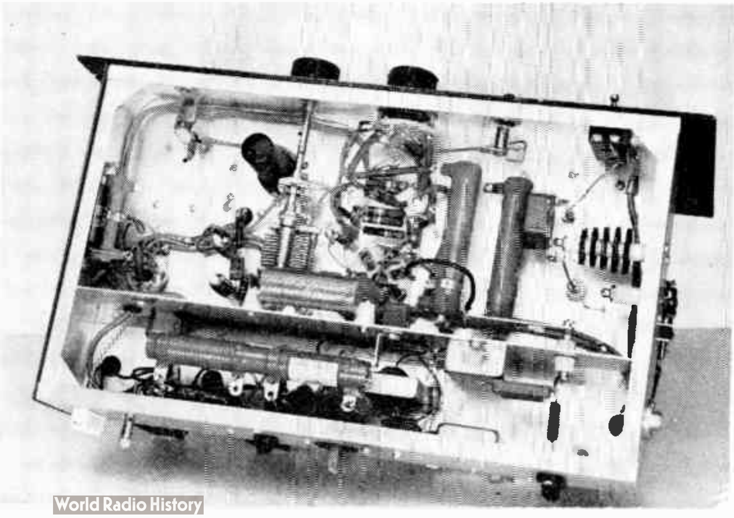


Fig. 6-64 — Bottom view of the 160-meter transmitter. R_4 is to the left. The inverted 6AG7 oscillator tube is just to the left of the buffer tuning-condenser shaft. In front of the 813 socket are the meter-shunting resistors and the meter switch. R_8 and R_9 are to the right of the 813 socket. The final plate choke is mounted on the right drop of the chassis. All power wiring is done with shielded wire to suppress v.h.f. harmonics.

A Simple VFO

The details of a simple VFO with output at 1.75, 3.5 or 7 Mc. are shown in Figs. 6-65 through 6-69. In the circuit, shown in Fig. 6-68, a Type 5763 miniature pentode in a series-tuned Colpitts oscillator circuit drives a similar tube as an amplifier or doubler. The output circuit of the oscillator stage is broadbanded through the use of self-resonant slug-tuned coils at L_2, and frequency may be doubled in this circuit, as well as in the output circuit, to obtain 7-Mc. output. For 3.5-Mc. output, frequency may be doubled in either stage. The nominal output is approximately 2 watts — sufficient for driving the usual crystal-oscillator stage of the transmitter.

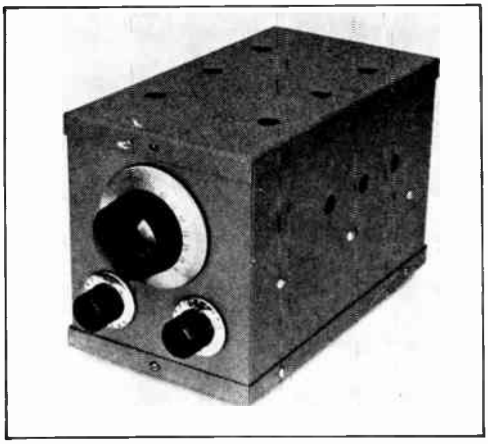


Fig. 6-65 — A simple VFO delivering output at 1.75, 3.5 or 7 Mc.

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

Construction

The unit is built in a 5 × 6 × 9-inch steel box with cap-type covers. The components are assembled on an aluminum-sheet base supported by sections of aluminum angle stock that hold the base halfway between the two covers. On top, the tuning condenser, C_1, is fastened directly to the base along the center line. The shaft is fitted with a National Type AM vernier dial. The two tubes and L_2 are in line to the right in Fig. 6-66 with the output tank coil, L_3, to the left of the amplifier tube. The L_2 coils are wound on Millen Type 74001 shielded slug-tuned forms.

Underneath, in Fig. 6-67, the band-set condenser, C_2, is mounted against the front of the box. A short lead through a feed-through point or clearance hole connects the stator of C_2 to the stator of C_1 above. L_1 is wound on a Millen 1-inch coil form and is placed immediately to the rear of C_2. The output tank condenser, C_{14}, is mounted on a bracket with its rear stator termi-

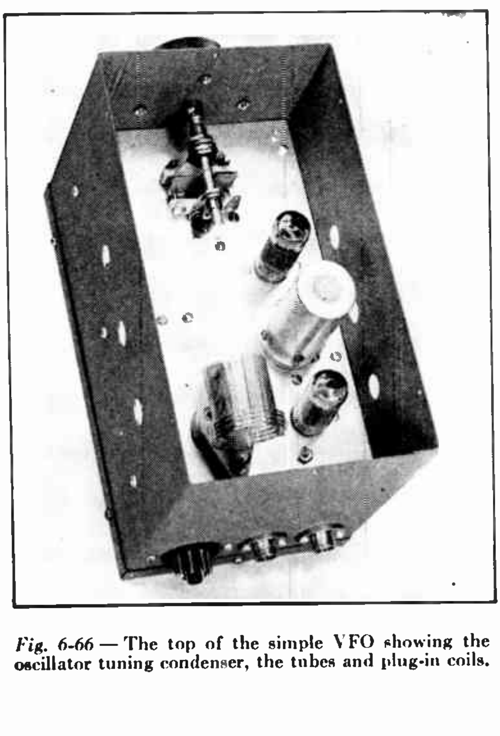


Fig. 6-66 — The top of the simple VFO showing the oscillator tuning condenser, the tubes and plug-in coils.

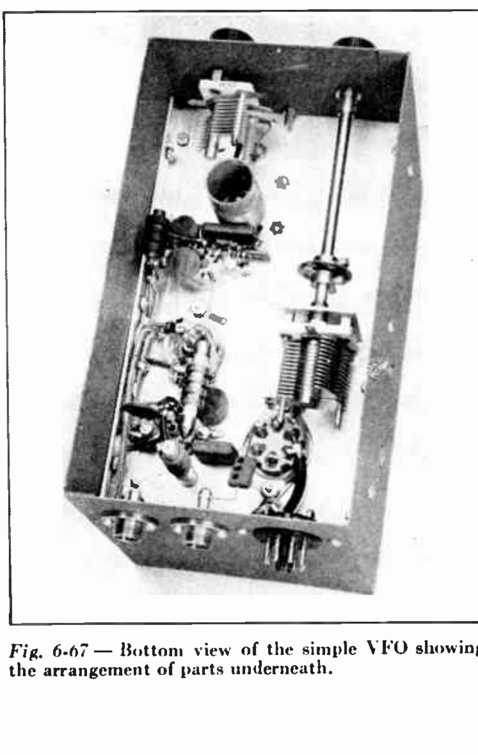


Fig. 6-67 — Bottom view of the simple VFO showing the arrangement of parts underneath.

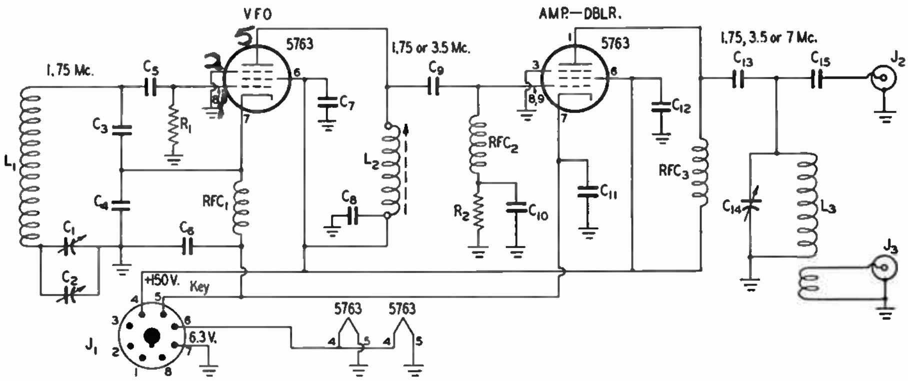


Fig. 6-68 — Circuit diagram of the simple VFO.

C_1 — Approx. 15-μμfd. variable (Millen 19025 with all but 1 rotor and 2 stators removed)
C_2 — 100-μμfd. variable (Millen 22100).
C_3, C_4 — 0.001-μfd. silvered mica.
C_5, C_9, C_{15} — 100-μμfd. mica.
C_6, C_7, C_8, C_{11}, C_{12} — 0.01-μfd. disc ceramic.
C_{10}, C_{13} — 0.001-μfd. disc ceramic.
C_{14} — 140-μμfd. variable (Millen 22140).
R_1, R_2 — 47,000 ohms, ½ watt.
L_1 — 62 turns No. 30 d.s.c., 1 inch diam., close-wound.
L_2 — 1.75 Mc. — 210 turns No. 36 d.s.c., ½ inch diam., close-wound (Millen 74001 form). (300 μh.)
— 3.5 Mc. — 126 turns No. 30 d.s.c., ½ inch diam., close-wound (Millen 74001 form). (75 μh.)
L_3 — 1.75 Mc. — 55 μh. — 45 turns No. 22 d.c.c., 1½ inches diam., close-wound (Bud OEL-160, 14 turns removed).
— 3.5 Mc. — 16 μh. — 20 turns No. 22 d.c.c., 1½ inches diam., close-wound (Bud OEL-80, 8 turns removed).
— 7 Mc. — 5 μh. — 12 turns No. 22 d.c.c., 1½ inches diam., ¾ inch long (Bud OEL-20).
J_1 — Chassis-mounting octal plug.
J_2, J_3 — Female coaxial connector (Jones S101-D).
RFC_1 — 2.5-mh. r.f. choke (National R-50).
RFC_2, RFC_3 — 2.5-mh. r.f. choke (standard type).

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

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

Adjustment

The unit requires a regulated 150-volt supply. The supply diagrammed in Fig. 6-69 is suitable. First adjust R_1, Fig. 6-69, to the maximum resistance that will permit the VR150 to stay ignited when the key is closed. Then, listening on a calibrated receiver, close the key, set C_1 at maximum capacitance and adjust C_2 until the oscillator signal is heard at 3500 kc. Tuning C_1 should then cover the band up to about 3750 kc. Mark the setting of C_2, set C_1 at maximum again and adjust C_2 until the signal is heard at 3750 kc. Then C_1 should cover the range from 3750 to approximately 4000 kc. Repeat the process, setting C_2 for about 3350 kc. to obtain the proper range for 11 meters.

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

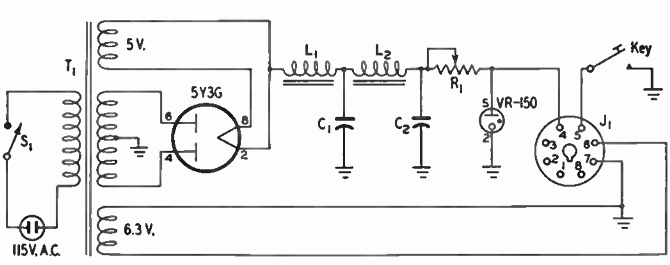


Fig. 6-69 — Circuit diagram of a power supply for the simple VFO.

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

A Silenced VFO for Break-In C.W.

Unfortunately, there is no known practical way in which an oscillator, particularly of the VFO type, can be keyed without a compromise in respect to clicks or chirps. Steps taken to eliminate one will aggravate the other. In the VFO unit shown in Figs. 6-70 through 6-74, the oscillator is not keyed, but allowed to run continuously while a subsequent amplifier is keyed. The signal from the oscillator is suppressed by proper shielding and circuit design, so that it does not interfere with reception on any frequency, including the operating frequency, even with the receiver r.f. gain control at maximum. Any desired shaping of the keyed signal can be applied to the amplifier without introducing chirps.

A diagram of the system is shown in Fig. 6-71. A very low-power high-C Hartley oscillator (15 to 20 volts at the plate), using a 6BD6 and operating in the region of 875 kc., drives a second 6BD6 as a strictly Class A isolating amplifier at the same frequency. The Class A stage, in turn, drives a 6AG5 doubler to 1750 kc. This stage is keyed by the blocked-grid method. Thus, until the key is closed, most of the signal is confined to 875 kc. Further supression of harmonics from the oscillator is obtained by omitting the cathode by-pass condenser in the Class A stage, thereby introducing a slight amount of degeneration.

The output circuits of both the oscillator and buffer are broadbanded, and require only initial adjustment. The output circuit contains a band-pass coupler, thus preserving single-control tuning throughout.

Construction

The photographs of Figs. 6-70 and 6-72 show one method of construction. The unit is housed in a standard 5 × 6 × 9-inch steel utility box. Small rubber shock mounts are bolted to the bottom cover of the box so that the entire assembly can be mounted on a chassis close to the input circuit of the transmitter it is used to drive. Parts layout within the box is not critical, and may be changed from the arrangement shown in the photographs to meet individual preferences, provided that certain considerations are kept in mind. It is desirable to have as much isolation as possible between stages to eliminate stray coupling of the oscillator harmonic to the output circuit. For this reason all heater and d.c. supply leads are made with shielded wire with the shield braid grounded at several points.

Most of the parts are mounted on an aluminum shelf cut to fit snugly inside the box, and spaced 1⅛ inches from the bottom. The interior of the bottom is divided into two compartments by a shield as shown in the photographs. The larger compartment contains the oscillator circuit, and the smaller the Class A and doubler stages. The coils in the smaller compartment should be mounted at right angles to one another.

In the top view, Fig. 6-70, the oscillator tube is at the right of the main tuning condenser, the Class A stage at the left, with the doubler centered about 1⅜ inches in from the left hand edge. The adjusting screws for L_2 and L_3 are visible between the tubes. Band-setting condenser C_2 is mounted at right angles to the main tuning condenser, with its adjustment shaft projecting through the right hand side of the case. The oscillator coil is mounted on a ceramic insulator adjacent to the tuning condensers. An extension shaft is brought out from the rear of C_1 so that additional stages may be ganged to the oscillator tuning condenser if desired.

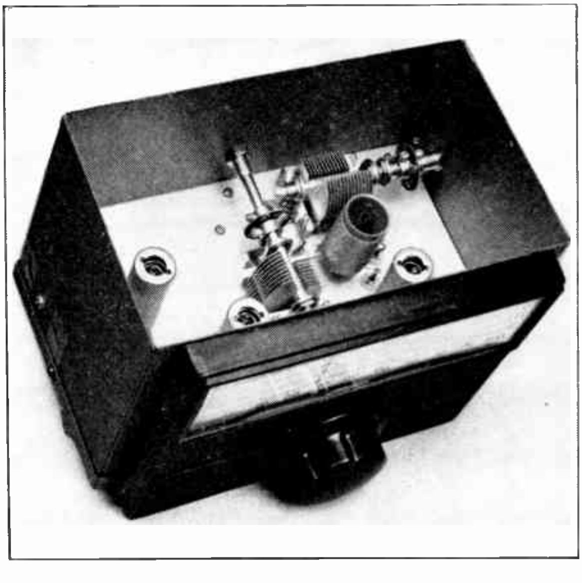


Fig. 6-70 — Top view of the silenced VFO with cover removed. The dial is a Millen type 10035.

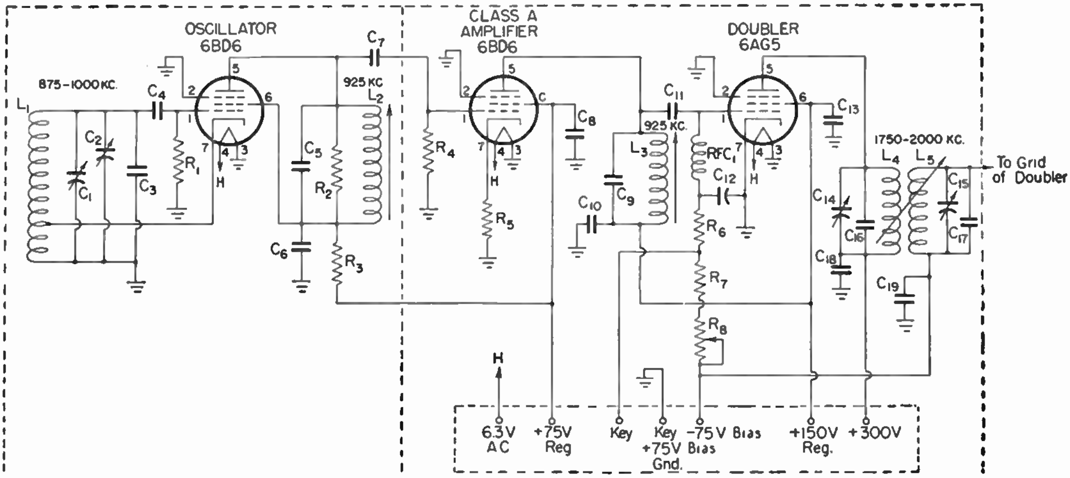


Fig. 6-71 — Circuit diagram of the silenced VFO.

C_1 — 140-μμfd. variable (Millen 19140).
C_2 — 200-μμfd. variable (Millen 19200).
C_3 — 680-μμfd. silvered mica.
C_4, C_7, C_{11} — 100-μμfd. mica.
C_5, C_9 — 68-μμfd. mica.
C_6, C_8, C_{10}, C_{13}, C_{18}, C_{19} — 0.005-μfd. disc ceramic.
C_{12} — 0.05-μfd. paper.
C_{14}, C_{15} — 30-μμfd. mica trimmer.
C_{16} — 25-μμfd. mica.
C_{17} — 33-μμfd. mica.
R_1 — 47,000 ohms, ½ watt.
R_2, R_4 — 10,000 ohms, ½ watt.
R_3 — 47,000 ohms, 1 watt.
R_5 — 330 ohms, 1 watt.
R_6 — 0.1 megohm, ½ watt.
R_7 — 0.22 megohm, ½ watt.
R_8 — 0.25-megohm potentiometer.
L_1 — 32 μh. — 47 turns No. 24 d.s.c., close-wound, 1 inch diam., tapped 14 turns from ground end.
L_2, L_3 — 430 μh. (CTC LS-3, 1 Mc.)
L_4 — 160 μh. — 75 turns No. 36 enam., 1 inch diam., close-wound.
L_5 — 175 μh. — 85 turns No. 36 enam., 1 inch diam., close-wound on same form as L_4, spaced approx. 3/16 inch.

Adjustment and Operation

Some adjustment of the amount of fixed capacitance used in the oscillator circuit may be required to permit tuning the range 875 kc. to 1000 kc. With the values shown, only the c.w. portion of the 3.5- to 4-Mc. band will be covered by C_1. This results in greater bandspread, but if full coverage is desired, the 200-μμfd. condenser should be used as C_1, with the 140-μμfd. unit for C_2. A wide range of frequencies including the 11-meter band, can be covered by readjustment of the band-setting condenser C_2.

The most important adjustment is to make sure that the Class A stage is operating true Class A, because if grid current flows in this stage, the oscillator harmonic will be heard in the receiver even when the key is opened, defeating the purpose of the unit. To do this, resonate the plate circuit of the oscillator in the center of the desired tuning range. Then do the same for the plate circuit of the Class A stage. If no wavemeter capable of tuning the required range is available, a receiver tuned to the broadcast band can be used. Connect a low-range voltmeter, through a 2.5-mh. r.f. choke, across cathode resistor R_5 of the Class A stage. About 3 volts bias should be indicated. Now pull the oscillator tube out of its socket. The voltage read across R_5 will remain the same if Class A conditions are being met. If they are not, reducing

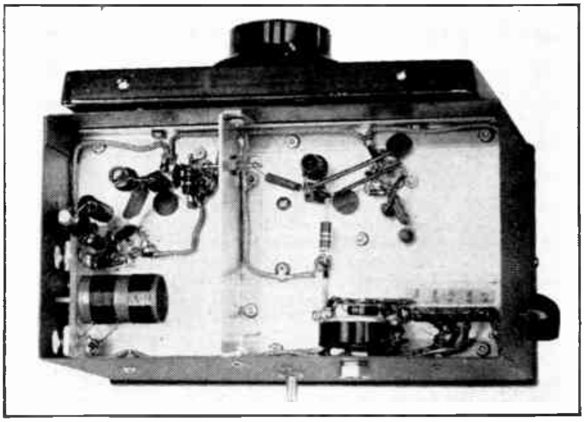
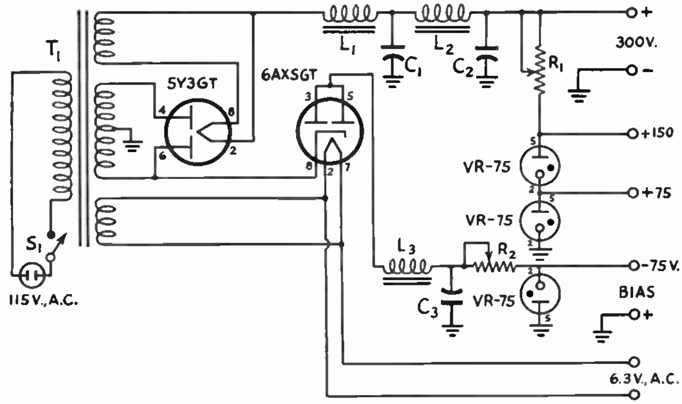


Fig. 6-72 — Bottom view of the silenced VFO.

Fig. 6-73 — Circuit diagram of a suitable power supply for the silenced VFO.

C_1, C_2, C_3 — 40-μfd. 450-volt electrolytic.
R_1 — 10,000 ohms, 25 watts, adjustable.
R_2 — 5,000 ohms, 10 watts, adjustable.
L_1, L_2, L_3 — 15 hy., 50 ma.
S_1 — S.p.s.t. toggle.
T_1 — Power transformer: 350-0-350 volts r.m.s., 70 ma.; 6.3 volts, 2.5 amp.; 5 volts, 2 amp.



oscillator output by increasing the size of R_3, or decreasing the size of either R_2 or R_4 should correct the trouble.

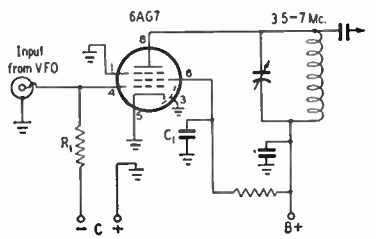
To adjust the bandpass coupler in the output circuit, it is first necessary to connect the unit to the stage it is to drive in the main portion of the transmitter. This should be done with as short a lead as possible. In the arrangement shown in the circuit diagram, direct connection of the output to the grid of the next stage is shown, so that the fixed bias applied to the keying circuit can also be applied to the following stage. This is a requirement if full advantage of the "silenced" feature of the design is to be gained, as explained below. Once connection to the grid of the following stage is made, open one side of the secondary circuit of the bandpass coupler, separate the two coils as far as possible, and resonate the primary circuit with the oscillator set to the center of the band. Reconnect the secondary, open the primary circuit, and resonate the secondary circuit, adjusting it for resonance in the center of the desired pass-band. A grid dip meter will be invaluable in making these adjustments, although they can be done, at a sacrifice of time, by other methods. Once both circuits are resonated properly, move one coil closer to the other a fraction of an inch at a time until the response of the coupler is flat across the band. Output should be observed by noting grid current in the following stage as the main tuning condenser is tuned through its range. If the output varies widely from one end of the band to the other, readjustment of the trimmer condensers, and the coupling between the windings, is required. Sufficient drive for the former crystal oscillator in almost any modern transmitter should be available across the entire band. To eliminate the last trace of signal from the oscillator, it is usually necessary to apply a certain amount of fixed bias to the grid of the stage into which the VFO works. When connected as indicated in Fig. 6-71, the 75 volts bias from the VFO power supply will be applied to the grid of the following stage. If the following stage has a grid blocking or coupling condenser, this should be removed. Any grid leak in this stage also should be eliminated.

Adjustment of the keying characteristics is made by changing the resistance and capacitance in the keying circuit, as described elsewhere in this book. A variable resistance, R_8, is included, but some experimentation with the value of C_{12} may be needed to suit individual tastes.

The diagram of a suitable power supply for this unit is shown in Fig. 6-73. R_1 should be adjusted until the two VR tubes operating from this branch stay ignited under load. R_2 should similarly be adjusted until the VR tube stays ignited under operating conditions.

(Originally described in *QST*, Feb. 1950.)

(A)



(B)

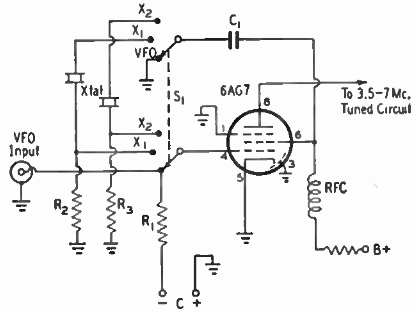


Fig. 6-74 — Two suggested methods of coupling the VFO unit to the transmitter. In both cases the 6AG7 is used as either a doubler or quadrupler from the output of the VFO. In A, a former crystal-oscillator stage has been revised to operate with fixed bias. In B, a switching system providing for either VFO or crystal control is shown.

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

A Beat-Frequency Exciter

Fig. 6-77 shows the circuit diagram of a transmitter frequency-generating unit employing the heterodyne principle. The output of the 6AK6 crystal oscillator at 6500 kc. and the output of the 6AK6 VFO, covering the range of 2650 to 3000 kc., are combined in a mixer of the balanced-modulator type. The output of the mixer, which makes use of a pair of 6BE6s, is tuned to the difference between these two frequencies to give the range of 3500 to 3850 kc. This range includes the c.w. portion of the 80-meter band and, by adding suitable frequency multipliers, all other bands up to and including the 28-Mc. band can be covered. With a change of crystal frequency, the unit will also cover the 80-meter 'phone band.

The advantage of such a system is that neither oscillator need be keyed for break-in operation, since the fundamental and harmonics of both oscillators fall outside amateur bands and therefore do not cause interference in the receiver. Both oscillators run continuously, while the mixer is keyed. Thus the keying characteristic can be shaped as desired to eliminate key clicks without the danger of introducing chirp.

The 6BE6s in the balanced-modulator circuit are connected with their plates in push-pull. The VFO drive is fed to the two No. 1 grids in parallel, while the crystal-oscillator signal is fed in push-pull to the No. 2 grids. The VFO fundamental and harmonics are out of phase in the push-pull output circuit and are cancelled to negligible amplitude, so that the only signal present is the desired difference beat to which the output circuit is tuned.

Amplifier Section

The output of the circuit shown in Fig. 6-77 will be quite low, and unless an adequate buffer-doubler section is already available, the addition of an amplifier will be necessary. Fig. 6-78 shows the circuit of a stable output section sufficient to drive a beam-tetrode final to rated input on the fundamental frequency. As a feature of convenience in tuning, a bandpass coupler is incorporated in the output of the mixer, thus making readjustment of this stage unnecessary over the range of operating frequencies. This coupler, consisting of C_1L_1 and C_2L_2, Fig. 6-78, is merely substituted for the output circuit C_7L_3 in Fig. 6-77 when the amplifier section is added. The 6AQ5 untuned buffer stage, although not strictly essential, provides a small amount of gain and, more important, eliminates the need for neutralizing the output stage, even when a poorly-screened tube, such as the 6L6, is used.

Construction

Figs. 6-75 and 6-76 show an example of the construction of a unit of this type. The exciter shown in the photographs is not the one whose circuit diagram appears here, although the circuit is essentially the same aside from the use of regular-size tubes. Mechanical stability of the variable oscillator, its drift characteristics and freedom from a.c. ripple are just as important in the beat-frequency unit as they are in a conventional VFO. Although a high-C Hartley VFO is shown in the diagram, a Clapp-type circuit can be used just as well, with a probable improvement in drift characteristics. It is suggested that the first step in construction be the building of the variable oscillator, followed by the crystal circuit and then the mixer and amplifier sections in that order. The proper functioning of each stage can be checked as construction progresses. Individual shielding of the variable-oscillator and mixer coils is recommended. The output tank of the amplifier section should be shielded from the preceding stages by a partition. In the rear-view photograph of Fig. 6-76, the VFO is in a separate shock-mounted box to the right. The tube is mounted externally in a horizontal position. The power-supply to the left is likewise a separate unit and is cushioned to prevent transmitting

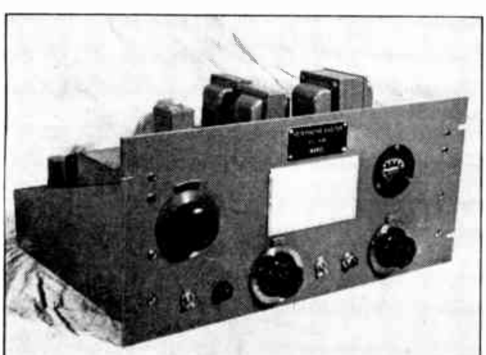


Fig. 6-75 — A beat-frequency exciter built by W6RZL. The dial at the left controls the frequency of the VFO and thereby the frequency of the exciter output. The other two dials are for the crystal-oscillator and amplifier-output tanks.

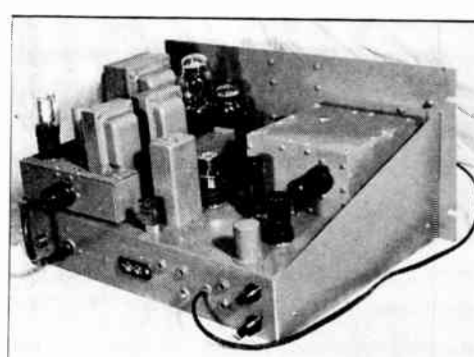


Fig. 6-76 — Rear view of W6RZL's exciter. The shielded compartment encloses the variable oscillator. The power supply is a detachable shock-mounted assembly. Octal, instead of miniature tubes, were used in this particular unit.

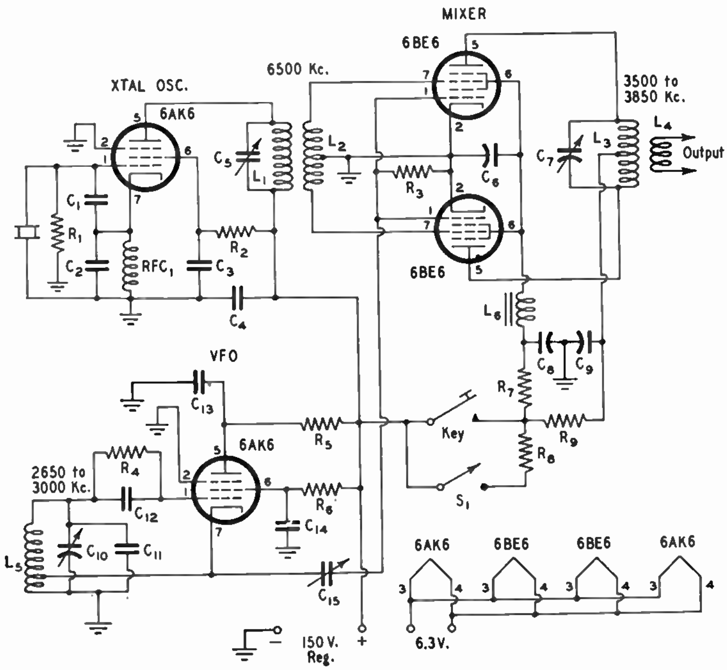


Fig. 6-77 — **Circuit diagram of W6OWP's basic beat-frequency source. Output from this unit will be low and an amplifier is recommended unless adequate buffer-doubler stages are already available. (See Fig. 6-78.)**

C_1, C_2 — 57-μμfd. mica.
C_3, C_4, C_{12}, C_{13} — 0.01-μfd. disk ceramic.
C_5, C_{15} — 30-μμfd. trimmer.
C_6, C_8, C_9 — 0.1-μfd. 600-volt paper.
C_7 — 100-μμfd. variable.
C_{10} — 140-μμfd. variable.
C_{11} — 240-μμfd. silvered mica.
C_{14} — 100-μμfd. mica.
R_1 — 47,000 ohms, ½ watt.
R_2, R_3, R_6 — 10,000 ohms, 1 watt.
R_4 — 47,000 ohms, 1 watt.
R_5, R_9 — 470 ohms, ½ watt.
R_7 — 10,000 ohms, 10 watts.
R_8 — 3.9 megohms, 1 watt.
L_1 — 26 turns No. 24 d.c.c., ⅞-inch diam., close wound.
L_2 — 15 turns No. 24 d.c.c., center-tapped, wound over L_1.
L_3 — 32 turns No. 24 d.c.c., 1½-inch diam., close wound, center-tapped.
L_4 — 3 turns No. 18 hook-up wire wound over center of L_3.
L_5 — 19 turns No. 20 enam., 1¼-inch diam., 1 inch long, tapped 5 turns from bottom.
L_6 — Primary of 50L6 output transformer.
RFC_1 — 2.5-mh. r.f. choke.

vibration to the VFO. The tuned circuits of the mixer and amplifier are mounted underneath the chassis, although the tubes of these stages are above.

Keying

The 6BE6 converter tubes present substantially constant loading to the variable oscillator. To preserve this condition, cathode keying cannot be employed. However, the tube design is such that Miller effect with changes in space current is negligible. Thus, interruption of the plate and screen supply offers an excellent method of keying. Since this places the "hot" side of the key at 150 volts above ground, a keying relay is recommended as a safety measure. The electrical circuit can be traced back from the key in Fig. 6-77. Shaping on both "make" and "break" is provided, with greater emphasis on the "break" characteristic. This gives the type of keying generally accepted as most desirable. The larger-than-usual (0.1 μfd.) plate and screen bypass condensers, as well as the plate decoupling and screen-dropping resistors, are all part of the shaping network. The "make" lag is introduced in the screen lead through L_6, which is the primary of a replacement-type 50L6 output transformer. Where the 6AQ5-6L6 output section of Fig. 6-78 is used, the screen of the 6AQ5 is connected to the key circuit as indicated. This is a further safeguard against the presence of residual key-up signal.

Adjustment

The circuit of a suitable power supply is shown in Fig. 6-79. With both oscillators on, and plate and screen voltage applied to the mixer, set the variable oscillator on 3000 kc. with the aid of a receiver. Now tune the receiver to 3500 kc. The desired beat should be clearly audible. Next adjust the mixer tuning condenser to give maximum signal strength. Repeat this procedure, setting the variable oscillator at 2650 kc., and tuning in the beat at 3850 kc. This establishes basic tuning ranges on the oscillator and mixer dials.

Now, connect the mixer output to the stage to be driven and make the necessary adjustments in the mixer tuning to show maximum drive, as indicated by the plate milliammeter of the driven stage.

The next step is to adjust the variable-oscillator coupling condenser, C_{15}, to the minimum value necessary to maintain full drive from the mixer. The plate tank circuit of the crystal oscillator should be detuned in like manner. Since the 6BE6s drive very easily, there is no object in overdriving; in fact, undesirable interaction could result if the drive were excessive.

The unit is now ready for final calibration and installation. In operation, the procedure is to set the oscillator tuning dial first and then trim up the mixer tuning as required to give necessary output. If the output section of Fig. 6-78 is used, the bandpass coupler is adjusted by varying the settings of C_1 and C_2 until fairly uniform drive over the 3500–3850 kc. band is secured. Oscillator tuning and output controls in this case are the

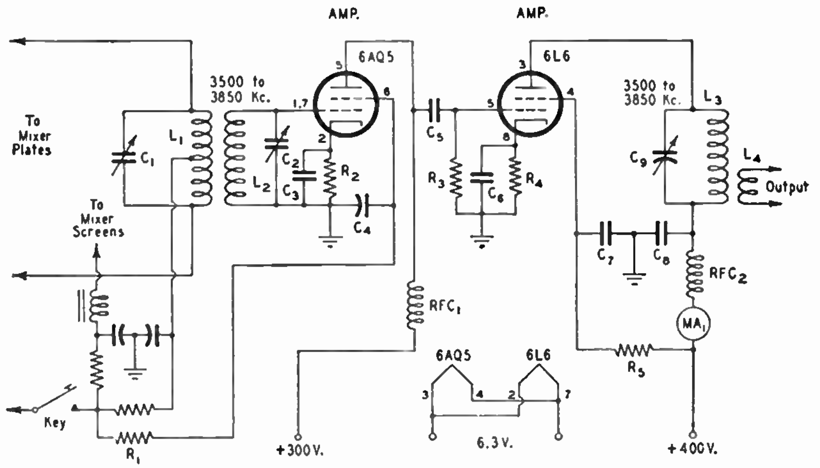


Fig. 6-78 — A stable output section giving sufficient boost in drive to handle a high-power beam final on the fundamental frequency. Note the bandpass coupler substituted for the mixer output circuit in Fig. 6-77. Use of the 6AQ5 untuned buffer is discussed in the text.

C_1, C_2 — 100-μμfd. trimmer.
C_3, C_6, C_7, C_8 — 0.01-μfd. disk ceramic.
C_4 — 0.1-μfd. 600-volt paper.
C_5 — 100-μμfd. mica.
C_9 — 100-μμfd. variable.
R_1 — 470 ohms, ½ watt.
R_2 — 470 ohms, 1 watt.
R_3 — 22,000 ohms, 1 watt.
R_4 — 600 ohms, 2 watts.
R_5 — 22,000 ohms, 2 watts.

L_1 — 32 turns No. 30 enam., in two equal sections 1½ inches apart on 1½-inch-diam. form.
L_2 — 26 turns No. 30 enam. wound in area between sections of L_1.
L_3 — 28 turns No. 14 enam., on 1¾-inch-diam. ceramic form.
L_4 — 3 turns No. 18 hook-up wire, wound at cold end of L_3.
MA_1 — 100-ma. milliammeter.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 1.5 mh. r.f. choke.

only ones requiring adjustment for QSY. To "zero in" on a station to be called, switch S_1 applies just enough voltage to the keyed circuit to provide a usable signal in the home-station receiver.

To maintain the keying characteristic through following transmitter stages, it will be necessary to observe two precautions. (1) In stages where fixed bias is used, the amount of this bias should be just sufficient to cut off plate current. Additional operating bias should be secured through grid resistance. (2) Adequate drive must be provided for each stage.

While the special problems of c.w. operation make the heterodyne exciter especially attractive in this field, it is equally adaptable to 'phone work. To cover the 75-meter 'phone band, a 6850-kc. crystal must be substituted for the 6500-kc. one used for c.w. The 'phone band will be covered in the 2850 to 3000-kc. range of the variable oscillator.

Narrow-band f.m. is readily obtained by connecting any of the standard reactance-tube circuits to the variable-oscillator circuit. An important advantage of the heterodyne unit is that deviation is unaffected by exciter loading. This factor also makes the unit ideally suited for frequency-shift transmission on bands where such operation is authorized.

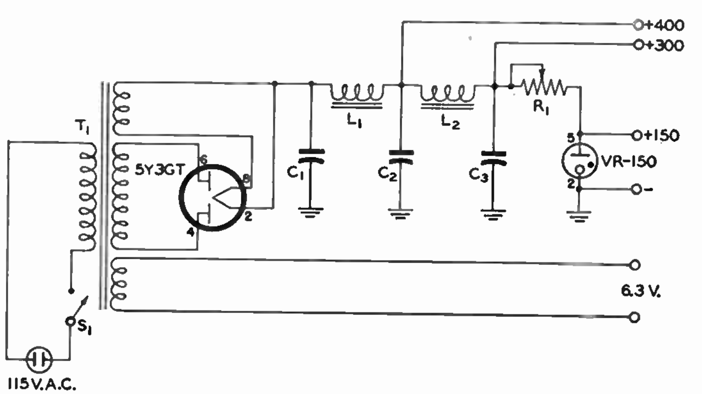


Fig. 6-79 — Circuit diagram of a power supply for the beat-frequency exciter.

C_1 — 8-μfd. 600-volt electrolytic.
C_2, C_3 — 8-μfd. 450-volt electrolytic.
R_1 — 10,000 ohms, 10 watts, adjustable.
L_1 — 2.3-hy. 150-ma. 60-ohm filter choke (Stancor 2304).
L_2 — 16-hy. 50-ma. 580-ohm filter choke (Stancor C-1003).
S_1 — S.p.s.t. toggle.
T_1 — Power transformer: 350–0–350 volts r.m.s., 110 ma.; 5 volts, 2 amp.; 6.3 volts, 3 amp. (Thordarson T-22R32).

A Remotely-Tuned VFO

The VFO shown in Figs. 6–80 through 6–84 is a series-tuned Colpitts (Clapp) circuit built in two sections. The large compartment contains only the tuned circuit (Fig. 6–81A), while the other contains the 5763 tube and a pair of OB2 voltage regulators (Fig. 6-81B). The two are connected with a piece of double-conductor coaxial cable that may be of any length up to 10 feet or so. The advantages of such a system are, first, that the tuned circuit is well removed from heat-generating equipment, including the oscillator tube itself, and second, that it forms a convenient means of remote frequency control. While this arrangement was designed primarily as a driver for the frequency-multiplier unit described later in this chapter, in many cases the existing crystal-oscillator tube of a transmitter can be substituted for the second unit mentioned, if the tube is a 6AG7 or 5763. If the grid-plate crystal-oscillator circuit is in use in the transmitter, it should be possible to feed the tuned circuit directly through the 2-conductor cable to the crystal terminals without modifying the crystal circuit in any way. RG-22/U is recommended for the connecting cable.

The oscillator operates in the 3.5-Mc. region and the bandspread tuning system, consisting of C_1, C_2 and C_3, is designed to cover the desired frequency ranges in three steps, when C_1 and C_2 are altered as described under Fig. 6–81. With one setting of C_2, the tuning condenser C_1 spreads the range of 3500 to 3750 kc. out over 95 per cent of the National ACN dial. Since this fundamental range covers the most-used 80-meter c.w. frequencies, and harmonics of this range cover all of the higher-frequency bands, excepting only the 11-meter band, this range will usually suffice for 90 per cent of all operating. By shifting the setting of C_2, the range of 3750 to 4000 kc. is spread out over about 75 per cent of the dial. The 11-meter band is provided for by a third setting of C_2.

Tuned-Circuit Unit

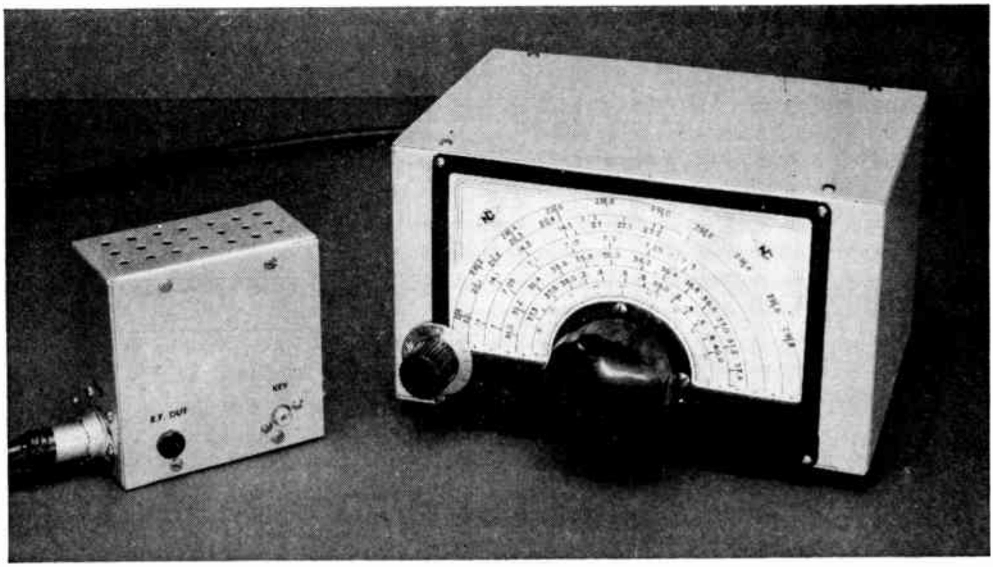
The tuned circuit is housed in a 5 × 6 × 9-inch aluminum box. An enclosure of this size is needed not only to provide mounting for an adequate dial, but also to permit spacing the coil well away from the sides of the box so that its Q will not be drastically reduced by the shielding in its field.

The dial is first mounted centrally on one of the 5 × 9-inch sides of the box. The tuning condenser, C_1, is then coupled to the dial and the mounting step at the rear of the condenser is supported against the bottom of the box with a heavy metal spacer cut to fit. The band-set condenser, C_2, is shaft-hole mounted 1 inch in from the left side and bottom of the box. This necessitates drilling the shaft hole through the edge of the dial frame. C_3 is soldered directly across the terminals of C_2. The knob is a National HRS-5.

The B & W coil is removed from its mounting by first drilling out the rivets in the plug-in base, leaving the metal angle pieces at each end attached to the coil, and unsoldering the leads from the pins. The link winding is carefully removed by snipping the turns and prying the spacing blocks loose with a knife. One turn is removed from the coil itself. The coil is then mounted on National GS-1 pillar insulators so that it will be centrally located in the box in both directions.

The three-contact jack for the remote-tuning

Fig. 6-80 — The remotely-tuned VFO. The large box contains the tuned circuit, the smaller one the oscillator and voltage-regulator tubes. The two terminals on the smaller box are for output and key connections. The power connector is at the end opposite the cable connection.



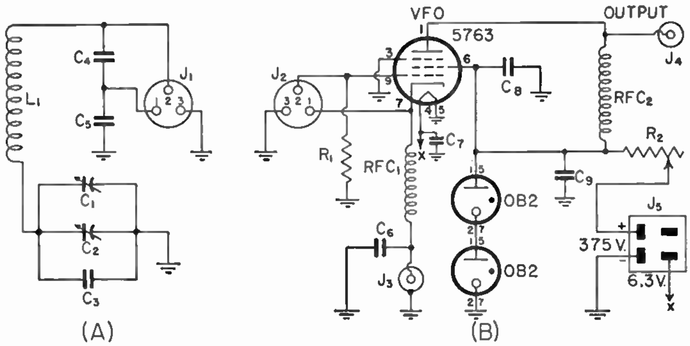


Fig. 6-81 — Circuit of the remotely-tuned VFO.

C_1 — Approx. 12-$\mu\mu$fd. variable (Hammarlund HF-15, rear stator plate removed, rear rotor plate bent; see text).
C_2 — Approx. 23-$\mu\mu$fd. variable (Hammarlund HF-35, last stator and last two rotor plates removed).
C_3 — 39-$\mu\mu$fd. silvered mica.
C_4, C_5 — 0.001-μfd. silvered mica.
C_6, C_7, C_8, C_9 — 0.001-μfd. disk ceramic.
R_1 — 47,000 ohms, ½ watt.
R_2 — 10,000 ohms, 10 watts, with slider.
L_1 — 35 μh. — 39 turns No. 18, 1⅞ inches long, 1½ inches diam. (B & W JEL-80, 1 turn and link removed).
J_1, J_2 — 3-contact female jack (78-PCG3F).
J_3 — Key jack — 'phono input jack.
J_4 — Insulated 'phone-tip jack.
J_5 — 4-contact male connector (C-J P-304-AB).
RFC_1, RFC_2 — 1-mh. r.f. choke (National R-50).
NOTE: RG-22/U remote cable is terminated at each end with Amphenol 91-MPM 36 male connector to fit J_1 and J_2.

cable is set in the back of the box, and C_4 and C_5 are soldered to its terminals.

Tube Unit

The photographs show the essential details of the assembly of the tube unit. The enclosure is a standard 2 × 2 × 4-inch aluminum box. The three tubes are mounted on a shelf spaced 1½ inches from the top of the box. This dimension is critical if the tubes are to be removed without difficulty. The keying and output jacks are mounted in one of the covers, below the shelf level, and the power connector is mounted at one end and the jack for the coax cable at the other. The resistor, R_2, is mounted on top of the shelf, alongside the tubes, on the same side of the box as the keying and output jacks. This makes it possible to remove the tubes and adjust the slider by removing the blank cover of the box. The resistor is supported between two small angle pieces joined with a piece of threaded rod (or a long 6-32 screw) through the resistor form.

All wiring, with the exception of the connections to the keying and output jacks and the cable connector, can be done before the shelf is placed in the box. This includes connections to the power connector which mounts from the inside. In the bottom view of Fig. 6-84, the plate choke, RFC_2 is to the lower left, soldered between Pin 6 of the 5763 socket and Pin 5 of the socket of the first 0B2 regulator. The cathode choke, RFC_1, is above, with one end fastened to Pin 7 of the 5763 socket, while the other end is left free until the cover plate carrying the key jack is ready to be put in place. C_6 is soldered directly across J_3. Leads of proper length are made for the jacks and cable connector, and these connections can be made after the shelf has been put in place, and just before the cover is put on. Care should be used in placing the tubes in their sockets, since there is little height to spare. If necessary, the tips of the tubes can be run up through the ventilating holes in the top of the box to allow the pins to clear the sockets.

Power Supply

Any power supply delivering between 250 and 400 volts at 50 ma. or more may be used to operate this VFO. If a 120-ma. transformer, instead of the 70-ma. unit specified for the power-supply diagram of Fig. 6-90, is provided, the VFO and the multiplier unit may be operated from the single supply.

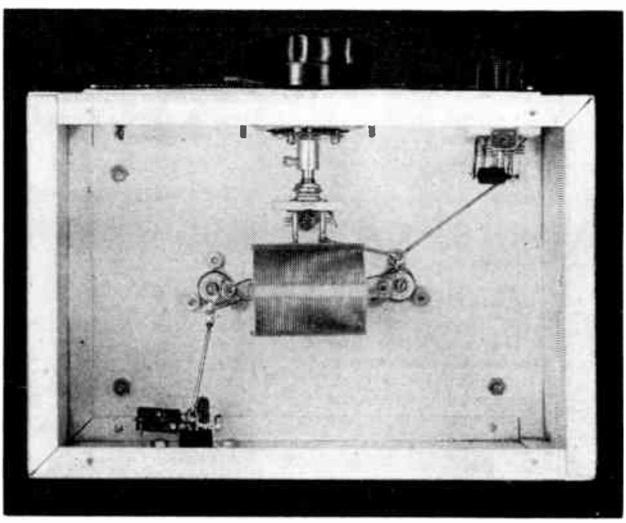


Fig. 6-82 — Interior of the tuned-circuit box. C_4 and C_5 are to the rear. C_3 is soldered across C_2 to the left in front.

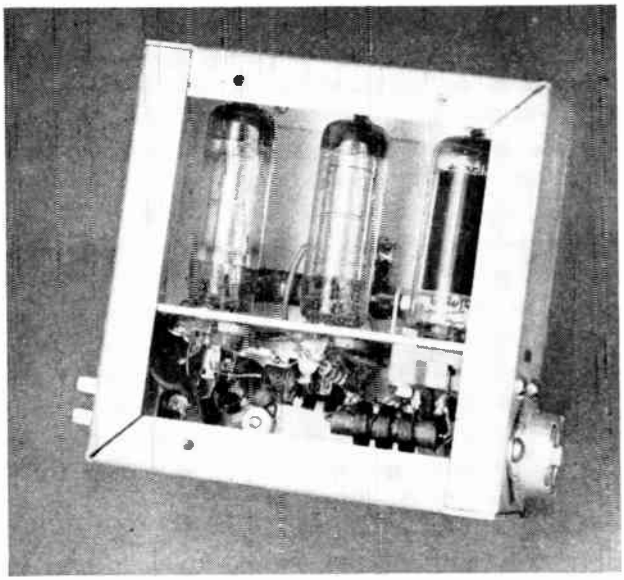


Fig. 6-83 — The completed tube section with the tubes in place. Ventilation holes are drilled in the top of the box and in the plate covering the free side.

Adjustment

Adjustment of the frequency range for maximum bandspread is quite simple. Set C_1 to a dial reading of 5. Then adjust C_2 until the oscillator signal is heard on the receiver at 3500 kc. Set the receiver to 3750 kc. and adjust C_1 until the signal is heard. If this occurs with the dial set at less than 100, carefully bend the rearmost rotor plate of C_1 away from the adjacent stator plate, making sure that the plates do not touch and short the condenser in any position of the rotor. Turn C_1 again to a dial reading of 5, reset C_2 for 3500 kc., and check again for the point where C_1 tunes to 3750 kc. By proper adjustment of the rotor plate on C_1, the 3500-to-3750-kc. range can be made to cover the entire dial, or as much of it as desired.

'Phone Band

After this initial range has been set, tune the receiver to 3875 kc. Set C_1 to midscale and adjust C_2 until the VFO signal is heard. Then the range of 3750 to 4000 kc. should be approximately centered on the dial with a coverage of about 75 divisions. The range can be shifted one way or the other by simply shifting C_2 slightly.

11-Meter Band

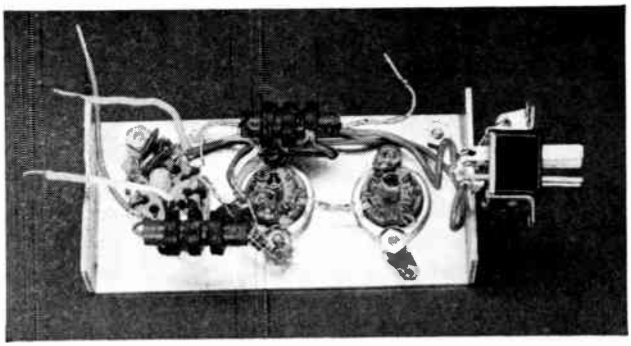
If it is desired to center the 11-meter band on the dial, set C_1 to midscale, set the receiver to 3387 kc. and adjust C_2 until the VFO is heard. All three settings of C_2 should be plainly marked so that they can be returned to when desired.

The cathode current may vary from about 28 ma. with both C_1 and C_2 set at maximum capacitance to 37 ma. with both at minimum.

In using the VFO, the tube unit should be placed close to the stage to be driven and fastened securely to the chassis. A short lead should be used to connect the output terminal to the grid of the stage to be driven. If the driven stage has no grid condenser, a 100-$\mu\mu$fd. mica condenser should be connected between the output terminal and the grid of the driven stage. If more than adequate drive is obtained, the screen of the oscillator tube can be connected to the junction between the two VR tubes, rather than to the end of R_2 as shown in Fig. 6-81. This unit is not a power device, and adequate gain in the way of a crystal-oscillator tube or other buffer amplifier should be provided.

(Originally described in *QST*, Jan. 1953.)

Fig. 6-84 — Bottom view of the tube-unit shelf. RFC_1 is above, RFC_2 below. C_6 is soldered to J_3 on the cover plate. The two leads going to the left solder to the cable connector. The one to the left above goes to J_4, the lead to the right to J_3.



A 6-Band Frequency-Multiplier Unit

The unit shown in Figs. 6-85 through 6-90 is a subassembly containing all tubes and circuits necessary for multiplying frequency from any low-power 1.75- or 3.5-Mc. VFO or crystal oscillator. It gives enough output on any of the six bands from 3.5 to 28 Mc. to drive any amplifier tube such as the 2E26, 807, or 6146. Changing from one band to another is simply a matter of clicking a switch and resonating with the single control for maximum grid current to a following amplifier.

The Circuit

The circuit diagram is shown in Fig. 6-86. The first stage, operating at 80 meters, uses a well-screened tube, the 6AK6, because it is called upon to work as a straight amplifier when the VFO output is in the same band. Type 6C4 triodes are used in the remaining stages which are always operated as frequency multipliers.

The 80-meter circuit is designed to cover 3500 to 4000 kc. C_8 is a bandspread padder. However, when the bandswitch is turned to the 7-Mc. and higher-frequency positions, C_{11} adds enough capacitance across the 80-meter tank circuit to shift its lowest frequency to about 3350 kc. so that the harmonics will include the 11-meter band. It is to this second range that the following stages are tracked. The 21-Mc. band is reached by tripling frequency in the stage otherwise used for 14 Mc. The bandswitch shorts out an appropriate portion of L_3 for 21 Mc.

The trimmers, C_{19} and C_{28}, are to compensate for the difference between the input capacitance of the 6C4s and the larger capacitance of the screen-grid tube to be used in the amplifier, thereby automatically maintaining proper conditions for tracking. C_{16}, C_{24} and C_{35} adjust the range over which the tuning condensers will tune.

All tubes are protected against excessive dissipation, when not being driven, by the use of cathode biasing resistors.

Construction

If dimensions are to be kept to a minimum, it will be necessary to make a special shielding enclosure of sheet aluminum. However, if size is not considered an important factor, a standard 5 × 6 × 9-inch box can be used.

The chassis shown is made from sheet aluminum about 1/16 inch thick. It is 4½ inches wide and 7½ inches long, with ½-inch lips bent down along the longer edges for fastening to the sides of the box. The box is made to fit the chassis as closely as possible and has an inside height of 4½ inches. The front and the two sides are made from a single piece, with ½-inch lips bent along both top and bottom edges. Similar lips are bent along all four edges of the removable back. The two rear corners of the chassis must be notched out for these lips.

The chassis is placed in the box with its top surface 2¼ inches down from the top of the box and a row of ¼-inch holes is drilled along each side of the box, just above the chassis level. The top cover also is perforated.

The bandswitch is made up from Centralab Switchkit parts. The index assembly is Type P-123 and the ceramic wafers are Type X having 6 positions, 5 of which are used. The switch is mounted on aluminum brackets (with the tie rods in a vertical plane) to bring the center of the shaft 1⅛ inches below the chassis. In the bottom-view photograph, the first wafer at the top (80) is

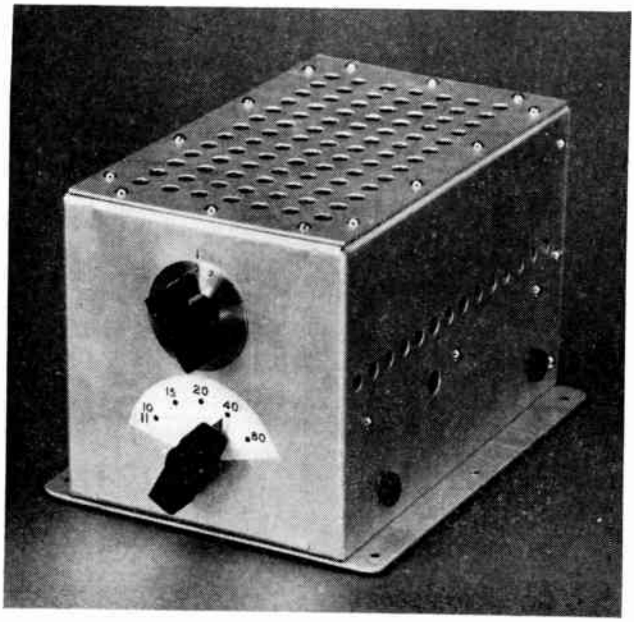


Fig. 6-85 — This small package contains the necessary frequency multipliers to give output on any of the six ham bands from 80 to 10 from any 1.75- or 3.5-Mc. VFO or crystal oscillator. The switch knob at the bottom selects the band, while the single tuning control resonates all circuits. Oscillator input is connected to the pin jack in front; output on the desired band is taken from the one to the rear. The large hole below the row of ventilating holes in the side is for adjusting the 14-Mc. grid trimmer. A single hole in the opposite side provides access to the 10-meter grid trimmer.

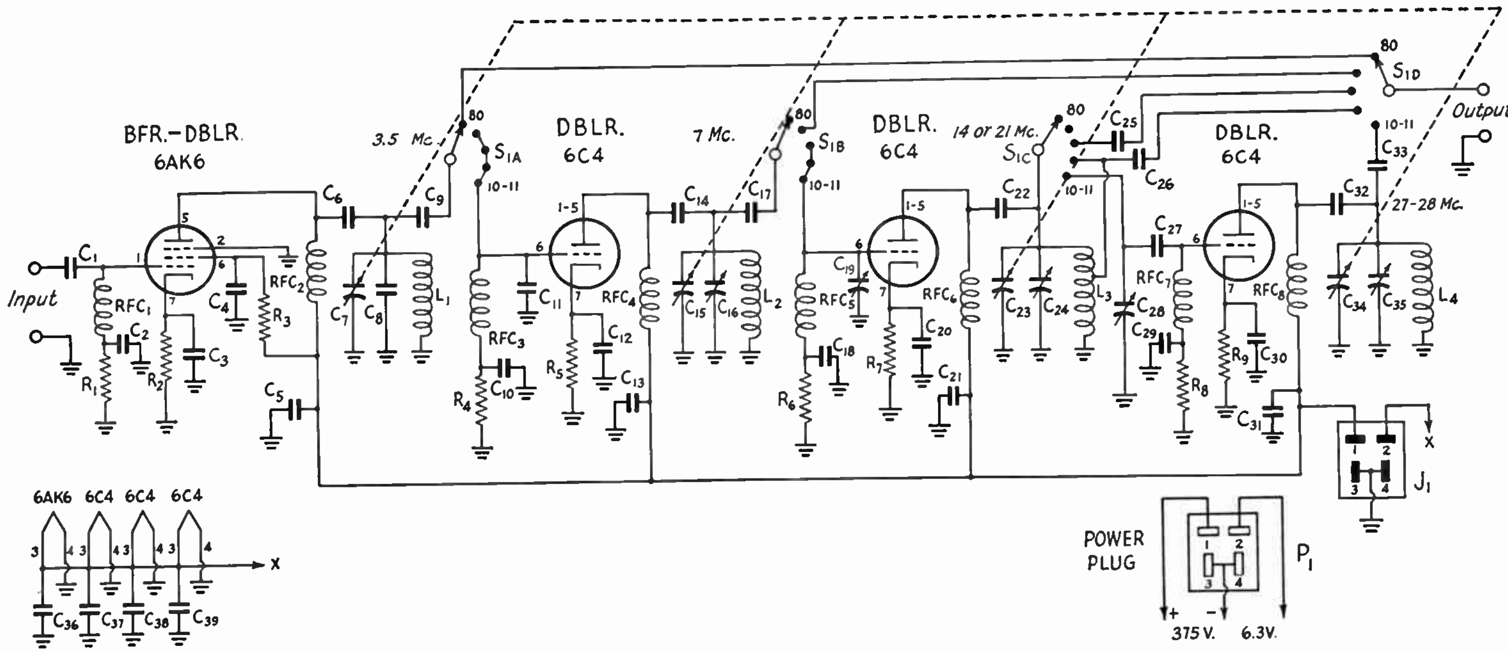


Fig. 6-86 — Circuit diagram of the single-control frequency multiplier.

C_1 — 470-μμfd. mica.

C_2, C_3, C_4, C_5, C_6, C_{10}, C_{12}, C_{13}, C_{14}, C_{18}, C_{20}, C_{21}, C_{22}, C_{29}, C_{30}, C_{31}, C_{32}, C_{36}, C_{37}, C_{38}, C_{39} — 0.001-μfd. disc ceramic.

C_7 — Approx. 65-μμfd. variable (see text).

C_8 — 100-μμfd. silvered mica.

C_9 — 220-μμfd. mica.

C_{11} — 47-μμfd. silvered mica.

C_{15} — Approx. 35-μμfd. variable (see text).

C_{16} — 150-μμfd. mica trimmer or 30-μμfd. mica trimmer and 47-μμfd. silvered mica in parallel.

C_{19}, C_{24}, C_{28}, C_{35} — 30-μμfd. mica trimmer (Millen 27030).

C_{17}, C_{25}, C_{26}, C_{27} — 100-μμfd. mica.

C_{23}, C_{34} — Approx. 25-μμfd. variable (see text).

C_{33} — 47-μμfd. mica.

R_1, R_4, R_6, R_8 — 22,000 ohms, ½ watt.

R_2 — 3300 ohms, 1 watt.

R_3 — 33,000 ohms, 1 watt.

R_5 — 2200 ohms, 1 watt.

R_7 — 2350 ohms, 2 watts (two 4700-ohm 1-watt in parallel).

R_9 — 1940 ohms, 2 watts (3300-ohm 1-watt and 4700-ohm 1-watt in parallel).

L_1 — Approx. 12 μh. — 24 turns No. 22 d.c.c., 1-inch diam., close-wound, or smaller wire spaced to length of ¾ inch (see text).

L_2 — Approx. 4.2 μh. — 17 turns, ¾-inch diam., 17/32 inch long (B & W 3012 Miniductor).

L_3 — Approx. 1.8 μh. — 12 turns, ¾-inch diam., ¾ inch long, tapped at 6½ turns from ground end; see text (B & W 3011 Miniductor).

L_4 — Approx. 0.4 μh. — 7 turns, ½-inch diam., 7/16 inch long (B & W 3003 Miniductor).

J_1 — Four-contact male power connector (Jones P-304-AB).

P_1 — Four-contact female cable connector (Jones S-304-CCD).

RFC_1, RFC_2, RFC_3, RFC_4, RFC_5, RFC_6, RFC_7, RFC_8 — 2.5-mh. r.f. choke (National R-100-S).

S_1 — 4-pole 6-contact rotary switch (see text for assembly procedure).

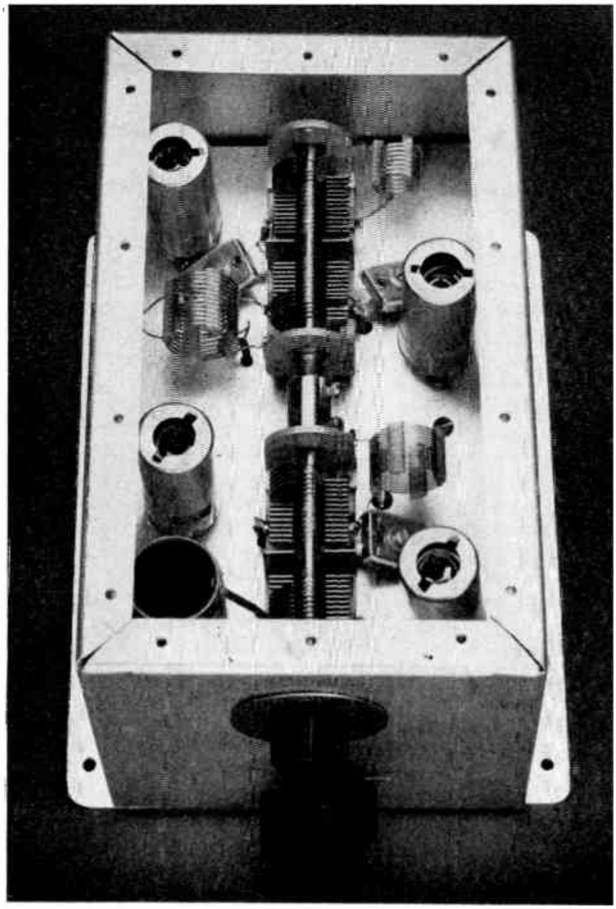


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Fig. 6-87 — **Top interior view of the frequency multiplier showing the tubes, coils and the tuning-condenser gang. The 80-meter coil is in the foreground with the 6AK6 to the right. The 40-meter coil and plate trimmer are behind the 6AK6 with the 7-Mc. 6C4 to the left. In the second section to the rear, the 14-Mc. coil with its 21-Mc. tap is to the left, followed by the 28-Mc. plate trimmer and tube. The 20-meter 6C4, its plate trimmer and the 28-Mc. coil are to the right. The lips along the top edges of the box are duplicated on the bottom. The condenser gang is made up of two Hammarlund type HFD-100 units.**

♦

spaced ½ inch from the index head, with its point contacts to the left. The second wafer (40) is spaced 1 inch from the first with its point contacts to the right. The third wafer (20 and 15) is spaced 2 inches from the second with its point contacts to the left. The last wafer (output) is spaced 1 inch from the preceding one with its point contacts also to the left. The rear mounting bracket is spaced ¼ inch behind the last wafer. The front mounting bracket is fastened to the index head at the shaft bushing.

The tube sockets are placed ⅞ inch in from the edges of the chassis. The 6AK6 and the 14-Mc. 6C4 are to the right, spaced 1¼ and 4¾ inches respectively back from the front edge of the chassis. The 7-Mc. and 28-Mc. tubes are to the left, spaced back 2⅝ and 6¼ inches respectively.

The shafts of the two tuning-condenser units are coupled together with a Millen type 39003 rigid coupling. It may be necessary to file down the front end of the coupling close to the setscrew hole to permit the setscrew to get a good grip on the short tail shaft of the front condenser. In the first condenser section at the front (80), the last 5 rotor plates are removed. In the second section (40), the first 9 rotor plates are removed. In the third section (20 and 15), the first 4 rotor plates are left in and the remainder are removed. The fourth stator plate of this section also is removed, but the rest of the stators are left in. In the last section, all rotors except the last four are removed.

The condenser gang is mounted on top of the chassis with its front mounting hole ½ inch from the front edge of the chassis. In assembling the unit, the condenser gang should be mounted first with screws at the two inner mounting holes only. Then the switch gang underneath should be positioned and the mounting holes in the brackets drilled to match the front and rear mounting holes of the condenser gang. In other words, the switch brackets should be fastened to the chassis by means of the front and rear condenser-mounting screws. After the holes have been drilled in the switch brackets, remove the front bracket, fasten it down with the front condenser-mounting screw, slide the front of the switch into the front bracket, fasten with the shaft nut, and then fasten the rear switch bracket with the rear condenser-mounting screw.

Mount the tube sockets with the plate terminals toward the nearest switch wafer.

The two grid trimmers, C_{19} and C_{28}, are mounted vertically underneath, C_{19} just to the rear of the second wafer and C_{28} immediately behind the third wafer. Half-inch holes are drilled in the sides of the box and the chassis lips are notched out so that these condensers can be adjusted from the outside. The three plate trimmers are fastened on top of the chassis, using the nearest choke-mounting screw to fasten the grounded side to the chassis. The other terminal of the trimmer is soldered directly to the appropriate tuning-condenser stator terminal.

Coils

Approximate inductance values for the coils are given under Fig. 6-86 for the benefit of those who must wind their own. However, the use of the B & W Miniductor coils has the advantage that the original coil dimensions can be duplicated closely. This is necessary if pruning of the coils for tracking is to be avoided. The 80-meter coil, L_1, is wound on a Millen bakelite 1-inch diameter form, fastened to the chassis. The other coils are supported by their leads which are soldered directly to the condenser terminals. The 21-Mc. tap on L_3 should be made with a piece of wire about 3 inches long. When the outer ends of the coil are soldered across the condenser terminals, this tap, which comes near the top of the seventh

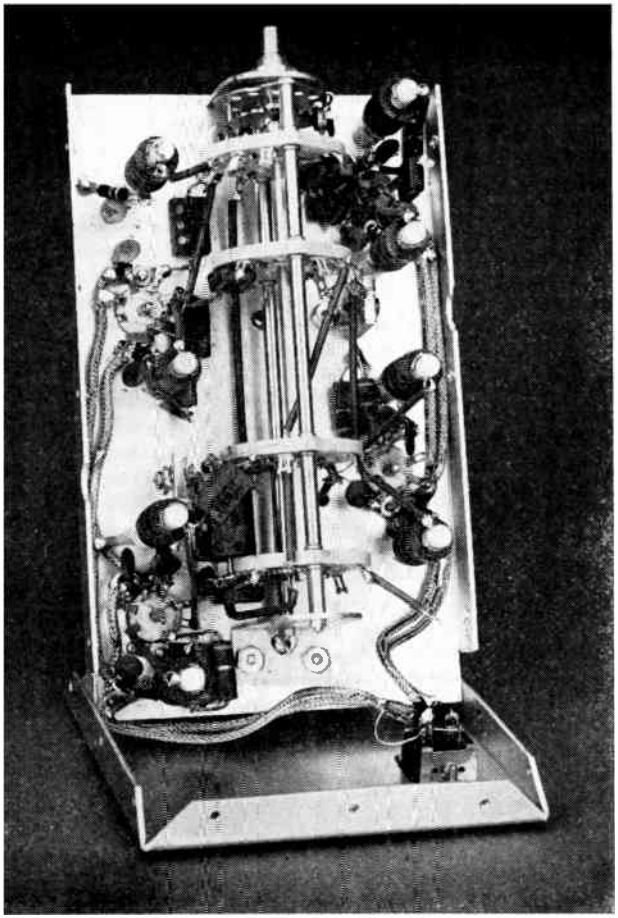


Fig. 6-88 — Bottom view of the multiplier chassis showing the bandswitch, r.f. chokes and other small components. The 80-meter circuit is at the top, the 10-meter circuit at the bottom. The 20-meter grid trimmer is to the right and behind the second switch wafer. The 10-meter grid trimmer is to the left of the third wafer. This view also shows how the removable back of the enclosure is made. The text describes a somewhat different and simpler method of mounting the switch.

turn, should be bent in a sweeping curve around the *outer* side of the coil (counterclockwise as viewed from the front) to the end of a wire from the bandswitch, coming up through a hole in the chassis drilled alongside the condenser frame. The tap is soldered to the end of this switch wire. Don't clip off the excess tap length until adjustments for tracking, described later, have been made.

The Centralab switches have two rotor contacts and C_9 and C_{17} are most conveniently mounted by opening up the lower rotor contact so that it does not make connection with the rotor, and then soldering the condenser between this terminal and the other rotor terminal above. The lower terminal is then used also as a tie point for the preceding 0.001-μfd. plate blocking condenser and a lead going through the chassis to the tuning-condenser stator terminal above. C_{25} and C_{26} are soldered directly between the contact terminals of the two switch sections, while C_{27} is soldered between the terminal of the switch and the top end of the near-by grid choke, RFC_7. C_1 is soldered between the input pin jack and the grid terminal of the 6AK6 socket.

Mounting the Unit in a Transmitter

In mounting the multiplier unit on a chassis with other stages, it is not necessary, of course, that it be placed close to the panel. By using extension shafts, it can be placed as far to the rear as desired. The unit should be fastened securely to the chassis and the amplifier tube mounted close to the output terminal. The grid of the amplifier should be connected to the output terminal of the multiplier unit with a short wire well spaced from the chassis, and the cathode of the amplifier should be grounded or by-passed immediately to the chassis. If the grid wire, or the path from the amplifier cathode to the multiplier box is much over 6 inches long, there may be a noticeable loss in output at 28 Mc., and it may not be possible to resonate the higher-frequency multiplier circuits.

It is preferable also to have the oscillator located on the same chassis as the multiplier unit so that the coupling leads will be short. However, if the oscillator has the power and tuning range to spare, a piece of coax cable can be used, as shown in Fig. 6-89. In order to do this, it must be possible to retune the oscillator output circuit to compensate for the capacitance of the cable.

Power Supply

A power supply delivering 375 to 380 volts at 60 or 70 ma. is required to operate the unit. To assure adequate output, the supply voltage should be close to this figure. A suitable circuit is shown in Fig. 6-90.

Adjustment

Until the unit has been tuned up, no plate or screen voltage should be applied to the amplifier. Means should be provided for checking the amplifier grid current, or the voltage across its grid leak. While it should be possible to make adjust-

TABLE 6-I
Typical Voltage Readings* (Supply Voltage 380)

Stage	*80*			*40*		*20/15*		*10*	
Switch Position	*Cathode*	*Grid Leak*	*Screen*	*Cathode*	*Grid Leak*	*Cathode*	*Grid Leak*	*Cathode*	*Grid Leak*
80	65	25	235	17	0	19	0	16	0
40	60	30	221	40	97	19	0	16	0
20	59	30	211	36	96	72	126	16	0
15	58	31	207	34	89	93	106	16	0
10	58	30	207	34	89	69	120	45	130

* By dividing these voltages by the associated resistance values, any desired current value may be easily calculated.

ments without metering the multiplier unit, the job will be a little easier if a milliammeter is inserted temporarily in the high-voltage lead to the power supply, at least.

With the switch in the 80-meter position, turn on the oscillator and tune it to 3500 kc. (1750 kc. if the oscillator output is at 160 meters). If the oscillator is crystal-controlled, use the lowest-frequency crystal at hand. Now resonate the multiplier for maximum drive to the amplifier. With the multiplier tuned to resonance, adjust the coupling to the oscillator to give maximum drive to the amplifier. Maximum drive should occur with the oscillator developing a bias of 15 to 30 volts across the grid leak of the 6AK6. If no other means is available, the drive to the 6AK6 can be reduced by reducing the size of C_1, Fig. 6-89. If a VFO is used, the multiplier should be checked at both 3500 and 4000 kc. to make sure it is covering the proper frequency range. (The multiplier must always be retuned, of course, for any appreciable change in oscillator frequency.) It may be necessary to spread out the last few turns of L_1 on the coil form to get the circuit to hit both ends of the band. Drive to the

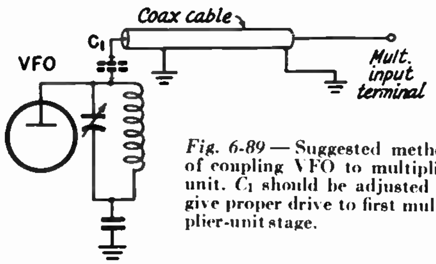


Fig. 6-89 — Suggested method of coupling VFO to multiplier unit. C_1 should be adjusted to give proper drive to first multiplier-unit stage.

amplifier should be essentially the same anywhere in the band, providing the output of the oscillator is reasonably constant.

With the 80-meter stage working properly, the switch should be turned to the 7-Mc. position. Set the VFO to 3500 kc. and resonate the multiplier. If there is no indication of drive to the amplifier, it may be necessary to adjust the 7-Mc. trimmer, C_{16}, a little bit at a time, retuning the gang, until an indication of output is obtained. As an aid, a milliammeter in the high-voltage lead should show a dip when C_{16} is tuned through resonance. When an indication is obtained, tune the gang for peak drive and then adjust C_{16} to increase the peak. The correct adjustment is the one where no readjustment of either the gang or the trimmer will increase the drive. Now turn the oscillator to 3750 kc. and retune the multiplier. The drive to the amplifier should be essentially unchanged.

Now tune the oscillator back to 3500 kc. and retune the multiplier for maximum output. Leave the multiplier and oscillator tuning at this point and turn the bandswitch to 14 Mc. Adjust first C_{24}, and then C_{19}, for maximum amplifier grid current. It may take a little juggling back and forth between these two before a maximum reading of drive is obtained. The milliammeter in the high-voltage lead should

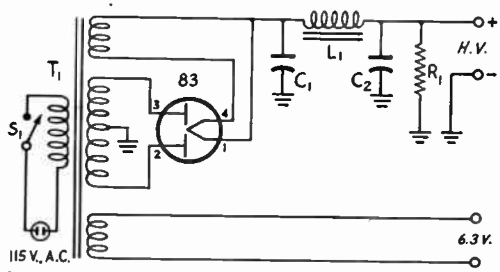


Fig. 6-90 — Circuit diagram of a suitable power supply for the frequency-multiplier unit.

C_1, C_2 — 16-μfd. 600-volt wkg. electrolytic.
R_1 — 50,000 ohms, 10 watts.
L_1 — 12-hy. 80-ma. filter choke.
S_1 — S.p.s.t. toggle switch.
T_1 — Power transformer: 350-0-350 volts r.m.s., 70 ma.: 6.3 volts, 2.5 amp.; 5 volts, 3 amp. (Stancor P-4078 or equivalent).

show a dip when C_{24} is tuned through resonance.

Leaving all tuning adjustments fixed, turn the switch to the 21-Mc. position. Now adjust C_{24} carefully and note whether an increase or decrease in capacitance causes an increase in drive to the amplifier. If it is an increase, lengthen the tap wire (see preceding section on coils) slightly. Then turn the switch back to 14 Mc. and readjust C_{24} for maximum drive. Then switch back to 21 Mc. and check carefully again. By adjusting the length of the tap wire carefully, it should be possible to arrive at a condition where maximum drive is obtained both at 14 and 21 Mc. with the same adjustment of C_{24}.

Adjustment for 28 Mc. is similar to that for 14 Mc., although it will be more critical. Careful adjustment of C_{28} and C_{35} will be necessary for maximum amplifier drive. The 11-meter band is covered by tuning the multiplier to resonance at the desired frequency with the switch in the 28-Mc. position. The various circuits should be checked with an absorption wavemeter to make sure that they are tuning to the right multiple.

When the above adjustments for the low-frequency ends of the various bands have been completed as described, it should be found that the output will be essentially the same at any point within a given band.

The accompanying tables show typical voltage readings taken with the unit in operation driving the grid of a 6146 amplifier.

(Originally described in *QST* for April 1952.)

TABLE 6-II
Typical Total Current and Output Readings*

	80	40	20	15	10
*Amplifier bias ** (volts)*	152	195	187	144	140
Total B ma. at resonance	41	47	53	60	60
Total B ma. off resonance	45	58	75	78	85
Total B ma., no excitation — 35					

* Average supply voltage 380.
** Voltage measured across 39,000-ohm grid leak of unloaded 6146 amplifier.

A 6146 Multiband Amplifier

Figs. 6-91 through 6-95 show the circuit and constructional details of a 6146 amplifier using a National multiband tuner (Type MB-40SL). It is capable of handling up to 90 watts input at 600 volts with plate-screen modulation, or up to 750 volts for c.w. operation. This unit, contained in a standard 5 × 6 × 9-inch aluminum box, was designed primarily as an amplifier to follow the Bandbox frequency-multiplier unit described in the preceding section.

The circuit diagram is shown in Fig. 6-91. To permit operation at 600 or 750 volts without exceeding the voltage breakdown rating of the tank condensers in the multiband tuner, the original series plate feed was changed to parallel feed, thereby removing the d.c. from the condenser. The center tap of the low-frequency coil, L_4, as well as the rotors of the low-frequency tank condenser are grounded. The low-frequency condenser section is the one at the rear.

A combination of fixed bias and a 6AQ5 clamp tube cuts the input to zero when excitation is removed. The 6AQ5 is connected as a tetrode with its screen operated from the voltage divider made up of R_2 and R_3. This provides a heavier clamper-tube current at low amplifier screen voltages than with the triode connection which permits the clamper-tube screen voltage to fall along with its plate voltage when excitation is removed from the amplifier.

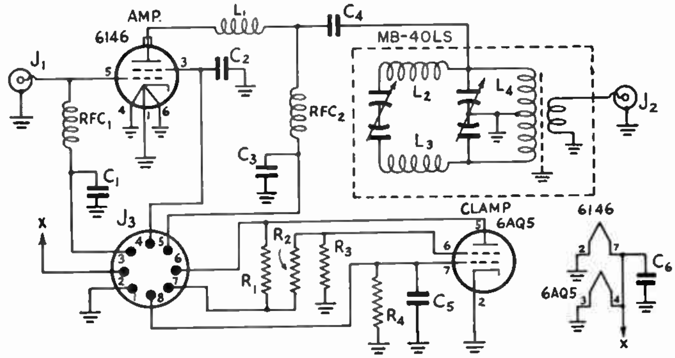


Fig. 6-91 — Circuit of the shielded 6146 multiband amplifier.

C_1, C_2, C_5, C_6 — 0.001-μfd. disk.
C_3, C_4 — 0.001-μfd. 1200-volt mica.
R_1, R_2 — 20,000 ohms, 10 watts.
R_3 — 11,000 ohms, 2 watts (two 22,000-ohm 1-watt units in parallel).
R_4 — 12,000 ohms, ½ watt.
L_1 — 5 turns No. 16, ¼-inch diam., length adjusted for v.h.f. parasitic, approx. 1½ inches long.
L_2, L_3, L_4 — Multiband-tuner coils (see text).
J_1, J_2 — Shielded 'phono jack.
J_3 — Octal power plug.

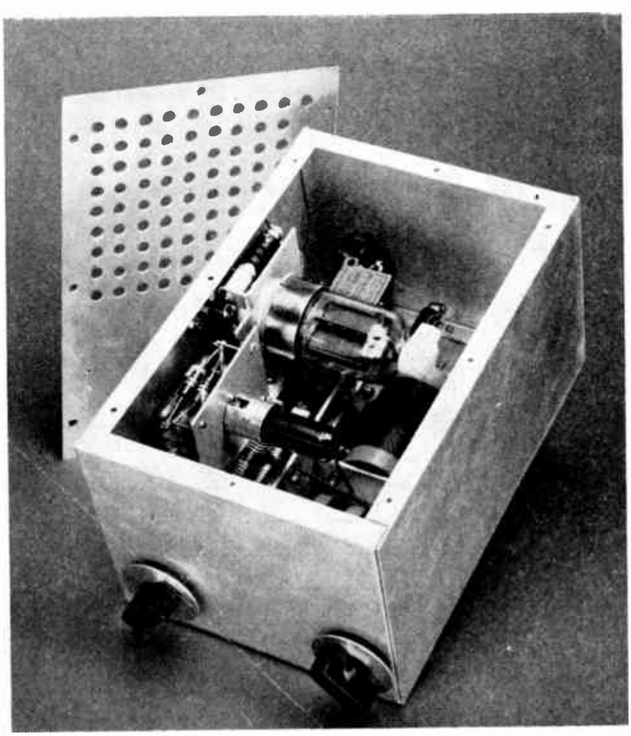


Fig. 6-92 — An 80-watt amplifier using a 6146 and a multiband tuner.

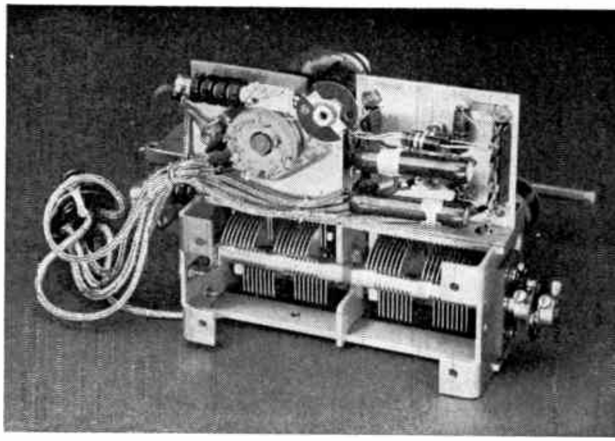


Fig. 6-93 — Bottom view of the multiband amplifier.

Construction

All assembly and wiring can be done before the unit is placed in the box. As shown in Fig. 6-94, the two tubes are mounted on aluminum brackets slung from the left side of the frame of the tuner condenser. These are each 2½ inches square, not including a half-inch mounting lip at the rear. (The two separate brackets might well be replaced with a single shelf running the length of the condenser frame.) In the photograph of Fig. 6-94, the lip of the 6146 bracket is bent upward, while the lip of the 6AQ5 bracket is bent downward to provide room for the components underneath. (If a single shelf is used, the lip should be bent downward.) The only other essential precaution is to place the socket of the 6146 far enough away from the tuning unit so that the tube will not interfere with the swinging of the output link. The brackets are fastened to the side strip of the condenser frame by drilling and tapping holes in the frame for 6-32 screws.

Before the brackets are fastened permanently, the holes for the two controls should be spotted on the front of the box. These should be centered 1⅛ inches from the bottom of the box in Fig. 6-92, and so that they will line up with the two control shafts when the bottom of the condenser frame is resting against the left-hand side of the box. (The stand-off insulators furnished with the unit should be removed and the tuner mounted directly by its metal feet.) The shafts are cut off so that they may be coupled to panel-bearing units. Holes for the power plug and coax connector should be punched at the rear of the box where they will not interfere with the placement of the unit in the box. Holes should also be drilled in the box for the output connector, J_2. This is placed near the junction of the two brackets, as seen in Fig. 6-94.

Three-terminal tie points should be fastened underneath the 6AQ5 bracket for mounting the four resistors. The grid choke, RFC_1, is mounted on a small angle fastened underneath the 6146 bracket.

In Fig. 6-95, the plate r.f. choke is threaded onto one of the assembly screws, while the plate by-pass condenser, C_3, is fastened by its lower terminal with a screw tapped into the condenser frame. The plate blocking condenser, C_4, is mounted by soldering one of its terminals to the rear stator terminal of the tank condenser. The parasitic choke, L_1, is wound in the lead from the outer end of the plate r.f. choke to the 6146 plate cap. A short lead connects one of the link terminals to J_2, while the other link terminal is grounded to one of the J_2 mounting screws.

All power wiring should be done with shielded wire. This includes all wiring to the clamper tube.

Depending on the most convenient arrangement in combining the amplifier with other units, it may be used either in a horizontal position, as shown in Fig. 6-92, or in a vertical position. In the latter case, ventilating holes should be drilled in the side that will be the top.

The following section discusses power supply for and operation of this amplifier.

(Originally described in *QST*, May, 1953.)

Fig. 6-94 — Side view of the multiband amplifier.

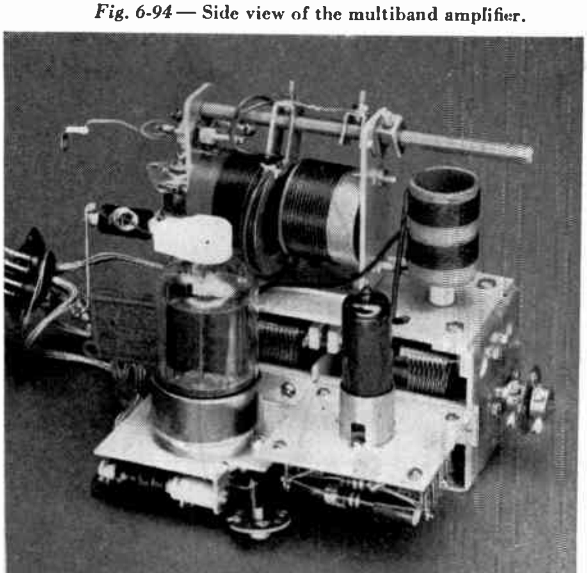
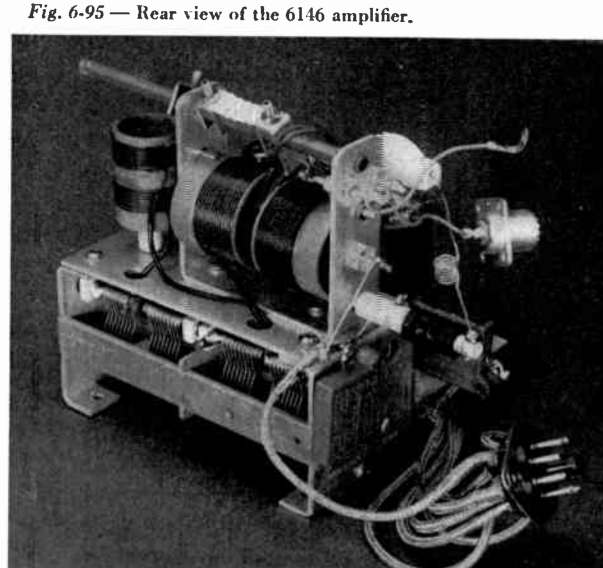


Fig. 6-95 — Rear view of the 6146 amplifier.



A Bandswitching 80-Watt Transmitter

Figs. 6-96 through 6-99 show how the three preceding units may be combined to form a bandswitching VFO transmitter covering 6 bands. Power supply for all units is included.

Circuit Considerations

Fig. 6-97 shows the circuit external to the individual units. High voltage for the 6146 amplifier is obtained from an inexpensive b.c. transformer, T_1, working into a bridge rectifier. The supply delivers 550 volts at a full load to the amplifier of 150 ma. on c.w. A choke-input filter is used with this supply, and the 6X5GT rectifier filaments are operated from the 6.3-volt winding of this transformer. The required filter-condenser voltage rating is obtained by connecting 500-volt electrolytics in series. A supply voltage of 350 for the VFO, frequency multiplier, and the screen of the 6146 is obtained from a second supply using a condenser-input filter. All transmitter-tube heaters are operated from the 6.3-volt winding of the low-voltage transformer.

Provision is made for the external connection of a plate modulator. There are Millen safety terminals at the rear for connecting in the output of the modulator, and an audio choke is included for the screen circuit. Also, an a.c. outlet and switch, S_1, are available for the modulator power supply.

The rotary switch, S_3, performs, in a single operation, the combined duties of power control and meter switching. In the mid-position, the meter is switched to read grid current, plate voltage is removed from the 6146, and the screen is grounded. While the switch is in this position, the VFO may be set to frequency and the frequency multiplier tuned for the desired amplifier grid current without putting a signal on the air. When the switch is thrown to the right, plate and screen voltages are applied to the amplifier, and the meter reads plate current with a 10-times shunt across the meter. This is the operating position for c.w.

For 'phone operation, the switch is thrown to the left, instead of to the right. In this position, the modulator-input terminals are connected in the plate circuit of the 6146, and the choke, L_1, is inserted in the screen lead.

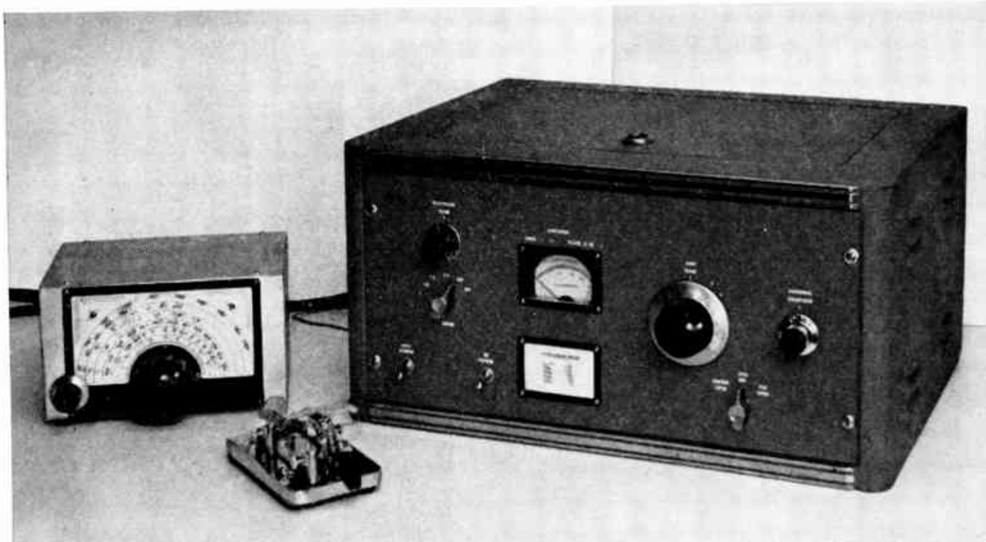
Construction

The components are assembled on a 13 × 17 × 3-inch steel chassis, with a 3/16-inch aluminum rack panel 8¾ inches high. An aluminum chassis could be used, but is not necessary since the units are already shielded in aluminum boxes. Also, since each of the r.f. units is shielded and filtered, the power wiring has been done with ordinary unshielded wire, although, of course, there is no objection to the use of shielded wire.

The frequency-multiplier unit is placed at the left-hand end of the chassis, far enough to the rear to allow space for shaft couplings between the panel and the shaft bearings. To facilitate the use of coax-cable r.f. connections between the units, the bakelite pin-jack input and output terminals of the frequency-multiplier unit were replaced with shielded 'phono jacks, similar to those suggested for the VFO and amplifier units.

The amplifier is mounted at the right-hand end of the chassis. If the shafts of the panel bearings for the amplifier unit are long enough, the unit can be spaced back of the panel the same distance as the multiplier unit. In this case, the link shaft is run through a ¼-inch hole in the

Fig. 6-96 — An 80-watt 6-band bandswitching transmitter built around VFO, frequency-multiplier and output-stage units described in preceding sections.



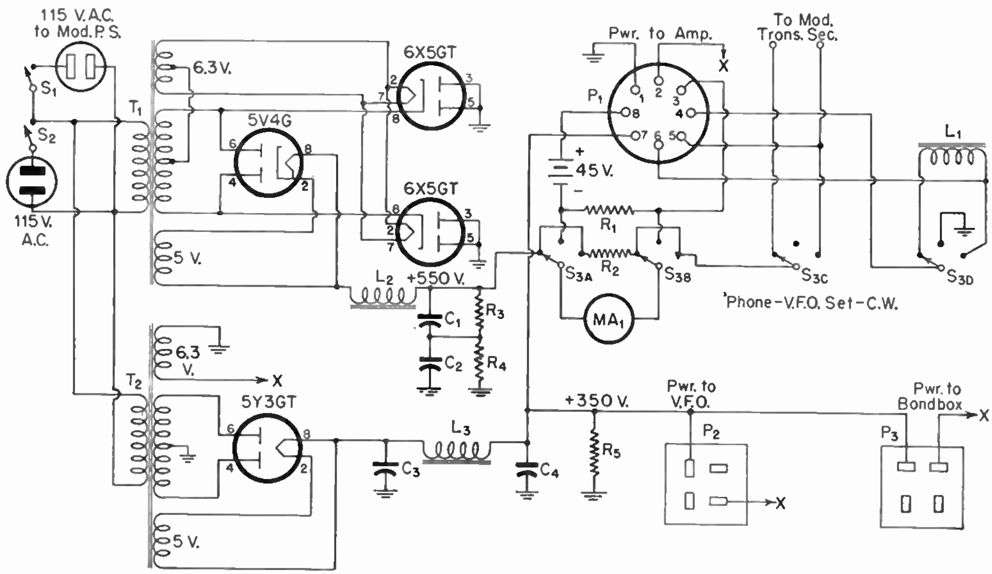


Fig. 6-97 — Power and control circuits for the 80-watt multiband transmitter.

C_1, C_2, C_3, C_4 — 16-μfd. 500-volt electrolytic (e.g., Aerovox PRS-16).
R_1 — 100 ohms, ½ watt.
R_2 — 10-times shunt for 25-ma. meter (No. 30 wire wound on 1-watt resistor of 50 ohms or more and connected to its terminals. Adjust turns to required multiplication, see instruments chapter).
R_3, R_4 — 20,000 ohms, 10 watts.
R_5 — 50,000 ohms, 10 watts.
L_1 — 35-hy. 15-ma. choke (e.g., Thordarson T-20C51).
L_2 — 7-hy. 150-ma. filter choke (e.g., Halldorson C-5027).
L_3 — 10.5-hy. 110-ma. filter choke (e.g., Stancor 1001).
MA_1 — 3-inch d.c. milliammeter, 25-ma. scale (e.g., Triplett 327A).
P_1 — Octal male connector (Amphenol 86-PF-8).
P_2 — Female cable connector to fit VFO connector (Jones S-304-CCT).
P_3 — Female cable connector to fit multiplier connector (Jones S-304-CCT).
S_1, S_2 — S.p.s.t. toggle switch.
S_3 — 4-wafer 3-position rotary switch, bakelite insulation (e.g., Centralab 1427, or assembled from Switchkit parts).
T_1 — Power transformer: 375–0–375 volts r.m.s., 150 ma.; 5 volts, 3 amp.; 6.3 volts, 4.7 amp. (e.g. Thordarson 24R06U).
T_2 — Power transformer: 360–0–360 volts r.m.s., 120 ma.; 5 volts, 3 amp.; 6.3 volts, 3.5 amp. (e.g., Halldorson P-9315).

panel, while the tuning-condenser shaft is cut to fit the insert of the National AM dial. If the shafts have been cut short, however, it will be necessary to move the unit farther back to make room for shaft couplings.

The VFO unit is mounted in the rear left-hand corner of the chassis, with the remote-tuning cable connector toward the left.

All three units are fastened down to the chassis with self-tapping screws from the bottom. If a steel chassis is used, the aluminum cover plates of the multiplier and amplifier units should be used between the chassis and the bottoms of the boxes, the self-tapping mounting screws going through both the chassis and the covers into the lips of the boxes. The paint on the chassis should be removed in the areas to be covered by the units, so that the shielding will make good electrical contact with the chassis. This operation can be made easier by using an application of paint remover, being careful not to allow the remover to creep beyond the limits of the edges of the

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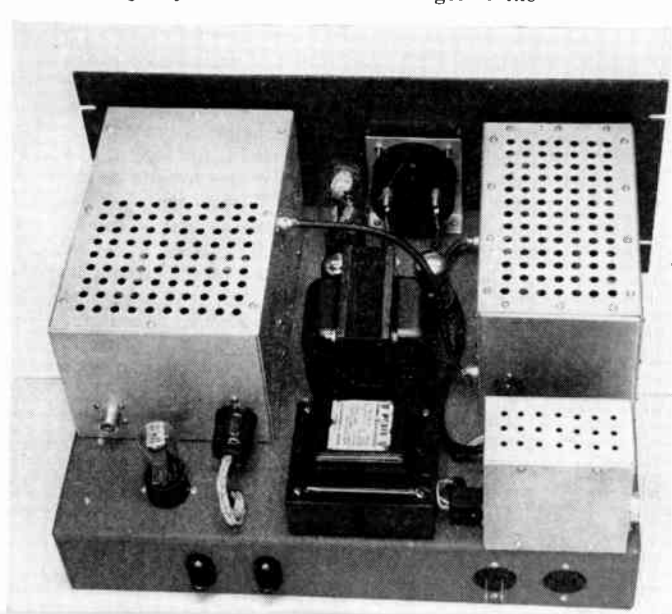
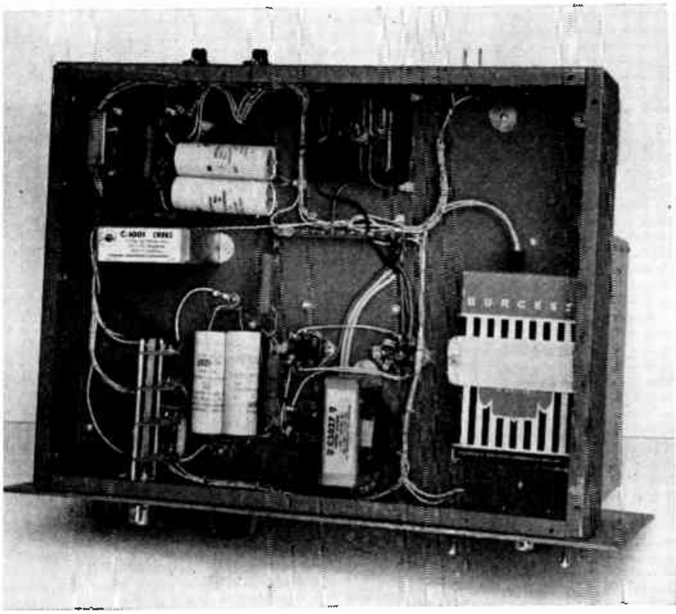


Fig. 6-98 — Inside view of the 80-watt multiband rig. Two power supplies are sandwiched in between the individually-shielded units. Along the rear are modulator-input terminals, a.c. input connector, and an outlet for an external modulator power supply.



CHAPTER 6

Fig. 6-99 — Bottom view of the 80-watt bandswitching transmitter. The low-voltage power-supply components and screen audio choke are in the upper left-hand corner. Below, at the center, are the filter components for the high-voltage supply. The biasing battery is held in place by an aluminum cleat and a pair of long machine screws. The power-control switch is at the lower left.

units and thus spoil the appearance of the chassis.

The low-voltage transformer, T_2, is centered between the two larger units, and placed as far to the rear as possible. A flush-mounting transformer happened to be on hand, but a vertical type is easier to mount, since it requires no large cutout on the chassis. The 5Y3GT is placed to the rear of the amplifier box, in such a position that it will not interfere with the power plug.

The high-voltage transformer, T_1, is mounted in front of the low-voltage unit with a space of about ¼ inch between the two. The two 6X5GTs are, in turn, mounted forward of the high-voltage transformer, spaced about 3 inches apart, center to center. The 5V4G is in front of the right-hand 6X5GT, leaving space for the meter.

The meter is mounted behind the panel as a safety measure. It is held in an aluminum bracket, with the face of the meter flush with the front edge of the chassis. The meter should be placed so that its center comes about 7 inches from the left-hand end of the chassis, and the center of its scale about 5½ inches up from the bottom edge. The meter-scale opening in the panel is cut out to fit the inside dimensions of a National CFA chart frame. If desired, a hole can be drilled in the panel to give access to the zero-adjust screw of the meter. A similar chart frame, centered 3½ inches below, helps to balance the panel layout and is very useful for logging the multiband-tuner settings for the various bands, since it is not too difficult to tune the amplifier up on a harmonic, instead of the desired fundamental, without realizing it.

The two power switches are placed either side of a line running through the two multiplier controls, and the rotary switch, S_3, is centered on a line between the two amplifier controls. These three switches are mounted 1½ inches up from the bottom edge.

Panel bearings for the controls are not used, the short extension shafts riding in ¼-inch holes, reamed out just enough to provide free turning without excessive play.

Holes lined with rubber grommets are drilled in the chassis to pass the power leads to the three units. R.f. connections between the VFO output and the multiplier input, and between the multiplier output and amplifier input are made with RG-59/U coax cable in lengths as short as possible to minimize capacitance and yet provide shielding. The plugs that fit the shielded 'phono jacks can be easily attached to the ends of the RG-59/U by baring the center conductor so that it will extend through to the tip of the plug, and fraying the braid out around the shell of the plug, and soldering. Be sure, however, to leave enough of the inner insulation so that the inner conductor does not short against the grounded shell. After the cables have been carefully made up, it would be well to check for short-circuits with an ohmmeter.

The filter and audio chokes can be seen in the bottom-view photograph. The filter condensers are also mounted under the chassis, supported at each end on terminal strips. The biasing battery is held in place with a simple clamping arrangement. A pair of 2-inch machine screws are spaced slightly greater than the width of the battery. They are fastened permanently in place with nuts. An aluminum cleat with holes to fit the machine-screw spacing is held down with nuts and lockwashers at the bottom ends of the screws.

The cabinet shown is a Par-Metal DL-128. A hole is cut in the left side, toward the rear, to line up with the connector in the VFO unit for the remote-tuning cable.

Adjustment

Adequate drive is obtained with the VFO screen operated from the tap between the two VR tubes in the VFO unit (approximately 108 volts). With a fixed bias of 45 volts and a 12,000-ohm grid leak, grid currents in excess of 3 ma. should be obtainable on all bands. It should be limited to 3 ma. by detuning the multiplier unit. If the power supply shown is duplicated, the high-voltage supply should deliver 550 volts under a load of 150 ma. plus bleeder current, making the operating input to the amplifier a little over 80 watts. For 'phone operation, the 40-watt 807 modulator shown in the chapter on speech amplifiers and modulators should be just about right for this transmitter. The modulation-transformer secondary should be set for 5000 ohms, and the plate current under modulation should be limited to 112 ma.

A High-Power Tetrode Amplifier

Figs. 6-100 through 6-105 show the construction of a high-power tetrode amplifier covering all bands from 3.5 to 29 Mc. It is capable of being operated at an input of 1 kw., although it will operate efficiently at less input.

The circuit is shown in Fig. 6-101. The tube is the type 4-250A. A National type MB-40L "all-band" tank is used in the grid circuit. This circuit is a combination of inductance and variable condensers that may be tuned to any of the above bands without switching or changing coils. A pi-section tank circuit is used in the output. It is designed to feed into a flat 52- or 75-ohm line, either feeding an antenna directly or through a conventional antenna coupler. A B & W rolling-type variable inductance makes coil switching unnecessary in this circuit also. L_2 is a separate inductance section for 28 Mc. S_1 selects the proper network output capacitance.

The amplifier is neutralized by the capacitive-bridge method. C_2 is the neutralizing condenser. L_1 and R_1 form a v.h.f. parasitic-suppressor circuit. The plate of the amplifier is parallel-fed through the special r.f. choke, RFC_4. All power leads are filtered for v.h.f. harmonics. B_1 is a small electric blower required as an aid in dissipating the heat developed inside the shielding enclosure. RFC_3 is a safety choke to provide a d.c. path to ground in case C_{27} breaks down. Otherwise, high voltage will appear on the output cable if the condenser fails.

Construction

The amplifier is assembled on a standard chassis, 17 × 10 × 3 inches, with a 10½-inch panel. The grid tuner is mounted in a separate shielding enclosure at the right-hand end of the chassis in Fig. 6-102. This box is 3½ inches wide, 5 inches high and 7 inches deep, made of 1/16-inch aluminum sheet. This same material is used throughout the construction. A coax fitting at the rear of the grid-tuner box is the input connector. The grid and neutralizing leads pass through the side of the box into the large compartment. The constructional details of the latter may be seen in Fig. 6-101. The over-all dimensions of this section are 13⅜ × 10 × 7⅛ inches high. Three-quarter-inch flanges are bent along all four edges of the side pieces. The front and back pieces have these lips only along the top edges, since they are made high enough to allow an overlap over the edge of the chassis at the bottom. All sides, except the top, are fastened together with 6-32 screws and nuts. The top lid is fastened down by tapping screw holes along the lips around the top edges, and is perforated with ¼-inch holes above the area of the tube.

It is important that the pieces for this enclosure be made accurately so as to leave no gap at any point. If necessary, the pieces can be made by a local sheet-metal worker.

The plate tank condenser is mounted centrally in the box, using sheet-aluminum brackets to space it from the bottom. The condenser is placed with its end plates running vertically, i.e., on its side. The variable inductance, L_3, is placed alongside the condenser with the small fixed coil, L_2, mounted by fastening one end to the forward right-hand terminal of the variable inductance and the other end to a lug under one of the rear condenser-stator nuts. A flexible strip of copper connects the coax output fitting to the rear terminal of the variable coil.

The output condensers, excepting C_4, are stacked up behind the variable coil and the selector switch. S_1, is mounted on a small bracket to the rear, so that a control shaft may be run to the panel in between the tank coil and condenser. C_4 is soldered directly across the output connector. It may be helpful to series-resonate this condenser at the frequency of a local TV station to minimize TVI. This can be done by adjusting the length of the condenser leads and checking with a grid-dip oscillator, as described in the

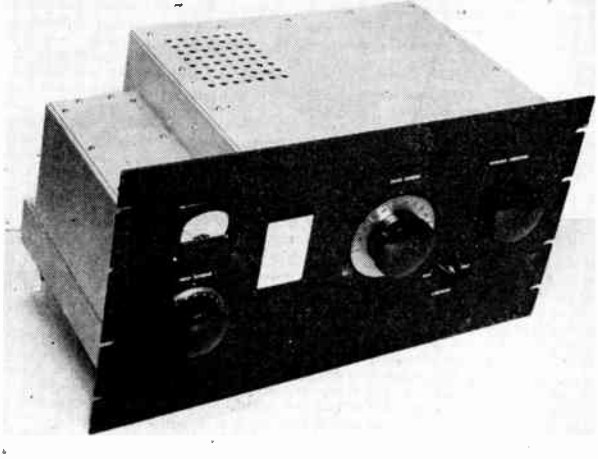


Fig. 6-100 — A high-power shielded tetrode amplifier. The small enclosure at the left contains an all-band tuner for the grid circuit. The dial near the center controls the input condenser of a pi-section output tank, while the knob at the right is the control for a roller-type variable inductance. The switch below selects the proper output capacitance.

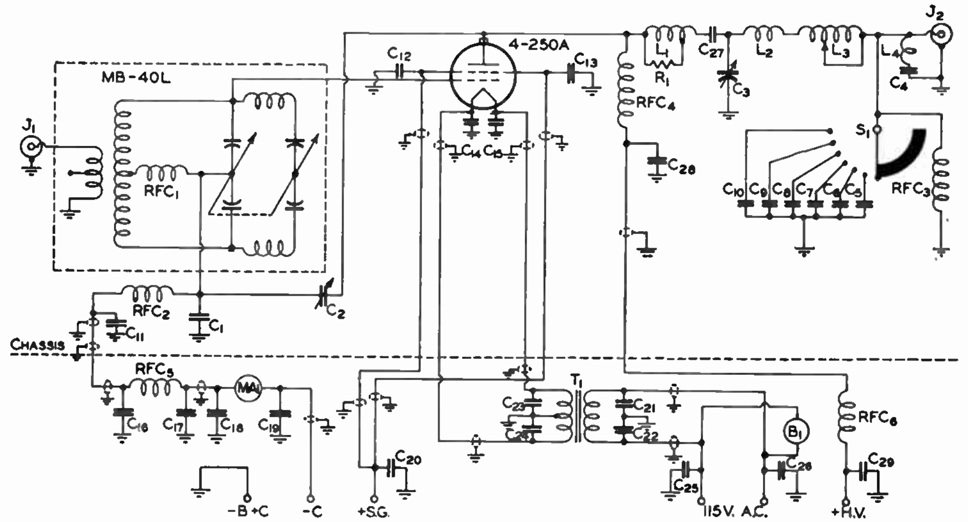


Fig. 6-101 — Circuit diagram of the amplifier. The broken line separates the above- and below-chassis wiring.

C_1 — 220-μμfd. mica.
C_2 — Disc-type neutralizing condenser, approx. 2 μμfd. with at least ¼-inch spacing (National NC-800A).
C_3 — 150-μμfd. variable, 6000 volts, 0.171-inch spacing (National TMA-150A).
C_4, C_5, C_6 — 100-μμfd. mica, 2500 volts.
C_7, C_8 — 220-μμfd. mica, 2500 volts.
C_9, C_{10} — 470-μμfd. mica, 2500 volts.
C_{11} to C_{22}, inc. — 0.001-μfd. disc ceramic, 600 volts.
C_{23}, C_{24}, C_{25}, C_{26} — 0.005-μfd. disc ceramic, 600 volts.
C_{27}, C_{28}, C_{29} — 500-μμfd. ceramic, 10,000 volts (Centralab TV3-501).
R_1 — Five 680-ohm 1-watt carbon resistors in parallel.
L_1 — Parasitic coil, 5½ turns No. 14, ¼-inch diam. R_1 tapped across 3 turns.
L_2 — 5 turns No. 10, 2½ inches long, 1½-inch diam.
L_3 — Variable inductor, 15 μh. max. (B & W 3852).
L_4 — To series-resonate with C_4 at desired TV frequency.
B_1 — Blower and motor, 115 v. a.c. (available from Allied Radio, Chicago, catalog No. 72-702 motor and 72-703 fan).
J_1, J_2 — Coaxial connectors, chassis-mounting type.
MA_1 — 0-50 ma. d.c. milliammeter.
RFC_1, RFC_2, RFC_3 — 2.5-mh. r.f. choke (NOTE: RFC_1 is not supplied with the National MB-40L multiband unit).
RFC_4 — National type R-175 choke modified as shown in Fig. 6-111.
RFC_5, RFC_6 — 2-μh. r.f. choke, 500 ma. (National R-60).
S_1 — Single-circuit 7-position ceramic switch, progressive shorting (Centralab type P-1-S wafer).
T_1 — Filament transformer, 5v. 13 amp. (UTC S-59).

chapter on TVI. At the lower TV frequencies, the condenser lead can be formed into a small coil of a turn or so.

The plate-feed r.f. choke, RFC_4, is placed to the rear of the tank condenser. To be effective on all bands, including the 21-Mc. band, it is necessary to alter the windings slightly, as shown in Fig. 6-103. It is a good idea to check the choke for resonances with a grid-dip oscillator after it has been placed in the position it is to occupy, but before it has been wired in, because proximity to surrounding components and shielding may affect the resonances. Performance of the choke will be poor at any frequency where the g.d.o. shows a resonance *with the terminals of the choke short-circuited.*

The tube socket is mounted above the chassis on spacers that are just long enough so that the shielded wires going to the screen and filament terminals, with their by-pass condensers, just span the distance between the socket terminals and lugs fastened to the chassis below each terminal. The lead then immediately passes through the chassis. Strips of copper sheet connect the plate terminal of the tube to the top terminal of the plate choke and the rotor terminal of the neutralizing condenser mounted on the right-hand wall of the enclosure, as shown in Fig. 6-102. The strips should be fitted carefully so as to avoid placing any strain on the cap terminal of the tube. The filament transformer is fastened down in the forward right-hand corner. Power terminals are lined up along the rear edge of the chassis. All r.f. grounds should be made directly to the chassis with the shortest possible lead length — even a half inch is worth saving.

Underneath, the d.c. and a.c. leads come out in shielded wire. A 0.001-μfd. disc ceramic by-pass is used across both ends of each lead excepting the high-voltage lead (see TVI chapter for method of connection). The high-voltage lead is by-passed with TV filter capacitors. RFC_6 is installed close to the high-voltage terminal. C_{20}, C_{25}, C_{26} and C_{29} likewise are fastened directly to the power terminals where the leads leave the chassis. The shielding of the power leads is grounded to the chassis by soldering to lugs wherever they pass through the chassis. The power wires are intentionally made to follow long paths around the edge of the chassis to provide additional harmonic attenuation. The braid is grounded at frequent intervals by soldering to lugs that also serve as hold-downs.

The blower is mounted on a bracket formed from a strip of aluminum. Air is forced through a set of holes in the chassis that duplicate in size and arrangement the holes in the 4-250A socket. The filament-transformer terminals project through clearance holes drilled in the chassis, and the four v.h.f. by-pass condensers, C_{21}, C_{22}, C_{23} and C_{24}, are connected directly from the terminals to grounding lugs.

Adjustment

The diagram of a suitable power supply for this amplifier is shown in Fig. 6-105. With 150 volts bias, a grid current of about 25 ma. is optimum, although the plate efficiency will change but little with any grid current between 15 and 30 ma. The single fixed link provided with the grid tuner will not provide uniform loading of the driver stage with coax input, so means should be provided in the output circuit of the driver for varying the coupling.

Optimum screen voltage is about 400 and the screen current should run between 50 and 75 ma., depending on the plate voltage used. At 2750 volts, a full kilowatt can be run to the amplifier, but it will work well at plate voltages as low as 1500, with a plate current of 350 ma.

It is important that the coaxial line into which the amplifier works be closely matched (see transmission-line chapter) at its terminating end, otherwise there is danger of damage to the mica output condensers. To protect the contacts on the variable inductance, adjustments should be made with litle or no power input to the amplifier. Experience will show where the tap should be placed for each band and thereafter it can be preset before applying full power. When reducing plate voltage, provision should also be made for reducing screen voltage, since otherwise the screen current may run to dangerous proportions.

It is advisable to set the tank condenser so as to operate the output circuit at a Q in the neighborhood of 12, as shown in the graph of Fig. 6-9, although it may not be possible to attain this figure at the extremes of the tuning range.

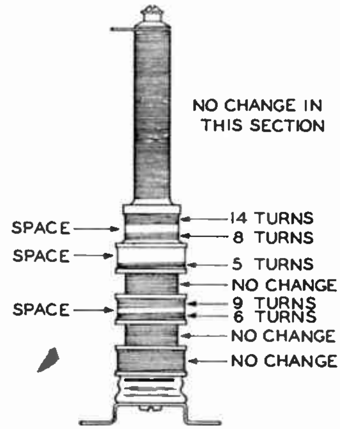
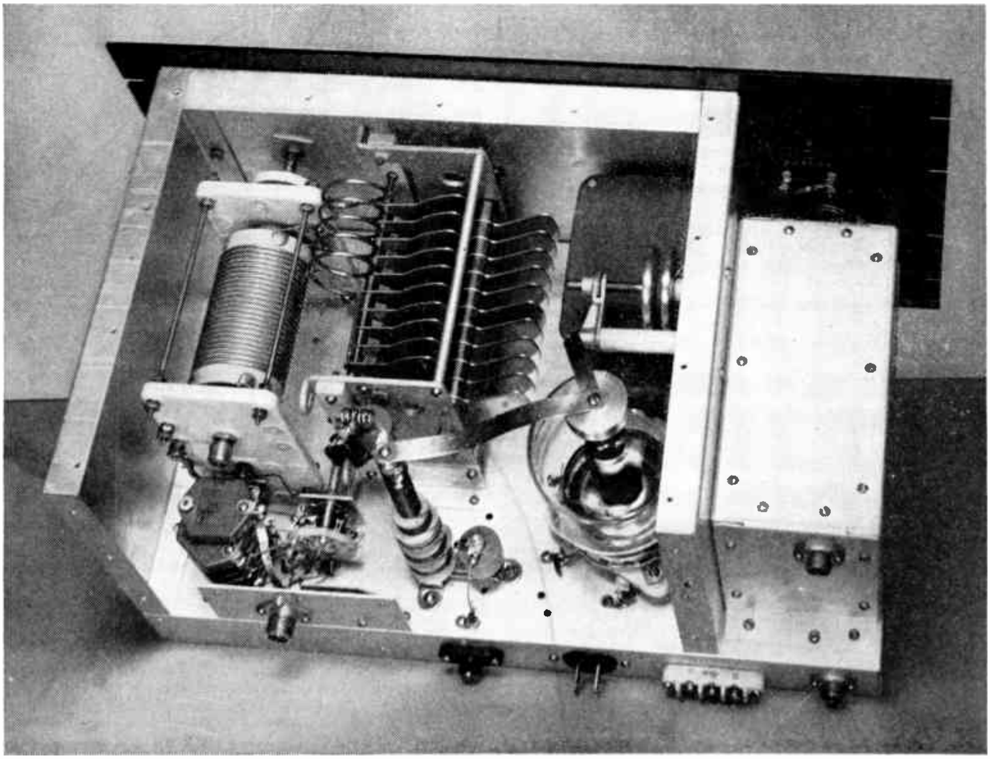


Fig. 6-103 — The R-175 choke as modified to work on all amateur bands in the 3.5- to 30-Mc. range, including 21 Mc.

Fig. 6-102 — Interior of the shielding compartment housing the 4-250A and its output circuit. The neutralizing condenser and filament transformer may be seen in the forward right-hand corner.



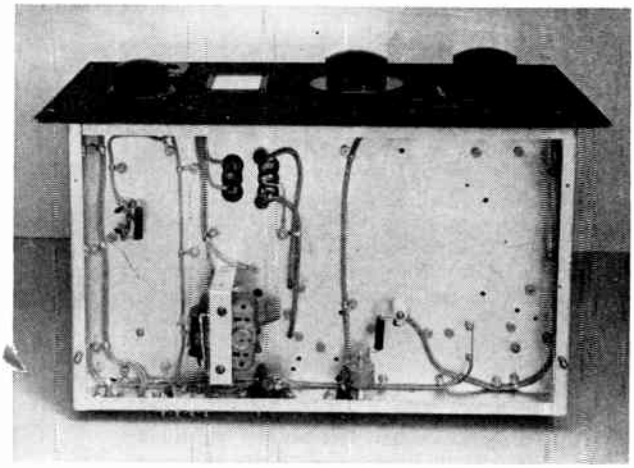


Fig. 6-104 — Bottom view of the high-power tetrode amplifier, showing the small ventilating fan and the shielded power wiring. No bottom plate on the chassis is necessary.

The neutralizing condenser should be adjusted for minimum reaction on the grid current under actual operating conditions. The approximate setting can be determined by the use of a grid-dip oscillator tuned to the operating frequency. All voltages should be removed and the g.d.o. coupled to the plate tank circuit. The neutralizing condenser should be adjusted for minimum r.f. in the grid tank circuit when both tanks are tuned to resonance. R.f. in the grid circuit can be checked with the aid of an indicating wavemeter of the type described in the measurements chapter. Final touching up can be done after checking the operation with voltages applied to the tube. In connection with the neutralizing circuit, the value of C_1 is fairly critical, but a capacitance within usual tolerance of the marked value should be satisfactory.

In adjusting the loading on the amplifier, increasing the output capacitance, or increasing the inductance, or both, while maintaining resonance with the tank condenser, will reduce the loading and vice versa.

(Originally described in *QST*, Oct. 1952.)

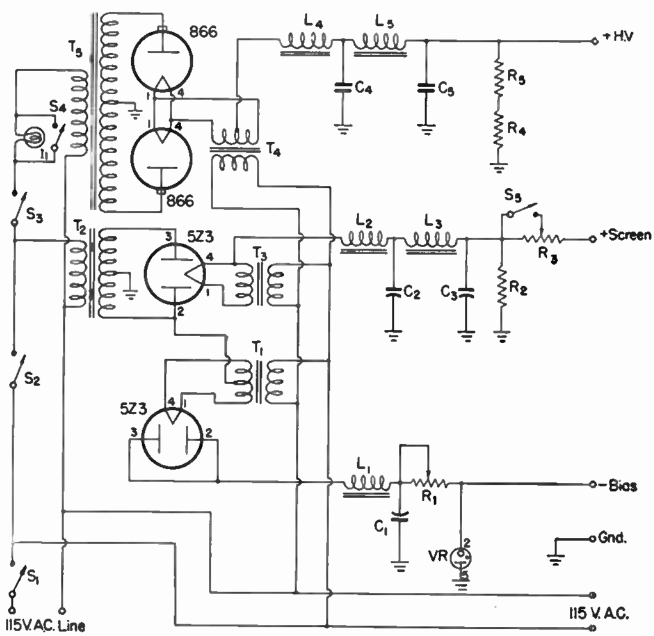


Fig. 6-105 — Circuit diagram of a power-supply system for the high-power tetrode amplifier.

C_1 — 8-μfd. 450-volt electrolytic.
C_2, C_3 — 4-μfd. 600-volt electrolytic.
C_4 — 2-μfd. oil-filled, voltage rating same as transformer r.m.s.
C_5 — 4-μfd. oil-filled, voltage rating same as transformer r.m.s.
R_1 — 25,000 ohms, 25 watts.
R_2 — 25,000 ohms, 50 watts.
R_3 — 50,000 ohms, 50 watts.
R_4, R_5 — 25,000 ohms, 100 watts.
L_1 — 30-hy. 50-ma. filter choke.
L_2 — 5/25-hy. 150-ma. swinging.
L_3 — 20-hy. 150-ma. smoothing.
L_4 — 5/25-hy. 500-ma. swinging.
L_5 — 20-hy. 500-ma. smoothing.
I_1 — 115-volt lamp of suitable size to reduce voltage for tune-up.
S_1 — 20-amp. s.p.s.t. switch.
S_2, S_3, S_4 — 15-amp. s.p.s.t. switch.
S_5 — Ceramic s.p.s.t. rotary switch.
T_1, T_3 — Filament transformer: 5 volts, 3 amp.
T_2 — Plate transformer: 400 volts d.c., 150 ma.
T_4 — Filament transformer: 2.5 volts, 10 amp., 10,000-volt insulation.
T_5 — Plate transformer: up to 2750 volts d.c., 350 ma.
VR — VR-150-30.

S_1 turns on all filaments and the bias supply. S_2 turns on the screen supply and S_3 the high-voltage supply. With S_4 open, a 115-volt lamp is inserted in series with the high-voltage-transformer primary to lower plate voltage for adjustment. Opening S_5 likewise reduces screen voltage. With all switches except S_2 closed, S_2 becomes the main control switch. The tap on R_3 should be adjusted to give the desired screen voltage under operating conditions with S_5 closed. Bias is obtained from the parallel-connected 5Z3 half-wave rectifier. The tap on R_1 should be adjusted until the VR tube just ignites without excitation to the amplifier.

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, 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.

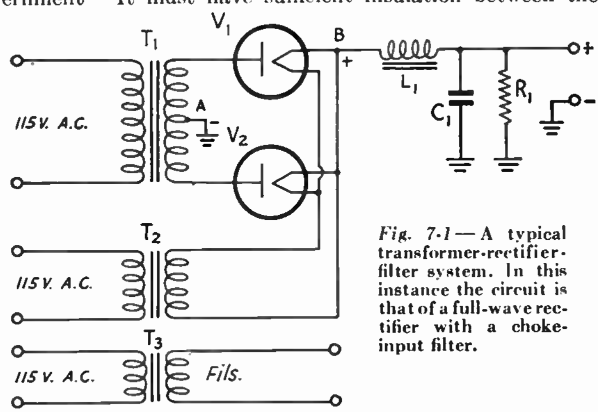


Fig. 7-1 — A typical transformer-rectifier-filter system. In this instance the circuit is that of a full-wave rectifier with a choke-input filter.

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 waveform 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 which may be drawn from this circuit is twice the rated load current of a single rectifier.

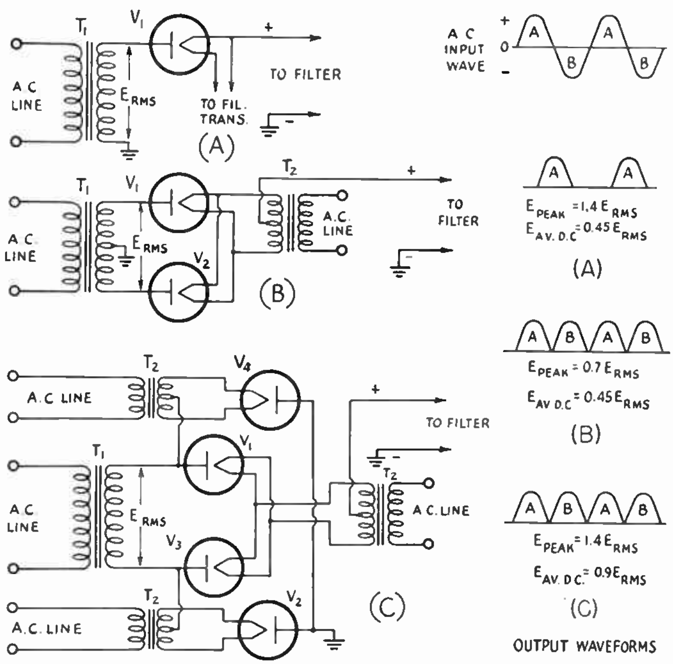


Fig. 7-2 — **Fundamental vacuum-tube rectifier circuits. A — Half-wave. B — Full-wave. C — Full-wave bridge. A.c.-input and pulsating-d.c. output wave forms are shown at the right. Output-voltage values indicated do not include rectifier drops. Other types of rectifiers may be substituted in these circuits.**

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.

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 waveshape (C), to the right, is the same as that from the simple center-tap rectifier circuit. The ouptut voltage obtainable with this circuit is 0.9 times the r.m.s. voltage delivered by the transformer secondary. For the same total transformer-secondary voltage, the average output voltage when using the bridge rectifier will be twice that obtainable with the center-tap rectifier circuit. However, when comparing rectifier circuits for use *with the same transformer*, it should be remembered that the *power* which a given transformer will handle remains the same regardless of the rectifier circuit used. If the output voltage is doubled by substituting the bridge circuit for the center-tap rectifier circuit, only half the rated load current can be taken from the transformer without exceeding its normal rating. The value of load current which may be drawn from the bridge rectifier circuit is twice the rated d.c. load current of a single rectifier.

Rectifiers

Cold-Cathode Rectifiers

Tube rectifiers fall into three general classifications as to type. The cold-cathode type is a diode which requires no cathode heating. Certain types will handle up to 350 ma. at 200 volts d.c. output. The internal drop in most types lies between 60 and 90 volts. Rectifiers of this kind are produced in both half-wave (single-diode) and full-wave (double-diode) types.

High-Vacuum Rectifiers

High-vacuum rectifiers depend entirely upon the thermionic emission from a heated 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 250 ma. at 400 to 500 volts d.c. output. Those in the higher-power class can be used to handle up to 500 ma. at 2000 volts d.c. in full-wave circuits. Most low-power high-vacuum rectifiers are produced in the full-wave type, while those for greater power are invariably of the half-wave type. A few of the lower-voltage types have indirectly heated cathodes, but are limited in heater-to-cathode voltage rating.

Mercury-Vapor Rectifiers

In mercury-vapor rectifiers the internal resistance is reduced by the introduction of a small amount of mercury which vaporizes under the heat of the filament, the vapor ionizing upon the application of voltage. The voltage drop through a rectifier of this type is practically constant at approximately 15 volts regardless of the load current. Tubes of this type are produced in sizes that will handle any voltage or current likely to be encountered in amateur transmitters. For high power they have the advantage of cheapness. Rectifiers of this type, however, have a tendency toward a type of oscillation which produces noise in near-by receivers. This can usually be eliminated by suitable filtering.

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

Selenium Rectifiers

Selenium rectifiers are available which make it possible to design a power supply capable of delivering up to 400 or 450 volts, 200 ma. These units have the advantages of compactness, low internal voltage drop (about 5 volts), and the fact that no filament transformer is needed. However, to limit the charging current with condenser input, a resistance of 25 to 100 ohms should be used in series with the rectifier. They may be substituted in any of the basic circuits shown in Fig. 7-2, the terminal marked "+" or "cathode" corresponding to the filament in these circuits. Circuits in which the selenium rectifier is particularly adaptable are shown later in Figs. 7-20 through 7-22. Since they develop little heat if operated within their ratings, they are especially suitable for use in equipment requiring minimum temperature variation.

Typical ratings are listed in the tube tables.

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 condenser-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.

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 condenser-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 a few per cent above 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

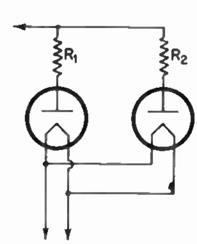


Fig. 7-3 — Connecting mercury-vapor rectifiers in parallel for heavier currents. R_1 and R_2 should have the same value, between 50 and 100 ohms, and corresponding filament terminals should be connected together.

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.

Rectifiers may be connected in parallel for current higher than the rated current of a single unit. This includes the use of the sections of a double diode for this purpose. Equalizing resistors of 50 to 100 ohms should be connected in series with each plate, as shown in Fig. 7-3, as a measure toward maintaining an equal division of current 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 *voltage regulation* of the power supply and the maximum load current that can be drawn from the supply without exceeding the peak-voltage rating of the rectifier.

Power-supply filters fall into two classifications, depending upon whether the first filter element following the rectifier is a condenser or a choke. Condenser-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 condenser-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 condenser. 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.

$$\text{Per cent regulation} = \frac{100\,(E_1 - E_2)}{E_2}$$

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 condenser-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 condensers 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 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-winding resistance and the rectifier resistance is called the input resistance.

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 condensers 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 condensers which short-circuit the a.c. component while not interfering with the flow of the d.c. component, and series chokes which pass d.c. readily but which impede the flow of the a.c. component.

The alternating component is called the ripple. The effectiveness of the filter can be expressed in terms of per cent ripple, which is the ratio of the r.m.s. value of the ripple to the d.c. value in terms of percentage. For c.w. transmitters, a reduction of the ripple to 5 per cent is considered adequate.

The ripple in the output of power supplies for voice transmitters and VFOs should be reduced to 0.25 per cent or less. High-gain speech amplifiers and receivers may require a reduction to as low as 0.1 per cent.

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 lower.

CONDENSER-INPUT FILTERS

Condenser-input filter systems are shown in Fig. 7-4. Disregarding voltage drops in the chokes, all have the same characteristics except

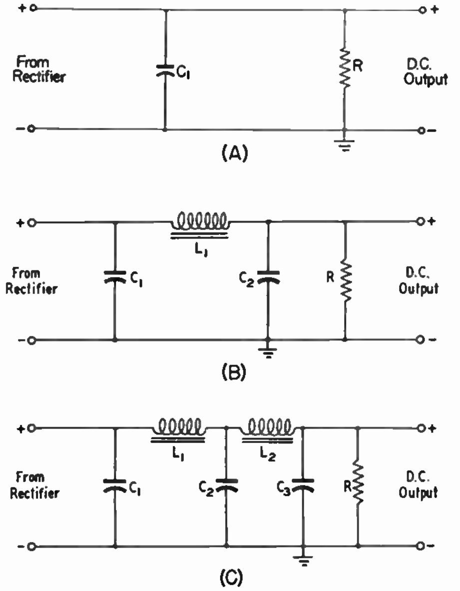


Fig. 7-4 — **Condenser-input filter circuits. A — Simple condenser. B — Single-section. C — Double-section.**

in respect to ripple. Better ripple reduction will be obtained when *LC* sections are added, as shown in Figs. 7-4B and C.

Output Voltage

To determine the approximate d.c. voltage output when a condenser-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.

$$\text{Load resistance} = \frac{350}{0.175} = 2000 \text{ ohms approx.}$$

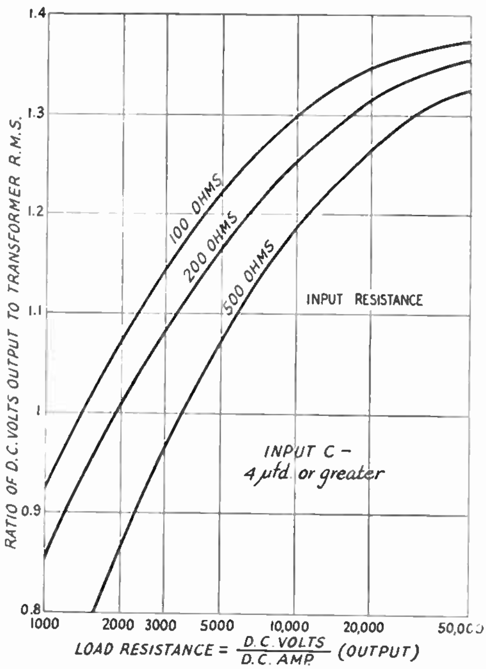


Fig. 7-5 — **Chart showing approximate ratio of d.c. output voltage across filter input condenser to transformer r.m.s. secondary voltage for different load and input resistances.**

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.

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 condenser-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 load current that can be drawn from a supply with a condenser-input 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

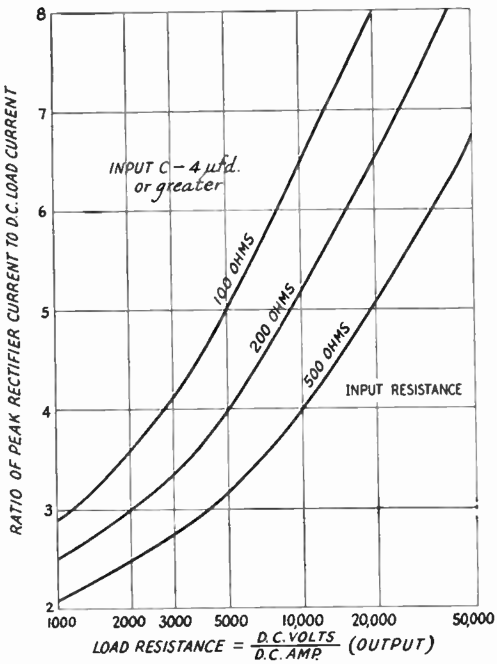


Fig. 7-6 — Graph showing the relationship between the d.c. load current and the rectifier peak plate current with condenser input for various values of load and input resistance.

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.

Ripple Filtering

The approximate ripple percentage after the simple condenser 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-μfd. condenser or 20% with a 4-μfd. condenser. For other capacitances, the ripple will be in inverse proportion to the capacitance, e.g., 5% with 16 μfd., 40% with 2 μfd., etc.

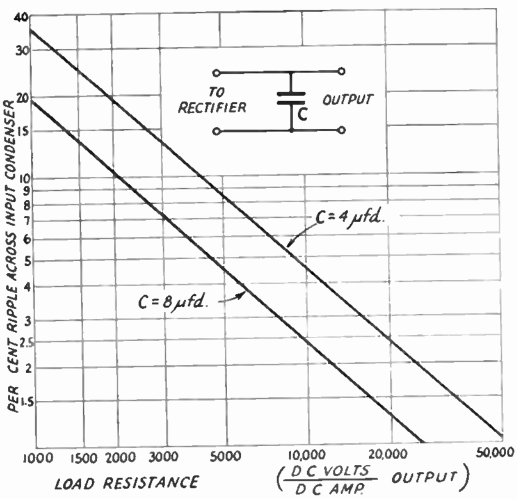


Fig. 7-7 — Showing approximate 120-cycle percentage ripple across filter input condenser for various loads.

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 hy. and a condenser of 4 μfd. were to be added to the simple condenser of Fig. 7-4A, the product is 4 × 5 = 20. Fig. 7-8 shows that the original ripple (10% as above with 8 μfd. for example) will be reduced by a factor of about 0.08. Therefore the ripple percentage after the new section will be

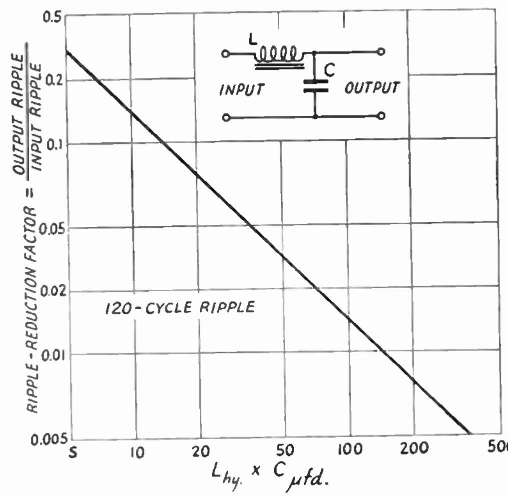


Fig. 7-8 — Ripple-reduction factor for various values of *L* and *C* in filter section. *Output ripple = input ripple × ripple factor.*

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, etc.

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.

If the first choke has a value equal to or greater than

$$L_{(hy.)} = \frac{\textit{Load resistance (ohms)}}{1000},$$

the output voltage will not soar above the average value of the rectified wave at the input of the choke when the load current is small. This is in contrast to the performance of the condenser-input filter where the output voltage tends to soar toward the peak value at light current loads. This value of inductance is known as the **critical** value.

If the first choke has a value equal to or greater than

$$L_{(hy.)} = \frac{\textit{Load resistance (ohms)}}{500}$$

the peak rectifier current will not exceed the d.c. load current by more than 10 per cent when the

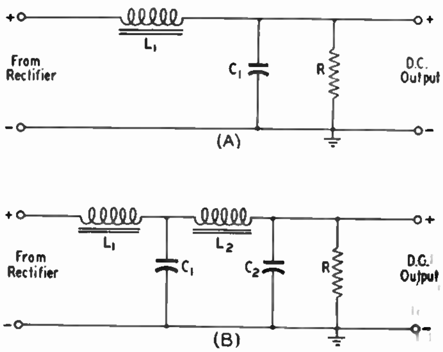


Fig. 7-9 — **Choke-input filter circuits. A — Single-section. B — Double-section.**

load current is large. This is in contrast to the condenser-input filter where the peak rectifier current may run 2 to 5 times the d.c. load current. This value of inductance is known as the **optimum** value.

Both of the above conditions will usually be satisfied for all values of load current drawn from the supply if the choke has at least the critical value of inductance for the minimum current load (usually the bleeder resistance only) and does not fall below the optimum value for the greatest current load to be drawn.

Specially-designed input chokes, called swinging chokes, are available. These chokes are usually rated in terms of maximum d.c. current and the range of inductance over which they are designed to "swing" with different load currents. For instance, a choke may have a rating of 5 to 25 hy., 250 ma. This means that the inductance is 5 hy. with 250 ma. d.c. flowing through it.

From the formula for optimum inductance, 5 hy. is *optimum* for a *minimum* load resistance of 5 × 500 = 2500 ohms. (At 250 ma., this resistance means a *minimum* voltage of 2500 × 0.250 = 625 volts — at higher voltages than 625, at the same current, the resulting load resistance will be higher. Therefore, the choke will have at least optimum inductance for all higher voltages.)

Bleeder Resistance

Also, 25 hy. is the *critical* inductance for 25 × 1000 = 25,000 ohms. Therefore the bleeder resistance should be *not greater than* 25,000 ohms.

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. A higher bleeder resistance drawing less current can be used, of course, but at a sacrifice in regulation. 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 to advantage in a c.w. transmitter is to use a very high-resistance bleeder for protective purposes and then use 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.

Output Voltage

Provided the input-choke inductance is at least the *critical* value, the output voltage may be calculated quite closely by the following equation:

$$E_o = 0.9E_t - \frac{(I_B + I_L)\,(R_1 + R_2)}{1000} - E_r$$

where E_o is the output voltage; E_t is the r.m.s. voltage applied to the rectifier (r.m.s. voltage between center-tap and one end of the secondary in the case of the center-tap rectifier); I_B and I_L are the bleeder and load currents, respectively, in milliamperes; R_1 and R_2 are the resistances of the first and second filter chokes; and E_r is the drop between rectifier plate and cathode. The various voltage drops are shown in Fig. 7-11. 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

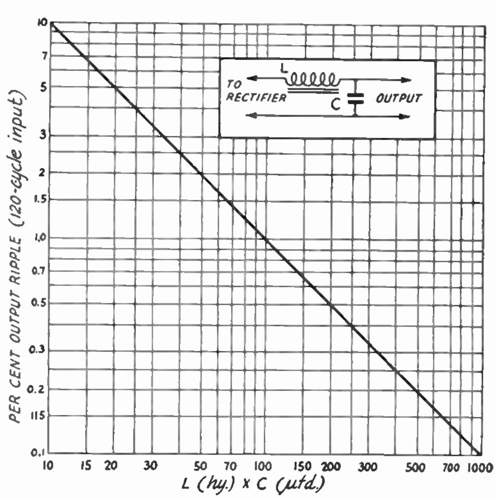


Fig. 7-10 — **Graph showing combinations of inductance and capacitance that may be used to reduce ripple with a single-section choke-input filter.**

a close approximation, for a ripple frequency of 120 cycles, from Fig. 7-10.

Example: $L = 5$ h., $C = 4$ μfd., $LC = 20$.

From Fig. 7-10, percentage ripple = 5 per cent.

Example: $L = 5$ hy. 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 $LC = 100$. $100/5 = 20$ μfd.

In selecting values for the first filter section, the inductance of the choke should be determined by the considerations discussed previously. Then the condenser 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 condenser-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.

OUTPUT CONDENSER

If the supply is intended for use with an audio-frequency amplifier, the reactance of the last filter condenser should be small (20 per cent or less) compared with the other 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 μfd., the higher value of capacitance being used in the case of lower tube and load resistances.

RESONANCE

Resonance effects in the series circuit across the output of the rectifier which is formed by the first choke (L_1) and first filter condenser (C_1) must be avoided, since the ripple voltage would build up to large values. This not only is the opposite action to that for which the filter is intended, but also may cause excessive rectifier peak currents and abnormally-high inverse peak voltages. For full-wave rectification the ripple frequency will be 120 cycles for a 60-cycle supply, and resonance will occur when the product of choke inductance in henrys times condenser capacitance in microfarads is equal to 1.77. The corresponding figure for 50-cycle supply (100-cycle ripple frequency) is 2.53, and for 25-cycle supply (50-cycle ripple frequency) 13.5. At least twice these products of inductance and capacitance should be used to ensure against resonance effects. With a swinging choke, the minimum rated inductance of the choke should be used.

RATINGS OF FILTER COMPONENTS

Although filter condensers in a choke-input filter are subjected to smaller variations in d.c. voltage than in the condenser-input filter, it is advisable to use condensers rated for the peak transformer voltage in case the bleeder resistor should burn out when there is no load on the power supply, since the voltage then will rise to the same maximum value as it would with a filter of the condenser-input type.

In a condenser-input filter, the condensers should have a working-voltage rating at least as high, and preferably somewhat higher, than the peak-voltage rating of the transformer. Thus, in the case of a center-tap rectifier having a transformer delivering 550 volts each side of the center-tap, the minimum safe condenser voltage rating will be 550×1.41 or 775 volts. An 800-volt condenser should be used, or preferably a 1000-volt unit.

Filter condensers are made in several different types. Electrolytic condensers, 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. Condensers 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 condensers 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). Condensers with higher-voltage ratings usually are made with a dielectric of thin paper impregnanted with oil. The **working voltage** of a condenser is the voltage that it will withstand continuously.

The input choke may be of the swinging type, the required minimum no-load and full-load inductance values being calculated as described above. For the second choke **(smoothing choke)** values of 10 to 20 henrys ordinarily are used. Since chokes usually are placed in the positive leads, the negative being grounded, the windings should be insulated from the core to withstand the full d.c. output voltage of the supply and be capable of handling the required load current.

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.

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:

$$E_t = 1.1\left[E_o + \frac{I(R_1 + R_2)}{1000} + E_r\right]$$

where E_o is the required d.c. output voltage, I is the load current (including bleeder current) in milliamperes, R_1 and R_2 are the d.c. resistances of the chokes, and E_r is the voltage drop in the rectifier. E_t is the full-load r.m.s. secondary voltage; the open-circuit voltage usually will be 5 to 10 per cent higher than the full-load value.

The approximate transformer output voltage required to give a desired d.c. output voltage

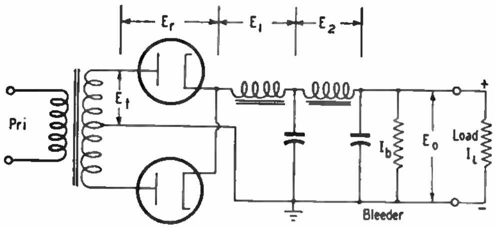


Fig. 7-11 — Diagram showing various voltage drops that must be taken into consideration in determining the required transformer voltage to deliver the desired output voltage.

with a given load with a condenser-input filter system can be calculated with the help of Fig. 7-11.

Example:
Required d.c. output volts — 500
Load current to be drawn — 100 ma.

$$\text{Load resistance} = \frac{500}{0.1} = 5000 \text{ ohms.}$$

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

$$E_t = \frac{E_o + I\left(\frac{R_1 + R_2 + R_r}{1000}\right)}{1.15}$$

$$= \frac{500 + 100\left(\frac{200 + 300 + 200}{1000}\right)}{1.15}$$

$$= \frac{570}{1.15} = 495 \text{ volts.}$$

Volt-Ampere Rating

The volt-ampere rating of the transformer depends upon the type of filter (condenser or choke input). With a condenser-input filter the heating effect in the secondary is higher because of the high ratio of peak to average current, consequently the volt-amperes consumed by the transformer may be several times the watts delivered to the load. With a choke-input filter, provided the input choke has at least the critical inductance, the secondary volt-amperes can be calculated quite closely by the equation:

$$\text{Sec. } V.A. = 0.00075EI$$

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.

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.

Rewinding Filament Transformers

Although the home winding of high-voltage transformers is a task that few amateurs undertake these days, the rewinding of a small-transformer secondary to give some desired filament voltage is not difficult. It involves a matter of only a small number of turns and the wire is large enough to be handled easily. Often a broadcast-receiver power transformer with a burned-out high-voltage winding, but with the primary winding intact, can be converted into an entirely satisfactory filament transformer without great effort.

The primary volt-ampere rating of a transformer to be rewound may be taken from the label on the transformer or from the manufacturer's catalogue. This will indicate whether or not the transformer will be capable of handling the necessary power. The secondary volt-ampere rating will be ten to twenty per cent less than the primary rating. The product of the voltage and the number of amperes required from the new filament winding, plus that for any other secondaries that may be kept in use, should not exceed the secondary volt-ampere rating, unless the builder is willing to accept a lower safety factor.

Before disconnecting the winding leads from their terminals, each should be marked for identification. In removing the core laminations, care should be taken to note the manner in which the core is assembled, so that the reassembling will be done in the same manner. Some transformers have secondaries wound over the primary, while in others the order is reversed. In case the secondaries are on the inside, the turns can be pulled out from the center after slitting and removing the fiber core.

The turns removed from one of the original filament windings of known voltage should be carefully counted as the winding is removed. This will give the number of turns per volt and the same figure should be used in determining the number of turns for the new secondary. For instance, if the old filament winding was rated at 5 volts and has 20 turns, this is $20/5 = 4$ turns per volt. If the new secondary is to deliver 7.5 volts, the required number of turns on the new winding will be $7.5 \times 4 = 30$ turns.

In winding a transformer, the size of wire is an important factor in the heat developed in operation. A cross-sectional area of 1000 circular mils per ampere is conservative. A value commonly used in amateur-service transformers is 700 c.m.p.a. The windings of some of the less-expensive broadcast-receiver transformers may run as low as 500 c.m.p.a. The larger the c.m.p.a. figure, the cooler the transformer will run. The current rating in amperes of each wire size shown in the miscellaneous data chapter at 1000 c.m.p.a. may be obtained by pointing off three decimal places from the right in the figures in the third column. Similar ratings at 700 c.m.p.a. are given in a separate column. Ratings at 500 c.m.p.a. will be twice the current rating at 1000 c.m.p.a. As an example, No. 18 has a current rating of 1.62 amperes at 1000 c.m.pa., 2.32 amperes at 700 c.m.p.a., or 3.25 amperes at 500 c.m.p.a. If the transformer being rewound is a filament transformer, it may be necessary to choose the wire size carefully to fit the small available space. On the other hand, if the transformer is a power unit, with the high-voltage winding removed, there should be plenty of room for a size of wire that will conservatively handle the required current.

The insulation to be used between the primary and secondary windings (and also between the secondary winding and the core if the secondary is on the inside) will depend on whether the transformer is to be used to supply r.f. tubes or rectifier tubes in a high-voltage supply. A few layers of linen paper should be sufficient for the former service, but insulating cambric sheet should be used if the voltage between primary and secondary runs over 1000 volts

Voltage Dropping

Series Voltage-Dropping Resistor

Certain plates and screens of the various tubes in a transmitter or receiver often require a variety of operating voltages differing from the output voltage of available power supplies. In most cases, it is not economically feasible to provide a separate power supply for each of the required voltages. If the current drawn by an electrode, or combination of electrodes operating at the same voltage, is reasonably constant under normal operating conditions, the required voltage may be obtained from a supply of higher voltage by means of a voltage-dropping resistor in series, as shown in Fig. 7-12A. The value of the series resistor, R_1, may be obtained from Ohm's Law, $R = \frac{E_d}{I}$, where E_d is the voltage *drop* required from the supply voltage to the desired voltage and I is the total rated current of the load.

Example: The plate of the tube in one stage and the screens of the tubes in two other stages require an operating voltage of 250. The nearest available supply voltage is 400 and the total of the rated plate and screen currents is 75 ma. The required resistance is

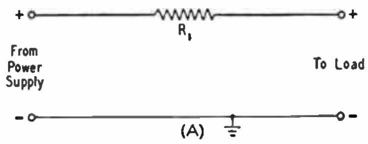
$$R = \frac{400 - 250}{0.075} = \frac{150}{0.075} = 2000 \text{ ohms.}$$

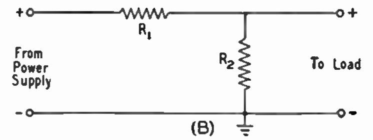
The power rating of the resistor is obtained from P (watts) $= I^2R = (0.075)^2\,(2000) = 11.2$ watts. A 25-watt resistor is the nearest safe rating to be used.

Voltage Dividers

The regulation of the voltage obtained in this manner obviously is poor, since any change in current through the resistor will cause a directly-proportional change in the voltage drop across the resistor. The regulation can be improved somewhat by connecting a second resistor from the low-voltage end of the first to the negative power-supply terminal, as shown in Fig. 7-12B. Such an arrangement constitutes a **voltage divider.** The second resistor, R_2, acts as a constant load for the first, R_1, so that any variation in current from the tap becomes a smaller percentage of the total current through R_1. The heavier the current drawn by the resistors when they alone are connected across the supply, the better will be the voltage regulation at the tap.

Such a voltage divider may have more than a single tap for the purpose of obtaining more than one value of voltage. A typical arrangement is shown in Fig. 7-12C. The terminal voltage is E, and two taps are provided to give





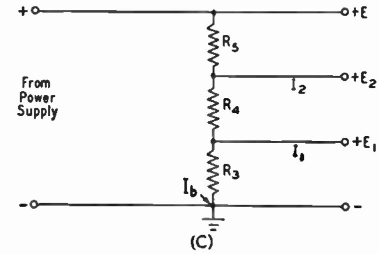


Fig. 7-12 — A — Series voltage-dropping resistor. B — Simple voltage divider. C — Multiple divider circuit.

$$R_3 = \frac{E_1}{I_b};\ R_4 = \frac{E_2 - E_1}{I_b + I_1};\ R_5 = \frac{E - E_2}{I_b + I_1 + I_2}$$

lower voltages, E_1 and E_2, at currents I_1 and I_2 respectively. The smaller the resistance between taps in proportion to the total resistance, the smaller the voltage between the taps. For convenience, the voltage divider in the figure is considered to be made up of separate resistances R_3, R_4, R_5, between taps. R_3 carries only the bleeder current, I_b; R_4 carries I_1 in addition to I_b; R_5 carries I_2, I_1 and I_b. To calculate the resistances required, a bleeder current, I_b, must be assumed; generally it is low compared with the total load current (10 per cent or so). Then the required values can be calculated as shown in Fig. 7-12C, I being in decimal parts of an ampere.

The method may be extended to any desired number of taps, each resistance section being calculated by Ohm's Law using the need voltage drop across it and the total current through it. The power dissipated by each section may be calculated either by multiplying I and E or I^2 and R.

Voltage Stabilization

Gaseous Regulator Tubes

There is frequent need for maintaining the voltage applied to a low-voltage low-current circuit at a practically constant value, regardless of the voltage regulation of the power supply or variations in load current. In such applications, gaseous regulator tubes (VR105-30, VR150-30, etc.) can be used to good advantage. The voltage drop across such tubes is constant over a moderately wide current range. Tubes are available for regulated voltages of 150, 105, 90 and 75 volts.

The fundamental circuit for a gaseous regulator is shown in Fig. 7-13A. The tube is connected in series with a limiting resistor, R_1, across a source of voltage that must be higher than the starting voltage. The starting voltage is about 30 per cent higher than the operating voltage. The load is connected in parallel with the tube. For stable operation, a minimum tube current of 5 to 10 ma. is required. The maximum permissible current with most types is 40 ma.; consequently, the load current cannot exceed 30 to 35 ma. if the voltage is to be stabilized over a range from zero to maximum load current.

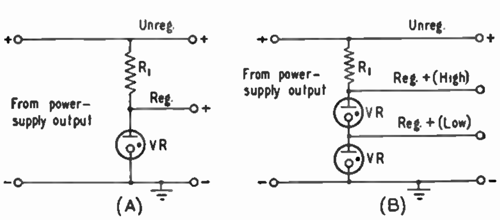


Fig. 7-13 — Voltage-stabilizing circuits using VR tubes.

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:

$$R = \frac{1000\ (E_s - E_r)}{I}$$

where R is the limiting resistance in ohms, E_s is the voltage of the source across which the tube and resistor are connected, E_r is the rated voltage drop across the regulator tube, and I is the maximum tube current in milliamperes (usually 40 ma.).

Fig. 7-13B shows how two tubes may be used in series to give a higher regulated voltage than is obtainable with one, and also to give two values of regulated voltage. The limiting resistor may be calculated as above, using the sum of the voltage drops across the two tubes for E_r. Since the upper tube must carry more current than the lower, the load connected to the low-voltage tap must take small current. The total current taken by the loads on both the high and low taps should not exceed 30 to 35 milliamperes.

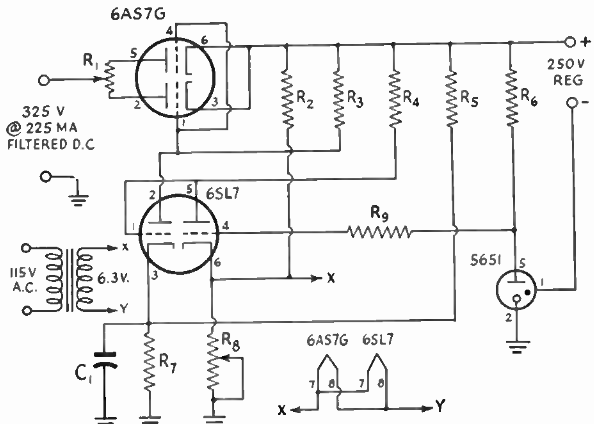


Fig. 7-14 — Electronic voltage-regulator circuit.

C_1 — 0.1-μfd. 400-volt paper.
R_1 — 160-ohm 10-watt potentiometer (balance).
R_2, R_5 — 12,000 ohms, 2 watts.
R_3, R_4 — 0.47 megohm, ½ watt.
R_6 — 68,000 ohms, 1 watt.
R_7 — 15,000 ohms, 2 watts.
R_8 — 10,000-ohm potentiometer (output control).
R_9 — 1 megohm, ½ watt.

Voltage regulation of the order of 1 per cent can be obtained with regulator circuits of this type.

Electronic Voltage Regulation

Several circuits have been developed for regulating the voltage output of a power supply electronically. 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. 7-14, the 5651 regulator tube supplies the grid (*4*) of the 6SL7 with a constant reference voltage. When the load connected across the output terminals increases, the output voltage tends to decrease. This decreases the plate (*5*) voltage. Since grid (1) is connected directly to plate (*5*), grid (*1*) becomes less positive and that triode draws less plate current. The voltage drop across R_3 being less, the bias on the grids of the 6AS7G is reduced, decreasing the voltage drop across the 6AS7G and thereby maintaining the original output voltage.

For a maximum regulated voltage output of 250, the filtered d.c. input voltage should be 325 volts at 225 ma. For a constant line voltage the output voltage will remain constant within 0.2 volt over a load-current range of 0 to 225 ma. With a line-voltage variation of plus or minus 10 per cent, the output voltage will vary less than 0.1 volt.

Another similar regulator circuit is shown in Fig. 7-15. 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 ob-

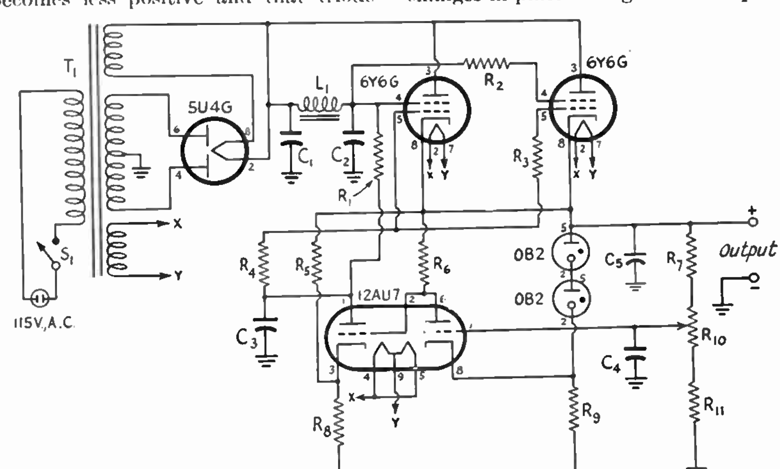


Fig. 7-15 — Circuit diagram of an electronically-regulated power supply rated at 300 volts max., 150 ma. max.

C_1, C_2, C_5 — 16-μfd. 600-volt electrolytic.
C_3 — 0.015-μfd. paper.
C_4 — 0.1-μfd. 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} — 13,000 ohms, ½ watt.
L_1 — 8-hy., 40-ma. filter choke.
S_1 — S.p.s.t. toggle.
T_1 — Power transformer: 375–375 volts r.m.s., 160 ma.; 6.3 volts, 3 amps.; 5 volts, 3 amps. (Thor. 22 R 33).

Table of Performance for Circuit of Fig. 7-15

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.

tain 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. 7-15. 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.

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.

Bias Supplies

As discussed in the chapter on high-frequency transmitters, the chief function of a bias supply for the r.f. stages of a transmitter is that of providing protective bias, although under certain circumstances, a bias supply, or pack, as it is sometimes called, can provide the operating bias if desired.

Simple Bias Packs

Fig. 7-16A 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.

A full-wave circuit is shown in Fig. 7-16C. 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. 7-16 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. 7-17A. A VR tube with a voltage rating anywhere between the

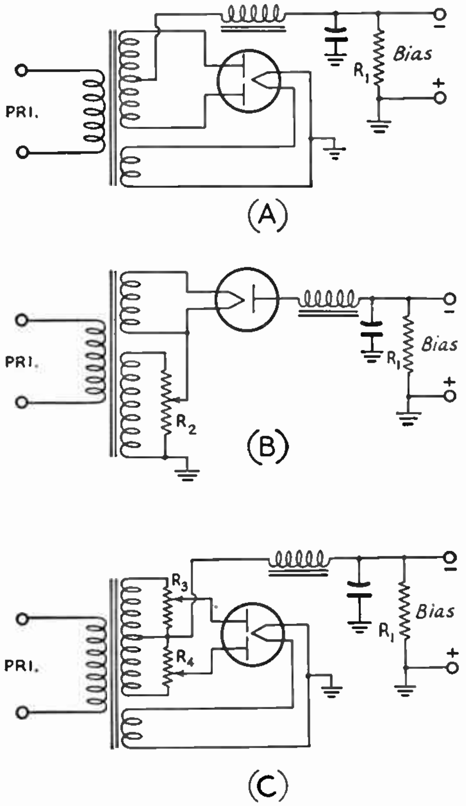
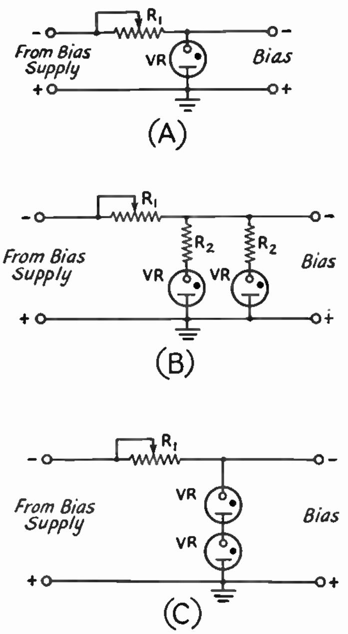


Fig. 7-16 — Simple bias-supply circuits. In A, the peak transformer voltage must not exceed the operating value of bias. The circuits of B (half-wave) and C (full-wave) may be used to reduce transformer voltage to the rectifier. R_1 is the recommended grid-leak resistance.



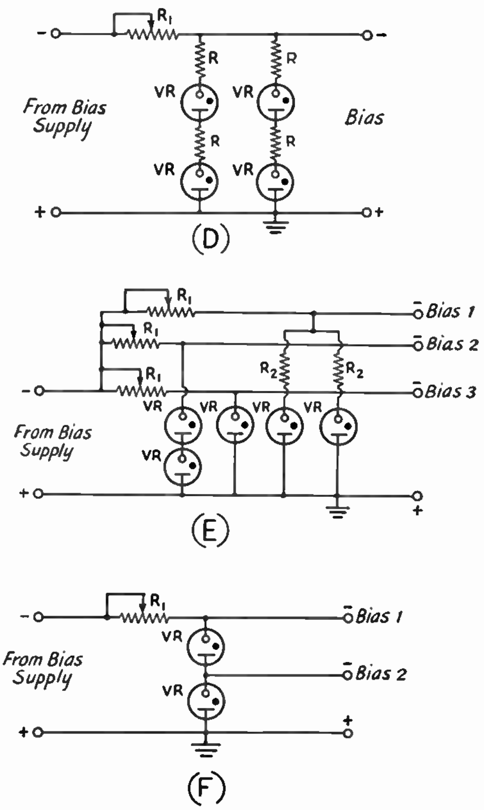


Fig. 7-17 — Illustrating the use of VR tubes in stabilizing protective-bias supplies. R_1 is a resistor whose value is adjusted to limit the current through each VR tube to 5 ma. before amplifier excitation is applied. R and R_2 are current-equalizing resistors of 50 to 1000 ohms.

♦

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. 7-17B, for each 40 ma., or less, of additional grid current. The resistors R_2 are for the purpose of helping to maintain equal currents through each VR tube, and should have a value of 50 to 1000 ohms or more as required.

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. 7-17C and D.

If a single value of fixed bias will serve for more than one stage, the biasing terminal of each such stage may be connected to a single supply of this type, provided only that the total grid current of all stages so connected does not exceed the current rating of the VR tube or tubes. Alternatively, other separate VR-tube branches may be added in any desired combination to the same supply, as shown

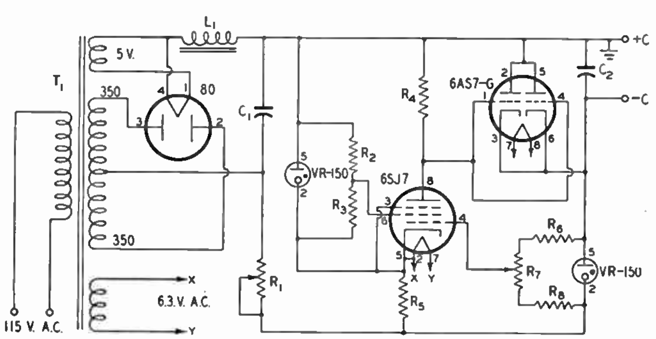


Fig. 7-18 — Circuit diagram of an electronically-regulated bias supply.

C_1 — 20-μfd. 450-volt electrolytic.
C_2 — 20-μfd. 150-volt electrolytic.
R_1 — 5000 ohms, 25 watts.
R_2 — 22,000 ohms, ½ watt.
R_3 — 68,000 ohms, ½ watt.
R_4 — 0.27 megohm, ½ watt.
R_5 — 3000 ohms, 5 watts.
R_6 — 0.12 megohm, ½ watt.
R_7 — 0.1-megohm potentiometer.
R_8 — 27,000 ohms, ½ watt.
L_1 — 20-hy. 50-ma. filter choke.
T_1 — Power transformer: 350 volts r.m.s. each side of center, 50 ma.; 5 volts, 2 amp.; 6.3 volts, 3 amp.

in Fig. 7-17E, 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. 7-18. The output voltage may be adjusted to any value between 20 volts and 80 volts and the unit will handle grid currents up to 200 ma. over the range of 30 to 80 volts, and 100 ma. over the remainder of the range. This will take care of the bias requirements of most tubes used in Class B amplifier service. The regulation will hold to about 0.001 volt per milliampere of grid current.

Other Sources of Biasing Voltage

In some cases, it may be convenient to obtain the biasing voltage from a source other than a separate supply. A half-wave rectifier may be connected with reversed polarization to obtain biasing voltage from a low-voltage plate supply, as shown in Fig. 7-19A. In another arrangement, shown at B, a spare filament winding can be used to operate a filament transformer of similar voltage rating in reverse to obtain a voltage of about 130 from the winding that is customarily the primary. This will be sufficient to operate a VR75 or VR90 regulator tube.

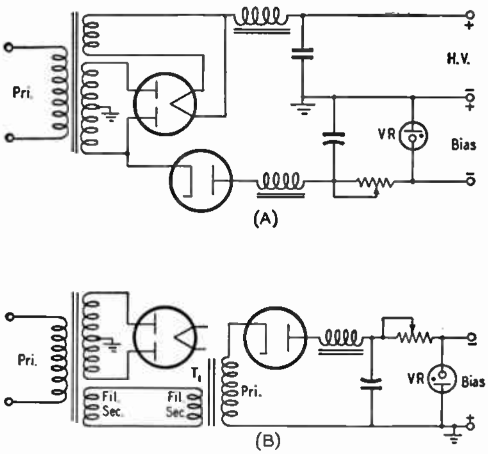


Fig. 7-19 — Convenient means of obtaining biasing voltage. A — From a low-voltage plate supply. B — From spare filament winding. T_1 is a filament transformer, of a voltage output similar to that of the spare filament winding, connected in reverse to give 115 volts r.m.s. output. If cold-cathode or selenium rectifiers are used, no additional filament supply is required.

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. 7-20, 7-21 and 7-22 may be used with any type of rectifier, they find their greatest advantage when used with selenium rectifiers which require no filament transformer.

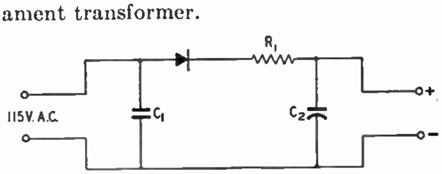


Fig. 7-20 — Simple half-wave circuit for selenium rectifier.

C_1 — 0.05-μfd. 600-volt paper.
C_2 — 40-μfd. 200-volt electrolytic.
R_1 — 25 to 100 ohms.

Fig. 7-20 is a straightforward half-wave rectifier circuit which may be used in applications where 115 to 130 volts d.c. is desired. It can be used for bias supply, for instance. In this, as well as other circuits, it will be observed that the negative side of the output is common with one side of the a.c. line and it is suggested that this side be fused with a ½-ampere fuse.

Fig. 7-21 shows several voltage-doubler circuits. Of the three, the one shown at A is the most desirable since there is no series condenser. It is a full-wave circuit and there will be very little ripple voltage appearing at the output. The arrangement of circuit B is such that one side of the output may be grounded. In circuit C, the point X is common to both condensers in the rectifier and filter, and a single-unit

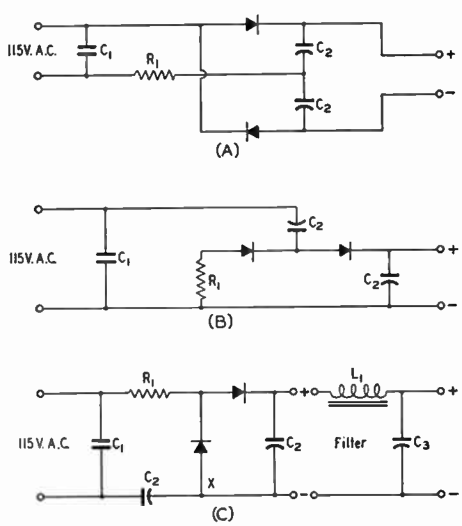


Fig. 7-21 — Voltage-doubling circuits for use with selenium rectifiers.

C_1 — 0.05-μfd. 600-volt paper.
C_2 — 40-μfd. 200-volt electrolytic.
C_3 — Filter condenser.
R_1 — 25 to 100 ohms.
L_1 — Filter choke.

3-section condenser can be used to save space. If the load current is less than 100 ma., this is the best circuit.

Fig. 7-22A shows a voltage tripler, and B and C quadruplers.

All components are standard. C_1 in all circuits is for "hash" filtering and its value is not critical. A 0.05-μfd. 600-volt-working condenser should serve. All other condensers should be 40-μfd. 200-volt units, except those in the tripler and quadrupler circuits. Those in the circuit of Fig. 7-22 should have a rating of 450 volts working. In the voltage multipliers and in other circuits where a condenser is passing the full current, good condensers should be used because the a.c. ripple mentioned above appears across the condenser and increases as the load increases. If the current is allowed to become too high, it will cause heating and deterioration of the condenser. This can be kept to a minimum by using a capacitor of high value and making sure it is of good make. R_1 should be 25 ohms, but if it is found that the rectifier units are running a little too warm, this value may be increased to as high as 100 ohms, with a corresponding drop in output voltage, of course. A single-section filter, as shown in Fig. 7-21C, will provide sufficient smoothing for most applications.

These circuits should be used with caution, since there is always a direct connection between power line and load.

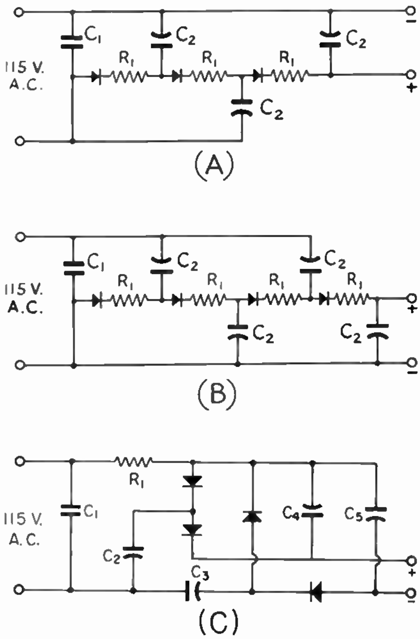


Fig. 7-22 — A — Tripler circuit. B — Half-wave quadrupler. C — Full-wave quadrupler.
C_1 — 0.05-μfd. 600-volt paper.
C_2 — 40-μfd. 450-volt electrolytic.
C_3 — 100-μfd. 150-volt electrolytic.
R_1 — 25 to 100 ohms.

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. 7-23A. 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 equip-

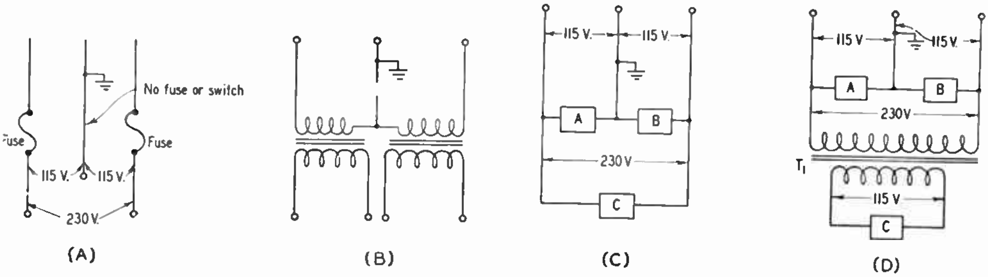


Fig. 7-23 — Three-wire power-line circuits. A — Normal 3-wire-line termination. No fuse should be used in the grounded (neutral) line. B — Showing that a switch in the neutral does not remove voltage from either side of the line. C — Connections for both 115- and 230-volt transformers. D — Operating a 115-volt plate transformer from the 230-volt line to avoid light blinking. T_1 is a 2-to-1 step-down transformer.

ment. 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. 7-23B. Furthermore, with the neutral open, the voltage will then be divided between the two sides in proportion to the load resistance, the voltage on one side dropping below normal, while it soars on the other side, unless the loads happen to be equal.

The usual line running to baseboard outlets is rated at 15 amperes. Considering the power consumed by filaments, lamps, modulator, receiver and other auxiliary equipment, it is not unusual to find this 15-ampere rating exceeded by the requirements of a station of only mod-

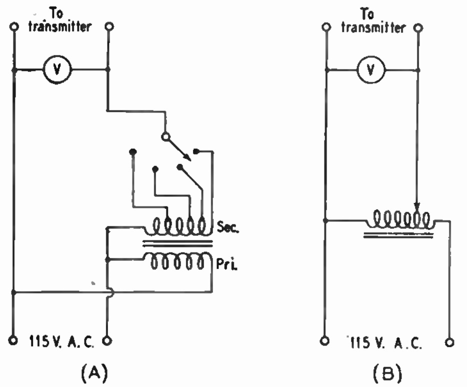


Fig. 7-24 — Two methods of transformer primary control. At A is a tapped toy transformer which may be connected so as to boost or buck the line voltage as required. At B is indicated a variable transformer or autotransformer (Variac) which feeds the transformer primaries.

erate 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. 7-23C. 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. 7-23D).

When a special heavy-duty line is to be installed, the local power company should be consulted as to local requirements. In some localities it is necessary to have such a job done by a licensed electrician, and there may be special requirements to be met in regard to fittings and the manner of installation. Some amateurs terminate the special line to the station at a switch box, while others may use electric-stove receptacles as the termination. The power is then distributed around the station by means of conventional outlets at convenient points. All circuits should be properly fused.

LINE-VOLTAGE ADJUSTMENT

In certain communities trouble is sometimes experienced from fluctuations in line voltage. Usually these fluctuations are caused by a variation in the load on the line and, since most of the variation comes at certain fixed times of the day or night, such as the times when lights are turned on at evening, they may be taken care of by the use of a manually-operated compensating device. A simple arrangement is shown in Fig. 7-24A. A toy transformer is used to boost or buck the line voltage as required. The transformer should have a tapped secondary varying between 6 and 20 volts in steps of 2 or 3 volts and its secondary should be capable of carrying the full load current of the entire transmitter, or that portion of it fed by the toy transformer.

The secondary is connected in series with the line voltage and, if the phasing of the windings is correct, the voltage applied to the primaries of the transmitter transformers can be brought

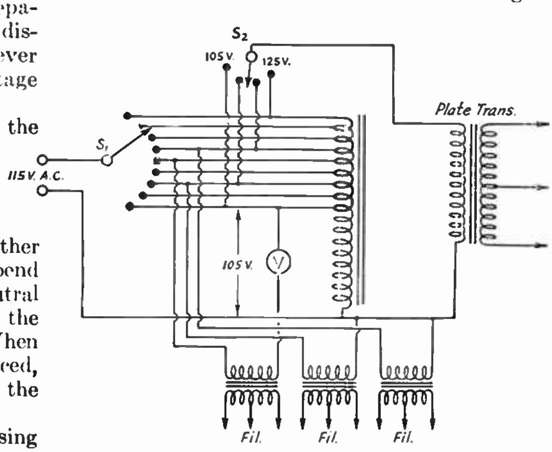


Fig. 7-25— With this circuit, a single adjustment of the tap switch S_1 places the correct primary voltage on all transformers in the transmitter. Information on constructing a suitable autotransformer at negligible cost is contained in the text. The light winding represents the regular primary winding of a revamped transformer, the heavy winding the voltage-adjusting section.

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 7-24B illustrates the use of a variable transformer (Variac) for adjusting line voltage to the desired value.

Another scheme by which the primary voltage of each transformer in the transmitter may be adjusted to give a desired secondary voltage, with a master control for compensating for changes in line voltage, is shown in Fig. 7-25.

This arrangement has the following features:

1) Adjustment of the switch S_1 to make the voltmeter read 105 volts automatically adjusts all transformer primaries to the predetermined correct voltage.

2) The necessity for having all primaries work at the same voltage is eliminated. Thus, 110 volts can be applied to the primary of one transformer, 115 to another, etc., 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 va. at 6.3 volts output, for small tube-heater demands, up to several thousand volt-amperes at 115 or 230 volts. In average figures, such transformers will hold their output voltages within one per cent under an input-voltage variation of 30 per cent.

Construction of Power Supplies

The length of most leads in a power supply is unimportant, so that the arrangement of components from this consideration is not a factor in construction. More important are the points of good high-voltage insulation, adequate conductor size for filament wiring, proper ventilation for rectifier tubes and — most important of all — safety to the operator. Exposed high-voltage terminals or wiring which might be bumped into accidentally should not be permitted to exist. They should be covered with adequate insulation or placed inaccessible to contact during normal operation and adjustment of the transmitter. Power-supply units should be fused individually. 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 condenser cases should also be grounded to the chassis.

Rectifier filament leads should be kept short to assure proper voltage at the rectifier socket, and the sockets should have good insulation and adequate contact surface. Plate leads to mercury-vapor tubes should be kept short to minimize the radiation of noise.

Where high-voltage wiring must pass through a metal chassis, grommet-lined clearance holes will serve for voltages up to 500 or 750, but ceramic feed-through insulators should be used for higher voltages. Bleeder and

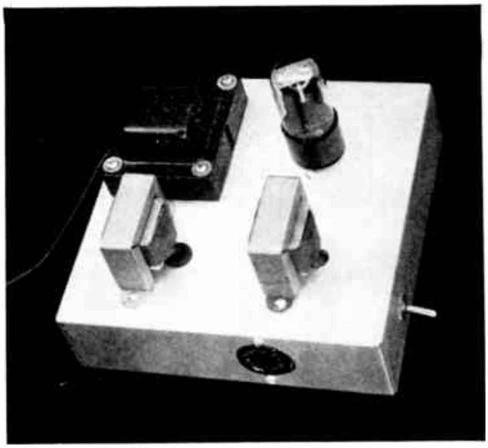


Fig. 7-26— A typical simple receiver power supply. Filament and plate voltages are taken from the multi-contact tube socket which serves as an outlet.

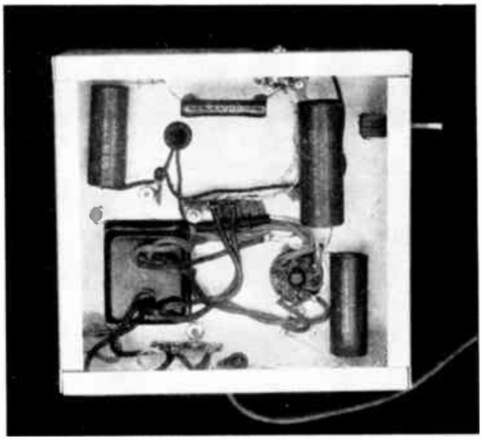


Fig. 7-27— Bottom view of the simple receiver power supply showing the cut-out for the flush-mounting transformer.

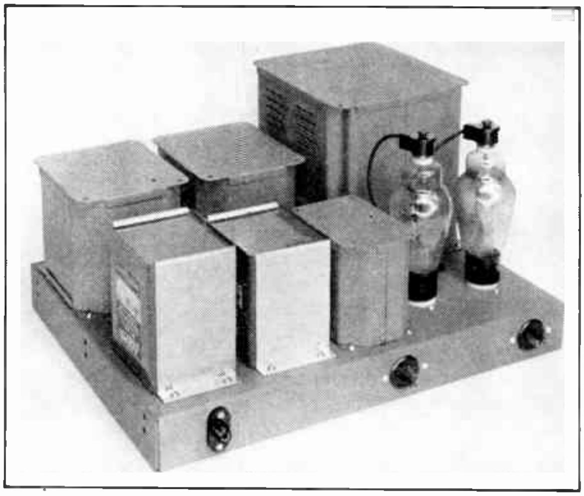


Fig. 7-28 — A typical high-voltage transmitter power supply. The transformers, chokes and condensers are inverted so that no terminals are exposed to accidental contact. The caps of the 866 rectifiers are the insulated type.

voltage-dropping resistors should be placed where they are open to air circulation. Placing them in confined space reduces the rating.

It is highly preferable from the standpoint of operating convenience to have separate filament transformers for the rectifier tubes, rather than to use combination filament and plate transformers, such as those used in receivers. This permits the transmitter plate voltage to be switched on without the necessity

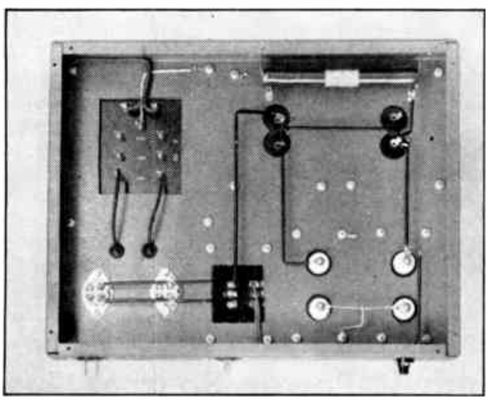


Fig. 7-29 — Bottom view of the transmitter power supply showing the cut-outs for the terminals. Separate power plugs are used for the rectifier-filament and plate transformers so that they may be switched independently from the control position.

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. 7-30. The arrangements shown in Fig. 7-30A 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

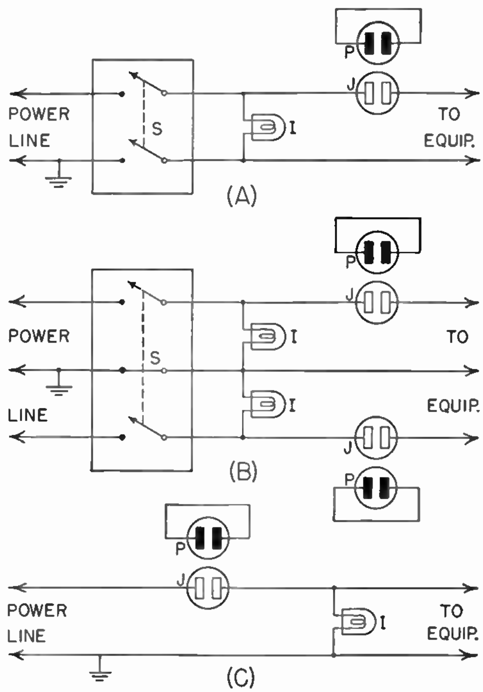


Fig. 7-30 — Reliable arrangements for cutting off all power to the transmitter. *S* is an enclosed double-pole knife-type switch, *J* a standard a.c. outlet, *P* a shorted plug to fit the outlet and *I* a red lamp.

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.

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 members 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.

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. 7-30C.

The test bench ought to be fed through the main power switch, or a similar arrangement at the bench, if the latter is 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 condensers will be discharged when the high-voltage transformer is turned off. To guard against the possibility of danger to the operator should the bleeder resistor burn out without his knowledge, and also to protect him in case he neglects to turn off the power supply before opening a cabinet transmitter enclosure, one of the devices shown in Fig. 7-31 is recommended. In A, a grounded pivoted metal lever drops by gravity against a contact connected to the positive high-voltage terminal when the cabinet door is opened, shorting the power supply. When the door is closed, it pushes against the end of the lever protruding through the door opening and the short is removed automatically. In another scheme, shown at B, a metal ball, suspended on a cord, drops into a triangle of contacts, one of which is grounded, while the other two go to positive terminals of power supplies. The wedge mounted on the door pushes against the suspending cord, lifting the ball when the door is closed. The power supplies should be equipped with suitable fuses to save the equipment in case the device is ever called upon to perform its duty.

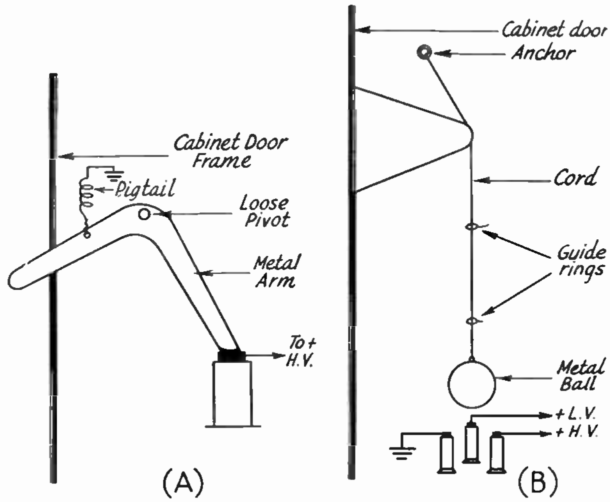


Fig. 7-31 — Two schemes for shorting the high-voltage supply automatically for safety purposes when the transmitter door is opened.

Emergency and Independent Power Sources

Emergency power supply which operates independently of a.c. lines is available, or can be built in a number of different forms, depending upon the requirements of the service for which it is intended.

The most practical supply for the average individual amateur is one that operates from a 6-volt car storage battery. Such a supply may take the form of a small motor generator (often called a genemotor), a rotary converter, or a vibrator-transformer-rectifier combination.

Dynamotors

A dynamotor differs from a motor generator in that it is a single unit having a double armature winding. One winding serves for the driving motor, while the output voltage is taken from the other. Dynamotors usually are operated from 6-, 12-, 28- or 32-volt storage batteries and deliver from 300 to 1000 volts or more at various current ratings.

Genemotor is a term popularly used when making reference to a dynamotor designed especially for automobile-receiver, sound-truck and similar applications. It has good regulation and efficiency, combined with economy of operation. Standard models of genemotors have ratings ranging from 250 volts at 50 ma. to 400 volts at 375 ma. or 600 volts at 250 ma. The normal efficiency averages around 50 per cent, increasing to better than 60 per cent in the higher-power units. The voltage regulation of a genemotor is comparable to that of well-designed a.c. supplies.

Successful operation of dynamotors and genemotors requires heavy direct leads, mechanical isolation to reduce vibration, and thorough r.f. and ripple filtration. The shafts and bearings should be thoroughly "run in" before regular operation is attempted, and thereafter the tension of the bearings should be checked occasionally to make certain that no looseness has developed.

In mounting the genemotor, the support should be in the form of rubber mounting blocks, or equivalent, to prevent the transmission of vibration mechanically. The frame of the genemotor should be grounded through a heavy flexible connector. The brushes on the high-voltage end of the shaft should be bypassed with 0.002-μfd. mica condensers to a common point on the genemotor frame, preferably to a point inside the end cover close to the brush holders. Short leads are essential. It may prove desirable to shield the entire unit, or even to remove the unit to a distance of three or four feet from the receiver and antenna lead.

When the genemotor is used for receiving, a filter should be used similar to that described for vibrator supplies. A 0.01-μfd. 600-volt (d.c.) paper condenser should be connected in shunt across the output of the genemotor, followed by a 2.5-mh. r.f. choke in the positive high-voltage lead. From this point the output should be run to the receiver power terminals through a smoothing filter using 4- to 8-μfd. condensers and a 15- or 30-henry choke having low d.c. resistance.

D.C.-A.C. Converters

In some instances it is desirable to utilize existing equipment built for 115-volt a.c. operation. To operate such equipment with any of the power sources outlined above would require a considerable amount of rebuilding. This can be obviated by using a rotary converter capable of changing the d.c. from 6-, 12- or 32-volt batteries to 115-volt 60-cycle a.c. Such converter units are built to deliver outputs ranging from 40 to 250 watts, depending upon the battery power available.

The conversion efficiency of these units averages about 50 per cent. In appearance and operation they are similar to genemotors of equivalent rating. The over-all efficiency of the converter will be lower, however, because of losses in the a.c. rectifier-filter circuits and the necessity for converting heater (which is supplied directly from the battery in the case of the genemotor) as well as plate power.

Vibrator Power Supplies

The vibrator type of power supply consists of a special step-up transformer combined with a vibrating interrupter **(vibrator)**. When the unit is connected to a storage battery, plate power is obtained by passing current from the battery through the primary of the transformer. The circuit is made and reversed rapidly by the vibrator contacts, interrupting the current at regular intervals to give a changing magnetic field which induces a voltage in the secondary. The resulting square-wave d.c. pulses in the primary of the transformer cause an alternating voltage to be developed in the secondary. This high-voltage a.c. in turn is rectified, either by a vacuum-tube rectifier or by an additional synchronized pair of vibrator contacts. The rectified output is pulsating d.c., which may be filtered by ordinary means. The smoothing filter can be a single-section affair, but the output capacitance should be fairly large — 16 to 32 μfd.

Fig. 7-32 shows the two types of circuits. At A is shown the **nonsynchronous** type of vibrator. When the battery is disconnected the reed is midway between the two contacts, touching neither. On closing the battery circuit the magnet coil pulls the reed into contact with one contact point, causing current to flow through the lower half of the transformer primary winding. Simultaneously, the magnet

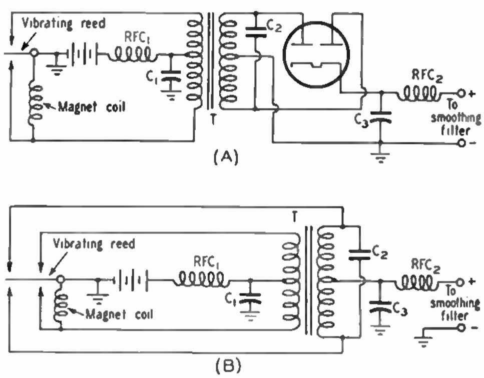


Fig. 7-32 — **Basic types of vibrator power-supply circuits. A—Nonsynchronous. B—Synchronous.**

coil is short-circuited, deënergizing it, and the reed swings back. Inertia carries the reed into contact with the upper point, causing current to flow through the upper half of the transformer primary. The magnet coil again is energized, and the cycle repeats itself.

The synchronous circuit of Fig. 7-32B is provided with an extra pair of contacts which rectify the secondary output of the transformer, thus eliminating the need for a separate rectifier tube. The secondary center-tap furnishes the positive output terminal when the relative polarities of primary and secondary windings are correct. The proper connections may be determined by experiment.

The buffer condenser, C_2, across the transformer secondary, absorbs the surges that occur on breaking the current, when the magnetic field collapses practically instantaneously and hence causes very high voltages to be induced in the secondary. Without this condenser excessive sparking occurs at the vibrator contacts, shortening the vibrator life. Correct values usually lie between 0.005 and 0.03 μfd., and for 250–300-volt supplies the condenser should be rated at 1500 to 2000 volts d.c. The exact capacitance is critical, and should be determined experimentally. The optimum value is that which results in least battery current for a given rectified d.c. output from the supply. In practice the value can be determined by observing the degree of vibrator sparking as the capacitance is changed. When the system is operating properly there should be practically no sparking at the vibrator contacts. A 5000-ohm resistor in series with C_2 will limit the secondary current to a safe value should the condenser fail.

Vibrator-transformer units are available in a variety of power and voltage ratings. Representative units vary from one delivering 125 to 200 volts at 100 ma. to others that have a 400-volt output rating at 150 ma. Most units come supplied with "hash" filters, but not all of them have built-in ripple filters. The requirements for ripple filters are similar to those for a.c. supplies. The usual efficiency of vibrator packs is in the vicinity of 70 per cent, so a 300-volt 200-ma. unit will draw approximately 15 amperes from a 6-volt storage battery. Special vibrator transformers are also available from transformer manufacturers so that the amateur may build his own supply if he so desires. These have d.c. output ratings varying from 150 volts at 40 ma. to 330 volts at 135 ma.

Vibrator-type supplies are also available for operating standard a.c. equipment from a 6-volt storage battery in power ratings up to 100 watts continuous or 125 watts intermittent.

"Hash" Elimination

Sparking at the vibrator contacts causes r.f. interference ("hash," which can be distinguished from hum by its harsh, sharper pitch) when used with a receiver. To minimize this, r.f. filters are incorporated, consisting of RFC_1 and C_1 in the battery circuit, and RFC_2 with C_3 in the d.c. output circuit.

Equally as important as the hash filter is thorough shielding of the power supply and its connecting leads, since even a small piece of wire or metal will radiate enough r.f. to cause interference in a sensitive receiver.

Testing in connection with hash elimination should be carried out with the supply operating a receiver. Since the interference usually is picked up on the receiving-antenna leads by radiation from the supply itself and from the battery leads, it is advisable to keep the supply and battery as far from the receiver as the connecting cables will permit. Three or four feet should be ample. The microphone cord likewise should be kept away from the power supply and its leads.

The power supply should be built on a metal chassis, with all unshielded parts underneath. A bottom plate to complete the shielding is advisable. The transformer case, vibrator cover and the metal shell of the tube all should be grounded to the chassis. If a glass tube is used it should be enclosed in a tube shield. The battery leads should be evenly twisted, since these leads are more likely to radiate hash than any other part of a well-shielded supply. Experimenting with different values in the hash filters should come *after* radiation from the battery leads has been reduced to a minimum. Shielding the leads is not often found to be particularly helpful.

● PRACTICAL VIBRATOR-SUPPLY CIRCUITS

A vibrator-type power supply may be designed to operate from a six-volt storage battery only, or in a combination unit which may be operated interchangeably from either battery or 115 volts a.c.

An example of the latter-type circuit is shown in Fig. 7-33. It consists essentially of two transformer-rectifier systems — one for 115 volts a.c. and the other a vibrator system to operate from a 6-volt storage battery. A common filter is used for the two systems. In interchanging between a.c. and d.c. operation, the rectifier tube (a 6X5 or 6W5G) is shifted to the appropriate socket, while the filament connections are made to the proper output termi-

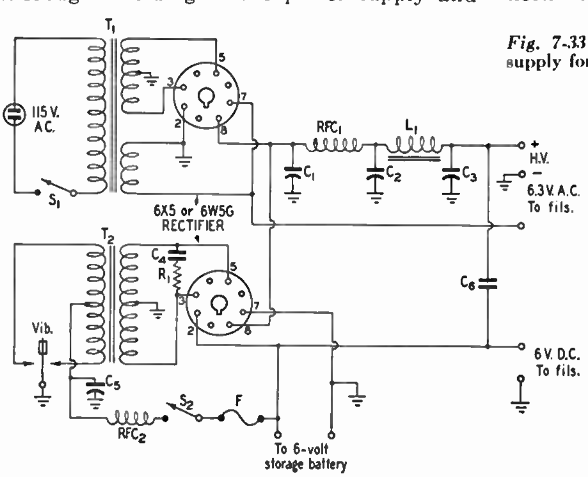


Fig. 7-33 — Circuit of a combination a.c.-d.c. power supply for emergency work.

C_1 — 0.01-μfd. 600-volt paper.
C_2 — 8-μfd. 450-volt electrolytic.
C_3 — 32-μfd. 450-volt electrolytic.
C_4 — 0.005- to 0.01-μfd. 1600-volt paper.
C_5 — 500-μfd. electrolytic, 25 volts or higher.
C_6 — 100-μμfd. 600-volt mica.
R_1 — 4700 ohms, 1 watt.
L_1 — 10- to 12-hy. filter choke, 100 ma. (not over 100 ohms) (Stancor C-2303 or equivalent).
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 55 turns No. 12 on 1-inch form, close-wound.
S_1, S_2 — Toggle switch.
T_1 — Power transformer: 275 to 300 volts r.m.s. each side of center tap, 100 to 150 ma., 6.3-volt filament winding.
T_2 — Vibrator transformer (Stancor P-6131 or similar).
VIB — Vibrator unit (Mallory 500P, 294, etc.).

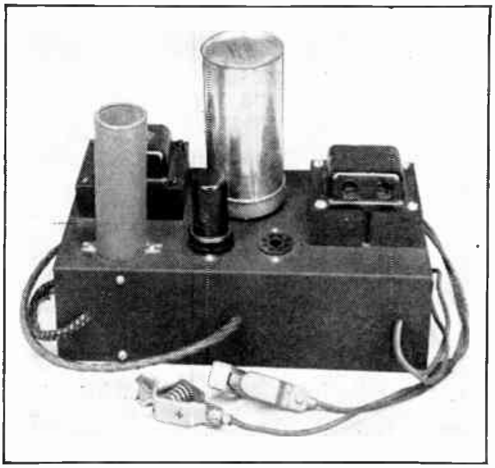


Fig. 7-34 — A typical combination a.c.-d.c. power pack for low-power emergency work. The two transformers are mounted at either end of the chassis. The filter condenser is at the left, the two rectifier sockets at the center and the vibrator to the rear.

nals. If desired, two rectifier tubes may be used and the changeover made through suitable switches.

R.f. filters for reducing hash are incorporated in both primary and secondary circuits. The secondary filter consists of a 0.01-μfd. paper condenser directly across the rectifier output, with a 2.5-mh. r.f. choke in series ahead of the smoothing filter. In the primary circuit a low-inductance choke and high-capacitance condenser are needed because of the low impedance of the circuit. A choke of the specifications given should be adequate, but if there is trouble with hash it may be beneficial to experiment with other sizes. The wire should be large — No. 12, preferably, or No. 14 as a minimum. Manufactured chokes such as the Mallory RF583 are more compact and give higher inductance for a given resistance because they are bank-wound, and may be substituted if obtainable. C_5 should be at least 500 μfd.; even more capacitance may help in bad cases of hash. The components are assembled on a 5 × 10 × 3-inch steel chassis. Three socket holes are required — one for the 4-prong socket for the vibrator and two octal sockets for the rectifier.

The compactness of selenium rectifiers and the fact that they do not require filament voltage make them particularly suited to compact lightweight power supplies for portable emergency work.

Fig. 7-35 shows the circuit of a vibrator pack that will deliver an output voltage of 400 at 200 ma. It will work with either 115-volt ac. or 6-volt battery input. The circuit is that of the familiar voltage tripler whose d.c. output voltage is, as a rough approximation, three times the peak voltage delivered by the transformer or line. An interesting feature of the circuit is the fact that the single transformer serves as the vibrator transformer when operating from 6-volt d.c. supply and as the filament transformer when operating from an a.c. line. This is accomplished without complicated switching.

The vibrator transformer, T_1, is a dual-secondary 6.3-volt filament transformer connected in reverse. In either event, the filament windings must have a rating of 10 amperes if the full load current of 200 ma. is to be used. The vibrator also must be capable of handling the current. The hash-filter choke, L_1, must carry a current of 20 amperes.

The following table shows the output voltage to be expected at various load currents, depending upon the size of condensers used at C_1, C_2 and C_3.

C_1, C_2, C_3 (μfd.)	Output Voltage at 50 ma.	100 ma.	150 ma.	200 ma.
60	455	430	415	395
40	425	390	360	330
20	400	340	285	225

In operating the supply from an a.c. line, it is always wise to determine the plug polarity with respect to ground. Otherwise the rectifier part of the circuit and the transformer circuit cannot be connected to actual ground except through by-pass condensers. Rectangular cutouts are also needed for the two flush-mounting transformers. The filter choke, L_1, and other small components can be fitted under the chassis. The clip leads to the battery should be no longer than necessary.

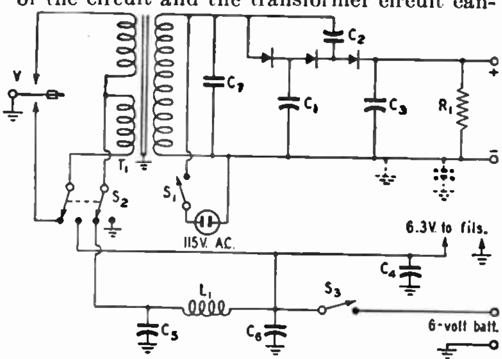


Fig. 7-35 — Circuit diagram of a compact vibrator-a.c. portable power supply using selenium rectifiers.

C_1 — 60-μfd. 200-volt electrolytic.
C_2 — 60-μfd. 400-volt electrolytic.
C_3 — 60-μfd. 600-volt electrolytic.
C_4 — 25-μfd. 25-volt electrolytic.
C_5, C_6 — 0.5-μfd. 25-volt paper.
C_7 — 0.007-μfd. 1500-volt paper.
R_1 — 25,000 ohms, 10 watts.
L_1 — 25-μhy. 20-amp. choke.
S_1 — 115-volt toggle switch.
S_2 — D.p.d.t. heavy-duty knife switch.
S_3 — 25-amp. s.p.s.t. switch.
T_1 — See text.
V — Heavy-duty vibrator (Cornell-Dub. 4123).

GASOLINE-ENGINE DRIVEN GENERATORS

For higher-power installations, such as for communications control centers during emergencies, the most practical form of independent power supply is the gasoline-engine driven generator which provides standard 115-volt 60-cycle supply.

Such generators are ordinarily rated at a minimum of 250 or 300 watts. They are available up to two kilowatts, or big enough to handle the highest-power amateur rig. Most are arranged to charge automatically an auxiliary 6- or 12-volt battery used in starting. Fitted with self-starters and adequate mufflers and filters, they represent a high order of performance and efficiency. Many of the larger models are liquid-cooled, and they will operate continuously at full load.

A variant on the generator idea is the use of fan-belt drive. The disadvantage of requiring that the automobile must be running throughout the operating period has not led to general popularity of this idea among amateurs. Such generators are similar in construction and capacity to the small gas-driven units.

The output frequency of an engine-driven generator must fall between the relatively narrow limits of 50 to 60 cycles if standard 60-cycle transformers are to operate efficiently from this source. A 60-cycle electric clock provides a means of checking the output frequency with a fair degree of accuracy. The clock is connected across the output of the generator and the second hand is checked closely against the second hand of a watch. The speed of the engine is adjusted until the two second hands are in synchronism. If a 50-cycle clock is used to check a 60-cycle generator, it should be remembered that one revolution of the second hand will be made in 50 seconds and the clock will gain 4.8 hours in each 24 hours.

Output voltage should be checked with a voltmeter since a standard 115-volt lamp bulb, which is sometimes used for this purpose, is very inaccurate. Tests have shown that what appears to be normal brilliance in the lamp may occur at voltages as high as 150 if the check is made in bright sunlight.

Noise Elimination

Electrical noise which may interfere with receivers operating from engine-driven a.c. generators may be reduced or eliminated by taking proper precautions. The most important point is that of grounding the frame of the generator *and* one side of the output. The ground lead should be short to be effective, otherwise grounding may actually increase the noise. A water pipe may be used if a short connection can be made near the point where the pipe enters the ground, otherwise a good separate ground should be provided.

The next step is to loosen the brush-holder locks and slowly shift the position of the brushes while checking for noise with the receiver. Usually a point will be found (almost always different from the factory setting) where there is a marked decrease in noise.

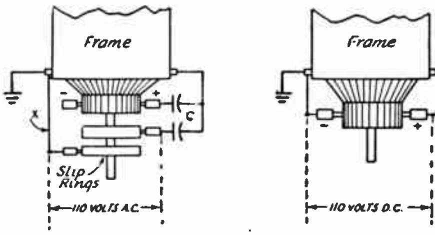


Fig. 7-36 — Connections used for eliminating interference from gas-driven generator plants. C should be 1 μfd., 300 volts, paper, while C_2 may be 1 μfd. with a voltage rating of twice the d.c. output voltage delivered by the generator. X indicates an added connection between the slip ring on the grounded side of the line and the generator frame.

From this point on, if necessary, by-pass condensers from various brush holders to the frame, as shown in Fig. 7-36, will bring the hash down to within 10 to 15 per cent of its original intensity, if not entirely eliminating it. Most of the remaining noise will be reduced still further if the high-power audio stages are cut out and a pair of headphones is connected into the second detector.

POWER FOR PORTABLES

Dry-cell batteries are the only practical source of supply for equipment which must be transported on foot. From certain considerations they may also be the best source of voltage for a receiver whose filaments may be operated from a storage battery, since no problem of noise filtering is involved.

Their disadvantages are weight, high cost, and limited current capability. In addition, they will lose their power even when not in use, if allowed to stand idle for periods of a year or more. This makes them uneconomical if not used more or less continuously.

Dry "B" batteries are made in a variety of sizes and shapes, from a 45-volt unit weighing about 1 lb. that has an intermittent service rating of 20 hours at a drain of 20 ma., to a 12-lb. unit rated at 130 hours at 40 ma. "A" batteries for filament service range from a 6-volt unit weighing 1½ lbs. delivering in intermittent service an average of 60 ma. for 150 hours, to a 6¼-lb. 1.5-volt unit having a service life of 870 hours at 200 ma. Miniature batteries, suitable for hand-portable use, are also available.

CHAPTER 8

Keying and Break-In

Offhand it would appear that keying a transmitter is a simple matter, since on the face of it nothing more is involved than turning the transmitter output on and off to correspond to the code characters being sent. Unfortunately, it is not this simple, and perfect keying of a c.w. rig is as difficult to come by as perfect voice quality is with a 'phone transmitter. The problem cannot be dismissed lightly.

Although the operation is basically that of turning the transmitter output power on and off, it is complicated by the fact that it must not be turned on and off *instantaneously*. Instead, the output must be made to rise to (and fall from) maximum in some finite period of time, if **key clicks** are to be avoided. These clicks are the inescapable result of changing the power level rapidly, and they appear in the radio spectrum adjacent to the signal proper. The more rapidly the output is varied, the farther the clicks will extend in frequency and the greater will be their amplitude. They interfere unnecessarily with other signals and, if severe enough, can be cause for a discrepancy report by the FCC.

Another effect of improper keying of a transmitter is the introduction of **chirp,** a change in frequency at the instant of making or breaking the signal. A chirp of 50 cycles is enough to make a signal unpleasant to copy, and a chirp of several hundred cycles may render the signal difficult to copy or a target for an FCC discrepancy report. Much depends, of course, upon the selectivity and beat note being used at the receiver, but the safest procedure is to aim for no detectable chirp.

A third keying fault is **backwave,** which consists of power leaking through and radiating when the key is "up." A strong backwave makes the signal unpleasant or difficult to copy.

In code transmission, there are intervals between dots and dashes, and slightly longer intervals between letters and words, when no power is being radiated by the transmitter. If the receiver can be made to operate at normal sensitivity during these intervals, it is possible for the receiving operator to signal the transmitting operator, by holding his key down. This is useful during the handling of messages, since the receiving operator can immediately signal the transmitting operator if he misses part of the message. It also reduces the time necessary for calling in answer to a "CQ." The ability to hear signals during the short "key-up" intervals is called **break-in operation.**

SELECTING THE STAGE TO KEY

It is often desirable from an operating standpoint to design the c.w. transmitter for break-in operation. In most cases this requires that the oscillator be keyed, since a continuously-running oscillator will create interference in the receiver and prevent break-in on or near one's own frequency, unless the oscillator stage is well shielded.[1] Chirpless and clickless keying of an oscillator is difficult to obtain, since the necessary slow turning on and off of the oscillator (for click elimination) shows up any oscillator frequency-*vs.*-voltage changes. It is easy to key an oscillator without chirps or without clicks but not without both. The effect of a chirp is multiplied with frequency, and it is difficult to obtain chirpless oscillator keying at an output frequency of 14, 21 or 28 Mc.

The best-sounding keying (and the simplest to adjust) is usually obtained by keying the output or driver stage, or both. With the oscillator running continuously and "buffered" by several intermediate stages, its frequency remains constant throughout all parts of the keying cycle. The only problem in keying then becomes that of properly "shaping" the keying to reduce or eliminate clicks. When keying several stages away from the output amplifier, it is necessary to bias the stages following the keyed stage so that they draw little or no plate current when the key is up, to avoid excessive plate dissipation. If the stages are biased too heavily, however, these subsequent amplifiers tend to shorten the rise and fall times and thus reintroduce clicks. This should always be borne in mind when a multistage transmitter is used with low-level keying.

The power broken by the key is an important consideration, both from the standpoint of safety to the operator and that of sparking and sticking at the key contacts. Keying of the oscillator or a low-power stage is favorable on both counts. The use of a keying relay or keyer tube is recommended when a high-power circuit is keyed.

Because transmitters vary widely in design, there is no specific recommendation that can be made about choosing the stage to key. If the oscillator alone keys satisfactorily (no chirps or clicks), even when listening to its

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[1] For a description of a well-shielded oscillator, see Smith, "A Solution to the Keyed-VFO Problem," *QST*, February, 1950.

harmonics on 21 or 28 Mc., the transmitter should be keyed there, but the effect of adding the additional multipliers and amplifiers should be carefully checked, to see that clicks are not reintroduced. Methods for checking will be given later. If the oscillator cannot be keyed satisfactorily by itself or with the following stage added, a stage near the output should be keyed and any thought of break-in operation should be discarded. A close approach to break-in operation can be obtained by using a convenient and fast "on-off" switch for the oscillator, or the break-in system described later in the chapter can be used. Foot-actuated switches are available for use as the "on-off" switch — they leave both hands free at all times.

Keying Circuits

The plate circuit is a good one to key in an oscillator or low-voltage amplifier, because it is easy to shape the keying properly in this circuit. When **plate-circuit keying** is used, it is usually done in the negative lead, since this permits one side of the key to be grounded. The stage can be keyed in the positive lead, but both sides of the keyed circuit will be "hot," and a keying relay is advisable. Fig. 8-1 shows the general circuit for negative-lead keying in either an oscillator or an amplifier. Two examples are shown using triodes, but screen-grid tubes can be used just as readily. Plate-circuit keying is recommended only for low-voltage circuits if no keying relay is used, since a large portion of the supply voltage can appear across the open key.

Shaping circuits applicable to this and later circuits will be discussed in this chapter under "Testing Your Keying."

Somewhat closely related to plate-circuit keying is **screen-grid keying**, shown in Fig. 8-2. The only basic difference is that the screen grid is pulled down to a negative voltage when the key is up, to avoid the backwave that may be present when the screen goes only to zero volts. The negative supply can be small, since its current demand is only a few milliamperes. If the screen voltage is taken from the plate supply, it should come from a voltage divider rather than a simple dropping resistor.

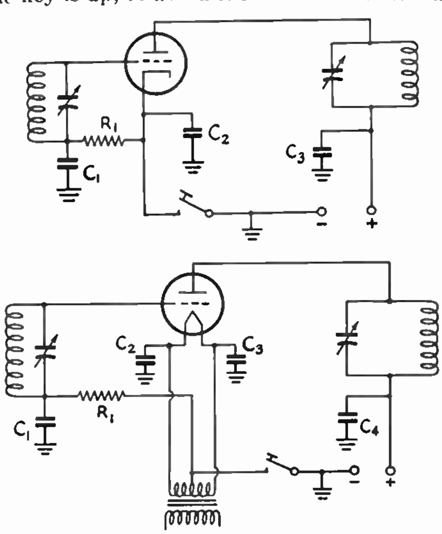


Fig. 8-1 — Negative plate-lead keying for cathode- or filament-type tubes. These circuits are useful for oscillator or low-power stages, where the voltage across the open key is not very dangerous. Tetrode or pentode stages can be keyed in this manner, but the screen circuit should be stabilized with VR tubes or a heavy voltage divider. R_1 is the normal grid leak, C_1, C_2, C_3 and C_4 are r.f. by-pass condensers.

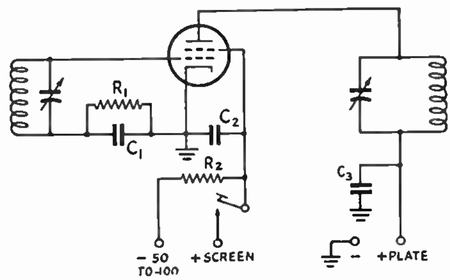


Fig. 8-2 — Screen-grid keying, suitable for oscillator or amplifier keying. R_1 is the normal grid leak, R_2 should be about 200 to 500 ohms per screen volt, and C_1, C_2 and C_3 are normal by-pass condensers.

Grid-circuit, or **blocked-grid,** keying is shown in Fig. 8-3. With the key up, a negative voltage is applied to the grid sufficient to cut off the tube and prevent current flow. With the key closed, the grid circuit develops normal grid bias through R_2. The drain on the negative-voltage supply is small, since it is limited by the size of R_1. Grid-circuit keying is generally used with low-power stages or where the voltage necessary to cut off the amplifier is only a few hundred volts. The value of C_1 determines the keying characteristic, together with the ratio of R_2 and R_1, and will be discussed later.

By placing the key in the cathode (or center tap) circuit of an oscillator or amplifier, both the grid and plate (and screen, if any) circuits are opened by the key. **Cathode keying** is good for use with amplifiers, because the proper

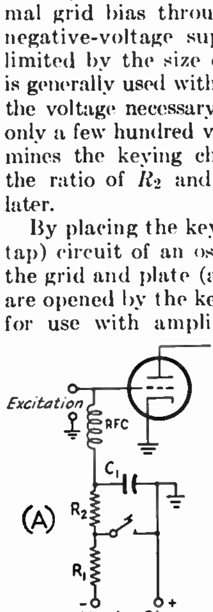


Fig. 8-3 — Blocked-grid keying. R_1, the current-limiting resistor, should have a value of about 50,000 ohms. C_1 may have a capacity of 0.1 to 1 μfd., depending upon the keying characteristic desired. R_2 is the normal value of grid leak for the tube.

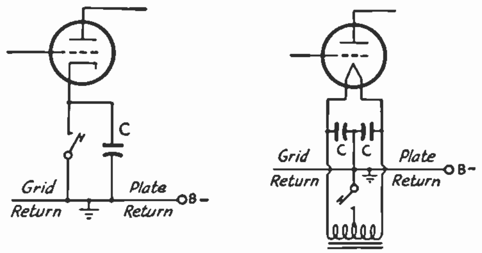


Fig. 8-4 — Cathode and center-tap keying. The condensers C are r.f. by-pass condensers. Their capacity is not critical, values of 0.001 to 0.01 μfd. ordinarily being used.

shaping can be accomplished readily. It is also widely used with oscillators, but here the shaping is often complicated by the grid-circuit time constant. Cathode keying is shown in Fig. 8-4. It is popular for use in low- and medium-power stages, although a keying relay or keyer tube should be used where the plate voltage is more than 300.

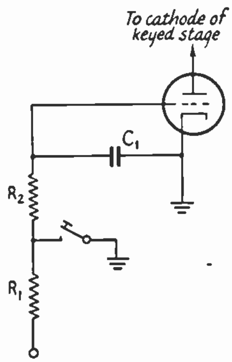


Fig. 8-5 — The basic keyer-tube circuit for cathode or negative-lead keying.

A popular method of keying involves using one or more tubes as **keyer tubes**, in place of a relay. A keyer tube (or tubes) can be used in the negative-lead or cathode-keying circuits of Figs. 8-1 and 8-4. One advantage of tube keying is that the voltage across the key is limited by large resistors, and so the operator has no chance for anything but the slightest electrical shock. A further advantage is that the shaping is done in the grid circuit of the keyer tube with inexpensive parts. The basic keyer tube circuit is shown in Fig. 8-5 — it is similar to the grid-circuit keying of Fig. 8-3.

A **keying relay** can be substituted for a key in any of the keying circuits shown in this chapter. Most keying relays operate from 6.3 or 115 volts a.c., and they should be selected for their speed of operation and adequate insulation for the job to be done. Adequate current-handling capability is also a factor. A typical circuit is shown in Fig. 8-6.

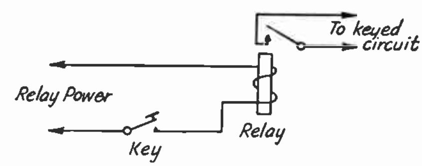


Fig. 8-6 — A keying relay can always be substituted for the key, to provide better isolation from the keyed circuit. An r.f. filter is generally required at the key, and the keying filter is connected in the keyed circuit at the relay contacts.

The relay-coil current that is broken by the key will cause clicks in the receiver, and an r.f. filter (see later in this chapter) is often necessary across the key. The normal keying filter connects at the relay armature contacts in the usual manner. Vibration effects of the keying relay upon the oscillator circuit should be avoided.

Testing Your Keying

The choice of a keying circuit is not as important as its complete testing. Any of the circuits shown in this section can be made to give satisfactory keying, but they must be adjusted properly.

The easiest way to find out what your keyed signal sounds like on the air is to trade stations with a near-by ham friend some evening for a short QSO. If he is a half mile or so away, that's fine, but any distance where the signals are still S9 will be satisfactory.

After you have found out how to work his rig, make contact and then have him send slow dashes, with dash spacing. (The letter "T" at about 5 w.p.m.) With the crystal filter out, cut the r.f. gain back just enough to avoid receiver overloading (the condition where you get crisp signals instead of mushy ones) and tune slowly from out of beat-note range on one side of the signal through to zero and out the other side. Knowing the tempo of the dashes, you can readily identify any clicks in the vicinity as yours or someone else's. A good signal will have a thump on "make" that is perceptible only where you can also hear the beat note, and the click on "break" should be practically negligible at any point. Fig. 8-7A shows how it should sound. If your signal is like that, it will sound good, provided there are no chirps. Then have him run off a string of 35- or 40-w.p.m. dots with the bug — if they are easy to copy, your signal has no "tails" worth worrying about and is a good one for any speed up to the limit of manual keying. If the receiver has poor selectivity with the crystal filter out, make one last check with the filter in (Fig. 8-7B), to see that the clicks off the signal are negligible even at high signal level.

If you don't have any convenient friends with whom to trade stations, you can still check your keying, although you have to be a little more careful The first step is to get rid of the r.f. click at the key, because if you don't you cannot make further observations. Locally (meaning in your own receiver) this click will coincide in time with clicks that may or may not be on your signal, so there is just no way to observe your signal without first eliminating the r.f. click. And unless you have a keying system that breaks no current, you have a

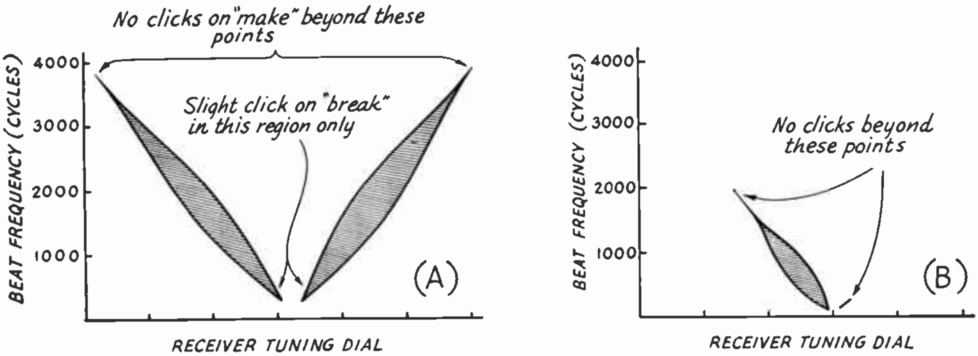


Fig. 8-7 — Representations of a clean c.w. signal as a receiver is tuned through it. (A) shows a receiver with no crystal filter and the b.f.o. set in the center of the passband, and (B) shows the crystal filter in and the receiver adjusted for single-signal reception. The variation in thickness of the lines represents the relative signal intensity. The audio frequency where the signal disappears will depend upon the receiver selectivity characteristic and the strength of the signal.

click at the key. Even the current broken by the key in a vacuum-tube keyer circuit (which is sometimes only 0.1 ma. or so) will cause r.f. clicks that can be heard in your receiver and often in the b.c. set. If you key with a relay, the key opens the relay-coil circuit and clicks are generated at the key as well as at the relay contacts. Don't make the very common mistake of thinking these clicks are the same as the on-the-air clicks discussed earlier — they are not! They are simply local clicks that you must eliminate before you can observe your signal in your receiver. These clicks are the same as the ones you get when you turn an electric light on or off — when you suddenly start or stop current flow, no matter how little, you generate r.f. and that's the click.

Getting rid of this little click is generally no trick at all, unless you're breaking a lot of current. All it requires is a small r.f. filter, as shown in Fig. 8-8. Sometimes just a small (0.001-μfd.) condenser mounted right at the key terminals will do it, and sometimes it will require the full treatment complete with r.f. chokes and second condenser. Measure the normal current through the key leads, remove the transmitter leads, and then connect a d.c. power supply and resistor to give the same current through the key. When your key will break this current with no click, as observed in your receiver and the b.c. set (tuned off any station), you have a suitable r.f. filter at the

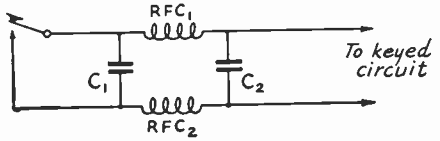


Fig. 8-8 — A filter for eliminating the r.f. click at the key. First try C_1, then add the two r.f. chokes, and then C_2. This filter does not eliminate on-the-air clicks, but it is necessary if you are trying to check keying in your own receiver. It should be mounted right at the key.

C_1, C_2 — 0.01 to 0.001 μfd., not critical.
RFC_1, RFC_2 — 1- to 2.5-mh. r.f. choke.

key and you can reconnect the transmitter. If you use a vacuum-tube keyer, just don't turn on the transmitter but key the normal keyer grid current. If you use a keying relay, first eliminate the click at the key by just keying the relay and adding filter across the key, and then eliminate the click at the relay contacts with another r.f. filter in the relay-keyed circuit. The filter should be mounted right at the key or relay contacts. The objective is to be able to make or break normal key current without generating a local click, and the filtering is usually so simple that the junk box will yield the parts and the process takes longer to describe than to apply.

So far you haven't done a thing for your signal on the air and you still don't know what it sounds like, but you may have cleaned up some clicks in the b.c. set. Now disconnect the antenna from your receiver and short the antenna terminals with a short piece of wire. Tune in your own signal and reduce the r.f. gain to the point where your receiver doesn't overload. Detune any antenna trimmer the receiver may have. If you can't avoid overload within the r.f. gain-control range, pull out the r.f. amplifier tube and try again. If you still can't avoid overload, listen to the second harmonic as a last resort. Since an overloaded receiver can generate clicks, it is easy to realize the importance of eliminating overload during any tests or observations.

Describing the volume level at which you should set your receiver for these "shack" tests is a little difficult. The r.f. filter should be effective with the receiver running wide open and with an antenna connected. When you turn on the transmitter and take the other steps mentioned to reduce the signal in the receiver, run the audio up and the r.f. down to the point where you can just hear a little "rushing" sound with the b.f.o. off and the receiver tuned to the signal. This is with the crystal filter in. At this level, a properly-adjusted keying circuit will show no clicks off the rushing-sound range. With the b.f.o. on and the same gain setting, there should be no clicks outside the beat-note range. When observing clicks, make the slow-dash and fast-dot tests outlined previously.

Now you know how your signal sounds on the air, with one exception. If keying your transmitter makes the house lights blink or the dial light in your receiver flicker, you may not be able to tell too accurately about any chirp on your signal. However, if you are satisfied with the absence of chirp when tuning *either side of zero beat*, it is safe to assume that your receiver isn't chirping with the light flicker and the observed signal is a true representation. No chirp either side of zero beat is fine — some chirp can be either in your transmitter or your receiver, when the lights flicker. But don't try to make these tests without first getting rid of the r.f. click at the key, because clicks can mask a chirp.

In some instances, particularly if the transmitter power is several hundred watts or more, you may find that a small click still persists on all frequencies. If such a click is observed, pull out the last i.f. amplifier tube in your receiver and listen again. If the click is still there, it indicates rectification in the audio system of your receiver, the same type of BCI we condemn cheap midget receivers for. You can cure it with the usual resistor-condenser filter used for curing such BCI cases, or you can leave it in and make mental compensation for it. Any click you hear on your signal should reduce to this minimum click immediately off the signal.

Another unavoidable click can be encountered by r.f. pick-up on the lead from a receiver i.f. amplifier to an "outrigger" selective i.f. amplifier ("Q5-er"). Here again the click will be present at any setting of the receiver tuning control. The solution here is to make your checks with the Q5-er disconnected and the lead removed from the receiver.

Key clicks are caused by the key turning your transmitter on and off too fast — and sometimes by parasitic oscillations in an amplifier — and all a key-click filter does is to slow down the turning-on and turning-off processes. Parasitic clicks occur at points 25 to 100 kc. either side of the signal, and are caused by low-frequency parasitic oscillations triggered by the keying. The cure consists of eliminating the oscillation, not adding key-click filters.

Plate, screen or cathode keying requires a key-click filter of the type shown in Fig. 8-9. Adjustment of such a filter is a simple matter. If the signal has too heavy a click or thump on "make," L should have more inductance. If the click is too heavy on "break," C should have more capacity. The "break" characteristic is also influenced by the value of L, so start with a value of C that reduces the clicks noticeably on "break," adjust the value of L for best "make" characteristic, and then clean up the "break" by further modification of C. Since you may have only a few stray inductances around the shack, you may not find just the value you want for L. In this case, use a value that gives too soft a "make" and then shunt the inductance with resistance to reduce its effect. Transformer windings will often serve as well as standard chokes in this application, so try everything around the shack until you find what you need. For a given voltage, high-current circuits will require more C and less L than will low-current ones.

In the screen-grid keying circuit (Fig. 8-2), the value of R_2 will also affect the "break" characteristic. If R_2 is too large the "break" will tail off too gradually, if it is too small it may introduce a click on "break." In general it is best to start with a value as suggested in Fig. 8-2 and adjust C (Fig. 8-9) for the proper "break" characteristic.

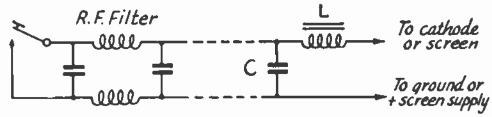


Fig. 8-9 — A key-click filter for cathode, negative-lead or screen keying. It can be located anywhere in the keying line. The values of L and C will vary widely with different currents and voltages, and must be found by cut-and-try. For screen keying, the resistor R_2 (Fig. 8-2) should connect to the junction of L and C.

C — 0.05 to 2.0 μfd.

L — 0.5 to 30 henrys.

Adjustment of control-grid or keyer-tube keying characteristics is simple, since the important components are C_1, R_1 and R_2 (Figs. 8-3 and 8-5). For a given value of C_1, increasing the value of R_2 will soften the "make" characteristic, and increasing the value of R_1 will soften the "break." The value of R_1 will be many times the value of R_2. With grid-block keying, the value of R_2 is determined already if the tube runs grid current, because this will be the normal grid leak, and so the value of C_1 must be adjusted for proper "make" characteristic and then the "break" made satisfactory by adjustment of R_1. Tubes running heavy grid current are not too suitable for grid-block keying because the value of R_1 generally ends up comparatively low and the negative supply must furnish too much current when the key is down.

If you are keying in a low-level stage, don't overlook the clipping action of subsequent stages that are fixed-biased beyond cut-off. It can reintroduce clicks.[2] And if you key your oscillator, don't be too disappointed in the chirp that shows up when you have clickless keying. Amplifier keying is the answer.

[2] For a more complete discussion of this effect, see Carter, "Reducing Key Clicks," *QST*, March, 1949.

Vacuum-Tube Keyers

The practical tube-keyer circuit of Fig. 8-10 can be used for keying any stage of any transmitter. Depending upon the power level of the keyed stage, more or fewer Type 45 tubes can be connected in parallel to handle the necessary current. The voltage drop through a single 45 varies from about 90 volts at 50 ma. to 50 volts at 20 ma. Tubes added in parallel will reduce the drop in proportion to the number of tubes used.

When connecting the output terminals of the keyer to the circuit to be keyed, the grounded output terminal of the keyer must be connected to the transmitter ground. Thus the keyer can be used only in negative-lead or cathode keying. When used in cathode keying, it will introduce cathode bias to the stage and reduce the output. This can be compensated for by a reduction in the grid-leak bias of the stage.

The negative-voltage supply (T_1, C_1, R_1 and the 80 rectifier) can be eliminated if a negative voltage is available from some other source, such as a bias supply. A simplified version of this circuit could eliminate S_1 and S_2 and their associated resistors and condensers, since they are incorporated only to allow the operator to select the combination he prefers. But once the values have been selected, they can be soldered permanently in place. The rule for adjusting the keying characteristic is the same as for blocked-grid keying.

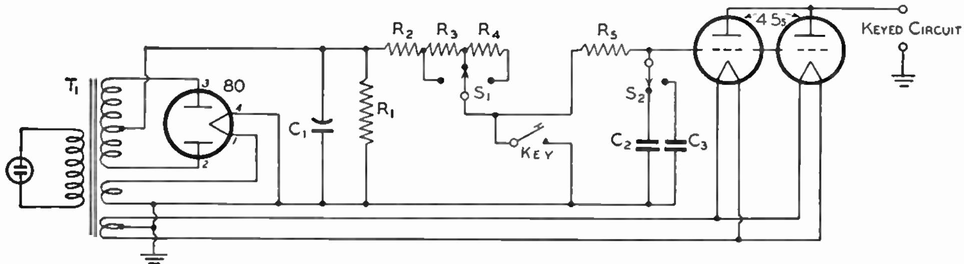


Fig. 8-10 — Wiring diagram of a practical vacuum-tube keyer.

C_1 — 2-μfd. 600-volt paper.
C_2 — 0.0033-μfd. mica.
C_3 — 0.0047-μfd. mica.
R_1 — 0.22 megohm, 1 watt.
R_2 — 50,000 ohms, 10 watts.
R_3, R_4 — 4.7 megohms, 1 watt.
R_5 — 0.47 megohm, 1 watt.
S_1, S_2 — 1-circuit rotary switch.
T_1 — 350-0-350 volts, 5 volts and 2.5 volts (Stancor P6003).

A Low-Power Keyer

If a low-level stage running only a few watts is to be keyed, the tube-keyer circuit of Fig. 8-11 offers a simple solution. By using a 117L7 type tube, which incorporates its own rectifier, it is only necessary to connect to some existing power supply at the point marked "X". The keying characteristic will vary with many factors, so the values of R_1 and R_2 only represent starting points for experimentation.

When the key or keying lead has poor insulation, the resistance may become low enough (particularly in humid weather) to reduce the blocking voltage and allow the keyer tube to pass some current. This may cause a slight backwave, but it can be cured by better insulation, or by reduced values of R_3 and R_4 in Fig. 8-10 or R_1 in Fig. 8-11.

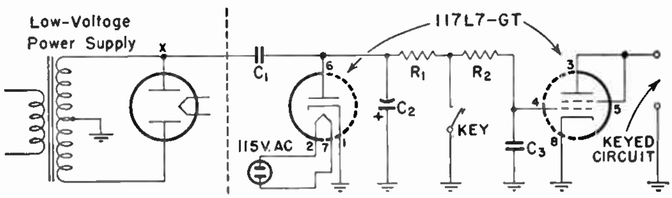


Fig. 8-11 — Simple low-power vacuum-tube keyer.

C_1 — 0.5-μfd. 600-volt paper.
C_2 — 8-μfd. 450-volt electrolytic.
C_3 — 0.01-μfd. ceramic.
R_1 — 1 megohm, ½ watt.
R_2 — 0.1 megohm, ½ watt.

Connect keyer to a low-voltage power supply at point "X".

Monitoring of Keying

In general, there are two common methods for monitoring one's "fist" and signal. The first, and perhaps more common type, involves the use of an audio oscillator that is keyed simultaneously with the transmitter.

The second method is one that permits receiving the signal through one's receiver, and this generally requires that the receiver be tuned to the transmitter (not always convenient unless working on the same frequency) and that some method be provided for preventing overloading of the receiver, so that a good replica of the transmitted signal will be received. Except where quite low power is used, this usually involves a relay for simultaneously shorting the receiver input terminals and reducing the receiver gain.

The Monitone—for C.W. and 'Phone

The "Monitone" is a useful device for monitoring c.w. or 'phone transmissions. When used for c.w. work, it furnishes an audio tone every time the transmitter key is closed, and it also blanks the receiver output at the same time. When used with a 'phone transmitter, it blanks the receiver when the transmitter carrier is turned on, and also furnishes an audio replica of the transmitted signal, at any desired volume level. The Monitone requires no direct connection to the transmitter or key, and no changes are needed in the receiver. The sidetone and blanking are keyed by the r.f. output of the transmitter, regardless of frequency.

Referring to Fig. 8-12, the 6SL7GT acts as a dual amplifier, for the receiver output and for the sidetone oscillator (consisting of the neon bulb NE-2, C_6 and $R_{10} + R_{11}$). When r.f. from the transmitter is fed in at J_1 it is rectified by XTAL and a negative voltage is developed across R_9. This negative voltage cuts off the 6J5 and one-half of the 6SL7GT. The neon-bulb oscillator goes into action and the resultant tone is amplified in the other half of the 6SL7GT. For 'phone work, S_{1B} is opened and S_{1A} is closed. This turns off the sidetone oscillator and feeds the rectified audio from the transmitter through volume control R_6.

The tone of the neon-bulb oscillator is varied by the position of R_{11}. Since the power drain of the Monitone is only about 5 ma. at 250 volts, a resistor is used instead of a filter choke in the power supply.

One method of construction of the Monitone is to use a 6-inch cube aluminum utility box (ICA No. 29843) for a cabinet, mounting the components on one removable wall and a small 2-inch chassis fastened to this wall. R_6, R_{11}, S_2, J_2 and NE-2 can be mounted on the panel, with NE-2 projecting through a rubber grommet. The 1N34 crystal and most of the neon-oscillator parts can mount on the 6J5 socket, and the audio components can be grouped around the 6SL7 socket. A tip jack for the r.f. pick-up lead can be mounted on the rear wall of the chassis, near where the 115-volt line cord and the shielded lead to P_1 are brought out. It is advisable to keep the power-supply wiring and components away from the audio.

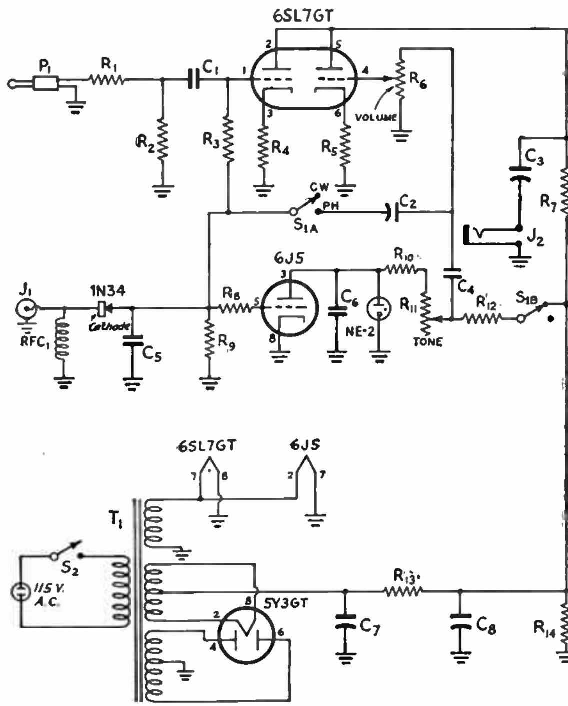


Fig. 8-12 — Wiring diagram of the Monitone.

C_1 — 0.005-μfd. disc ceramic.
C_2, C_3 — 0.1-μfd. 400-volt paper.
C_4 — 250-μμfd. ceramic.
C_5 — 100-μμfd. ceramic.
C_6 — 0.001-μfd. disc ceramic.
C_7, C_8 — 8-μfd. 450-volt electrolytic.
R_1 — 6800 ohms, ½ watt.
R_2 — 1000 ohms, ½ watt.
R_3 — 0.56 megohm, ½ watt.
R_4, R_5 — 1200 ohms, ½ watt.
R_6 — 1-megohm potentiometer (Mallory U-53).
R_7 — 22,000 ohms, 1 watt.
R_8 — 68,000 ohms, ½ watt.
R_9, R_{10} — 1 megohm, ½ watt.
R_{11} — 3-megohm potentiometer (Mallory U-59).
R_{12} — 2.2 megohms, ½ watt.
R_{13} — 47,000 ohms, 1 watt.
R_{14} — 0.1 megohm, 1 watt.
J_1 — Tip jack.
J_2 — Open-circuit jack.
P_1 — 'Phone plug.
RFC_1 — 2.5-mh. r.f. choke.
S_{1A}, S_{1B} — S.p.d.t. switch; see text. (Mallory US-28.)
S_2 — S.p.s.t. toggle switch.
T_1 — Replacement transformer (Stancor P-6010).

Changeover switch $S_{1A}S_{1B}$ is mounted on the tone potentiometer, R_{11}, and is wired so that S_{1A} is closed when the control arm for the potentiometer is rotated to the extreme counterclockwise position. S_{1B} should open at this setting of the tone control. $S_{1A}S_{1B}$, labeled by the manufacturer as a s.p.d.t. switch, is actually a pair of s.p.s.t. switches built into a single assembly.

Installation & Operation

The Monitone is used by plugging the audio plug, P_1, into the headphone jack of the receiver, the headphones into J_2 of the Monitone, and applying 115 volts a.c. A length of wire must be run from the r.f. input jack, J_1, to a point where it can pick up r.f. from the transmitter antenna system. With S_{1B} and the power switch, S_2, closed, the transmitter may be turned on and the position of the r.f. pick-up lead (Caution! High voltage!) adjusted for a sustained oscillation of the neon tube circuit. Sufficient r.f. coupling between the transmitter and the monitor is indicated by a glow in the bulb and by the side-tone as heard in the headphones.

The r.f. field around the antenna system may vary in strength as the transmitter is switched from one band to another. Usually, however, a coupling adjustment made at one frequency will suffice for all other frequencies as long as the pick-up line is coupled to one side of the antenna tuner and not the transmission line.

Break-In Operation

Break-in operation requires a separate receiving antenna, since none of the available antenna change-over relays is fast enough to follow keying. The receiving antenna should be installed as far as possible from the transmitting antenna. It should be mounted at right angles to the transmitting antenna and fed with low pick-up lead-in material such as coaxial cable or 300-ohm Twin-Lead, to minimize pick-up.

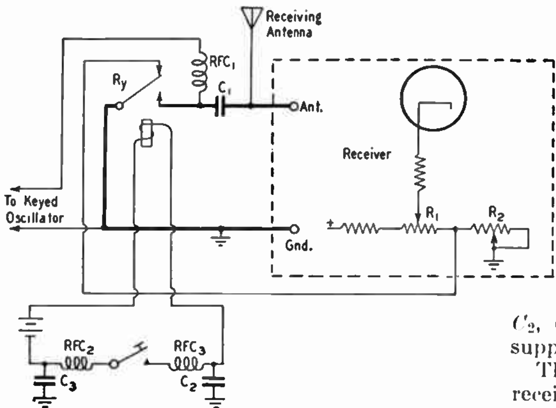


Fig. 8-13 — Wiring diagram for smooth break-in operation. The leads shown as heavy lines should be kept as short as possible, for minimum pick-up of the transmitter signal.

C_1, C_2, C_3 — 0.001 μfd.
R_1 — Receiver manual gain control.
R_2 — 5000- or 10,000-ohm wire-wound potentiometer.
RFC_1, RFC_2, RFC_3 — 2.5-mh. r.f. choke.
Ry — S.p.d.t. keying relay.

If a low-powered transmitter is used, it is often quite satisfactory to use no special equipment for break-in operation other than the separate receiving antenna, since the transmitter will not block the receiver too seriously. Even if the transmitter keys without clicks, some clicks will be heard when the receiver is tuned to the transmitter frequency because of overload in the receiver. An output limiter, as described in Chapter Five, will wash out these clicks and permit good break-in operation even on your transmitter frequency.

When powers above 25 or 50 watts are used, special treatment is required for quiet break-in on the transmitter frequency. A means should be provided for shorting the input of the receiver when the code characters are sent, and a means for reducing the gain of the receiver at the same time is often necessary. The system shown in Fig. 8-13 permits quiet break-in operation for higher-powered stations. It requires a simple operation on the receiver but otherwise is perfectly straightforward. R_1 is the regular receiver r.f. and i.f. gain control. The ground lead is lifted on this control and run to a rheostat, R_2, that goes to ground. A wire from the junction runs outside the receiver to the keying relay, R_y. When the key is up, the ground side of R_1 is connected to ground through the relay arm, and the receiver is in its normal operating condition. When the key is closed, the relay closes, which breaks the ground connection from R_1 and applies additional bias to the tubes in the receiver. This bias is controlled by R_2. When the relay closes, it also closes the circuit to the transmitter oscillator. C_2, C_3, RFC_2 and RFC_3 compose a filter to suppress the clicks caused by the relay current.

The keying relay should be mounted on the receiver as close to the antenna terminals as possible, and the leads shown heavy in the diagram should be kept short, since long leads will allow too much signal to get through into the receiver. A good high-speed keying relay should be used. If a two-wire line is used from the receiving antenna, another r.f. choke, RFC_4, will be required. The revised portion of the schematic is shown in Fig. 8-14.

A DE LUXE BREAK-IN SYSTEM

In many instances it is quite difficult to key an oscillator without clicks *and* chirps. Most oscillators will key without apparent chirp if the rise and decay times are made very short, but this introduces key clicks that cannot be

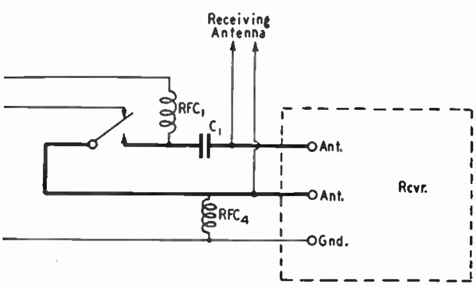


Fig. 8-14 — Necessary circuit revision of Fig. 8-13 if a two-wire lead from the receiving antenna is used. RFC_4 is a 2.5-mh. r.f. choke — other values are the same as in Fig. 8-13.

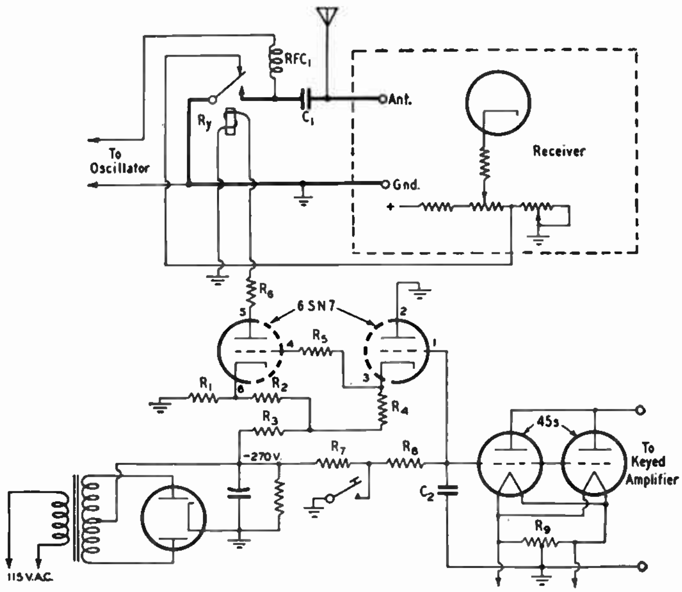


Fig. 8-15 — A de luxe break-in system that holds the oscillator circuit closed (and the receiver input shorted) during a string of fast dots but opens between letters or words.

C_1 — 0.001-μfd. mica.
C_2 — 0.0047-μfd. mica.
R_1 — 20,000 ohms, 10 watts, wire-wound.
R_2 — 1800 ohms.
R_3 — 1500 ohms.
R_4, R_5 — 1.0 megohm.
R_6 — 4700 ohms.
R_7 — 6.8 megohm.
R_8 — 0.47 megohm.
R_9 — 50-ohm center-tapped resistor, 2 watts.
All resistors 1-watt composition unless otherwise noted.
RFC_1 — 2.5-mh. r.f. choke.
Ry — High-speed relay, 1400-ohm 18-volt coil (Stevens-Arnold Type 172 Millisec relay).

avoided. The system shown in Fig. 8-15 avoids this trouble by turning on the oscillator quickly, keying an amplifier with a vacuum-tube keyer, and turning off the oscillator after the amplifier keying is finished. The oscillator is turned on and off without lag, but the resultant clicks are not passed through the transmitter. Actually, with keying speeds faster than about 15 w.p.m., the oscillator will stay turned on for a letter or even a word, but it turns off between words and allows the transmitting station to hear the "break" signal of the other station. It requires one tube more than the ordinary vacuum-tube keyer and a special high-speed relay.

As can be seen from Fig. 8-15, the circuit is a combination of the break-in system of Fig. 8-13 and the tube keyer of Fig. 8-11, with a 6SN7 tube and a few resistors added. Normally the left-hand portion of the 6SN7 is biased to a low value of plate current by the drop through R_2 (part of the bleeder $R_1R_2R_3$) and the relay is open. When the key is closed and C_2 starts to discharge, the right-hand portion of the 6SN7 draws current and this in turn puts a less-negative voltage on the grid of the left-hand portion. The tube draws current and the relay closes. The relay will stay closed until the negative voltage across C_2 is close to the supply voltage, and consequently a string of dots or dashes (which doesn't give C_2 a chance to charge to full negative) will keep the relay closed. In adjusting the system, R_2 controls the amount of idling current through the relay and R_6 determines the voltage across the relay. R_7, R_8 and C_2 are the normal resistors and condenser for the tube keyer. When adjusted properly, the relay will close without delay on the first dot and open quickly during the spaces between words or slower letters. When idling, the voltage across the relay should be one or two volts — with the key down it should be 18 volts.

The oscillator should be designed to key as fast as possible, which means that series resistances and shunt capacitances should be held to a minimum. Negative plate-lead keying is slightly faster than cathode keying and should be used in the oscillator. The keyer tubes are connected in the cathode circuit of an amplifier stage to avoid reaction on the oscillator. By using blocked-grid keying of the amplifier stage, the keyer tubes can be eliminated.

Full descriptions of allied systems for break-in operation can be found in the following *QST* articles:

Miller and Meichner, "TVG — An Aid to Break-In," March, 1953.
Puckett, "'DeLuxe' Keying Without Relays," September, 1953; Part II, Dec., 1953.

● ELECTRONIC KEYS

Electronic keys, as contrasted with mechanical automatic keys, use vacuum tubes or relays (or both) to form automatic dashes as well as automatic dots. Full descriptions of electronic keys can be found in the following *QST* articles:

Brann, "In Search of the Ideal Electronic Key," Feb., 1951.
Turrin, "Debugging the Electronic Bug," Jan., 1950.
Montgomery, "'Corkey' — A Tubeless Automatic Key," November, 1950.
Bartlett, "Compact Automatic Key Design," Dec., 1951.
Turrin, "The 'Tur-Key'", December, 1952. Correction, February, 1953.
Kaye, "The 'Ultimatic' — The Key with a Memory," February, 1953. Note on suitable relays, April, 1953.
Brann, "A Dot Anticipator for the Electronic Key," July, 1953.

Speech Amplifiers and Modulators

The audio amplifiers used in radiotelephone transmitters operate on the principles outlined earlier in this book in the chapter on vacuum tubes. The design requirements are determined principally by the type of modulation system to be used and by the type of microphone to be employed. It is necessary to have a clear understanding of modulation principles before the problem of laying out a speech system can be approached successfully. Those principles are discussed under appropriate chapter headings.

The present chapter deals with the design of audio amplifier systems for communication purposes. In voice communication the primary objective is to obtain the most *effective* transmission; i.e., to make the message be understood at the receiving point in spite of adverse conditions created by noise and interference. The methods used to accomplish this do not necessarily coincide with the methods used for other purposes, such as the reproduction of music or other program material. In other words, "naturalness" in reproduction is distinctly secondary to intelligibility.

The fact that satisfactory intelligibility can be maintained in a relatively narrow band of frequencies is particularly fortunate, because the width of the channel occupied by a 'phone transmitter is directly proportional to the width of the audio-frequency band. If the channel width is reduced, more stations can occupy a given band of frequencies without mutual interference.

In speech transmission, amplitude distortion of the voice wave has very little effect on intelligibility. Its importance in communication lies almost wholly in the fact that many of the audio-frequency harmonics caused by such distortion lie outside the channel needed for intelligible speech, and thus will create unnecessary interference to other stations.

Speech Equipment

In designing speech equipment it is necessary to know (1) the amount of audio power the modulation system must furnish and (2) the output voltage developed by the microphone when it is spoken into from normal distance (a few inches) with ordinary loudness. It then becomes possible to choose the number and type of amplifier stages needed to generate the required audio power without overloading or distortion anywhere in the system.

MICROPHONES

The **level** of a microphone is its electrical output for a given sound intensity. Level varies greatly with microphones of different types, and depends on the distance of the speaker's lips from the microphone. Only approximate values based on averages of "normal" speaking voices can be given. The values given later are based on close talking; that is, with the microphone about an inch from the speaker's lips.

The **frequency response** or **fidelity** of a microphone is its relative ability to convert sounds of different frequencies into alternating current. For understandable speech transmission only a limited frequency range is necessary, and intelligible speech can be obtained if the output of the microphone does not vary more than a few decibels at any frequency within a range of about 200 to 2500 cycles. When the variation expressed in terms of decibels is small between two frequency limits, the microphone is said to be **flat** between those limits.

Carbon Microphones

The **carbon microphone** consists of a metal diaphragm placed against an insulating cup containing loosely-packed carbon granules (**microphone button**). Current from a battery flows through the granules, the diaphragm being one connection and the metal backplate the other. Fig. 9-1A shows connections for carbon microphones. A variable resistor is included for adjusting the button current to the value as specified with the microphone. The primary of a transformer is connected in series with the battery and microphone.

As the diaphragm vibrates, its pressure on the granules alternately increases and decreases, causing a corresponding increase and decrease of current flow through the circuit, since the pressure changes the resistance of the mass of granules. The resulting change in the current flowing through the transformer primary causes an alternating voltage, of corresponding frequency and intensity, to be set up in the transformer secondary.

Good-quality carbon microphones give outputs ranging from 0.1 to 0.3 volt across 50 to 100 ohms; that is, across the primary winding of the microphone transformer. With the step-up of the transformer, a peak voltage of between 3 and 10 volts can be assumed to be available at the grid of the amplifier tube. The usual button current is 50 to 100 ma.

Crystal Microphones

The **crystal microphone** makes use of the piezoelectric properties of Rochelle salts crystals. This type of microphone requires no battery or transformer and can be connected directly to the grid of an amplifier tube. It is the most popular type of microphone among amateurs, for these reasons as well as the fact that it has good frequency response and is available in inexpensive models. The input circuit for the crystal microphone is shown in Fig. 9-1B.

Although the level of crystal microphones varies with different models, an output of 0.03 volt or so is representative for communication types. The level is affected by the length of the cable connecting the microphone to the first amplifier stage; the above figure is for lengths of 6 or 7 feet. The frequency characteristic is unaffected by the cable, but the load resistance (amplifier grid resistor) does affect it; the lower frequencies are attenuated as the value of load resistance is lowered. A grid-resistor value of at least 1 megohm should be used for reasonably flat response, 5 megohms being a customary figure.

Velocity and Dynamic Microphones

In a **velocity** or **"ribbon" microphone,** the element acted upon by the sound waves is a thin corrugated metallic ribbon suspended between the poles of a magnet. When vibrating, the ribbon cuts the lines of force between the poles, first in one direction and then the other, thus generating an alternating voltage.

Velocity microphones are built in two types, high impedance and low impedance, the former being used in most applications. A high-impedance microphone can be directly connected to the grid of an amplifier tube, shunted by a resistance of 0.5 to 5 megohms (Fig. 9-1C). Low-impedance microphones are used when a long connecting cable (75 feet or more) must be employed. In such a case the output of the microphone is coupled to the first amplifier stage through a suitable step-up transformer, as shown in Fig. 9-1D.

The level of the velocity microphone is about 0.03 to 0.05 volt. This figure applies directly to the high-impedance type, and to the low-impedance type when the voltage is measured across the secondary of the coupling transformer.

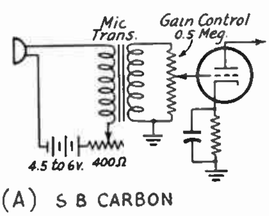
The **dynamic microphone** somewhat resembles a dynamic loudspeaker. A light-weight voice coil is rigidly attached to a diaphragm, the coil being suspended between the poles of a permanent magnet. Sound causes the diaphragm to vibrate, thus moving the coil back and forth between the magnet poles and generating an alternating voltage.

The dynamic microphone usually is built with high-impedance output, suitable for working directly into the grid of an amplifier tube. If the connecting cable must be unusually long, a low-impedance type should be used, with a step-up transformer at the end of the cable.

A small permanent-magnet 'speaker can be used as a dynamic microphone, although the fidelity is not as good as is obtainable with a properly-designed microphone.

THE SPEECH AMPLIFIER

The audio-frequency amplifier stage that causes the r.f. carrier output to be varied is called the **modulator,** and all the amplifier stages preceding it comprise the **speech amplifier.** Depending on the modulator used, the speech amplifier may be called upon to deliver a power output ranging from practically zero (only voltage required) to 20 or 30 watts.



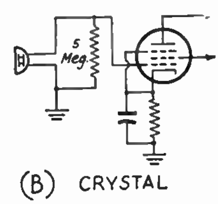
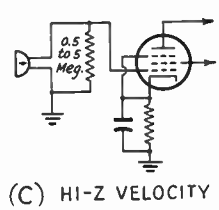
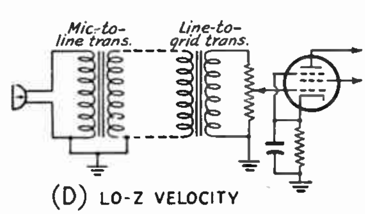


Fig. 9-1 — Speech input circuits used with various types of microphones.





Before starting the design of a speech amplifier, therefore, it is necessary to have selected a suitable modulator for the transmitter. This selection must be based on the power required to modulate the transmitter, and this power in turn depends on the type of modulation system selected, as described in other chapters. With the modulator picked out, its **driving-power** requirements (audio power required to excite the modulator to full output) can be determined from the tube tables in the last chapter. Generally speaking, it is advisable to choose a tube or tubes for the last stage of the speech amplifier that will be capable of

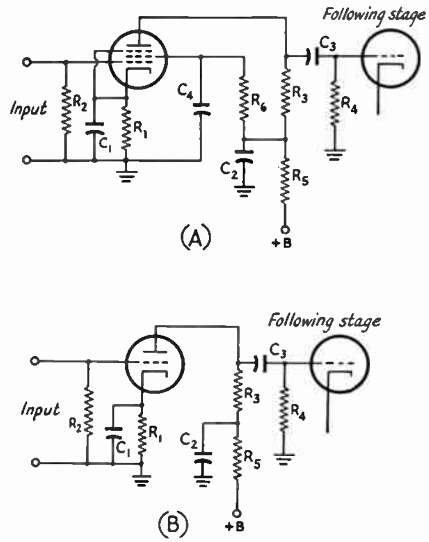


Fig. 9-2 — Resistance-coupled voltage-amplifier circuits. A, pentode; B, triode. Designations are as follows:
C_1 — Cathode by-pass condenser.
C_2 — Plate by-pass condenser.
C_3 — Output coupling condenser (blocking condenser).
C_4 — Screen by-pass condenser.
R_1 — Cathode resistor.
R_2 — Grid resistor.
R_3 — Plate resistor.
R_4 — Next-stage grid resistor.
R_5 — Plate decoupling resistor.
R_6 — Screen resistor.
Values for suitable tubes are given in Table 9-I. Values in the decoupling circuit, C_2R_5, are not critical. R_5 may be about 10% of R_3; an 8- or 10-μfd. electrolytic condenser is usually large enough at C_2.

developing at least 50 per cent more power than the rated driving power of the modulator. This will provide a factor of safety so that losses in coupling transformers, etc., will not upset the calculations.

Voltage Amplifiers

If the last stage in the speech amplifier is a Class AB_2 or Class B amplifier, the stage ahead of it must be capable of sufficient power output to drive it. However, if the last stage is a Class AB_1 or Class A amplifier the preceding stage can be simply a voltage amplifier. From there on back to the microphone, all stages are voltage amplifiers.

The important characteristics of a voltage amplifier are its **voltage gain,** maximum undistorted **output voltage,** and its **frequency response.** The voltage gain is the voltage-amplification ratio of the stage. The output voltage is the maximum a.f. voltage that can be secured from the stage without distortion. The amplifier frequency response should be adequate for voice reproduction; this requirement is easily satisfied.

The voltage gain and maximum undistorted output voltage depend on the operating conditions of the amplifier. Data on the popular types of tubes used in speech amplifiers are given in Table 9-I, for resistance-coupled amplification. The output voltage is in terms of *peak* voltage rather than r.m.s.; this makes the rating independent of the waveform. Exceeding the peak value causes the amplifier to distort, so it is more useful to consider only peak values in working with amplifiers.

Resistance Coupling

Resistance coupling generally is used in voltage-amplifier stages. It is relatively inexpensive, good frequency response can be secured, and there is little danger of hum pick-up from stray magnetic fields associated with heater wiring. It is the only type of coupling suitable for the output circuits of pentodes and high-μ triodes, because with transformers a sufficiently high load impedance cannot be obtained without considerable frequency distortion. Typical circuits are given in Fig. 9-2 and design data in Table 9-I.

Transformer Coupling

Transformer coupling between stages ordinarily is used only when power is to be transferred (in such a case resistance coupling is very inefficient), or when it is necessary to couple between a single-ended and a push-pull stage. Triodes having an amplification factor of 20 or less are used in transformer-coupled voltage amplifiers. With transformer coupling, tubes should be operated under the Class A conditions given in the tube tables at the end of this book.

Representative circuits for coupling single-ended to push-pull stages are shown in Fig. 9-3. The circuit at A combines resistance and transformer coupling, and may be used for exciting the

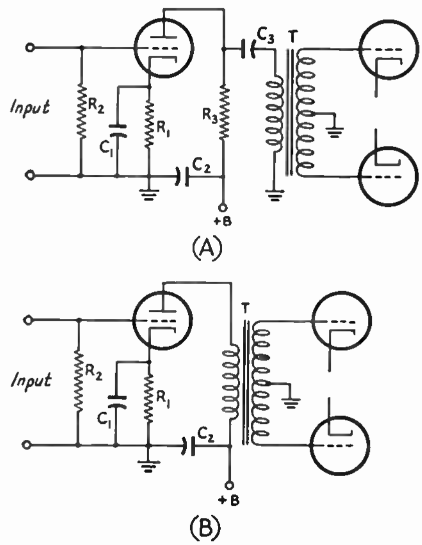


Fig. 9-3 — Transformer-coupled amplifier circuits for driving a push-pull amplifier. A is for resistance-transformer coupling; B for transformer coupling. Designations correspond to those in Fig. 9-2. In A, values can be taken from Table 9-I. In B, the cathode resistor is calculated from the rated plate current and grid bias as given in the tube tables for the particular type of tube used.

TABLE 9-I — RESISTANCE-COUPLED VOLTAGE-AMPLIFIER DATA

Data are given for a plate supply of 300 volts. Departures of as much as 50 per cent from this supply voltage will not materially change the operating conditions or the voltage gain, but the output voltage will be in proportion to the new voltage. Voltage gain is measured at 400 cycles; condenser values given are based on 100-cycle cut-off. For increased low-frequency response, all condensers may be made larger than specified (cut-off frequency in inverse proportion to condenser values provided all are changed in the same proportion). A variation of 10 per cent in the values given has negligible effect on the performance.

	Plate Resistor Megohms	Next-Stage Grid Resistor Megohms	Screen Resistor Megohms	Cathode Resistor Ohms	Screen By-pass µfd.	Cathode By-pass µfd.	Blocking Condenser µfd.	Output Volts (Peak)[1]	Voltage Gain [2]
6SJ7, 12SJ7	0.1	0.1	0.35	500	0.10	11.6	0.019	72	67
		0.25	0.37	530	0.09	10.9	0.016	96	98
		0.5	0.47	590	0.09	9.9	0.007	101	104
	0.25	0.25	0.89	850	0.07	8.5	0.011	79	139
		0.5	1.10	860	0.06	7.4	0.004	88	167
		1.0	1.18	910	0.06	6.9	0.003	98	185
	0.5	0.5	2.0	1300	0.06	6.0	0.004	64	200
		1.0	2.2	1410	0.05	5.8	0.002	79	238
		2.0	2.5	1530	0.04	5.2	0.0015	89	263
6J7, 7C7, 12J7-GT	0.1	0.1	0.44	500	0.07	8.5	0.02	55	61
		0.25	0.5	450	0.07	8.3	0.01	81	82
		0.5	0.53	600	0.06	8.0	0.006	96	94
	0.25	0.25	1.18	1100	0.04	5.5	0.008	81	104
		0.5	1.18	1200	0.04	5.4	0.005	104	140
		1.0	1.45	1300	0.05	5.8	0.005	110	185
	0.5	0.5	2.45	1700	0.04	4.2	0.005	75	161
		1.0	2.9	2200	0.04	4.1	0.003	97	200
		2.0	2.95	2300	0.04	4.0	0.0025	100	230
6AU6, 6SH7, 12AU6, 12SH7	0.1	0.1	0.2	500	0.13	18.0	0.019	76	109
		0.22	0.24	600	0.11	16.4	0.011	103	145
		0.47	0.26	700	0.11	15.3	0.006	129	168
	0.22	0.22	0.42	1000	0.1	12.4	0.009	92	164
		0.47	0.5	1000	0.098	12.0	0.007	108	230
		1.0	0.55	1100	0.09	11.0	0.003	122	262
	0.47	0.47	1.0	1800	0.075	8.0	0.0045	94	248
		1.0	1.1	1900	0.065	7.6	0.0028	105	318
		2.2	1.2	2100	0.06	7.3	0.0018	122	371
6AQ6, 6AQ7, 6AT6, 6Q7, 6SL7GT, 6SZ7, 6T8, 12AT6, 12Q7-GT, 12SL7-GT (one triode)	0.1	0 1	—	1500	—	4.4	0.027	40	34
		0.22	—	1800	—	3.6	0.014	54	38
		0.47	—	2100	—	3.0	0.0065	63	41
	0.22	0.22	—	2600	—	2.5	0.013	51	42
		0.47	—	3200	—	1.9	0.0065	65	46
		1.0	—	3700	—	1.6	0.0035	77	48
	0.47	0.47	—	5200	—	1.2	0.006	61	48
		1.0	—	6300	—	1.0	0.0035	74	50
		2.2	—	7200	—	0.9	0.002	85	51
6AV6, 12AV6, 12AX7 (one triode)	0.1	0.1	—	1300	—	4.6	0.027	43	45
		0.22	—	1500	—	4.0	0.013	57	52
		0.47	—	1700	—	3.6	0.006	66	57
	0.22	0.22	—	2200	—	3.0	0.013	54	59
		0.47	—	2800	—	2.3	0.006	69	65
		1.0	—	3100	—	2.1	0.003	79	68
	0.47	0.47	—	4300	—	1.6	0.006	62	69
		1.0	—	5200	—	1.3	0.003	77	73
		2.2	—	5900	—	1.1	0.002	92	75
6SC7, 12SC7 [3] (one triode)	0.1	0.1	—	750	—	—	0.033	35	29
		0.25	—	930	—	—	0.014	50	34
		0.5	—	1040	—	—	0.007	54	36
	0.25	0.25	—	1400	—	—	0.012	45	39
		0.5	—	1680	—	—	0.006	55	42
		1.0	—	1840	—	—	0.003	64	45
	0.5	0.5	—	2330	—	—	0.006	50	45
		1.0	—	2980	—	—	0.003	62	48
		2.0	—	3280	—	—	0.002	72	49
6J5, 7A4, 7N7, 6SN7GT, 12J5-GT, 12SN7-GT (one triode)	0.05	0.05	—	1020	—	3.56	0.06	41	13
		0.1	—	1270	—	2.96	0.034	51	14
		0.25	—	1500	—	2.15	0.012	60	14
	0.1	0.1	—	1900	—	2.31	0.035	43	14
		0.25	—	2440	—	1.42	0.0125	56	14
		0.5	—	2700	—	1.2	0.0065	64	14
	0.25	0.25	—	4590	—	0.87	0.013	46	14
		0.5	—	5770	—	0.64	0.0075	57	14
		1.0	—	6950	—	0.54	0.004	64	14
6C4, 12AU7 (one triode)	0.047	0.047	—	870	—	4.1	0.065	38	12
		0.1	—	1200	—	3.0	0.034	52	12
		0.22	—	1500	—	2.4	0.016	68	12
	0.1	0.1	—	1900	—	1.9	0.032	44	12
		0.22	—	3000	—	1.3	0.016	68	12
		0.47	—	4000	—	1.1	0.007	80	12
	0.22	0.22	—	5300	—	0.9	0.015	57	12
		0.47	—	800	—	0.52	0.007	82	12
		1.0	—	11000	—	0.46	0.0035	92	12

[1] Voltage across next-stage grid resistor at grid-current point.
[2] At 5 volts r.m.s. output.
[3] Cathode-resistor values are for phase-inverter service.

grids of a Class A or AB_1 following stage. The resistance coupling is used to keep the d.c. plate current from flowing through the transformer primary, thereby preventing a reduction in primary inductance below its no-current value; this improves the low-frequency response. With low-μ triodes (6C5, 6J5, etc.), the gain is equal to that with resistance coupling multiplied by the secondary-to-primary turns ratio of the transformer.

In B the transformer primary is in series with the plate of the tube, and thus must carry the tube plate current. When the following amplifier operates without grid current, the voltage gain of the stage is practically equal to the μ of the tube multiplied by the transformer ratio. This circuit also is suitable for transferring power (within the capabilities of the tube) to a following Class AB_2 or Class B stage.

Phase Inversion

Push-pull output may be secured with resistance coupling by using "phase-inverter" or "phase-splitter" circuits as shown in Fig. 9-4.

The circuits shown in Fig. 9-4 are of the "self-balancing" type. In A, the amplified voltage from V_1 appears across R_5 and R_7 in series. The drop across R_7 is applied to the grid of V_2, and the amplified voltage from V_2 appears across R_6 and R_7 in series. This voltage is 180 degrees out of phase with the voltage from V_1, thus giving push-pull output. The part that appears across R_7 from V_2 opposes the voltage from V_1 across R_7, thus reducing the signal applied to the grid of V_2. The negative feed-back so obtained tends to regulate the voltage applied to the phase-inverter tube so that the output voltages from both tubes are substantially equal. The gain is slightly less than twice the gain of a single-tube amplifier using the same operating conditions.

In the single-tube circuit shown in Fig. 9-4B the plate load resistor is divided into two equal parts, R_9 and R_{10}, one being connected to the plate in the normal way and the other between cathode and ground. Since the voltages at the plate and cathode are 180 degrees out of phase, the grids of the following tubes are fed equal a.f. voltages in push-pull. The grid return of V_3 is made to the junction of R_8 and R_{10} so normal bias will be applied to the grid. This circuit is highly degenerative because of the way R_{10} is connected. The voltage gain is less than 2 even when a high-μ triode is used at V_3.

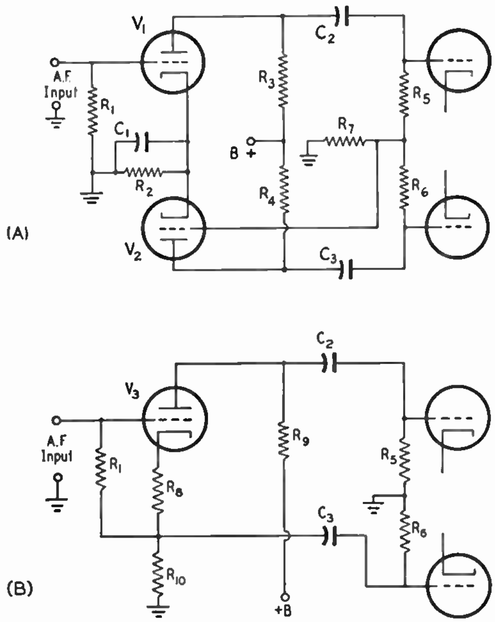


Fig. 9-4 — Self-balancing phase-inverter circuits. V_1 and V_2 may be a double triode such as the 6SN7GT or 6SL7GT. V_3 may be any of the triodes listed in Table 9-I, or one section of a double triode.

R_1 — Grid resistor (1 megohm or less).
R_2 — Cathode resistor; use one-half value given in Table 9-I for tube and operating conditions chosen.
R_3, R_4 — Plate resistor; select from Table 9-I.
R_5, R_6 — Following-stage grid resistor (0.22 to 0.47 megohm).
R_7 — 0.22 megohm.
R_8 — Cathode resistor; select from Table 9-I.
R_9, R_{10} — Each one-half of plate load resistor given in Table 9-I.
C_1 — 10-μfd. electrolytic.
C_2, C_3 — 0.01- to 0.1-μfd. paper.

Gain Control

A means for varying the over-all gain of the amplifier is necessary for keeping the final output at the proper level for modulating the transmitter. The common method of gain control is to adjust the value of a.c. voltage applied to the grid of one of the amplifiers by means of a voltage divider or potentiometer.

The gain-control potentiometer should be near the input end of the amplifier, at a point where the signal voltage level is so low there is no danger that the stages ahead of the gain control will be overloaded by the full microphone output. With carbon microphones the gain control may be placed directly across the microphone-transformer secondary. With other types of microphones, however, the gain control usually will affect the frequency response of the microphone when connected directly across it. Also, in a high-gain amplifier it is better to operate the first tube at maximum gain, since this gives the best signal-to-hum ratio. The control therefore is usually placed in the grid circuit of the second stage.

DESIGNING THE SPEECH AMPLIFIER

The steps in designing a speech amplifier are as follows:

1) Determine the power needed to modulate the transmitter and select the modulator. In the case of plate modulation, a Class B amplifier may be required. Select a suitable tube type and determine from the tube tables at the end of this book the grid driving power required, if any.

2) As a safety factor, multiply the required driver power by at least 1.5.

3) Select a tube, or pair of tubes, that will deliver the power determined in the second step. This is the last or output stage of the speech-amplifier. Receiver-type power tubes can be used (beam tubes such as the 6L6 may be needed in some cases) as determined from the receiving-tube tables. If the speech amplifier is to drive a Class B modulator, use a Class A or AB_1 amplifier, in preference to Class AB_2, if it will give enough power output.

4) If the speech-amplifier output stage must operate Class AB_2, use a medium-μ triode (such as the 6J5 or corresponding types) to drive it. In the extreme case of driving 6L6s to maximum output, two triodes should be used in push-pull in the driver. In either case transformer coupling will have to be used, and transformer manufacturers' catalogs should be consulted for a suitable type.

5) If the speech-amplifier output stage operates Class A or AB_1, it may be driven by a voltage amplifier. If the output stage is push-pull, the driver may be a single tube coupled through a transformer with a balanced secondary, or may be a dual-triode phase inverter. Determine the signal voltage required for full output from the last stage. If the last stage is a single-tube Class A amplifier, the peak signal is equal to the grid-bias voltage; if push-pull Class A, the peak signal voltage is equal to twice the grid bias; if Class AB_1, twice the bias voltage when fixed bias is used; if cathode bias is used, twice the bias figured from the cathode resistance and the no-signal plate current.

6) From Table 9-I, select a tube capable of giving the required output voltage and note its rated voltage gain. A double-triode phase inverter (Fig. 9-4A) will have approximately twice the output voltage and twice the gain of one triode operating as an ordinary amplifier. If the driver is to be transformer-coupled to the last stage, select a medium-μ triode and calculate the gain and output voltage as described earlier in this chapter.

7) Divide the voltage required to drive the output stage by the gain of the preceding stage. This gives the peak voltage required at the grid of the next-to-the-last stage.

8) Find the output voltage, under ordinary conditions, of the microphone to be used. This information should be obtained from the manufacturer's catalog. If not available, the figures given in the section on microphones in this chapter will serve.

9) Divide the voltage found in (7) by the output voltage of the microphone. The result is the over-all gain required from the microphone to the grid of the next-to-the-last stage. To be on the safe side, double or triple this figure.

10) From Table 9-I, select a combination of tubes whose gains, when multiplied together, give approximately the figure arrived at in (9). These amplifiers will be used in cascade. If high gain is required, a pentode may be used for the first speech-amplifier stage, but it is *not* advisable to use a second pentode because of the possibility of feed-back and self-oscillation. In most cases a triode will give enough gain, as a second stage, to make up the total gain required. If not, a medium-μ triode, may be used as a third stage.

A high-μ double triode with the sections in cascade makes a good low-level amplifier, and will give somewhat greater gain than a pentode — medium-μ triode combination. With resistance-coupled input to the first section the cathode of that section may be grounded, which is helpful in reducing hum.

SPEECH-AMPLIFIER CONSTRUCTION

Once a suitable circuit has been selected for a speech amplifier, the construction problem resolves itself into avoiding two difficulties — excessive hum, and unwanted feed-back. For reasonably humless operation, the hum voltage should not exceed about 1 per cent of the maximum audio output voltage — that is, the hum should be at least 40 db. below the output level.

Unwanted feed-back, if negative, will reduce the gain below the calculated value; if positive, is likely to cause self-oscillation or "howls." Feed-back can be minimized by isolating each stage with "decoupling" resistors and condensers, by avoiding layouts that bring the first and last stages near each other, and by shielding of "hot" points in the circuit, such as grid leads in low-level stages.

Speech-amplifier equipment, especially voltage amplifiers, should be constructed on steel chassis, with all wiring kept below the chassis to take advantage of the shielding afforded. Exposed leads, particularly to the grids of low-level high-gain tubes, are likely to pick up hum from the electric field that usually exists in the vicinity of house wiring. Even with the chassis, additional shielding of the input circuit of the first tube in a high-gain amplifier usually is necessary. In addition, such circuits should be separated as much as possible from power-supply transformers and chokes and also from any audio transformers that operate at fairly-high power levels; this will minimize magnetic coupling to the grid circuit and thus reduce hum or audio-frequency feed-back. It is always safe, although not absolutely necessary, to separate the speech amplifier and its power supply, building them on separate chassis.

If a low-level microphone such as the crystal type is used, the microphone, its connecting cable, and the plug or connector by which it is attached to the speech amplifier, all should be shielded. The microphone and cable usually are constructed with suitable shielding; this should be connected to the speech-amplifier chassis, and it is advisable — as well as usually necessary — to connect the chassis to a ground such as a water pipe. With the top-cap tubes, complete shielding of the grid lead and grid cap is a necessity.

Heater wiring should be kept as far as possible from grid leads, and either the center-tap or one side of the heater-transformer secondary winding should be connected to the chassis. If the center-tap is grounded, the heater leads to each tube should be twisted together to reduce the magnetic field from the heater current. With either type of connection, it is advisable to lay heater leads in the corner formed by a fold in the chassis, bringing them out from the corner to the tube socket by the shortest possible path.

When metal tubes are used, always ground the shell connection to the chassis. Glass tubes used in the low-level stages of high-gain amplifiers must be shielded; tube shields are obtainable for that purpose. It is a good plan to enclose the entire amplifier in a metal box, or at least provide it with a cane-metal cover, to avoid feed-back difficulties caused by the r.f. field of the transmitter. R.f. picked up on exposed wiring, leads or tube elements causes overloading, distortion, and self-oscillation of the amplifier.

When using paper condensers as by-passes, be sure that the terminal marked "outside foil" is connected to ground. This utilizes the outside foil of the condenser as a shield around the "hot" foil. When paper condensers are used for coupling between stages, always connect the outside-foil terminal to the side of the circuit having the lowest impedance to ground. Usually, this will be the plate side rather than the following-grid side.

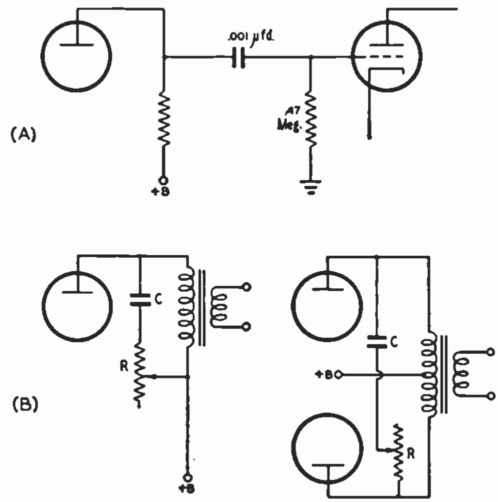


Fig. 9-5 — **A, use of a small coupling condenser to reduce low-frequency response; B, tone-control circuits for reducing high-frequency response. Values for *C* and *R* are discussed in the text; 0.01 μfd. and 25,000 ohms are typical.**

● INCREASING THE EFFECTIVENESS OF THE 'PHONE TRANSMITTER

The effectiveness of an amateur 'phone transmitter can be increased to a remarkable extent by taking advantage of speech characteristics. Measures that may be taken to make the modulation more effective include band compression (filtering), volume compression, and speech clipping.

Compressing the Frequency Band

Most of the intelligibility in speech is contained in the medium band of frequencies; that is, between about 500 and 2500 cycles. On the other hand, the major portion of speech power is normally concentrated below 500 cycles. It is these low frequencies that modulate the transmitter most heavily. If they are eliminated, the frequencies that carry most of the actual communication can be increased in amplitude without exceeding 100-per-cent modulation, and the effectiveness of the transmitter is correspondingly increased.

One simple way to reduce low-frequency response is to use small values of coupling capacitance between resistance-coupled stages, as shown in Fig. 9-5A. A time constant of 0.0005 second for the coupling condenser and following-stage grid resistor will have little effect on the amplification at 500 cycles, but will practically halve it at 100 cycles. In two cascaded stages the gain will be down about 5 db. at 200 cycles and 10 db. at 100 cycles. When the grid resistor is ½ megohm a coupling condenser of 0.001 μfd. will give the required time constant.

The high-frequency response can be reduced by using "tone control" methods, utilizing a condenser in series with a variable resistor connected across an audio impedance at some point in the speech amplifier. The best spot for the tone control is across the primary of the output transformer of the speech amplifier, as in Fig. 9-5B. The condenser should have a reactance at 1000 cycles about equal to the load resistance required by the amplifier tube or tubes, while the variable resistor in series may have a value equal to four or five times the load resistance. The control can be adjusted while listening to the amplifier, the object being to cut the high-frequency response as much as possible without unduly sacrificing intelligibility.

Restricting the frequency response not only puts more modulation power in the optimum frequency band but also reduces hum, because the low-frequency response is reduced, and helps reduce the width of the channel occupied by the transmission, because of the reduction in the amplitude of the high audio frequencies.

Volume Compression

Although it is obviously desirable to modulate the transmitter as completely as possible, it is difficult to maintain constant voice intensity when speaking into the microphone. To overcome this variable output level, it is possible to use automatic gain control that follows the *average* (not instantaneous) variations in speech amplitude. This can be done by rectifying and filtering some of the audio output and applying the rectified and filtered d.c. to a control electrode in an early stage in the amplifier

A practical circuit for this purpose is shown in Fig. 9-6. The rectifier must be connected, through the transformer, to a tube capable of delivering some power output (a small part of the output of the power stage may be used) or else a separate power amplifier for the rectifier circuit alone may have its grid connected in parallel with that of the last voltage amplifier.

Resistor R_4, in series with R_5 across the plate supply, provides an adjustable positive bias on the rectifier cathodes. This prevents the limiting action from beginning until a desired microphone input level is reached. R_2, R_3, C_2, C_3 and C_4 filter the audio frequencies from the rectified output. The output of the rectifier may be connected to the suppressor grid of a pentode first stage of the speech amplifier.

A transformer with a turns ratio such as to give about 50 volts when its primary is connected to the output circuit of the power stage should be used. If a transformer having a center-tapped secondary is not available, a half-wave rectifier may be used instead of the full-wave circuit shown, but it will be harder to get satisfactory filtering.

The over-all gain of the system must be high enough so that full output can be secured at a moderately low voice level.

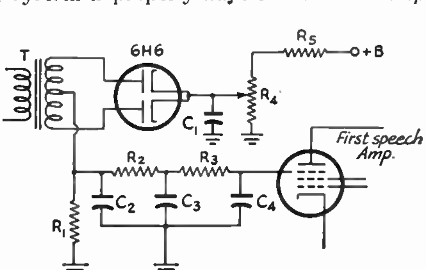


Fig. 9-6 — Speech-amplifier output-limiting circuit. C_1, C_2, C_3, C_4 — 0.1-μfd.; R_1, R_2, R_3 — 0.22 megohm; R_4 — 25,000-ohm pot.; R_5 — 0.1 megohm; T — see text.

Speech Clipping and Filtering

In speech waveforms the average power content is considerably less than in a sine wave of the same peak amplitude. Since modulation percentage is based on peak values, the modulation or sideband power in a transmitter modulated 100 per cent by an ordinary voice waveform will be considerably less than the sideband power in the same transmitter modulated 100 per cent by a sine wave. In Fig. 9-7 the upper drawing, A, represents a sine wave having a maximum amplitude that just modulates a given transmitter 100 per cent. The speech wave at B also represents 100-per-cent modulation.

If the amplitude of the wave shown at B is increased so that its power is comparable with or higher than the power in a sine wave, but with everything above 100-per-cent modulation cut off, it will appear as shown at C. This signal will not modulate the transmitter more than 100 per cent, but the voice power is several times greater than B. The wave is not exactly like the one at B, so the result will not sound exactly like the original. However, "clipping" of this type can be used to secure a worth-while increase in modulation power without sacrificing *intelligibility*. Once the system is properly adjusted *it will be impossible to overmodulate the transmitter* because the maximum output amplitude is held to the same value no matter what the amplitude of the signal applied.

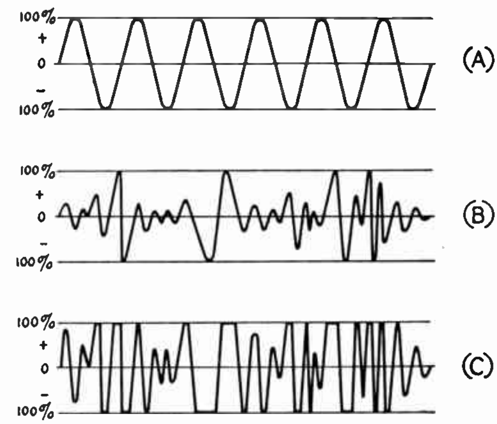


Fig. 9-7 — The normal speech wave (B) has high peaks but low average energy content. When the peaks are clipped the signal may be increased to a considerably higher power level without causing overmodulation (C).

By itself, clipping generates the same high-order harmonics that overmodulation does, and a signal modulated by the clipped waveform shown in Fig. 9-7 would "splatter". To prevent this, the audio frequencies above those needed for intelligible speech must be filtered out, *after* clipping and *before* modulation. The filter required for this purpose should have relatively little attenuation at frequencies below about 2500 cycles, but high attenuation for all frequencies above 3000 cycles.

It is possible to use as much as 25 db. of clipping before intelligibility suffers; that is, if the original peak amplitude is 10 volts, the signal can be clipped to such an extent that the resulting maximum amplitude is less than one volt. If the original 10-volt signal represented the amplitude that caused 100-per-cent modulation on peaks, the clipped and filtered signal can then be amplified up to the same 10-volt peak level for modulating the transmitter, with a very considerable increase in modulation power.

There is a loss in naturalness with "deep" clipping, even though the voice is highly intelligible. With moderate clipping levels (6 to 12 db.) there is almost no perceptible change in "quality" but the voice power is four to sixteen times as great as in ordinary modulation.

Before drastic clipping can be used, the speech signal must be amplified several times more than is necessary for normal modulation. Also, the hum and noise must be much lower than the tolerable level in ordinary amplification, because the noise in the output of the amplifier increases in proportion to the gain.

One type of clipper-filter system is shown in block form in Fig. 9-8. The clipper is a peak-limiting rectifier of the same general type that is used in receiver noise limiters. It must clip both positive and negative peaks. The gain or clipping

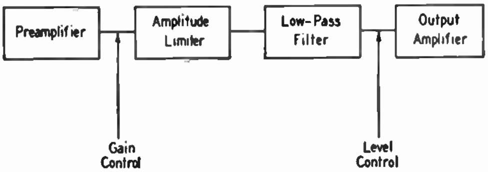


Fig. 9-8 — **Block diagram of speech-clipping and filtering amplifier.**

control sets the amplitude at which clipping starts. Following the low-pass filter for eliminating the harmonic distortion frequencies is a second gain control, the "level" or modulation control. This control is set initially so that the amplitude-limited output of the clipper-filter cannot modulate the transmitter more than 100 per cent.

It should be noted that the peak amplitude of the audio waveform actually applied to the modulated stage in the transmitter is not necessarily held at the same relative level as the peak amplitude of the signal coming out of the clipper stage. When the clipped signal goes through the filter, the relative phases of the various frequency components that pass through the filter are shifted, particularly those components near the cut-off frequency. This may cause the peak amplitude out of the filter to exceed the peak amplitude of the clipped signal applied to the filter input terminals. Similar phase shifts can occur in amplifiers following the filter, especially if these amplifiers, including the modulator, do not have good low-frequency response. With poor low-frequency response the more-or-less "square" waves resulting from clipping tend to be changed into triangular waves having higher peak amplitude. Best practice is to cut the low-frequency response *before* clipping and to make all amplifiers following the clipper-filter as flat and distortion-free as possible.

The best way to set the modulation control in such a system is to check the actual modulation percentage with an oscilloscope connected as described in the chapter on modulation. With the gain control set to give a desired clipping level with normal voice intensity at the microphone, the level control should be adjusted so that the maximum modulation does not exceed 100 per cent no matter how much sound is applied to the microphone.

Practical circuits for clipping and filtering are illustrated in a speech amplifier described in this chapter.

High-Level Clipping and Filtering

Clipping and filtering also can be done at high level — that is, at the point where the modulation is applied to the r.f. amplifier — instead of in the low-level stages of the speech amplifier. In one rather simple but effective arrangement of this type the clipping takes place in the Class-B modulator itself. This is accomplished by carefully adjusting the plate-to-plate load resistance for the modulator tubes so that they saturate or clip peaks at the amplitude level that represents 100 per cent modulation. The load adjustment can be made by choice of output transformer ratio or by adjusting the plate-voltage/plate-current ratio of the modulated r.f. amplifier. It is best done by examining the output waveform with an oscilloscope.

The filter for such a system consists of a choke and condensers as shown in Fig. 9-9. The values of L and C should be chosen to form a low-pass filter section having a cut-off frequency of about 2500 cycles, using the modulating impedance of the r.f. amplifier as the load resistance. For this cut-off frequency the formulas are

$$L_1 = \frac{R}{7850}$$

$$C_1 = C_2 = \frac{63.6}{R}$$

Where R is in ohms, L_1 in henrys, and C_1 and C_2 in microfarads. For example, with a plate modulated amplifier operating at 1500 volts and 200 ma. (modulating impedance 7500 ohms) L would be 7500/7850 = 0.96 henry and C_1 or C_2

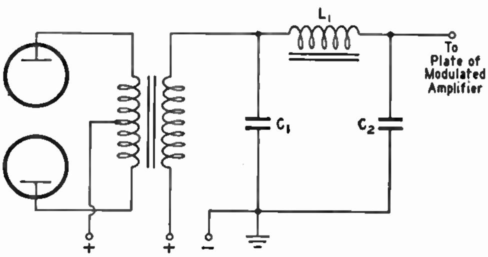


Fig. 9-9 — **Splatter-suppression filter for use at high level, shown here connected between a Class B modulator and plate-modulated r.f. amplifier. Values for L_1, C_1 and C_2 are determined as described in the text.**

would be 63.6/7500 = 0.0085 μfd. By-pass condensers in the plate circuit of the r.f. amplifier should be included in C_2. Voltage ratings for C_1 and C_2 when connected as shown must be the same as for the plate blocking condenser — i.e., at least twice the d.c. voltage applied to the plate of the modulated amplifier. L and C values can vary 10 per cent or so without seriously affecting the operation of the filter.

Besides simplicity, the high-level system has the advantage that high-frequency components of the audio signal fed to the modulator grids, whether present legitimately or as a result of amplitude distortion in lower-level stages, are suppressed along with the distortion components that arise in clipping. Also, the undesirable effects of poor low-frequency response following clipping and filtering, mentioned in the preceding section, are avoided. Phase shifts can still occur in the high-level filter, however, so adjustments preferably should be made by using an oscilloscope to check the actual modulation percentage under all conditions of speech intensity. (For further discussion see Bruene, "High-Level Clipping and Filtering", *QST*, November, 1951.)

A Clipper-Filter Speech Amplifier-Driver

The speech amplifier shown in Figs. 9-10 to 9-11, inclusive, uses push-pull triodes to obtain a power output of 13 watts with negligible distortion — sufficient to drive most of the commonly-used Class-B modulator tubes. It includes a clipper-filter for increasing the effectiveness of modulation and for confining the channel width to frequencies needed for intelligible speech. The over-all gain is ample for use with communications-type crystal microphones when using clipping of the order of 12–15 db. Miniature tubes are used in the voltage-amplifier stages. The output tubes are 6B4Gs, operated Class AB1 with fixed bias. Two power supplies are included, one for the voltage amplifier stages and the other for the output tube plates.

As shown in Fig. 9-11, the first two stages are voltage amplifiers of ordinary design, using a 6AU6 pentode in the first stage and a 6C4 triode in the second. The output of the second stage can be switched either to the 12AU7 double-triode clipper or to the 6C4 voltage amplifier that drives the 6B4G grids. In the latter case the amplifier operation is conventional. The clipper, when operative, provides additional voltage gain as well as clipping. Its output goes through a simple low-pass filter ($L_1C_{11}C_{12}$) so that harmonics generated by clipping will be attenuated before the signal reaches the grid of the second 6C4. The frequency response of the amplifier with the filter in circuit, but with the signal below the clipping level, drops at the rate of roughly 6 db. per octave below 500 cycles; above 4000 cycles the response is down 25 db. compared with the medium audio range.

A two-section filter is used in the plate supply for the voltage-amplifier stages. The hum level must be kept low because of the high gain required when using clipping. A single-section filter is sufficient for the output stage. Bias for the 6B4G grids is obtained from the low-voltage supply by means of R_{16}, by-passed by C_{14}.

Two gain controls are included, one (R_6) for setting the level into the clipper circuit and thus determining the amount of clipping, and the second (R_{13}) for setting the output level after clipping. With the clipper in use, proper setting of R_{13} will keep the modulation level high but will prevent overmodulation.

Construction

As shown in Fig. 9-10, the voltage amplifiers occupy the left front section of the chassis. The 6AU6 first amplifier is at the left, followed in order to the right by the first 6C4, the 12AU7, and the second 6C4. The 6B4Gs and their output transformer are at the right front. The cylindrical unit just behind the second 6C4 is the interstage audio transformer, T_1.

Power supply components are grouped along the rear edge of the chassis, with the low-voltage supply at the left. The power transformers should be kept well separated from the voltage amplifiers, particularly the first two stages, in order to minimize hum difficulties.

On the front panel, the microphone input connector is at the lower left. Next to it is the clipping control, then the clipper in-out switch, and then the modulation control. The two toggle switches at the right are S_2 and S_3. The a.c. input socket is by-passed by C_{15} and C_{16}, to reduce the possibility that r.f. picked up on the line cord will get into the low-level speech stages.

The wiring underneath the chassis is relatively simple, as shown by Fig. 9-12. The microphone input circuit, including RFC_1 and C_1, is enclosed in a National jack shield, and the lead from RFC_1 to the 6AU6 grid also is shielded.

Adjusting the Clipper-Filter Amplifier

The good effect of the low-pass filter in eliminating splatter can be entirely nullified if the amplifier stages following the filter can introduce appreciable distortion. Amplifier stages following the unit must be operated well within their capabilities; in particular, the Class B output transformer (if a Class B modulator is to be driven) should be shunted by condensers to reduce the high-frequency response as described in the section on Class B modulators.

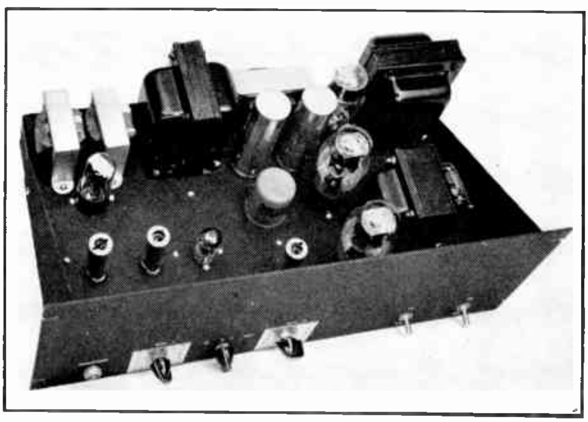


Fig. 9-10 — This speech-amplifier and driver has ample gain for a crystal microphone and is complete with power supply. The measured undistorted output is 13 watts. It incorporates a clipper-filter system for increasing modulation effectiveness and decreasing channel width.

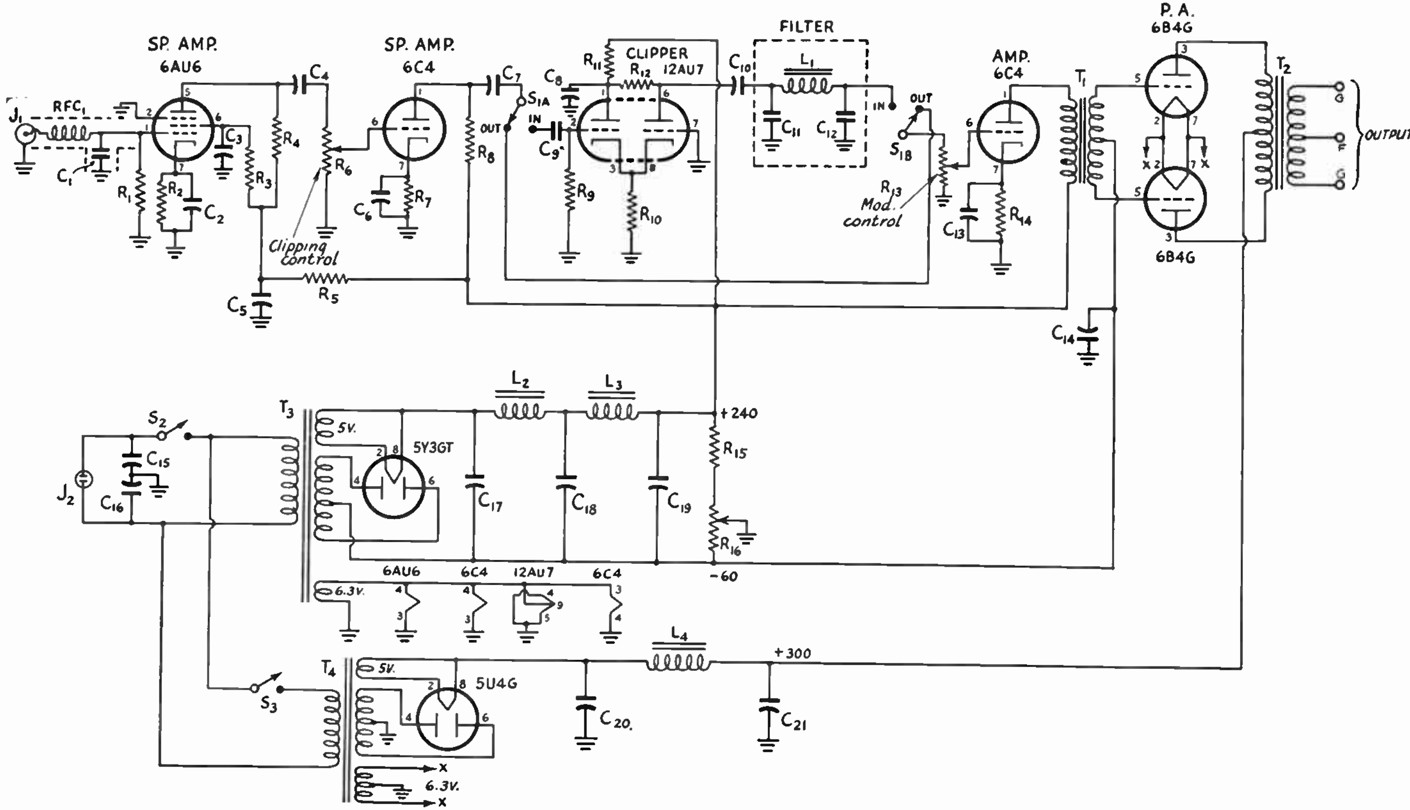


Fig. 9-11 — Circuit diagram of the clipper-filter speech amplifier.

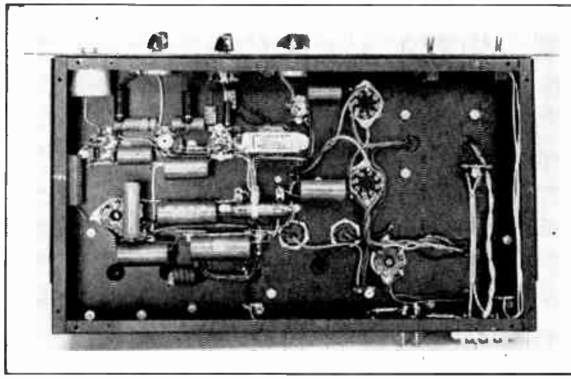


Fig. 9-12 — Below-chassis view of the clipper-filter speech amplifier. The relatively small number of components below the chassis makes wiring simple.

The setting of R_{13} is most important. It is most easily done with the aid of an oscilloscope (one having a linear sweep) and an audio oscillator, using the test set-up shown in the section on testing of speech equipment. Use a resistance load on the output transformer to reflect the proper load resistance (3000 ohms) at the plates of the 6B4Gs. First set R_{13} at about ¼ the resistance from the ground end, switch in the clipper-filter, and apply a 500-cycle sine-wave signal to the microphone input. Increase the signal amplitude until clipping starts, as shown by flattening of both the negative and positive peaks of the wave. To check whether the clipping is taking place in the clipper or in the following amplifiers, throw S_1 to the "normal" or "out" position; the waveshape should return to normal. If it does not, return S_1 to the "in" position and reduce the setting of R_{13} until it does. Then reduce the amplifier gain by means of R_6 until the signal is just below the clipping level. At this point the signal should be a sine wave. Increase R_{13}, without touching R_6, until the wave starts to become distorted, and then back off R_{13} until distortion disappears.

Next, change the input-signal frequency to 2500 cycles, without changing the signal level. Slowly increase R_6 while observing the pattern. At this frequency it should be almost impossible to get anything except a sine wave through the filter, so if distortion appears it is the result of overloading in the amplifiers *following* the filter. Reduce the setting of R_{13} until the distortion disappears, even when R_6 is set at maximum and the maximum available signal from the audio oscillator is applied to the amplifier. The position of R_{13} should be noted at this point and the observed setting should never be exceeded.

To find the *operating* setting of R_{13}, leave the audio-oscillator signal amplitude at the value just under the clipping level and set up the complete transmitter for a modulation check, using the oscilloscope to give the trapezoidal pattern. With the Class C amplifier and modulator running, find the setting of R_{13} (keeping the audio signal just under the clipping level) that just gives 100-per-cent modulation. This setting should be below the maximum setting of R_{13} as previously determined; if it is not, the driver and modulator are not capable of modulating the transmitter 100 per cent and must be redesigned — or the Class C amplifier input must be lowered. Assuming a satisfactory setting is found, connect a microphone to the amplifier and set the amplifier gain control, R_6, so that the transmitter is modulated 100 per cent. Observe the pattern closely at different settings of R_6 to see if it is possible to overmodulate. If overmodulation does not occur at any setting of R_6, the transmitter is ready for operation and R_{13} may be locked in position; it need never be touched subsequently. If some overmodulation does occur, R_{13} should be backed off until it disappears and then locked.

In the absence of an oscilloscope the other methods of checking distortion described in the section on speech-amplifier testing may be used. The object is to prevent distortion in stages following the filter, so that when the clipping level is exceeded the following stages will be working within their capabilities.

C_1 — 100-μμfd. mica.
C_2, C_6, C_{13} — 20-μfd. 25-volt electrolytic.
C_3 — 0.1-μfd. 400-volt paper.
C_4, C_7, C_{15}, C_{16} — 0.01-μfd. 400-volt paper.
C_5, C_8 — 8-μfd. 450-volt electrolytic.
C_9, C_{11} — 470-μμfd. mica.
C_{10} — 0.002-μfd. mica or paper.
C_{12} — 330-μμfd. mica.
C_{14} — 30-μfd. 150-volt electrolytic.
C_{17}, C_{18}, C_{19} — 16-μfd. 450-volt electrolytic.
C_{20}, C_{21} — 8-μfd. 450-volt electrolytic (can type).
R_1 — 2.2 megohms, ½ watt.
R_2, R_{14} — 2200 ohms, ½ watt.
R_3 — 1 megohm, ½ watt.
R_4, R_9 — 0.47 megohm, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_6 — 2-megohm volume control.
R_7 — 3900 ohms, ½ watt.
R_8 — 0.1 megohm, ½ watt.
R_{10} — 1500 ohms, 1 watt.
R_{11} — 47,000 ohms, 1 watt.
R_{12} — 56,000 ohms, ½ watt.
R_{13} — 0.5-megohm volume control.
R_{15} — 10,000 ohms, 20 watts.
R_{16} — 2000-ohm 25-watt adjustable.
L_1 — 20 henrys, 900 ohms (Stancor C-1515).
L_2, L_3 — 15 henrys, 75 ma. (Stancor C-1002).
L_4 — 10.5 henrys, 110 ma. (Stancor C-1001).
J_1 — Microphone cable receptacle (Amphenol PC1M).
J_2 — Chassis-mounting 115-volt plug.
S_1 — D.p.d.t. rotary switch (Mallory 3122-J).
S_2, S_3 — S.p.s.t. toggle.
T_1 — Audio transformer, single plate to p.p. grids, ratio 2:1 (Thordarson T20A17).
T_2 — Driver transformer, variable ratio, p.p. driver to Class-B grids, pri. rating 120 ma. per side (Stancor A-4763).
T_3 — Power transformer: 700 v. c. t., 90 ma.; 5 v., 2 amp.; 6.3 v. 3.5 amp. (Stancor P-4079).
T_4 — Power transformer: 700 v. c. t., 110 ma.; 5 v., 3 amp.; 6.3 v. 4.5 amp. (Stancor P-4080).
RFC_1 — 2.5 mh. r.f. choke.

A Simple Grid Modulator

The modulator circuit shown in Fig. 9-14 is capable of modulating any transmitter, up to the maximum power limit, to about 80 per cent with low distortion. It requires no power supply other than heater power for the tubes, since it gets plate power from the cathode circuit of the r.f. amplifier with which it is used. Although the modulator output is connected in series with the r.f. amplifier cathode, the modulation is essentially of the grid-bias type (see chapter on amplitude modulation). A useful characteristic of the system is that it does not require a fixed source of grid bias for the amplifier.

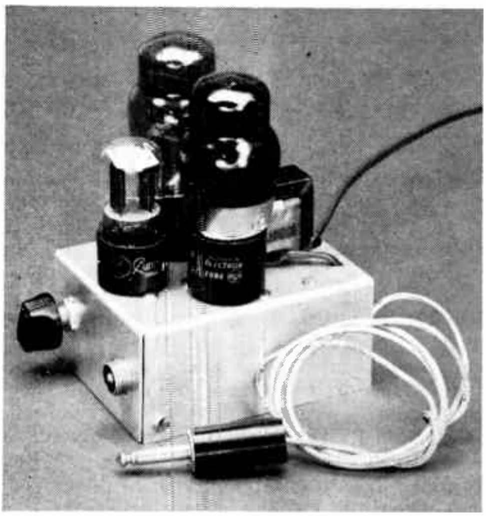


Fig. 9-13 — A simple modulator of the grid-bias type, usable with transmitters having c.w. plate inputs up to a kilowatt. Plate power for the unit is obtained automatically from the r.f. amplifier supply.

The speech amplifier uses a high-μ double triode to give two stages of resistance-coupled amplification. This gives sufficient gain for a crystal microphone. Resistors R_3, R_7 and R_{10}, together with C_1 and C_3, provide decoupling and additional fitering of the d.c. obtained from the r.f. amplifier cathode circuit.

The output stage uses one or more 6Y6Gs in parallel; in determining the number of tubes required to modulate a particular amplifier, use one 6Y6G for each 200 ma. of amplifier plate current based on the operating conditions for c.w. work. The audio output voltage is developed across L_1 and R_{11} in series; R_{11} may be omitted if the d.c. voltage between the screen and cathode of the 6Y6G does not exceed the rated value of 135 volts.

No special constructional precautions need be observed in laying out the amplifier. The unit shown in Fig. 9-13 is built on a homemade chassis folded from a sheet of aluminum, but a small standard chassis may be used instead. A filament transformer may be included in the unit in case the heater power cannot conveniently be obtained from the transmitter itself.

To use the modulator, first tune up the transmitter for ordinary c.w. operation with the modulator disconnected. Then connect the modulator output terminals in series with the amplifier cathode as indicated in Fig. 9-14. (Make certain that the modulator cathodes are up to operating temperature before applying plate voltage to the r.f. amplifier.) The amplifier plate current should drop to approximately one-half the c.w. value. If the plate current is too high, increase the value of R_9 until it is in the proper region; if too low, decrease the resistance at R_9. Once this adjustment is made the system is ready for 'phone operation. The r.f. amplifier plate current should show no change with speech input, except for a slight upward kick on voice peaks.

The carrier power output with this system is somewhat less than would be obtained with conventional grid modulation because the d.c. voltage drop in the 6Y6G modulators subtracts from the amplifier plate voltage. The difference is small with r.f. tubes operating at 1000 volts or more.

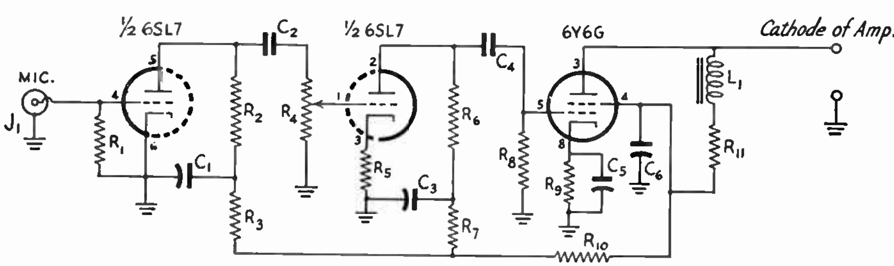


Fig. 9-14 — Circuit diagram of the speech amplifier and modulator.

C_1, C_3, C_6 — 8-μfd. electrolytic, 450 volts.
C_2 — 0.005 μfd. 400 volts.
C_4 — 0.01 μfd. 400 volts.
C_5 — 50-μfd. electrolytic, 50 volts.
R_1 — 2.2 megohms, ½ watt.
R_2 — 0.22 megohm, ½ watt.
R_3, R_7, R_{10} — 22,000 ohms, ½ watt.
R_4 — 0.5-megohm volume control.
R_5 — 2200 ohms, ½ watt.
R_6, R_8 — 0.1 megohm, ½ watt.
R_9 — 50 ohms, 2 watts (see text).
R_{11} — 2000 ohms, 2 watts (see text).
L_1 — Small filter choke, "a.c.-d.c." type satisfactory.

Screen Modulator Circuit

Fig. 9-15 is a representative circuit for a modulator for the screen grid of a beam tetrode. Most r.f. tubes of this type require very little modulating power in the screen circuit, so a receiving-type audio power amplifier usually is sufficient. The circuit shown has ample gain for a crystal microphone and will fully modulate a screen grid that does not require an average audio power of more than three or four watts. It can also be used for modulating a pair of r.f. tubes where these requirements are not exceeded. The chapter on amplitude modulation should be consulted for information on determining the voltage swing and modulating power for a particular tube type. The turns ratio required in T_1, primary to secondary, will range from 1 to 1 to 0.8 to 1 for various r.f. tubes, since the peak output voltage of the tube across the primary of the transformer is about 200 volts. An inexpensive driver transformer, of the type used for coupling a triode or pentode to Class AB_2 tetrodes of the 6L6 class, will be satisfactory. It should preferably have two or three primary taps so the turns ratio can be adjusted. Transformer coupling is used in preference to direct coupling (i.e., "clamp-tube" modulation of the screen) because of simpler adjustment, ease of modulating 100 per cent, and because it permits using a low-voltage supply for the screen grid of the modulated r.f. amplifier.

The speech input stage uses a 6SJ7 pentode and is followed by a 6J5 voltage amplifier. The 6V6 output stage uses negative feed-back, the feed-back voltage being taken from the plate circuit by means of the voltage divider $R_{10}R_{11}$ and applied in series with the plate resistor, R_7, of the preceding stage. Negative feed-back in the modulator is very desirable when a screen or control grid is to be modulated because the load on the modulator varies over the audio-frequency cycle, and feed-back reduces the distortion that arises from this cause. In this circuit the percent feed-back is chosen to be as large as possible while still retaining enough voltage gain for normal voice intensity into a crystal microphone.

The lead between the microphone connector and the 6SJ7 grid should be shielded, as should also the first-stage grid-resistor, R_1. Such shielding prevents hum pick-up on the grid lead. Aside from this, no special precautions need be observed in constructing the amplifier, beyond keeping the heater leads well away from the plate and grid leads of the tubes.

The heater requirement for the unit is 1 ampere at 6.3 volts. Plate-supply requirements vary from about 70 to 85 ma. at 250 to 300 volts, depending on the screen current taken by the tube being modulated. R_{13} should be adjusted, by means of the slider, to give the proper d.c. voltage at the screen of the modulated stage. This voltage will, in general, be approximately half the d.c. screen voltage recommended for c.w. operation, as described in the chapter on amplitude modulation. The method of adjustment for linear modulation is also covered in that chapter.

The same circuit may be used for control-grid modulation of either triode or tetrode r.f. amplifiers. The method of adjustment is described in the chapter on amplitude modulation.

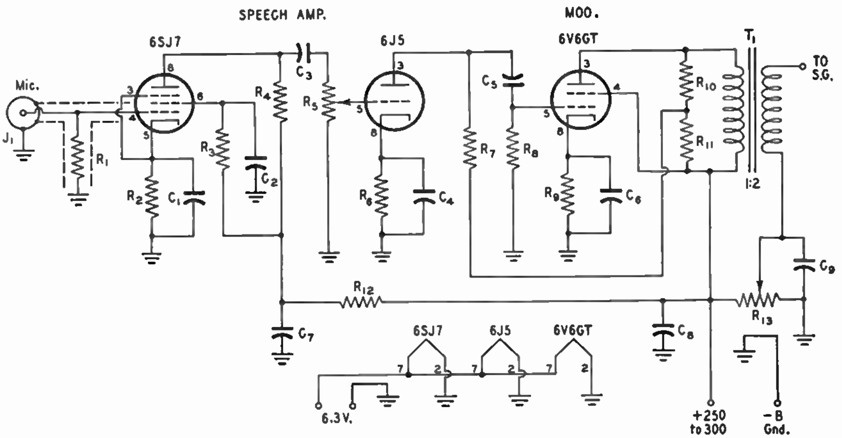


Fig. 9-15 — Modulator circuit for screen or control grid modulation.

C_1, C_4 — 10-μfd. 25-volt electrolytic.
C_2 — 0.1-μfd. 400-volt paper.
C_3, C_5 — 0.01-μfd. 400-volt paper.
C_6 — 50-μfd. 50-volt electrolytic.
C_7, C_8, C_9 — 10-μfd. 450-volt electrolytic.
R_1 — 2.2 megohms, ½ watt.
R_2, R_6 — 1500 ohms, ½ watt.
R_3 — 1 megohm, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5 — 1-megohm potentiometer, audio taper.
R_7, R_8 — 0.1 megohm, ½ watt.
R_9 — 235 ohms, 2 watts. (Two 470-ohm 1-watt units in parallel.)
R_{10}, R_{12} — 47,000 ohms, 1 watt.
R_{11} — 27,000 ohms, 1 watt.
R_{13} — 25,000-ohm adjustable, 25 watts.
J_1 — Microphone jack.
S_1 — 4-pole 2-position rotary switch (see text).
T_1 — Audio driver transformer (see text).

40-Watt Class AB_1 Modulator

The modulator unit shown in Figs. 9-16 to 9-18, inclusive, has an undistorted power output of somewhat better than 40 watts. Although designed as a companion unit for the 75-watt transmitter described in the chapter on transmitters, it may be used with any transmitter operating at a d.c. plate power input of 80 watts or less.

Speech Circuit

The speech amplifier uses a high-μ dual triode as a two-stage resistance-coupled amplifier, followed by a medium-μ triode. The latter is transformer-coupled to the modulator grids. The gain from the microphone input to the 807 grids is more than ample for crystal and other microphones of similar output level. Battery bias is used for the modulator grids since it is the simplest method and a small battery such as those made for hearing-aids can be used. Since no current is taken from the battery, its life is the same as the normal shelf life.

The frequency response of the amplifier is adjusted to put maximum energy in the range where it contributes most to speech intelligibility; that is, the output is highest between 500 and 1200 cycles and drops off gradually on either side. The lower frequencies are reduced by low values of capacitance at C_3 and C_4, and the high-frequency end is attenuated by C_6 and C_7.

Power Supply

The power supply uses a replacement-type transformer with a bridge rectifier to obtain dual output voltages. A single-section filter is used on the high-voltage output and a double-section filter on the low voltage. With the values shown in Fig. 9-17 the hum level is 40 db. below full output of the modulator.

Control Circuits

The switching arrangement in the control section provides an on-off control for 'phone operation, disables the modulator when c.w. is to be used, and includes a "Test" position similar to that in the 75-watt transmitter. In the "'Phone" position the control switch disconnects the 6146 r.f. amplifier screen from the supply in the transmitter unit and connects it to the screen-dropping resistor, R_{18}. Simultaneously, the secondary of the modulation transformer is connected in series with the 6146 d.c. plate lead, and the cathodes of the 807 modulators are connected to ground through R_{16}. The drop in voltage across R_{16} is used for measuring the modulator plate current. In the "C.W." position the modulator cathode circuit is opened, the secondary of the modulation transformer is short-circuited, and the 6146 screen is connected to the screen supply in the transmitter unit. In both the "Test" and "Off" positions the 6146 screen is disconnected from its supply and grounded, and the modulation transformer secondary is shorted. If a key is plugged into the jack on the transmitter, either position can be used for testing; but the key must be left open when using the "Off" position for on-off 'phone switching.

A few changes in the original transmitter circuit are required. Referring to the circuit diagram in the transmitter chapter, these are:

1) Disconnect the lead between the arm of S_{4A} and J_7. This section of S_4 is no longer needed.

2) Remove R_9 from the circuit. (This resistor is replaced by R_{17} in Fig. 9-17.)

3) Connect Pin 2 on the auxiliary socket, J_8, to the top contact of J_7. This connects the 5763 and 6146 cathodes to the auxiliary socket and thence through the connecting cable to the modulator.

A six-wire cable is used for making connections between the two units. The same pin numbers are used for corresponding circuits at each end, so it is merely necessary to connect Pin 1 in one plug to Pin 1 in the other, and so on. Pins 5 and 6 must be connected by a jumper in order to complete the heater circuit in the transmitter.

The meter in the transmitter is used for making measurements on the modulator by means of a cord with pin jacks running between the "External Voltmeter" jack on the transmitter and the jacks (J_2 to J_5, inclusive) shown in Fig. 9-17.

Fig. 9-16 — The 40-watt modulator alongside the 75-watt transmitter described in the transmitting chapter. It is completely self-contained, with power supply and control circuits, on a 5 × 10 × 3-inch chassis.

The control switch, *center*, has four positions — off, test, 'phone, and c.w. Microphone connector and gain control are at the left; a.c. switches at the lower right. The two speech amplifier tubes are at the left front, followed by the 807s and the 5V4G to the rear. The modulation transformer is at the right front and the power transformer is at the right rear.

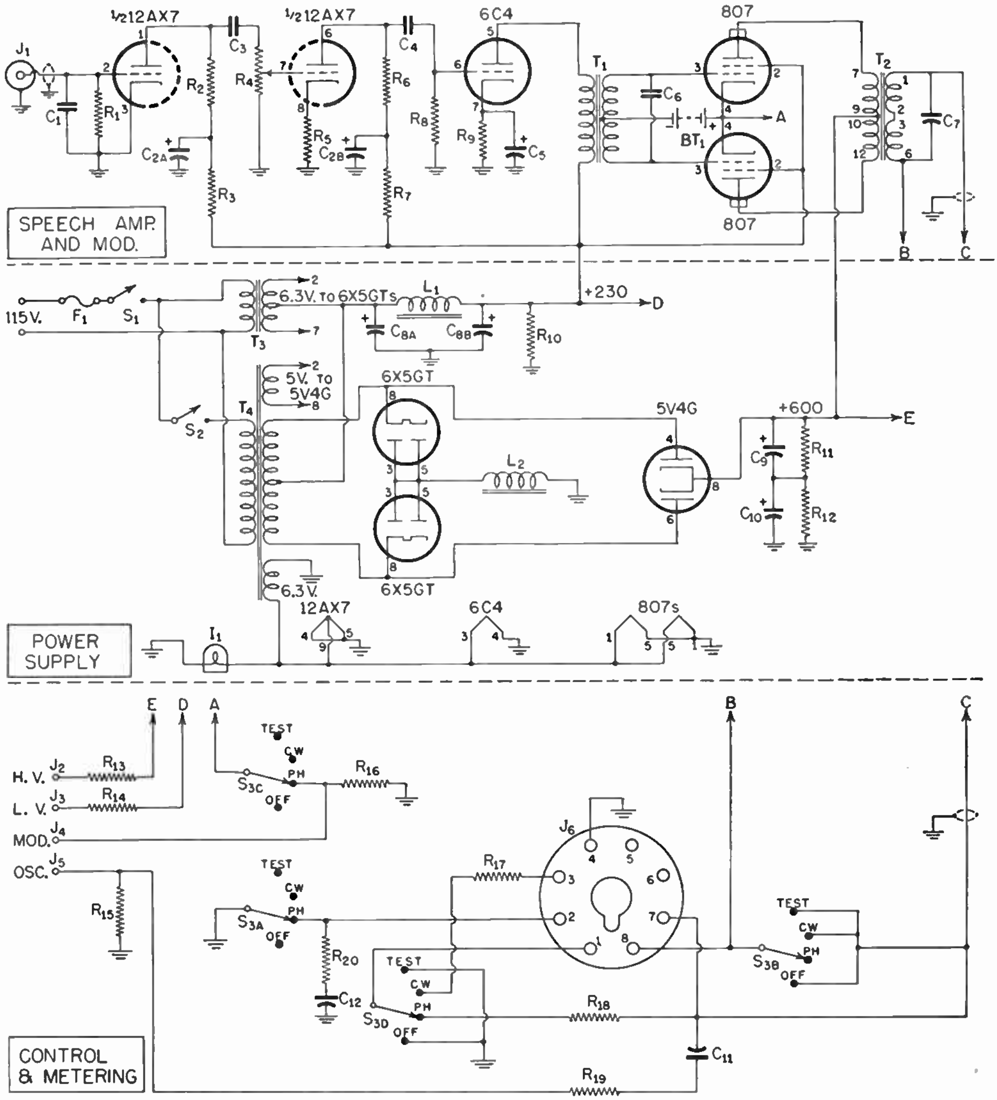


Fig. 9-17 — Circuit diagram of the Class AB_1 modulator using 807s.

C_1 — 100-μμfd. ceramic.
C_2 — Dual 8-μfd. electrolytic, 450 volts.
C_3, C_4 — 0.0015-μfd. ceramic.
C_5 — 10-μfd. electrolytic, 25 volts.
C_6 — 470-μμfd. ceramic.
C_7 — 0.002- to 0.004-μfd. paper, 600 volts.
C_8 — Dual electrolytic, 8 (A) and 16 (B) μfd., 450 volts.
C_9, C_{10} — 30-μfd. electrolytic, 450 volts.
C_{11} — 0.004-μfd. paper, 1600 volts.
C_{12} — 0.1-μfd. paper, 600 volts.
R_1 — 2.2 megohms, ½ watt.
R_2, R_6 — 0.1 megohm, ½ watt.
R_3 — 47,000 ohms, ½ watt.
R_4 — 1-megohm volume control, preferably log taper.
R_5 — 1500 ohms, ½ watt.
R_7 — 10,000 ohms, ½ watt.
R_8 — 1 megohm, ½ watt.
R_9 — 1000 ohms, ½ watt.
R_{10} — 0.1 megohm, 1 watt.
R_{11}, R_{12} — 20,000 ohms, 10 watts.
R_{13}, R_{19} — 1 megohm, 1 watt.
R_{14} — 0.47 megohm, ½ watt.
R_{15} — 15,000 ohms, ½ watt.
R_{16} — 50 ohms, ½ watt.
R_{17} — 4700 ohms, 1 watt.
R_{18} — 35,000 ohms, 10 watts.
R_{20} — 1000 ohms, ½ watt (value not critical).
F_1 — 2-amp. fuse.
I_1 — Pilot light, 6.3 v., 150 ma.
J_1 — Panel-type microphone connector (Amphenol PC1M).
J_2, J_3, J_4, J_5 — Insulated tip jack.
J_6 — Octal socket.
L_1 — 4.5 henrys, 50 ma., 300 ohms.
L_2 — 8 henrys, 100 ma., 375 ohms.
S_1, S_2 — S.p.s.t. toggle.
S_3 — 4-section, 5-position ceramic wafer switch (2 wafers), 4 positions used.
T_1 — Interstage audio transformer, center-tapped secondary, 10-ma. primary, total secondary-to-primary turns ratio 3 to 1.
T_2 — Adjustable-ratio modulation transformer, app. 30 watts (UTC CVM-1).
T_3 — Filament transformer, 6.3 v. at 1.2 amp.
T_4 — Power transformer, 350 v. each side c.t., 90 ma.; 5 v. at 2 amp.; 6.3 v. at 3 amp.
BT_1 — 22.5-volt battery (hearing-aid type used in modulator shown in photographs).

Modulator plate current is read through J_4; the full scale range is 100 ma. with a 50-ohm resistor at R_{16}. A.f. voltage for an oscilloscope can be taken from J_5, through the voltage divider formed by R_{15} and R_{19}. C_{11} is a blocking condenser for the voltage divider. R_{15} and the total resistance are such as to give about 10 volts peak.

C_{12} and R_{20} are used to suppress sparking at the control switch.

When the transmitter and modulator are connected by the cable all the control functions, except keying, are performed by the switch on the modulator unit. The "Test-Operate" switch on the transmitter should be left permanently in the "Operate" position. The key may be left permanently in the jack on the transmitter. Both power supplies run continuously. The 6X5GT heater transformer in each unit should be turned on sufficiently ahead of the power transformer to allow the 6X5GT heaters to attain temperature before the 5V4GT.

Construction

Although the unit shown is complete on a 5 × 10 × 3-inch chassis, such compact construction is not ordinarily necessary. A larger chassis will provide more freedom for placement of components and will make wiring easier.

In choosing a layout, it is advisable to keep the output transformer, T_2, well separated from the low-level speech amplifier circuits. This will tend to reduce stray coupling and feed-back and thus prevent any tendency toward self-oscillation. To prevent such oscillation in the layout shown, it was necessary to install a small aluminum shield between the speech amplifier circuits and the output transformer terminals (Fig. 9-18), and to use a shielded lead from the "hot" terminal (C) of the transformer to the terminal socket at the rear of the chassis.

Operating Values

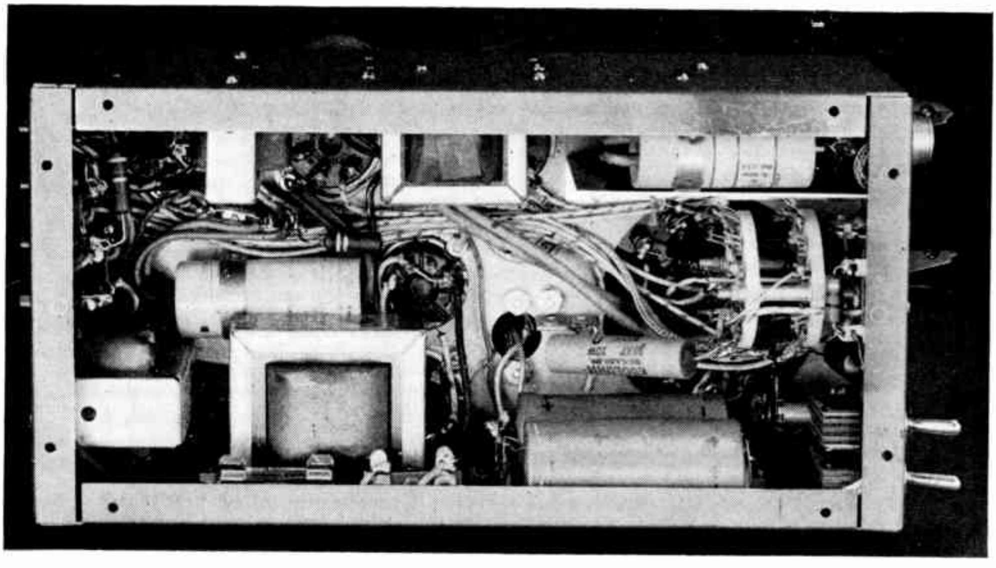
The optimum plate-to-plate load resistance for 807s operating Class AB_1 with the voltages indicated is approximately 13,000 ohms. For modulating the 75-watt transmitter at full rated 'phone input of 67.5 watts for the 6146 the proper transformer taps are indicated in Fig. 9-17. The antenna loading should be adjusted to make the 6146 plate current 106 ma., at which load the screen current should be 12 ma. and the plate voltage 640. The 6146 grid current should be adjusted to be 3 to 4 ma.

For other tubes or different voltages and currents, or for a different type of modulation transformer, the load resistance should be calculated as described in the chapter on amplitude modulation and the transformer taps chosen accordingly.

The d.c. power supply voltages in the modulator unit (line voltage 120) should measure 690 and 260 for the high and low supplies with no audio input. The modulator idling current is about 50 ma. under these conditions with a new 22.5-volt (actual voltage 24.5 volts) battery for bias. With tone input and the gain adjusted for maximum undistorted output, the modulator cathode current is about 100 ma. However, with speech the modulator plate current should not kick beyond 60 to 65 ma. on voice peaks; this represents 100 per cent modulation.

Fig. 9-18 — The principal components along the lower wall of the chassis, from left to right, are the filament transformer for the 6X5GT's, T_3, common power-supply choke, L_2 (with the low-voltage supply dual filter condenser C_8, just above it), filter condensers C_9 and C_{10}, and the a.c. switches. R_{19} and C_{11}, for the oscilloscope audio take-off are mounted on a tie-point strip on the front wall (at the right in this view) near the control switch, S_3. The screen-dropping resistor, R_{18}, for the r.f. amplifier is mounted between the S_3 and a tie point on the chassis.

The components visible along the upper wall are the low-voltage filter choke, L_1, at the left, transformer T_1 at the center, and C_8 at the right. The bias battery is mounted by a bracket on the chassis wall in the space between T_1 and the chassis deck.



Push-Pull 807 Modulator and Speech Amplifier

The speech amplifier and modulator shown in Fig. 9-19 is capable of modulating a power input to the modulated amplifier of approximately 200 watts when the maximum rated voltage of 750 is applied to the 807 plates. The maximum undistorted audio power output is 100 watts at that plate voltage, after allowing for losses in the output transformer. The 807s are operated as Class AB_2 amplifiers.

As shown in Fig. 9-20, the first speech amplifier tube is a 6SJ7, with its input circuit arranged for use with a crystal microphone. The second stage, also a resistance-coupled voltage amplifier, uses a 6J5. The third stage, which must deliver power to the grids of the Class AB_2 modulator tubes, uses a 6K6 pentode. Negative feed-back is incorporated in this stage as a means for improving its output voltage regulation and reducing distortion. The 6K6 is coupled to the modulator grids through a transformer.

In the modulator stage small chokes, RFC_1 and RFC_2, are connected in the grid leads and 100-ohm resistors are connected in the screen leads to prevent the parasitic oscillations that frequently occur with 807s. Each screen resistor is separately by-passed to ground with a mica condenser for the same reason.

A filament transformer capable of handling all tube heaters is included as part of the unit.

Circuit constants have been selected so that the overall frequency response is sufficiently flat in the normal range of voice frequencies, but drops off above 3000 cycles and below 150 cycles.

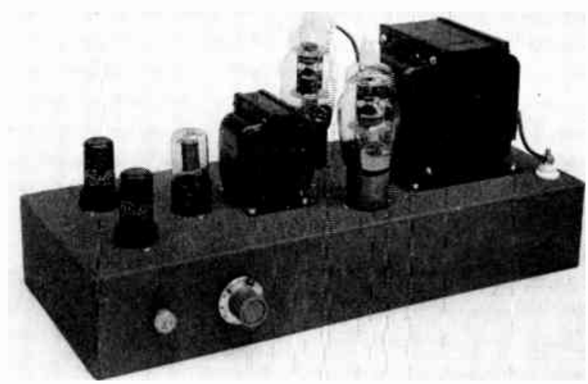


Fig. 9-19 — Modulator unit using push-pull 807s with speech amplifier designed for crystal-microphone input. It is built on a 7 by 17 by 3 steel chassis and can be mounted on a standard 8¾ inch relay-rack panel. The audio power output obtainable varies from 50 to 100 watts depending on the plate voltage supplied to the 807s.

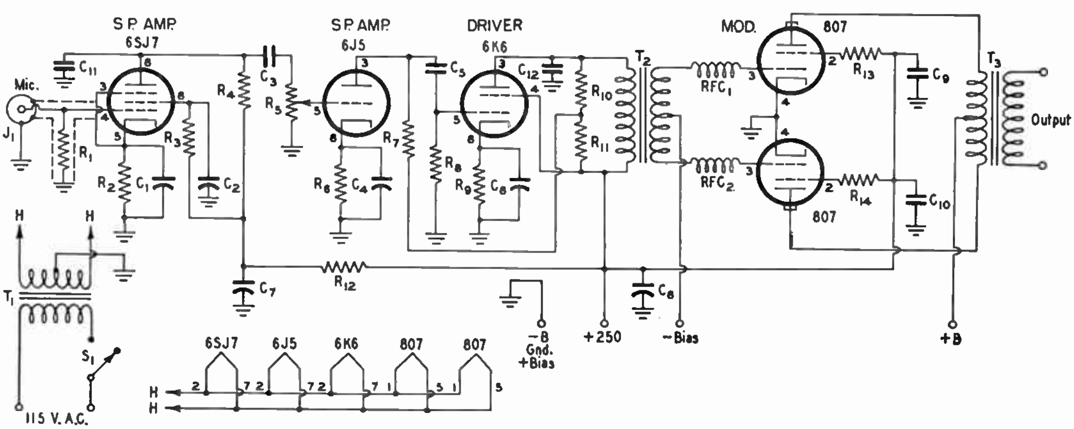


Fig. 9-20 — Circuit diagram of the push-pull 807 modulator

C_1, C_4 — 10-μfd. 25-volt electrolytic.
C_2 — 0.1-μfd. 400-volt paper.
C_3, C_5 — 0.0015-μfd. mica.
C_6 — 50-μfd. 50-volt electrolytic.
C_7, C_8 — 10-μfd. 450-volt electrolytic.
C_9, C_{10}, C_{12} — 0.002-μfd. mica.
C_{11} — 680-μμfd. mica.
R_1 — 2.2 megohms, ½ watt.
R_2, R_6 — 1500 ohms, ½ watt.
R_3 — 1 megohm, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5 — 1-megohm potentiometer, audio taper.
R_7, R_8 — 0.1 megohm, ½ watt.
R_9 — 680 ohms, 1 watt.
R_{10} — 0.1 megohm, 1 watt.
R_{11} — 27,000 ohms, 1 watt.
R_{12} — 47,000 ohms, 1 watt.
R_{13}, R_{14} — 100 ohms, ½ watt.
RFC_1, RFC_2 — 0.7 microhenry (Ohmite Z-50).
J_1 — Microphone jack.
S_1 — S.p.s.t. switch (part of gain-control assembly).
T_1 — 6.3 volts a.c., 3 amp.
T_2 — Class AB_2 driver transformer, single plate to p.p. grids, turns ratio 2 to 1, pri. to ½ sec.
T_3 — Output transformer (see text).

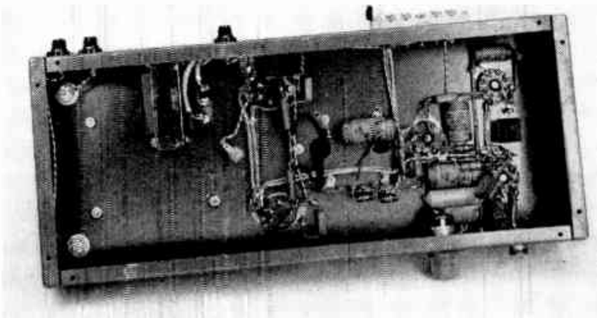


Fig. 9-21 — Bottom view of the push-pull 807 modulator. In this view the microphone connector is at the lower right, with the gain control just to its left. The filament transformer is in the upper left corner. Ceramic feed-through insulators are used to carry the output transformer connections through the chassis, and safety terminals are used for the high-voltage d.c. lead and the output transformer secondary terminals.

The general layout of the unit is shown in Figs. 9-19 and 9-21. The metal tube nearest the front of the chassis is the 6SJ7 and the 6J5 is toward the rear. The layout is not critical, except that it is advisable to keep the filament transformer well separated from the low-level stages and the input transformer, T_2.

To prevent hum pick-up the lead from the microphone connector to the grid of the 6SJ7 should be shielded, as should also the grid resistor, R_1. A satisfactory shield for the grid resistor may be made by slipping a short piece of spaghetti tubing over the resistor and then covering the tubing with shield braid. The braid should be grounded to the chassis. The leads to the gain control, R_5, should be made from shielded wire.

The type of output transformer to use will depend on the modulating impedance of the Class C r.f. stage. At maximum ratings the 807s require a plate-to-plate load of 6950 ohms, so the output transformer turns ratio must be selected accordingly.

In case the input to the modulated stage is less than 200 watts, the 807s may be operated at a reduced plate voltage to obtain the necessary audio power output. Typical operating conditions at various plate voltages are given below:

Plate voltage	400	500	600	750	volts
Screen voltage	300	300	300	300	volts
Grid bias	−25	−29	−30	−32	volts
Plate current, max. sig.	240	240	200	240	ma.
Plate current, no sig.	90	72	60	52	ma.
Load resistance	3200	4240	6400	6950	ohms
Power output	55	75	80	120	watts

The output figures given above are tube output only, and do not include transformer losses. They should be reduced by about 15 per cent to obtain the actual power available for modulating the transmitter. For example, with a plate-supply voltage of 500 the actual output can be expected to be about 65 watts, sufficient for modulating 130 watts input.

The table above gives the power supply requirements for the 807s at various plate voltages. The fixed bias may be supplied by batteries or a bias supply such as is described in the chapter on power supplies. The screen voltage may be between 250 and 300 in the practical case; at 250 volts somewhat less bias is needed and the driving power required is slightly increased but the power output is approximately the same.

The first three stages of the unit may be operated from a small power supply giving approximately 70 ma. at 250 to 300 volts. A suitable circuit diagram is given in Fig. 9-22. This circuit also supplies the fixed bias for the 807 grids, by utilizing the voltage drop between the negative side of the high-voltage output and ground through the tap on resistor R_2. The slider on R_2 should be adjusted so that the proper bias voltage, as given by the table on this page, is obtained. It is advisable to check the 807 screen current, with no plate voltage on the 807s, to be sure that the rated screen dissipation of 3.5 watts per tube is not exceeded. If it is, the bias should be increased to keep the dissipation within rating. This will prevent damage to the screens during stand-by periods.

Such a power supply can be incorporated in the modulator unit, if desired. The principal precaution to be observed is that the power transformer should not be mounted near the low-level stages. A slightly deeper chassis may be required.

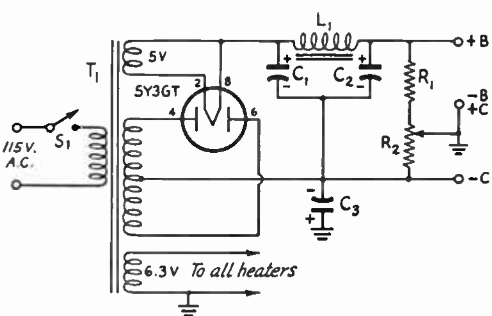


Fig. 9-22 — Power supply for speech-amplifier stages of 807 modulator. The unit also supplies fixed bias for the 807 grids.

C_1, C_2 — 8-μfd. electrolytic, 450 volts.
C_3 — 50-μfd. electrolytic, 50 volts.
L_1 — Filter choke, 30 henrys, 75 ma.
R_1 — 15,000 ohms, 10 watts.
R_2 — 1000-ohm adjustable, 10 watts.
S_1 — S.p.s.t. toggle.
T_1 — Power transformer, 350 volts each side c.t., 70 ma.; 5 v. 3 amp.; 6.3 v. 3 amp.

Modulators and Drivers

CLASS-AB AND -B MODULATORS

Class AB or B modulator circuits are basically identical no matter what the power output of the modulator. The diagrams of Fig. 9-23 therefore will serve for any modulator of this type that the amateur may elect to build. The triode circuit is given at A and the circuit for tetrodes at B. When small tubes with indirectly-heated cathodes are used, the cathodes should be connected to ground.

Modulator Tubes

The audio ratings of various types of transmitting tubes are given in the chapter containing the tube tables. Choose a pair of tubes that is capable of delivering sine-wave audio power equal to somewhat more than half the d.c. input to the modulated Class C amplifier. It is sometimes convenient to use tubes that will operate at the same plate voltage as that applied to the Class C stage, because one power supply of adequate current capacity may then suffice for both stages.

In estimating the output of the modulator, remember that the figures given in the tables are for the *tube* output only, and do not include output-transformer losses. To be adequate for modulating the transmitter, the modulator should have a theoretical power capability about 25 per cent greater than the actual power needed for modulation.

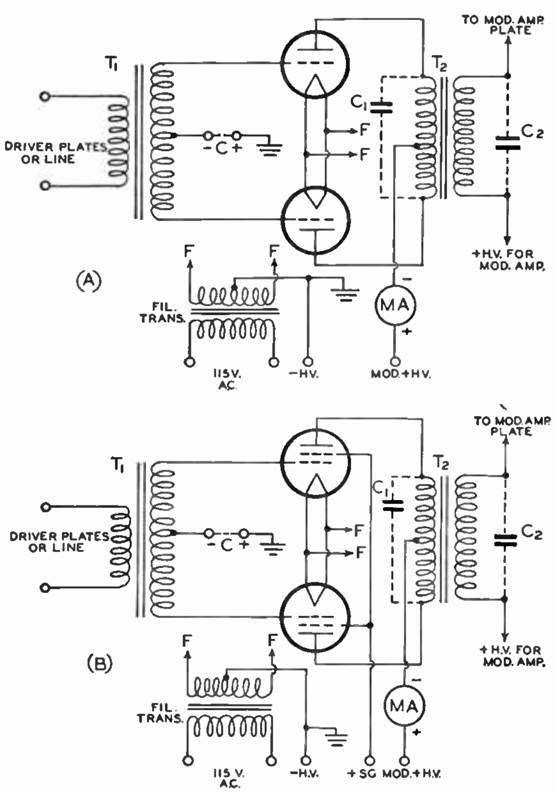


Fig. 9-23 — Modulator circuit diagrams. Tubes and circuit considerations are discussed in the text.

Matching to Load

In giving audio ratings on power tubes, manufacturers specify the plate-to-plate load impedance into which the tubes must operate to deliver the rated audio power output. This load impedance seldom is the same as the modulating impedance of the Class C r.f. stage, so a match must be brought about by adjusting the turns ratio of the coupling transformer. The required turns ratio, primary to secondary, is

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

where N = Turns ratio, primary to secondary

Z_m = Modulating impedance of Class C r.f. amplifier

Z_p = Plate-to-plate load impedance for Class B tubes

Example: The modulated r.f. amplifier is to operate at 1250 volts and 250 ma. The power input is

$$P = EI = 1250 \times 0.25 = 312 \text{ watts}$$

so the modulating power required is 312/2 = 156 watts. Increasing this by 25% to allow for losses and a reasonable operating margin gives 156 × 1.25 = 195 watts. The modulating impedance of the Class C stage is

$$Z_m = \frac{E}{I} = \frac{1250}{0.25} = 5000 \text{ ohms.}$$

From the tube tables a pair of Class B tubes is selected that will give 200 watts output when working into a 6900-ohm load, plate-to-plate. The primary-to-secondary turns ratio of the modulation transformer therefore should be

$$N = \sqrt{\frac{Z_p}{Z_m}} = \sqrt{\frac{6900}{5000}} = \sqrt{1.38} = 1.175{:}1.$$

The required transformer ratios for the ordinary range of impedances are shown graphically in Fig. 9-24.

Many modulation transformers are provided with primary and secondary taps, so that various turns ratios can be obtained to meet the requirements of particular tube combinations.

It may be that the exact turns ratio required cannot be secured, even with a tapped modulation transformer. *Small* departures from the proper turns ratio will have no serious effect if the modulator is operating well within its capabilities; if the actual turns ratio is within 10 per cent of the ideal value the system will operate satisfactorily. Where the discrepancy is larger, it is usually possible to choose a new set of operating conditions for the Class C stage to give a modulating impedance that

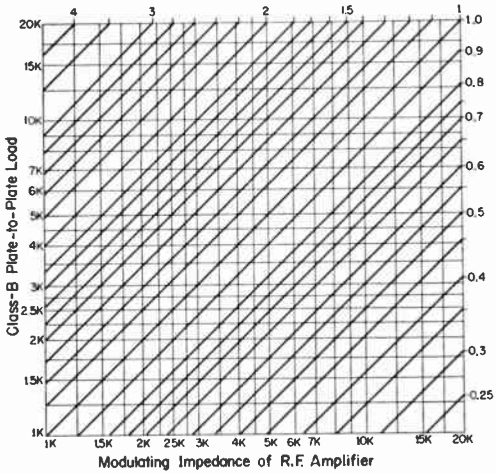


Fig. 9-24 — Transformer ratios for matching a Class C modulating impedance to the required plate-to-plate load for the Class B modulator. The ratios given on the curves are from total primary to secondary. Resistance values are in kilohms.

can be matched by the turns ratio of the available transformer. This may require operating the Class C amplifier at higher voltage and less plate current, if the modulating impedance must be increased, or at lower voltage and higher current if the modulating impedance must be decreased. However, this process cannot be carried very far without exceeding the ratings of the Class C tubes for either plate voltage or plate current, even though the power input is kept at the same figure.

Suppressing Audio Harmonics

Distortion in either the driver or Class B modulator will cause a.f. harmonics that may lie outside the frequency band needed for intelligible speech transmission. While it is almost impossible to avoid some distortion, it *is* possible to cut down the amplitude of the higher-frequency harmonics.

The purpose of condensers C_1 and C_2 across the primary and secondary, respectively, of the Class B output transformer in Fig. 9-23 is to reduce the strength of harmonics and unnecessary high-frequency components existing in the modulation. The condensers act with the leakage inductance of the transformer winding to form a rudimentary low-pass filter. The values of capacitance required will depend on the load resistance (modulating impedance of the Class C amplifier) and the leakage inductance of the particular transformer used. In general, capacitances between about 0.001 and 0.01 μfd. will be required; the larger values are necessary with the lower values of load resistance. The voltage rating of each condenser should at least be equal to the d.c. voltage at the transformer winding with which it is associated. In the case of C_2, part of the total capacitance required will be supplied by the plate by-pass or blocking capacitor in the modulated amplifier.

A still better arrangement is to use a low-pass filter as shown in Fig. 9-9, even though clipping is not deliberately employed.

Grid Bias

Certain triodes designed for Class B audio work can be operated without grid bias. Besides eliminating the grid-bias supply, the fact that grid current flows over the whole audio cycle represents a more constant load resistance for the driver. With these tubes the grid-return lead from the center-tap of the input transformer secondary is simply connected to the filament center-tap or cathode.

When the modulator tubes require bias, it should always be supplied from a *fixed* voltage source. Cathode bias or grid-leak bias cannot be used with a Class B amplifier; with both types the bias changes with the amplitude of the signal voltage, whereas proper operation demands that the bias voltage be unvarying no matter what the strength of the signal. When only a small amount of bias is required it can be obtained conveniently from a few dry cells. When greater values of bias are required, a heavy-duty "B" battery may be used if the grid current does not exceed 40 or 50 milliamperes on voice peaks. Even though the batteries are charged by the grid current rather than discharged, a battery will deteriorate with time and its internal resistance will increase. When the increase in internal resistance becomes appreciable, the battery tends to act like a grid-leak resistor and the bias varies with the applied signal. Batteries should be checked with a voltmeter occasionally while the amplifier is operating. If the bias varies more than 10 per cent or so with voice excitation the battery should be replaced.

As an alternative to batteries, a regulated bias supply may be used. This type of supply is described in the power supply chapter.

Plate Supply

The plate supply for a Class B modulator should be sufficiently well filtered to prevent hum modulation of the r.f. stage. An additional requirement is that the output condenser of the supply should have low reactance, at 100 cycles or less, compared with the load into which each tube is working. A 4-μfd. output condenser with a 1000-volt supply, or a 2-μfd. condenser with a 2000-volt supply, usually will be satisfactory. With other plate voltages, condenser values should be in inverse proportion to the plate voltage.

To keep distortion at a minimum, the voltage regulation of the plate supply should be as good as it can be made. If the d.c. output voltage of the supply varies with the load current, the voltage at *maximum* current determined the amount of power that can be taken from the modulator without distortion. A supply whose voltage drops from 1500 at no load to 1250 at the full modulator plate current is a 1250-volt supply, so far as the modulator is concerned, and any estimate of the power output available should be based on the lower figure.

It is particularly important, in the case of a tetrode Class B stage, that the screen-voltage power-supply source have excellent regulation, to prevent distortion. The screen voltage should be set as exactly as possible to the recommended value for the tube. The audio impedance between screen and cathode also must be low.

Overexcitation

When a Class B amplifier is overdriven in an attempt to secure more than the rated power, distortion increases rapidly. The high-frequency harmonics which result from the distortion modulate the transmitter, producing spurious sidebands which can cause serious interference over a band of frequencies several times the channel width required for speech. (This can happen even though the modulation percentage, as defined in the chapter on amplitude modulation, is less than 100 per cent, if the modulator is incapable of delivering the audio power required to modulate the transmitter.)

As stated earlier, such a condition may be reached by deliberate design, in case the modulator is to be adjusted for peak clipping. But whether it happens by accident or intention, the splatter and spurious sidebands can be eliminated by inserting a low-pass filter (Fig. 9-9) between the modulator and the modulated amplifier, and then taking care to see that the actual modulation of the r.f. amplifier does not exceed 100 per cent.

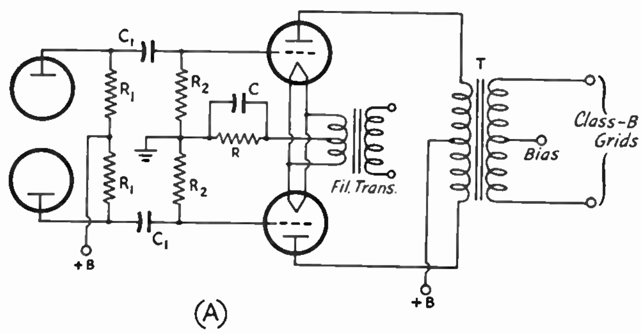
Operation Without Load

Excitation should never be applied to a Class B modulator until after the Class C amplifier is turned on and is drawing the value of plate current required to present the rated load to the modulator. With no load to absorb the power, the primary impedance of the transformer rises to a high value and excessive audio voltages are developed across it — frequently high enough to break down the transformer insulation. If the modulator is to be tested separately from the transmitter, a resistance of the same value as the modulating impedance, and capable of dissipating the full power output of the modulator, should be connected across the secondary.

DRIVERS FOR CLASS-B MODULATORS

Class AB_2 and Class B amplifiers are driven into the grid-current region, so power is consumed in the grid circuit. The preceding stage or driver must be capable of supplying this power at the required peak audio-frequency grid-to-grid voltage. Both of these quantities are given in the manufacturer's tube ratings. The grids of the Class B tubes represent a variable load resistance over the audio-frequency cycle, because the grid current does not increase directly with the grid voltage. To prevent distortion, therefore, it is necessary to have a driving source that will maintain the waveform of the signal without distortion even though the load varies. That is, the driver stage must have good regulation. To this end, it should be capable of delivering somewhat more power than is consumed by the Class B grids, as previously described in the discussion on speech amplifiers.

The driver transformer, T or T_2 in Fig. 9-25, may couple directly between the driver tube and the modulator grids or may be designed to work into a low-impedance (200- or 500-ohm) line. In the latter case, a tube-to-line output transformer must be used at the output of the driver stage. This type of coupling is recommended only when the driver must be at a considerable distance from the modulator; the second transformer not only introduces additional losses but also impairs the voltage regulation of the driver stage.



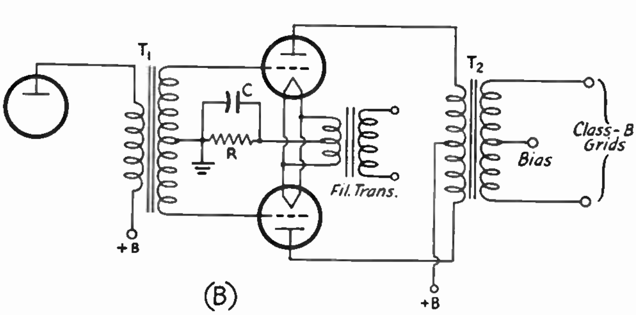


Fig. 9-25 — Triode driver circuits for Class B modulators. A, resistance coupling to grids; B, transformer coupling. R_1 in A is the plate resistor for the preceding stage, value determined by the type of tube and operating conditions as given in Table 9-I. C_1 and R_2 are the coupling condenser and grid resistor, respectively; values also may be taken from Table 9-I.

In both circuits the output transformer, T, T_2, should have the proper turns ratio to couple between the driver tubes and the Class B grids. T_1 in B is usually a 2:1 transformer, secondary to primary. R, the cathode resistor, should be calculated for the particular tubes used. The value of C, the cathode by-pass, is determined as described in the text.

Driver Tubes

To secure good voltage regulation the internal impedance of the driver, as seen by the modulator grids, must be low. The principal component of this impedance is the plate resistance of the driver tube or tubes as reflected through the driver transformer. Hence for low driving-source impedance the effective plate resistance of the driver tubes should be low and the turns ratio of the driver transformer, primary to secondary, should be as large as possible. The maximum turns ratio that can be used is that value which just permits developing the modulator grid-to-grid a.f. voltage required for the desired power output.

Low-μ triodes such as the 6B4G have low plate resistance and are therefore good tubes to use as drivers for Class AB_2 or Class B modulators. Tetrodes such as the 6L6 make very poor drivers in this respect when used without negative feed-back, but with such feed-back the effective plate resistance can be reduced to a value comparable with low-μ triodes.

In selecting a driver stage always choose Class A or AB_1 operation in preference to Class AB_2. This not only simplifies the speech-amplifier design but also makes it easier to apply negative feed-back to tetrodes for reduction of plate resistance. It is possible to obtain a tube power output of approximately 25 watts from 6L6s without going beyond Class AB_1 operation; this is ample driving power for the popular Class B modulator tubes, even when a kilowatt transmitter is to be modulated.

The rated tube output as shown by the tube tables should be reduced by about 20 per cent to allow for losses in the Class B input transformer. If two transformers are used, tube-to-line and line-to-grids, allow about 35 per cent for transformer losses. Another 25 per cent should be allowed, if possible, as a safety factor and to improve the voltage regulation.

Fig. 9-25 shows representative circuits for a push-pull triode driver using cathode bias. If the amplifier operates Class A the cathode resistor need not be by-passed, because the a.f. currents from each tube flowing in the cathode resistor are out of phase and cancel each other. However, in Class AB operation this is not true; considerable distortion will be generated at high signal levels if the cathode resistor is not by-passed. The by-pass capacitance required can be calculated by a simple rule: the cathode resistance in ohms multiplied by the by-pass capacitance in microfarads should equal at least 25,000. The voltage rating of the condenser should be equal to the maximum bias voltage. This can be found from the maximum-signal plate current and the cathode resistance.

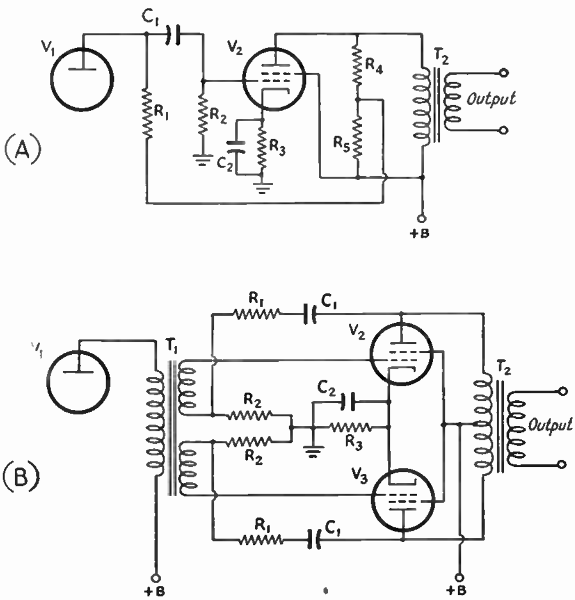


Fig. 9-26 — Negative feed-back circuits for drivers for Class B modulators. A — Single-ended beam-tetrode driver. If V_1 and V_2 are a 6J5 and 6V6, respectively, the following values are suggested: R_1, 47,000 ohms; R_2, 0.47 megohm; R_3, 250 ohms; R_4, R_5, 22,000 ohms; C_1, 0.01 μfd.; C_2, 50 μfd.

B — Push-pull beam-tetrode driver. If V_1 is a 6J5 and V_2 and V_3 6L6s, the following values are suggested: R_1, 0.1 megohm; R_2, 22,000 ohms; R_3, 250 ohms; C_1, 0.1 μfd.; C_2, 100 μfd.

Example: A pair of 6B4Gs is to be used in Class AB_1 self-biased. From the tube tables, the cathode resistance should be 780 ohms and the maximum-signal plate current 120 ma. From Ohm's Law,

$$E = RI = 780 \times 0.12 = 93.6 \text{ volts}$$

From the rule mentioned previously, the by-pass capacitance required is

$$C = 25{,}000/R = 25{,}000/780 = 32\ \mu\text{fd.}$$

A 40- or 50-μfd. 100-volt electrolytic condenser would be satisfactory.

Negative Feed-Back

Whenever tetrodes or pentodes are used as drivers for Class B modulators, negative feed-back should be used in the driver stage, for the reason discussed above.

Suitable circuits for single-ended and push-pull tetrodes are shown in Fig. 9-26. Fig. 9-26A shows resistance coupling between the preceding stage and a single tetrode, such as the 6V6, that operates at the same plate voltage as the preceding stage. Part of the a.f. voltage across the primary of the output transformer is fed back to the grid of the tetrode, V_2, through the plate resistor of the preceding tube, V_1. The total resistance of R_4 and R_5 in series should be ten or more times the rated load resistance of V_2. Instead of the voltage divider, a tap on the transformer primary can be used to supply the feed-back voltage, if such a tap is available.

The amount of feed-back voltage that appears at the grid of tube V_2 is determined by R_1, R_2 and the plate resistance of V_1, as well as by the relationship between R_4 and R_5. Circuit values for a typical tube combination are given in detail in Fig. 9-26.

The push-pull circuit in Fig. 9-26B requires an audio transformer with a split secondary. The feed-back voltage is obtained from the plate of each output tube by means of the voltage divider, R_1, R_2. The blocking condenser, C_1, prevents the d.c. plate voltage from being applied to R_1R_2; the reactance of this condenser should be low, compared with the sum of R_1 and R_2, at the lowest audio frequency to be amplified. Also, the sum of R_1 and R_2 should be high (ten times or more) compared with the rated load resistance for V_2 and V_3.

In this circuit the feed-back voltage that is developed across R_2 appears at the grid of V_2 (or V_3) through the transformer secondary and grid-cathode circuit of the tube, provided the tubes are not driven to grid current. The per cent feed-back is

$$n = \frac{R_2}{R_1 + R_2} \times 100$$

where n is the feed-back percentage, and R_1 and R_2 are connected as shown in the diagram. The higher the feed-back percentage, the lower the effective plate resistance. However, if the percentage is made too high the preceding tube, V_1, may not be able to develop enough voltage, through T_1, to drive the push-pull stage to maximum output without itself generating harmonic distortion. Distortion in V_1 is not compensated for by the feed-back circuit.

If V_2 and V_3 are 6L6s operated self-biased in Class AB_1 with a load resistance of 9000 ohms, V_1 is a 6J5, and T_1 has a turns ratio of 2-to-1, total secondary to primary, it is possible to use over 30 per cent feed-back without going beyond the output-voltage capabilities of the 6J5. Twenty per cent feed-back will reduce the effective plate resistance to the point where the output voltage regulation is better than that of 6B4Gs or 2A3s without feed-back.

If the grid-cathode impedance of the tubes is relatively low, as it is when grid current flows, the feed-back voltage decreases because of the voltage drop through the transformer secondary. The circuit should not be used with tubes that are operated Class AB_2.

SPEECH-AMPLIFIER CIRCUIT WITH NEGATIVE FEED-BACK

A circuit for a speech amplifier suitable for driving a Class B modulator is given in Fig. 9-27. In this amplifier the 6L6s are operated Class AB_1 and will deliver up to 20 watts to the grids of the Class B amplifier. The feed-back circuit requires no adjustment, but does require an interstage transformer with two separate secondary windings (split secondary).

Any convenient chassis layout may be used for the amplifier provided the principles outlined in the section on speech-amplifier construction are observed. The over-all gain is ample for a communications-type crystal microphone.

The output transformer, T_2, should be selected to work between a 9000-ohm plate-to-plate load and the grids of whatever Class B tubes will be used. The power-supply requirements for this amplifier are 145 ma. at 360 volts and 2.7 amp. at 6.3 volts.

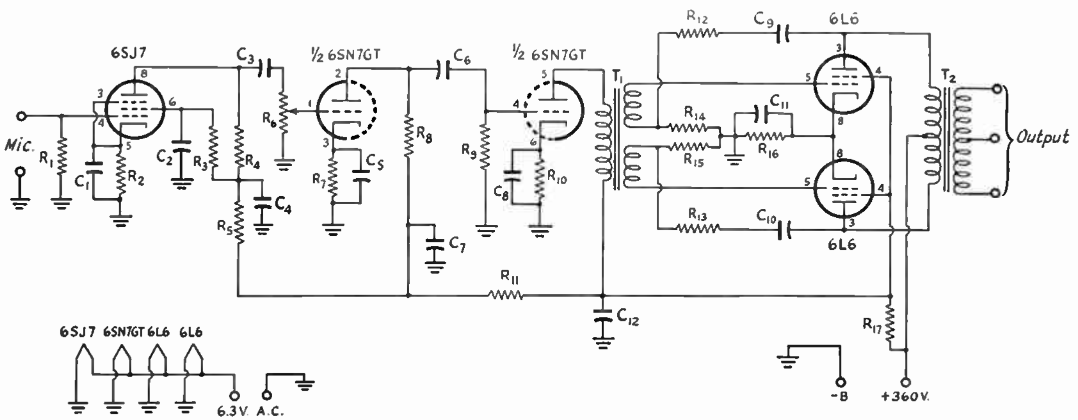


Fig. 9-27 — Circuit diagram of speech amplifier using 6L6s with negative feed-back, suitable for driving Class B modulators up to 500 watts output.

C_1, C_5, C_8 — 20-μfd. 25-volt electrolytic.
C_2, C_9, C_{10} — 0.1-μfd. 400-volt paper.
C_3, C_6 — 0.01-μfd. 600-volt paper.
C_4, C_7, C_{12} — 10-μfd. 450-volt electrolytic.
C_{11} — 100-μfd. 50-volt electrolytic.
R_1 — 2.2 megohms, ½ watt.
R_2, R_7 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5, R_8 — 47,000 ohms, ½ watt.
R_6 — 1-megohm volume control.
R_9 — 0.47 megohm, ½ watt.
R_{10} — 1500 ohms, 1 watt.
R_{11} — 10,000 ohms, ½ watt.
R_{12}, R_{13} — 0.1 megohm, 1 watt.
R_{14}, R_{15} — 22,000 ohms, ½ watt.
R_{16} — 250 ohms, 10 watts.
R_{17} — 2000 ohms, 10 watts.
T_1 — Interstage audio, 2:1 secondary (total) to primary, with split secondary winding.
T_2 — Class B input transformer to suit modulator tubes.

Class B Modulator with Filter

Representative Class B modulator construction is illustrated by the unit shown in Figs. 9-28 and 9-30. This modulator includes a splatter filter, $C_1C_2L_1$ in the circuit diagram, Fig. 9-29, and also has provision for short-circuiting the modulation transformer secondary when c.w. is to be used.

Fig. 9-28 — **A typical Class B modulator arrangement. This unit uses a pair of 811As, capable of an audio power output of 340 watts, and includes a splatter filter. The modulation transformer is at the left and the splatter choke at the right. All high-voltage terminals are covered so they cannot be touched accidentally.**

The audio input transformer is not built into this unit, it being assumed that this transformer will be included in the driver assembly as is customary. If the modulator and speech amplifier-driver are mounted in the same rack or cabinet, the length of leads from the driver to the modulator grids presents no problem. The bias required by the modulator tubes at their higher plate-voltage ratings should be fed through the center-tap on the secondary of the driver transformer. At a plate voltage of 1000 or less no bias is needed and the center-tap connection on the transformer can be grounded.

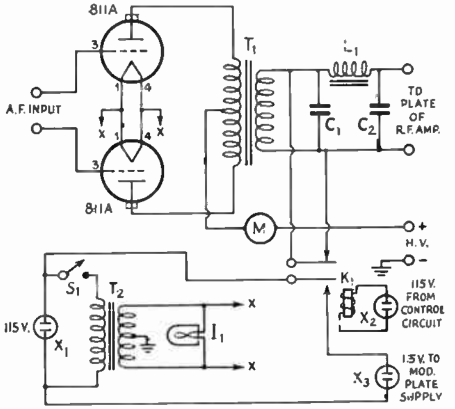


Fig. 9-29 — **Circuit diagram of the Class B modulator.**

C_1, C_2, L_1 — See text. (L_1 is Chicago Transformer type SR-300.)
K_1 — D.p.d.t. relay, high-voltage insulation (Advance type 400).
M — 0–500 d.c. milliammeter, bakelite case.
T_1 — Variable-ratio modulation transformer (Chicago Transformer type CMS-1).
T_2 — Filament transformer, 6.3 v., 8 amp.
I_1 — 6.3-volt pilot light.
X_1, X_2 — Chassis-type 115-volt plugs, male.
X_3 — Chassis-type 115-volt receptacle, female.
S_1 — S.p.s.t. toggle.

The values of C_1, C_2 and L_1 depend on the modulating impedance of the Class C r.f. amplifier. They can be determined from the formulas given in this chapter in the section on high-level clipping and filtering. The splatter filter will be effective regardless of whether the modulator operating conditions are chosen to give high-level clipping, but it is worth-while to design the system for clipping at 100 per cent modulation if the tube curves are available for that purpose. The voltage ratings for C_1 and C_2 should at least equal the d.c. voltage applied to the modulated r.f. amplifier.

A relay with high-voltage insulation (actually an antenna relay) is used to short-circuit the secondary of T_1 when the relay coil is not energized. A normally-closed contact is used for this purpose. The other arm is used to close the primary circuit of the modulator plate supply when the relay is energized. Shorting the transformer secondary is necessary when the r.f. amplifier is keyed, to prevent an inductive discharge from the transformer winding that would put "tails" on the keyed characters and, with cathode keying of the amplifier, would cause excessive sparking at the key contacts. The control circuit should be arranged in such a way that K_1 is not energized during c.w. operation but is energized by the send-receive switch during 'phone operation.

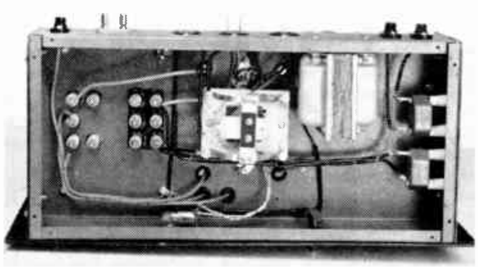


Fig. 9-30 — **The filament transformer is mounted below the chassis. The relay is used as described in the text. C_1 and C_2 are mounted on small stand-off insulators on the chassis wall.**

Careful attention should be paid to insulation since the instantaneous voltages in the secondary circuit of the transformer will be at least twice the d.c. voltage on the r.f. amplifier. Stand-off insulators are used in this unit wherever necessary, including the mounting for the relay.

Checking Amplifier Operation

An adequate job of checking speech amplifiers can be done with equipment that is neither elaborate nor expensive. A simple set-up is shown in Fig. 9-31. The construction of a simple audio oscillator is described in the chapter on measurements. The audio-frequency voltmeter can be either a vacuum-tube voltmeter or a multirange volt-ohm-milliammeter that has a rectifier-type a.c. range. The headset is included for aural checking of the amplifier performance.

An audio oscillator usually will have an output control, but if the maximum output voltage is in excess of a volt or so the output setting may be rather critical when a high-gain speech amplifier is being tested. In such cases an attenuator such as is shown in Fig. 9-31 is a convenience. Each of the two voltage dividers reduces the voltage by a factor of roughly 10 to 1, so that the over-all attenuation is about 100 to 1. The relatively low value of resistance, R_4, across the input terminals of the amplifier also will minimize stray hum pick-up on the connecting leads.

As a preliminary check, cover the microphone input terminals with a metal shield (with the audio oscillator and attenuator disconnected) and, while listening in the headset, note the hum level with the amplifier gain control in the off position. The hum should be very low under these conditions. Then increase the gain-control setting to maximum and observe the hum; it will no doubt increase. Next connect the audio oscillator and attenuator and, starting from minimum signal, increase the audio input voltage until the voltmeter indicates full power output. (The voltage should equal $\sqrt{PR}$, where P is the expected power output in watts and R is the load resistance — R_6 in the diagram.) While increasing the input, listen carefully to the tone to see if there is any change in its character. When it begins to sound like a musical octave instead of a single tone, distortion is beginning. Assuming that the output is substantially without audible distortion at full output, substitute the microphone for the audio oscillator and speak into it in a normal tone while watching the voltmeter. Reduce the gain-control setting until the meter "kicks" nearly up to the full-power reading on voice peaks. Note the hum level, as read on the voltmeter, at this point; the hum level should not exceed one or two per cent of the voltage at full output.

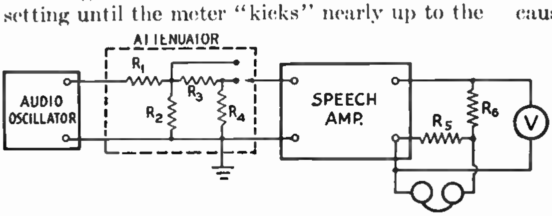


Fig. 9-31 — **Simple test set-up for checking a speech amplifier. It is not necessary that the frequency range of the audio oscillator be continuously variable; a number of "spot frequencies, or even one such frequency, will be satisfactory. Suitable resistor values are: R_1 and R_3, 10,000 ohms; R_2 and R_4, 1000 ohms; R_6, rated load resistance for amplifier output stage; R_5 determine by trial for comfortable headphone level (25 to 100 ohms, ordinarily). V is a high-resistance a.c. voltmeter, multirange rectifier type.**

If the hum level is too high, the amplifier stage that is causing the trouble can be located by temporarily short-circuiting the grid of each tube to ground, starting with the output amplifier. When shorting a particular grid makes a marked decrease in hum, the hum presumably is coming from a *preceding* stage, although it is possible that it is getting its start in that particular grid circuit. If shorting a grid does *not* decrease the hum, the hum is originating either in the plate circuit of that tube or the grid circuit of the next. Aside from wiring errors, a defective tube, or inadequate plate-supply filtering, objectionable hum usually originates in the first stage of the amplifier.

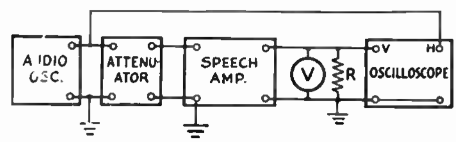


Fig. 9-32 — **Test set-up using the oscilloscope to check for distortion. These connections will result in the type of pattern shown in Fig. 9-33, the horizontal sweep being provided by the audio input signal. For waveform patterns, omit the connection between the audio oscillator and the horizontal amplifier in the 'scope, and use the horizontal linear sweep.**

If distortion occurs below the point at which the expected power output is secured, the stage in which it is occurring can be located by working from the last stage toward the front end of the amplifier, applying a signal to each grid in turn from the audio oscillator and adjusting the signal voltage for maximum output. In the case of push-pull stages, the signal may be applied to the primary of the interstage transformer — *after* disconnecting it from the plate-voltage source. Assuming that normal design principles have been followed and that all stages are theoretically working within their capabilities, the probable causes of distortion are wiring errors (such as accidental short-circuit of a cathode resistor), defective components, or use of wrong values of resistance in cathode and plate circuits.

Using the Oscilloscope

Speech-amplifier checking is facilitated considerably if an oscilloscope of the type having amplifiers and a linear sweep circuit is available. A typical set-up for using the oscilloscope is shown in Fig. 9-32. With the connections shown, the sweep circuit is not required but horizontal and vertical amplifiers are necessary. Audio voltage from the oscillator is fed directly to one oscilloscope amplifier (horizontal in this case) and the output of the speech amplifier is connected to the other. The 'scope amplifier gains should be adjusted so that each signal gives the same line length with the other signal shut off.

Under these conditions, when the input and output signals are applied simultaneously they are compared directly. If the speech amplifier is distortion-free and introduces no phase shift, the resulting pattern is simply a straight line, as shown at the upper left in Fig. 9-33, making an angle of about 45 degrees with the horizontal and vertical axes. If there is no distortion but there is phase shift, the pattern will be a smooth ellipse, as shown at the upper right. The greater the phase shift the greater the tendency of the ellipse to grow into a circle. When there is even-harmonic distortion in the amplifier one end of the line or ellipse becomes curved, as shown in the second row in Fig. 9-33. With odd-harmonic distortion such as is characteristic of overdriven push-pull stages, the line or ellipse is curved at both ends.

Patterns such as these will be obtained when the input signal is a fairly good sine wave. They will tend to become complicated if the input waveform is complex and the speech amplifier introduces appreciable phase shifts. It is therefore advisable to test for distortion with an input signal that is as nearly as possible a sine wave. Also, it is best to use a frequency in the 500–1000 cycle range, since improper phase shift in the amplifier is usually least in this region. Phase shift in itself is not of great importance in an audio amplifier of ordinary design because it does not change the character of speech so far as the ear is concerned. However, if a complex signal is used for testing, phase shift may make it difficult to detect distortion in the oscilloscope pattern.

In amplifiers having negative feed-back, excessive phase shift within the feed-back loop may cause self-oscillation, since the signal fed back may arrive at the grid in phase with the applied signal voltage instead of out of phase with it. Such a phase shift is most likely to be associated with the output transformer. Oscillation usually occurs at some frequency above 10,000 cycles, although occasionally it will occur at a very low frequency. If the pass-band in the stage in which the phase shift occurs is deliberately restricted to the optimum voice range, as described earlier, the gain at both very high and very low frequencies will be so low that self-oscillation is unlikely, even with large amounts of feed-back.

Generally speaking, it is easier to detect small amounts of distortion with the type of pattern shown in Fig. 9-33 than it is with the waveform pattern obtained by feeding the output signal to the vertical plates and making use of the linear sweep in the 'scope. However, the waveform pattern can be used satisfactorily if the signal from the audio oscillator is a reasonably good sine wave. One simple method is to examine the output of the oscillator alone and trace the pattern on a sheet of transparent paper. The pattern

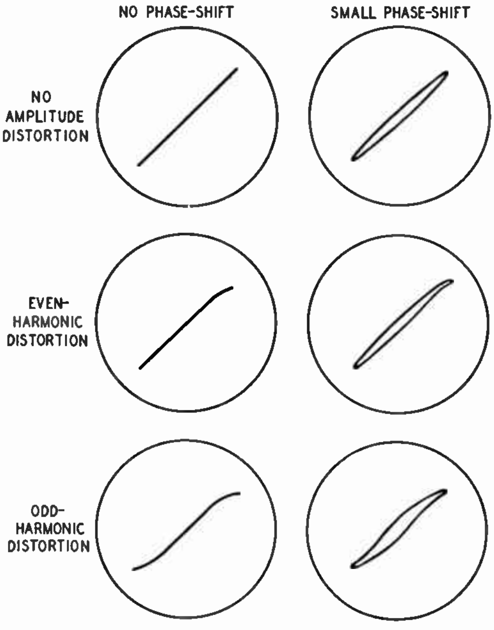


Fig. 9-33 — Typical patterns obtained with the connections shown in Fig. 9-32. Depending on the number of stages in the amplifier, the pattern may slope upward to the right, as shown, or upward to the left. Also, depending on where the distortion originates, the curvature in the second row may appear either at the top or bottom of the line or ellipse.

given by the output of the amplifier can then be compared with the "standard" pattern by adjusting the oscilloscope gain to make the two patterns coincide as closely as possible. The pattern discrepancies are a measure of the distortion.

In using the oscilloscope care must be taken to avoid introducing hum voltages that will upset the measurements. Hum pick-up on the 'scope leads or other exposed parts such as the amplifier load resistor or the voltmeter can be detected by shutting off the audio oscillator and speech amplifier and connecting first one and then the other to the vertical plates of the 'scope, setting the internal horizontal sweep to an appropriate width. The trace should be a straight horizontal line when the vertical gain control is set at the position used in the actual measurements. Waviness in the line indicates hum. If the hum is not in the 'scope itself (check by disconnecting the leads at the instrument) make sure that there is a good ground connection on all the equipment and, if necessary, shield the hot leads.

The oscilloscope can be used to good advantage in stage-by-stage testing to check waveforms at the grid and plate of each stage and thus to determine rapidly where a source of trouble may be located. When the 'scope is connected to circuits that are not at ground potential for d.c., a condenser of about 0.1 μfd. should be connected in series with the hot oscilloscope lead. The probe lead should be shielded so that it will not pick up hum.

Amplitude Modulation

The type of modulation most commonly employed in amateur radiotelephony is called **amplitude modulation (AM)**. The name arises from the fact that the methods of generating a modulated wave of a particular type all accomplish the desired result by varying the instantaneous amplitude of the r.f. output of the transmitter. As described in the chapter on circuit fundamentals, the process of modulating a signal sets up groups of frequencies called **sidebands**, these sidebands appearing both above and below the frequency of the unmodulated signal or carrier. An amplitude-modulated signal actually consists of a carrier which does *not* vary in amplitude plus sets of side frequencies or sidebands which in turn may or may not vary in amplitude. Modulation by a single-frequency, constant-amplitude tone, for example, sets up side frequencies that do not vary in amplitude. Modulation by voice sets up bands of side frequencies that do vary with the amplitude of the speech.

Amplitude modulation is frequently described as a process of "varying the amplitude of the carrier." A variation in amplitude does take place, when the *composite signal as a whole* is viewed in a circuit that accepts equally well all frequencies, carrier and sidebands, contained in the signal. The total r.f. output amplitude varies at the modulation-frequency rate because it is the *resultant* of the instantaneous amplitudes of the carrier and all side frequencies, which continually vary (at radio frequency) in both amplitude and phase relationships. Misunderstanding often occurs because commonly no distinction is made between the carrier, which does not vary in amplitude at modulation frequency, and the signal as a whole, which does vary in amplitude with modulation. In this chapter the term "signal" is used for the composite effect of carrier plus sidebands.

It is illuminating to consider amplitude modulation as a process of frequency conversion or mixing, in which case the relationship between the carrier, modulating frequencies, and sidebands is straightforward (see chapter on fundamentals). The amplitude variations in the signal arise as a result of the mixing process. These amplitude variations are highly important from a design standpoint, since they set up certain power requirements that must be met, so they are considered in detail in this chapter.

AM Sidebands and Channel Width

As described in the chapter on fundamentals, combining or mixing two frequencies in an appropriate circuit gives rise to sum and difference frequencies. Speech can be electrically reproduced, with high intelligibility, in a band of frequencies lying between approximately 100 and 3000 cycles. When these frequencies are combined with a radio-frequency carrier, the sidebands occupy the frequency spectrum from about 3000 cycles below the carrier frequency to 3000 cycles above — a total band or "channel" of about 6 kilocycles. Actual speech frequencies extend up to 10,000 cycles or so, so it is possible to occupy a 20-kc. channel if no provision is made for reducing its width. For communication purposes such a channel width represents a waste of valuable spectrum space, since a 6-kc. channel is fully adequate for intelligibility. Occupying more than the minimum channel creates unnecessary interference, so speech equipment and transmitter adjustment and operation should be pointed toward maintaining the channel width at the minimum.

THE MODULATED SIGNAL

In Fig. 10-1, the drawing at A shows the unmodulated r.f. signal, assumed to be a sine wave of the desired radio frequency. The graph can be taken to represent either voltage or current.

In B, the signal is assumed to be modulated by the audio-frequency shown in the small drawing above. This frequency is much lower than the carrier frequency, a necessary condition for good modulation, and always the case in radiotelephony because the audio frequencies used are very low compared with the radio frequency of the carrier. When the modulating voltage is "positive" (above its axis) the signal amplitude is increased *above* its unmodulated amplitude; when the modulating voltage is "negative" the signal amplitude is *decreased*. Thus the signal grows larger and smaller with the polarity and amplitude of the modulating voltage.

The drawings at C shows what happens with stronger modulation. The amplitude is doubled at the instant the modulating voltage reaches its positive peak. On the negative peak of the modulating voltage the amplitude just reaches zero; in other words, the signal is completely modulated.

Percentage of Modulation

When a modulated signal is detected in a receiver, the detector eliminates the carrier and takes from it the modulation. The stronger the modulation, therefore, the greater is the useful receiver output. Obviously, it is desirable to make the modulation as strong or "heavy" as possible. A wave modulated as in Fig. 10-1C would produce considerably more useful audio output than the one shown at B.

The "depth" of the modulation is expressed as a percentage of the unmodulated carrier amplitude. In either B or C, Fig. 10-1, X represents the unmodulated carrier amplitude, Y is the *maximum* amplitude on the modulation up-peak, and Z is the *minimum* amplitude on the modulation downpeak.

The outline of the modulated wave is called the **modulation envelope**. It is shown by the thin line outlining the patterns in Fig. 10-1. In a properly-operating modulation system either side of this outline is an accurate reproduction

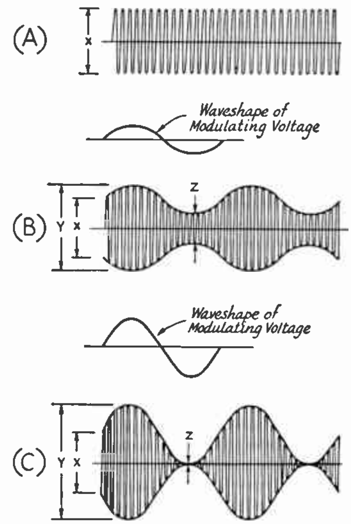


Fig. 10-1 — Graphical representation of (A) r.f. output unmodulated, (B) modulated 50%, (C) modulated 100%.

of the modulating wave, as can be seen in Fig. 10-1 at B and C by comparing the upper outline of the modulation envelope with the waveshape of the modulating wave. The lower outline duplicates the upper, but simply appears upside down in the drawing.

The **percentage of modulation** is

$$\% \text{ Mod.} = \frac{Y - X}{X} \times 100 \text{ (upward modulation), or}$$

$$\% \text{ Mod.} = \frac{X - Z}{X} \times 100 \text{ (downward modulation)}$$

If the waveshape of the modulation is such that its peak positive and negative amplitudes are equal, then the modulation percentage will be the same both up and down. If the two percentages differ, the larger of the two is customarily specified.

Power in Modulated Wave

The amplitude values shown in Fig. 10-1 correspond to current or voltage, so the drawings may be taken to represent instantaneous values of either. Now power varies as the *square* of either the current or voltage, so at the peak of the modulation up-swing the instantaneous power in the signal of Fig. 10-1C is four times the unmodulated carrier power (because the current and voltage both are doubled). At the peak of the down-swing the power is zero, since the amplitude is zero. These statements are true of 100 per cent modulation no matter what the waveform of the modulation. The instantaneous power in the modulated signal is proportional to the square of its amplitude at every instant. This fact is highly important in the operation of every method of amplitude modulation.

It is convenient, and customary, to describe the operation of modulation systems in terms of sine-wave modulation. Although this waveshape is seldom actually used in practice (voice waveshapes depart very considerably from the sine form) it lends itself to simple calculations and its use as a standard permits comparison between systems on a common basis. With sine-wave modulation the power in the modulated signal averaged over any number of full cycles of the modulation frequency is found to be 1½ times the power in the unmodulated carrier. In other words, the power output increases 50 per cent with 100 per cent modulation by a sine wave. This relationship is very useful in the design of modulation systems and modulators, since any such system that is capable of increasing the *average* power output by 50 per cent with sine-wave modulation automatically fulfills the requirement that the *instantaneous* power at the modulation up-peak be four times the carrier power. No such simple relationship exists with complex waveforms, consequently systems in which the additional power is supplied from outside the modulated r.f. stage (e.g., plate modulation) usually are designed on a sine-wave basis as a matter of convenience. Modulation systems in which the additional power is secured from the modulated r.f. amplifier (e.g., grid modulation) usually are more conveniently designed on the basis of peak power rather than average power.

The extra power that is contained in a modulated signal goes entirely into the sidebands, half in the upper sideband and half in the lower. As a numerical example, full modulation of a 100-watt carrier by a sine wave will add 50 watts of sideband power, 25 in the lower and 25 in the upper sideband. Supplying this additional power for the sidebands is the object of all of the various systems devised for amplitude modulation.

Complex waveforms such as speech do not, as a rule, contain as much average power as a sine wave. Ordinary speech waveforms have about half as much average power as a sine wave, for the same *peak* amplitude in both waveforms. Since it is the peak amplitude, not the average power, that determines the percentage of modulation, the sideband power with ordinary speech averages only about half the power with sine-wave modulation, for the same modulation percentage in both cases.

Unsymmetrical Modulation

In an ordinary electric circuit it is possible to increase the amplitude of current flow indefinitely, up to the limit of the power-handling capability of the components, but it cannot very well be decreased to less than zero. The same

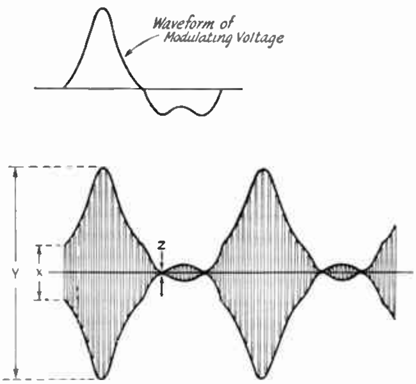


Fig. 10-2 — Modulation by an unsymmetrical waveform. This drawing shows 100% downward modulation along with 300% upward modulation. There is no distortion, since the modulation envelope is an accurate reproduction of the waveform of the modulating voltage.

thing is true of the amplitude of an r.f. signal; it can be modulated *upward* to any desired extent, but it cannot be modulated *downward* more than 100 per cent.

When the modulating waveform is unsymmetrical it is possible for the upward and downward modulation percentages to be different. A simple case is shown in Fig. 10-2. The positive peak of the modulating signal is about 3 times the amplitude of the negative peak. If, as shown in the drawing, the modulating amplitude is adjusted so that the peak downward modulation is just 100 per cent ($Z = 0$) the peak upward modulation is 300 per cent ($Y = 4X$). The carrier amplitude is represented by X, as in Fig. 10-1. The modulation envelope reproduces the waveform of the modulating signal accurately, hence there is no distortion. In such a modulated signal the increase in power output with modulation is considerably greater than when the modulation is symmetrical and has to be limited to 100 per cent both up and down. However, the peak amplitude, Y, is four times the carrier amplitude, X, so the peak *power* is 16 times the carrier power. When the upward modulation is more than 100 per cent the peak power capacity of the modulating system obviously must be increased sufficiently to take care of the much larger peak amplitudes.

Overmodulation

If the amplitude of the modulation on the downward swing becomes too great, there will be a period of time during which the output is entirely cut off. This is shown in Fig. 10-3. The shape of the downward half of the modulating wave is no longer accurately reproduced by the modulation envelope, consequently the modulation is distorted. Operation of this type is called **overmodulation.** The distortion of the modulation envelope causes new frequencies to be generated (harmonics of the modulating frequency, which combine with the carrier to form new sidebands correspondingly spaced from the carrier frequency) that widen the channel occupied by the modulated signal. These spurious frequencies are commonly called "splatter."

It is important to realize that the channel occupied by an amplitude-modulated signal is dependent *on the waveshape of the modulation envelope.* If this waveshape is complex and can be resolved into a wide band of audio frequencies, then the channel occupied will be correspondingly large. The modulation-envelope waveshape shown in Fig. 10-3 will contain a large number of harmonics of the original sine-wave frequency of the modulating wave because of the sharp corners in the waveshape when it is "clipped" at the zero axis. However, if the original modulating wave had had exactly this same shape the channel occupied by the modulated signal would be exactly the same. Basically, it is not the fact that the signal cannot be modulated more than 100 per cent downward that causes splatter, but the fact that *any* distorted waveshape contains higher frequencies than were present in the original undistorted wave. A wave that is efficiently clipped, as is the case with the waveshape shown in Fig. 10-3, will contain a wider range of spurious frequencies than one in which there are no highly abrupt changes in amplitude.

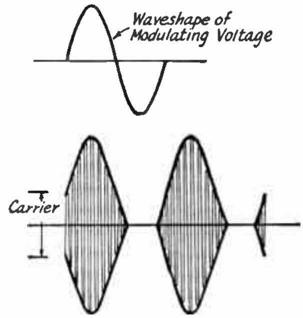


Fig. 10-3 — An overmodulated signal. The modulation envelope is not an accurate reproduction of the waveform of the modulating voltage. This or any type of distortion occurring during the modulation process generates spurious sidebands or "splatter."

Because of this clipping action at zero amplitude, it is important that care be taken to prevent applying too large a modulating signal in the downward direction. Overmodulation results in more splatter than is caused by most other types of distortion in a 'phone transmitter.

GENERAL REQUIREMENTS

For proper operation of an amplitude-modulated transmitter there are a few general requirements that must be met no matter what particular method of modulation may be used. Failure to meet them is accompanied by undesirable effects, principally distortion of the modulation envelope that increases the channel width as compared with that required by the legitimate frequencies contained in the original modulating wave.

Frequency Stability

For satisfactory amplitude modulation, the carrier *frequency* must be entirely unaffected by modulation. If the application of modulation causes a change in the carrier frequency, the frequency will wobble back and forth with the modulation. This causes distortion and widens the channel taken by the signal. Thus unnecessary interference is caused to other transmissions.

In practice, this undesirable frequency modulation is prevented by applying the modulation to an r.f. amplifier stage that is isolated from the frequency-controlling oscillator by a **buffer amplifier.** Amplitude modulation applied directly to an oscillator always is accompanied by frequency modulation. Under existing FCC regulations amplitude modulation of an oscillator is permitted only on frequencies above 144 Mc. Below that frequency the regulations require that an amplitude-modulated transmitter be completely free from frequency modulation.

Linearity

At least up to the limit of 100 per cent upward modulation, the amplitude of the r.f. output should be directly proportional to the amplitude of the modulating wave. Fig. 10-4 is a graph of an ideal **modulation characteristic,** or curve showing the relationship between r.f. output amplitude and instantaneous modulation amplitude. The modulation swings the r.f. amplitude back and forth along the curve *A*, as the modulating voltage alternately swings positive and negative. Assuming that the negative peak of the modulating wave is just sufficient to reduce the r.f. output to zero (modulating voltage equal to −1 in the drawing), the same modulating voltage peak in the *positive* direction (+1) should cause the r.f. amplitude to reach twice its unmodulated value. The ideal is a straight line, as shown by curve *A*. Such a modulation characteristic is perfectly **linear.**

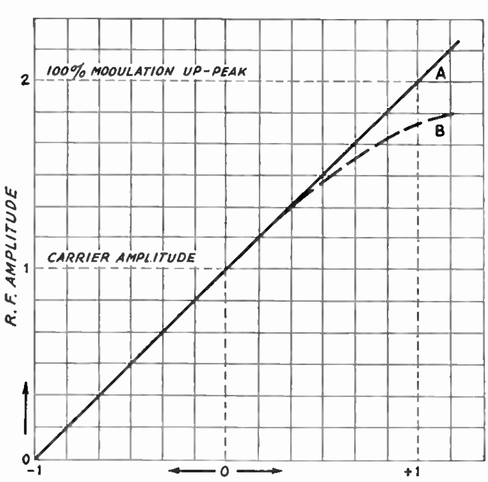


Fig. 10-4 — The modulation characteristic shows the relationship between the instantaneous amplitude of the r.f. output current (or voltage) and the instantaneous amplitude of the modulating voltage. The ideal characteristic is a straight line, as shown by curve *A*.

A **nonlinear** characteristic is shown by curve *B*. The r.f. amplitude does not reach twice the unmodulated carrier amplitude when the modulating voltage reaches its positive peak. A modulation characteristic of this type gives a modulation envelope that is "flattened" on the up-peak; in other words, the modulation envelope is not an exact reproduction of the modulating wave. It is therefore distorted and harmonics are generated, causing the transmitted signal to occupy a wider channel than is necessary. A nonlinear modulation characteristic can easily result when a transmitter is not properly designed or is misadjusted.

The **modulation capability** of the transmitter is the maximum percentage of modulation that is possible without objectionable distortion from nonlinearity. The maximum capability can never exceed 100 per cent on the down-peak, but it is possible for it to be higher on the up-peak. The modulation capability should be as close to 100 per cent as possible, so that the most effective signal can be transmitted.

Plate Power Supply

The d.c. power supply for the plate or plates of the modulated amplifier should be well filtered; if it is not, plate-supply ripple will modulate the carrier and cause annoying hum. The ripple voltage should not be more than about 1 per cent of the d.c. output voltage.

In amplitude modulation the plate current varies at an audio-frequency rate; in other words, an alternating current is superimposed on the d.c. plate current. The output filter condenser in the plate supply must have low reactance, at the lowest audio frequency in the modulation, if the transmitter is to modulate equally well at all audio frequencies. The condenser capacitance required depends on the ratio of d.c. plate current to plate voltage in the modulated amplifier. The requirements will be met satisfactorily if the capacitance of the output condenser is at least equal to

$$C = 25\frac{I}{E}$$

where C = Capacitance of output condenser in μfd.

I = D.c. plate current of modulated amplifier in milliamperes

E = Plate voltage of modulated amplifier

Example: A modulated amplifier operates at 1250 volts and 275 ma. The capacitance of the output condenser in the plate-supply filter should be at least

$$C = 25\frac{I}{E} = 25 \times \frac{275}{1250} = 25 \times 0.22 = 5.5 \ \mu\text{fd.}$$

Modulation Systems

An amplitude-modulated signal can be generated by a variety of methods, the only presently-used ones being those in which a modulating voltage is applied to one or more tube elements in an r.f. amplifier. The proper object of all methods is to generate an r.f. signal having a modulation envelope which reproduces the waveform of the modulating voltage with as little distortion as possible.

The methods described in this chapter are the basic ones. There are many specialized variations, usually involving some form of grid modulation with the object of increasing the rather low plate efficiency that is an inherent characteristic of grid modulation. Such systems, when they actually achieve substantially distortionless modulation, are rather complicated circuitwise, are difficult to adjust and are not well adapted to rapid frequency change. They have so far had little or no lasting application in amateur communication.

Amplitude Modulation Methods

PLATE MODULATION

The most popular system of amplitude modulation is plate modulation. It is the simplest to apply, gives the highest efficiency in the modulated amplifier, and is the easiest to adjust for proper operation.

Fig. 10-5 shows the most widely-used system of plate modulation, in this case with triode r.f. tubes. A balanced (push-pull Class A, Class AB or Class B) **modulator** is transformer-coupled to the plate circuit of the modulated r.f. amplifier. The audio-frequency power generated by the modulator is combined with the d.c. power in the modulated-amplifier plate circuit by transfer through the coupling transformer, *T*. For 100 per cent modulation the audio-frequency output of the modulator and the turns ratio of the coupling transformer must be such that the voltage at the plate of the modulated amplifier varies between zero and twice the d.c. operating plate voltage, thus causing corresponding variations in the amplitude of the r.f. output.

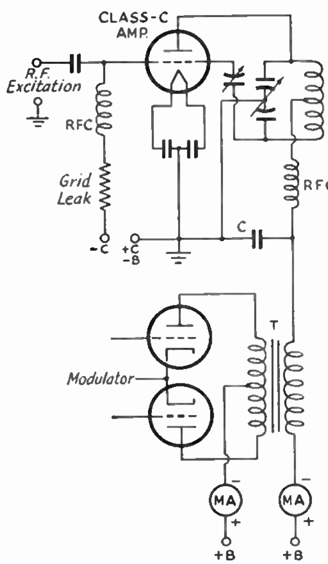


Fig. 10-5 — Plate modulation of a Class C r.f. amplifier. The r.f. plate by-pass condenser, *C*, in the amplifier stage should have reasonably high reactance at audio frequencies. A value of the order of 0.001 μfd. to 0.005 μfd. is satisfactory in practically all cases. (See chapter on modulators.)

Audio Power

As stated earlier, the average power output of the modulated stage must increase during modulation. The modulator must be capable of supplying to the modulated r.f. stage sine-wave audio power equal to 50 per cent of the d.c. plate input. For example, if the d.c. plate power input to the r.f. stage is 100 watts, the sine-wave audio power output of the modulator must be 50 watts.

Modulating Impedance; Linearity

The **modulating impedance,** or load resistance presented to the modulator by the modulated r.f. amplifier, is equal to

$$Z_m = \frac{E_b}{I_p} \times 1000 \text{ ohms}$$

where E_b = D.c. plate voltage
I_p = D.c. plate current (ma.)

E_b and I_p are measured without modulation.

The power output of the r.f. amplifier must vary as the square of the instantaneous plate voltage (the r.f. voltage must be proportional to the plate voltage) in order for the modulation to be linear. This will be the case when the amplifier operates under Class C conditions. The linearity depends upon having sufficient grid excitation and proper bias, and upon the adjustment of circuit constants to the proper values.

Adjustment of Plate-Modulated Amplifiers

The general operating conditions for Class C operation are described in the chapter on transmitters. The grid bias and grid current required for plate modulation usually are given in the operating data supplied by the tube manufacturer; in general, the bias should be such as to give an operating angle of about 120 degrees at the d.c. plate voltage used, and the grid excitation should be great enough so that the amplifier's plate efficiency will stay constant when the plate voltage is varied over the range from zero to twice the unmodulated value. For best linearity, the grid bias should be obtained partly from a fixed source of about the cut-off value, and then supplemented by grid-leak bias to supply the remainder of the required operating bias.

The maximum permissible d.c. plate power input for 100 per cent modulation is twice the sine-wave audio-frequency power output available from the modulator. This input is obtained by varying the loading on the amplifier (keeping its tank circuit tuned to resonance) until the product of d.c. plate voltage and plate current is the desired power. The modulating impedance under these conditions must be transformed to the proper value for the modulator by using the correct output-transformer turns ratio. This point is considered in detail in the chapter on modulator design.

Neutralization, when triodes are used, should be as nearly perfect as possible, since regeneration may cause nonlinearity. The amplifier also must be completely free from parasitic oscillations.

Although the total power input (d.c. plus audio-frequency a.c.) increases with modulation, the d.c. plate current of a plate-modulated amplifier should not change when the stage is modulated. This is because each increase in plate voltage and plate current is balanced by an equivalent decrease in voltage and current on the next half-cycle of the modulating wave. D.c. instruments cannot follow the a.f. variations, and since the average d.c. plate current and plate voltage of a properly-operated amplifier do not change, neither do the meter readings. A change in plate current with modulation indicates nonlinearity. On the other hand, a thermocouple r.f. ammeter connected in the antenna or transmission line will show an increase in r.f. current with modulation, because instruments of this type respond to power rather than to current or voltage.

Screen-Grid Amplifiers

Screen-grid tubes of the pentode or beam-tetrode type can be used as Class C plate-modulated amplifiers by applying the modulation to both the plate and screen grid. The usual method of feeding the screen grid with the necessary d.c. and modulation voltage is shown in Fig. 10-6. The dropping resistor, *R*, should be of the proper value to apply normal d.c. voltage to the screen under steady carrier conditions. Its value can be calculated by taking the difference between plate and screen voltages and dividing it by the rated screen current.

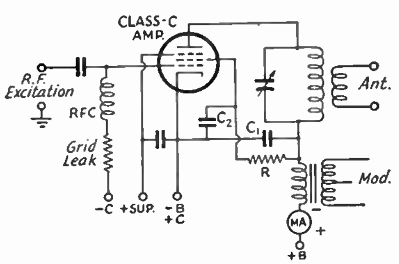


Fig. 10-6 — Plate and screen modulation of a Class C r.f. amplifier using a screen-grid tube. The plate r.f. by-pass condenser, C_1, should have reasonably high reactance at all audio frequencies; a value of 0.001 to 0.005 μfd. is generally satisfactory. The screen by-pass, C_2, should be 0.002 μfd. or less in the usual case.

When the modulated amplifier is a beam tetrode the suppressor connection shown in this diagram may be ignored. If a base terminal is provided on the tube for the beam-forming plates, it should be connected as recommended by the manufacturer.

The modulating impedance is found by dividing the d.c. plate voltage by the sum of the plate and screen currents. The plate voltage multiplied by the sum of the two currents gives the power input to be used as the basis for determining the audio power required from the modulator.

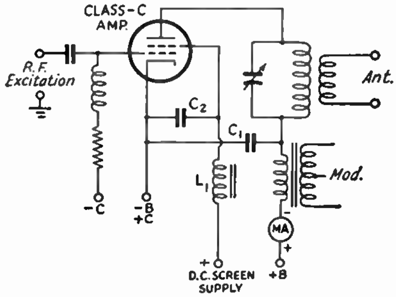


Fig. 10-7 — Plate modulation of a beam tetrode, using an audio impedance in the screen circuit. The value of L_1 is discussed in the text. See Fig. 10-6 for data on bypass capacitors C_1 and C_2.

Modulation of the screen along with the plate is necessary because the screen voltage has a much greater effect on the plate current than the plate voltage does. Very little modulation takes place and the modulation characteristic is nonlinear if the plate alone is modulated. However, beam tetrodes can be modulated satisfactorily by applying the modulating power to the plate circuit alone, *provided* the screen is "floating" at audio frequencies — that is, is not grounded for a.f. but is connected to its d.c. supply through an audio impedance. The circuit is shown in Fig. 10-7. The choke coil L_1 is the audio impedance in the screen circuit; its inductance should be large enough to have a reactance (at the lowest desired audio frequency) that is not less than the impedance of the screen. The latter can be taken to be approximately equal to the d.c. screen voltage divided by the d.c. screen current.

Choke-Coupled Modulator

One of the oldest types of modulation system is the choke-coupled Class A modulator shown in Fig. 10-8. Because of the relatively low power output and plate efficiency of a Class A amplifier, the method is seldom used now except for a few special applications. The audio power output of the modulator is combined with the d.c. power in the plate circuit, just as in the case of the transformer-coupled modulator. However, there is considerably less freedom in adjustment, since no transformer is available for matching impedances.

The modulating impedance of the r.f. amplifier must be adjusted to the value of load impedance required by the particular modulator tube used, and the power input to the r.f. stage must not exceed twice the rated a.f. power output of the modulator. A complication is the fact that the plate voltage on the modulator must be higher than the plate voltage on the r.f. amplifier, for 100 per cent modulation. This is because the a.f.

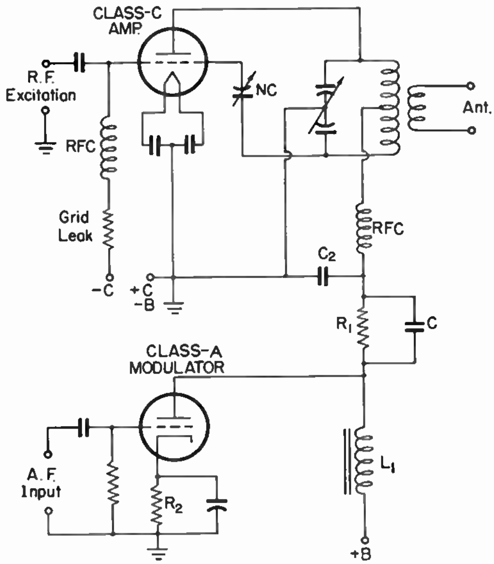


Fig. 10-8 — Choke-coupled Class A modulator. The cathode resistor, R_2, should have the normal value for operation of the modulator tube as a Class A power amplifier. The modulation choke, L_1, should be 5 henrys or more. A value of 0.001 to 0.005 μfd. is satisfactory at C_2, the r.f. amplifier plate by-pass condenser. See text for discussion of C_1 and R_1.

voltage developed by the modulator cannot swing to zero without a great deal of distortion. R_1 provides the necessary d.c. voltage drop between the modulator and r.f. amplifier, but its value cannot be calculated without using the published plate family of curves for the modulator tube used. The voltage drop through R_1 must equal the minimum instantaneous plate voltage on the modulator tube under normal operating conditions. C_1, an audio-frequency by-pass across R_1, should have a capacitance such that its reactance at 100 cycles is not more than about one-tenth the resistance of R_1. Without R_1C_1 the percentage of modulation is limited to 70 to 80 per cent in the average case.

● GRID MODULATION

The principal disadvantage of plate modulation is that a considerable amount of audio power is required. This requirement can be avoided by applying the modulation to a grid element in the modulated amplifier. However, the convenience and economy of the low-power modulator must be paid for, since no modulation system gives something for nothing. The increased power output that accompanies modulation is paid for, in the case of grid modulation, by a reduction in the carrier power output obtainable from a given r.f. amplifier tube, and by more rigorous operating requirements and more complicated adjustment.

The term "grid modulation" as used here applies to all types — control grid, screen, or suppressor — since the operating principles are exactly the same no matter which grid is actually modulated. With grid modulation the plate voltage is constant, and the increase in power output with modulation is obtained by making both the plate current and plate efficiency vary with the modulating signal as shown in Fig. 10-9. For 100 per cent modulation, both plate current and efficiency must, at the peak of the modulation up-swing, be twice their carrier values. Thus at the modulation peak the power input is doubled, and since the plate efficiency also is doubled at the same instant the peak output power will be four times the carrier power. The efficiency obtainable at the peak depends on how carefully the modulated amplifier is adjusted, and sometimes can be as high as 80 per cent. It is generally less when the amplifier is adjusted for good linearity, and under average conditions a round figure of ⅔, or 66 per cent, is representative. Since the carrier efficiency is only half the peak efficiency, the efficiency for carrier conditions, without modulation, is only about 33 per cent. Thus the carrier output is about one-fourth the power obtainable from the same tube in c.w. operation, and about one-third the carrier output obtainable from the tube with plate modulation.

The modulator is required to furnish only the audio power dissipated in the modulated grid under the operating conditions chosen. A speech amplifier capable of delivering 3 to 10 watts is usually sufficient.

Generally speaking, grid modulation does not give as linear a modulation characteristic as plate modulation, even under optimum operating conditions. When misadjusted the nonlinearity may be severe, resulting in bad distortion and splatter. However, with careful adjustment it is capable of quite satisfactory results.

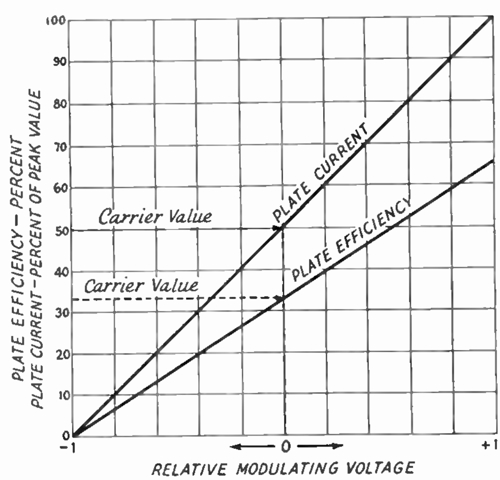


Fig. 10-9 — In a perfect grid-modulated amplifier both plate current and plate efficiency would vary with the instantaneous modulating voltage as shown. When this is so the modulation characteristic is as given by curve *A* in Fig. 10-4, and the peak output power is four times the unmodulated carrier power. The variations in plate current with modulation, indicated above, do not register on a d.c. meter, so the plate meter shows no change when the signal is modulated.

Plate-Circuit Operating Conditions

The d.c. plate power input to the modulated amplifier, assuming a round figure of ⅓ (33 per cent) for the plate efficiency, should not exceed 1½ times the plate dissipation rating of the tube or tubes used in the modulated stage. It is generally best to use the maximum plate voltage permitted by the manufacturer's ratings, because the optimum operating conditions are more easily achieved with high plate voltage and the linearity also is improved.

Example: Two tubes having plate dissipation ratings of 55 watts each are to be used with grid modulation.

The maximum permissible power input, at 33% efficiency, is

$P = 1.5 \times (2 \times 55) = 1.5 \times 110 = 165$ watts

The maximum recommended plate voltage for these tubes is 1500 volts. Using this figure, the average plate current for the two tubes will be

$$I = \frac{P}{E} = \frac{165}{1500} = 0.11 \text{ amp.} = 110 \text{ ma.}$$

At 33% efficiency, the carrier output to be expected is 55 watts.

The plate-voltage/plate-current ratio at *twice* carrier plate current is

$$\frac{1500}{220} = 6.8$$

The tank-circuit L/C ratio should be chosen on the basis of *twice* the average or carrier plate current. If the L/C ratio is based on the plate voltage/plate current ratio under carrier conditions the Q may be too low for good coupling to the output circuit.

Control-Grid Modulation

Control-grid modulation may be used with any type of r.f. amplifier tube. A typical triode circuit is given in Fig. 10-10. The same circuit can be used with screen-grid tubes merely by supplying the normal value of screen voltage by any convenient means; however, the screen should be by-passed for audio (1 μfd. or more) as well as radio frequencies. The audio signal is inserted, by means of transformer T, in series with the grid-bias lead. In a push-pull amplifier the transformer is connected in the common bias lead.

In control-grid modulation the d.c. grid bias is the same as in normal Class C amplifier service, but the r.f. grid excitation is somewhat smaller. The audio voltage superimposed on the d.c. bias changes the instantaneous grid bias at an audio rate, thus varying the operating conditions in the grid circuit and controlling the output and efficiency of the amplifier.

The change in instantaneous bias voltage with modulation causes the rectified grid current of the amplifier to vary, which places a variable load on the modulator. To reduce distortion, resistor R in Fig. 10-10 is connected in the output circuit of the modulator as a constant load, so that the over-all load variations will be minimized. This resistor should be equal to or somewhat higher than the load into which the modulator tube is rated to work at normal audio output. It is also recommended that the modulator circuit incorporate as much negative feed-back as possible, as a further aid in reducing the internal resistance of the modulator and thus improving the "regulation" — that is, reducing the effect of load variations on the audio output voltage. The turns ratio of transformer T should be about 1 to 1 in most cases.

The load on the r.f. driving stage also varies with modulation. This in turn will cause the excitation voltage to vary which may cause the modulation characteristic to be nonlinear. To overcome it, the driver should be capable of two or three times the r.f. power output actually required to drive the amplifier. The excess power may be dissipated in a dummy load (such as an incandescent lamp of appropriate power rating) that then performs the same function in the r.f. circuit that resistor R does in the audio circuit.

The d.c. bias source in this system should have low internal resistance. Batteries or a voltage-regulated supply are suitable. Grid-leak bias should not be used.

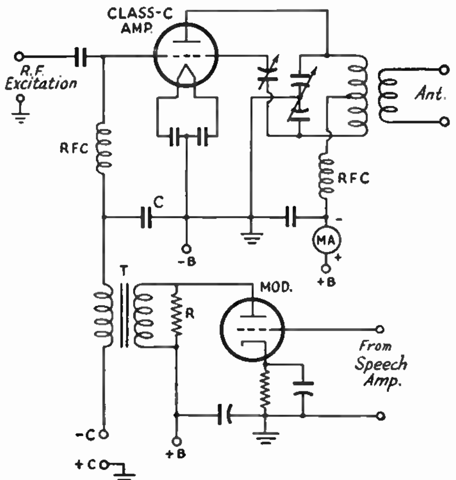


Fig. 10-10 — Control-grid modulation of a Class C amplifier. The r.f. grid by-pass condenser, C, should have high reactance at audio frequencies (0.005 μfd. or less).

Adjustment

A control-grid modulated amplifier should be adjusted with the aid of an oscilloscope connected as shown in Fig. 10-11. A tone source for modulating the transmitter is a convenience, since a steady tone will give a steady pattern on the oscilloscope. A steady pattern is easier to study than one that flickers with voice modulation.

Having determined the permissible carrier plate current as previously described, apply r.f. excitation and plate voltage and, without modulation, adjust the plate loading to give the required plate current (keeping the plate tank circuit tuned to resonance). Next, apply modulation and increase the modulating voltage until the modulation characteristic shows curvature (see later section in this chapter for use of the oscilloscope). If curvature occurs well below 100 per cent modulation, the plate efficiency is too

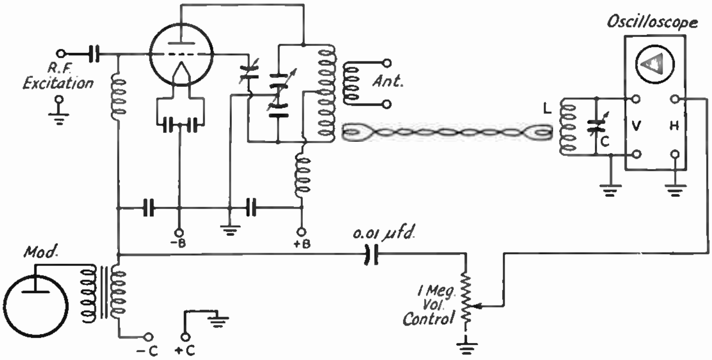


Fig. 10-11 — Using the oscilloscope for adjustment of a grid-modulated amplifier. The connections shown are for grid-bias modulation. With screen or suppressor modulation the connection to the horizontal plates of the 'scope should be taken from the grid being modulated; the r.f. pick-up arrangement remains unchanged. *L* and *C* should tune to the operating frequency, and may be coupled to the transmitter tank circuit through a twisted pair or coax, using single-turn links at each end. The 0.01-µfd. blocking condenser that couples the audio voltage to the horizontal plates of the oscilloscope should have a voltage rating equal to at least twice the d.c. voltage on the grid that is being modulated.

high. Increase the plate loading slightly and reduce the excitation to maintain the same plate current; then apply modulation and check the characteristic again. Continue this process until the characteristic is as linear as possible from the horizontal axis to twice the carrier amplitude.

Screen Modulation

Power tubes of the beam tetrode type have very good modulation characteristics when the modulating voltage is superimposed on the d.c. screen-grid voltage. The efficiency and plate current should vary with the modulating voltage as shown in Fig. 10-9.

In many ways screen modulation is more satisfactory than control-grid modulation, since the system does not require a fixed-bias supply for the control grid, and is not highly critical as to excitation voltage. However, the operating principles are identical, and the carrier output is limited to about one-half the plate dissipation rating of the tube or tubes used in the modulated amplifier.

The most satisfactory way to apply the modulating voltage to the screen is through a transformer, as shown in Fig. 10-12. In an ideal beam tetrode the plate current and output should be completely cut off with zero screen voltage, but in practical tubes it is necessary to drive the screen somewhat negative with respect to the cathode to get complete cut-off. For this reason the peak modulating voltage required for 100 per cent modulation is usually 10 per cent or so greater than the d.c. screen voltage. The latter, in turn, is approximately half the rated screen voltage under maximum ratings for c.w. operation.

The audio power required is approximately one-fourth the d.c. power input to the screen under c.w. operation, but varies somewhat with the operating conditions. A receiving-type audio power amplifier will suffice as the modulator for most transmitting tubes. Because the relationship between screen voltage and screen current is not linear (a typical curve giving this relationship is shown in Fig. 10-13) the load on the modulator varies over the audio-frequency cycle, and it is therefore highly advisable to use negative feed-back in the modulator circuit. If excess audio power is available, it is also advisable to load the modulator with a resistance corresponding to *R* in Fig. 10-10, the value of *R* being adjusted to dissipate the excess power. Unfortunately, there is no simple way to determine the proper resistance except experimentally, by observing the effect of different values on the waveshape with the aid of an oscilloscope.

On the assumption that the modulator will be fully loaded by the screen plus the additional load resistor *R*, the turns ratio required in the

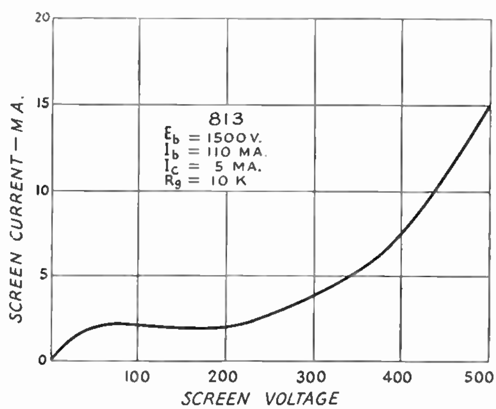


Fig. 10-13 — A typical screen voltage-current curve of a beam tetrode adjusted for optimum conditions for screen modulation.

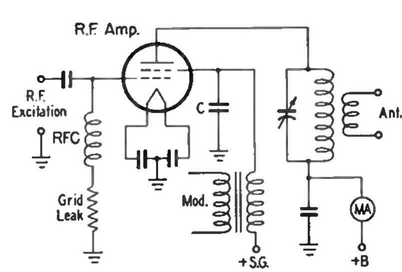


Fig. 10-12 — Screen-grid modulation of beam tetrode. Condenser *C* is an r.f. by-pass condenser and should have high reactance at audio frequencies. A value of 0.002 µfd. is satisfactory. The grid leak can have the same value that is used for c.w. operation of the tube.

coupling transformer may be calculated as follows:

$$N = \frac{E_d}{2.5\sqrt{PR_L}}$$

where N is the turns ratio, secondary to primary; E_d is the rated screen voltage for c.w. operation; P is the rated audio power output of the modulator; and R_L is the rated load resistance for the modulator.

The best method of adjustment is to use an oscilloscope (the connections of Fig. 10-11 may be used, except that the audio sweep voltage is taken from the screen instead of the control grid) and adjust plate loading, grid excitation, and modulating voltage for the greatest output compatible with good linearity at 100 per cent modulation. The amplifier should be loaded heavily and the grid current should be kept at the point where a further reduction decreases the r.f. output. Under proper operating conditions the plate-current dip as the amplifier plate circuit is tuned through resonance will be little more than just discernible.

In an alternative adjustment method not requiring an oscilloscope the r.f. amplifier is first tuned up for maximum output without modulation and the rated d.c. screen voltage (from a fixed-voltage supply) for c.w. operation applied. Use heavy loading and reduce the grid excitation until the output just starts to fall off, at which point the resonance dip in plate current should be small. Note the plate current and, if possible, the r.f. antenna or feeder current, and then reduce the d.c. screen voltage until the plate current is one-half its previous value. The r.f. output current should also be one-half its previous value at this screen voltage. The amplifier is then ready for modulation, and the modulating voltage may be increased until the plate current just starts to shift upward, which indicates that the amplifier is modulated 100 per cent. With voice modulation the plate current should remain steady, or show just an occasional small upward kick on intermittent peaks.

It is desirable to operate with the grid current as low as possible, since this reduces the screen current and thus reduces the amount of power required from the modulator. With proper adjustment the linearity is good up to about 90 per cent modulation. When the screen is driven negative for 100 per cent modulation there is a kink in the modulation characteristic at the zero-voltage point that introduces a small amount of distortion. The kink can be removed and the over-all linearity improved by applying a small amount of modulating voltage to the control grid simultaneously with screen modulation, but this requires adjustment with the oscilloscope.

"Clamp-Tube" Modulation

A method of screen-grid modulation that is convenient in transmitters provided with a screen protective tube ("clamp" tube) is shown in Fig. 10-14. Basically, the idea is that an audio-frequency signal is applied to the grid of the clamp tube, which then becomes a modulator. The simplicity of the circuit is somewhat deceptive, since it is considerably more difficult from a design standpoint than the transformer-coupled arrangement of Fig. 10-12.

For proper modulation the clamp tube must be operated as a triode Class A amplifier, and it will be recognized that the method is essentially identical with the choke-coupled Class A plate modulator of Fig. 10-8 with a resistance, R_2, substituted for the choke. R_2 in the usual case is the screen dropping resistor normally used for c.w. operation. Its value should be at least two or three times the load resistance required by the Class A modulator tube for optimum audio-frequency output. Unfortunately, relatively little information is available on the triode operation of the tubes most frequently used for screen-protective purposes.

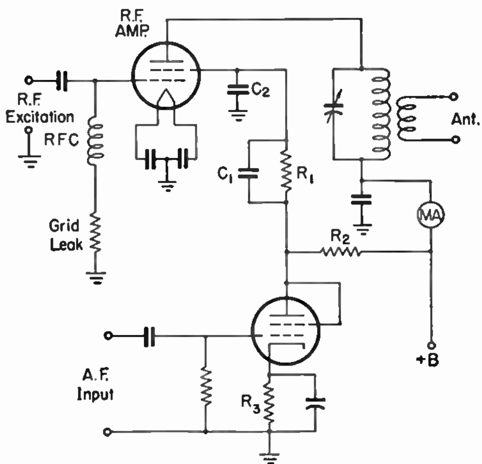


Fig. 10-14 — Screen modulation by a "clamp" tube. The grid leak is the normal value for c.w. operation and C_2 should be 0.002 μfd. or less. See text for discussion of C_1, R_1, R_2 and R_3. R_3 should have the proper value for Class A operation of the modulator tube, but cannot be calculated unless triode curves for the tube are available.

Like the choke-coupled modulator, the clamp-tube modulator is incapable of modulating the r.f. stage 100 per cent unless the dropping resistor, R_1, and audio by-pass, C_1, are incorporated in the circuit. The same design considerations hold, with the addition of the fact that the screen must be driven negative, not just to zero voltage, for 100 per cent modulation. The modulator tube must thus be operated at a voltage ranging from 20 to 40 per cent higher than the screen that it modulates. Proper design requires knowledge of the screen characteristics of the r.f. amplifier and a set of plate-voltage plate-current curves on the modulator tube as a triode.

Adjustment with this system, once the design voltages have been determined, is carried out in the same way as with transformer-coupled screen modulation, preferably with the oscilloscope. Without the oscilloscope, the amplifier may first be adjusted for c.w. operation as described earlier, but with the modulator tube removed from its socket. The modulator is then replaced, and the cathode resistance, R_3, adjusted to reduce the amplifier plate current to one-half its c.w. value. The amplifier plate current should remain constant with modulation, or show just a small upward flicker on occasional voice peaks.

Controlled Carrier

As explained earlier, a limit is placed on the output obtainable from a grid-modulation system by the low r.f. amplifier plate efficiency (approximately 33 per cent) under unmodulated carrier

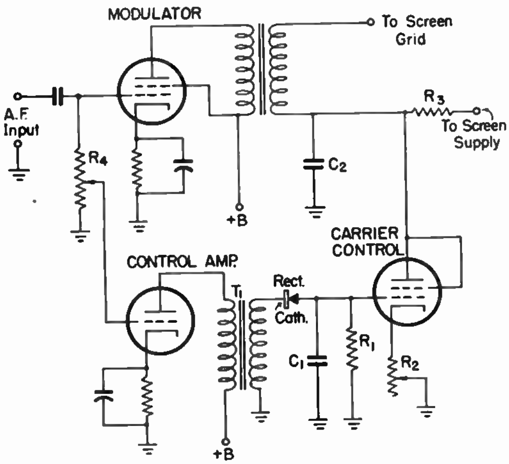


Fig. 10-15 — **Circuit for carrier control with screen modulation. A small triode such as the 6J5 can be used as the control amplifier and a 6Y6G is suitable as a carrier-control tube. T_1 is an interstage audio transformer having a 1-to-1 or larger turns ratio. R_4 is a 0.5-megohm volume control and also serves as the grid resistor for the modulator. A germanium crystal may be used as the rectifier. Other values are discussed in the text.**

conditions. The plate efficiency increases with modulation, since the output increases while the d.c. input remains constant, and reaches a maximum in the neighborhood of 50 per cent with 100 per cent sine-wave modulation. If the power input to the amplifier can be reduced during periods when there is little or no modulation, thus reducing the plate loss, advantage can be taken of the higher efficiency at full modulation to obtain higher effective output. This can be done by varying the power input to the modulated stage, in accordance with *average* variations in voice intensity, in such a way as to maintain just sufficient carrier power to keep the modulation high, but not exceeding 100 per cent, under all conditions. Thus the carrier amplitude is controlled by the voice intensity. Properly utilized, controlled carrier permits increasing the effective carrier output at maximum level to a value equal to the rated plate dissipation of the tube, or twice the output obtainable with constant carrier.

It is desirable to control the power input just enough so that the plate loss, without modulation, is safely below the tube rating. Excessive control is disadvantageous because the receiver's a.v.c. system must continually follow the variations in average signal level. The circuit of Fig. 10-15 permits adjustment of both the maximum and minimum power input, and although somewhat more complicated than some circuits that have been used is actually simpler to operate because it separates the functions of modulation and carrier control. A portion of the audio voltage at the modulator grid is applied to a Class A "control amplifier" which drives a rectifier circuit to produce a d.c. voltage negative with respect to ground. C_1 filters out the audio variations, leaving a d.c. voltage proportional to the average voice level. This voltage is applied to the grid of a "clamp" tube to control the d.c. screen voltage and thus the r.f. carrier level. Maximum output is obtained when the carrier-control tube grid is driven to cut-off, the voice level at which this occurs being determined by the setting of R_4. Minimum input is set to the desired level (usually about equal to the plate dissipation rating of the modulated stage) by adjusting R_2. R_3 may be the normal screen-dropping resistor for the modulated beam tetrode, but in case a separate screen supply is used it need be just large enough to give sufficient voltage drop to reduce the no-modulation power input to the desired value.

C_1R_1 should have a time constant of about 0.1 second. The time constant of C_2R_3 should be no larger. Further details may be found in *QST* for April, 1951, page 64. An oscilloscope is required for proper adjustment.

Suppressor Modulation

Pentode-type tubes do not, in general, modulate well when the modulating voltage is applied to the screen grid. However, a satisfactory modulation characteristic can be obtained by applying the modulation to the suppressor grid. The circuit arrangement for suppressor-grid modulation of a pentode tube is shown in Fig. 10-16.

The method of adjustment closely resembles that used with screen-grid modulation. If an oscilloscope is not available, the amplifier is first adjusted for optimum c.w. output with zero bias on the suppressor grid. Negative bias is then applied to the suppressor and increased in value until the plate current and r.f. output current drop to half their original values. When this condition has been obtained the amplifier is ready for modulation.

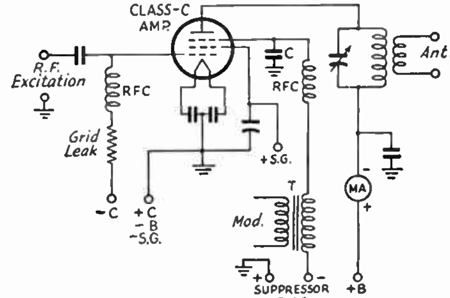


Fig. 10-16 — **Suppressor-grid modulation of an r.f. amplifier using a pentode-type tube. The suppressor-grid r.f. by-pass condenser, C, should be the same as the grid by-pass condenser in control-grid modulation.**

Since the suppressor is always negatively biased, the modulator is not required to furnish any power, so a voltage amplifier can be used. The suppressor bias will vary with the type of pentode and the operating conditions, but usually will be of the order of −100 volts. The peak a.f. voltage required from the modulator is equal to the suppressor bias.

CATHODE MODULATION

Circuit

The fundamental circuit for cathode modulation is shown in Fig. 10-17. It is a combination of the plate and grid methods, and permits a carrier efficiency midway between the two. The audio power is introduced in the cathode circuit, and both grid bias and plate voltage are modulated.

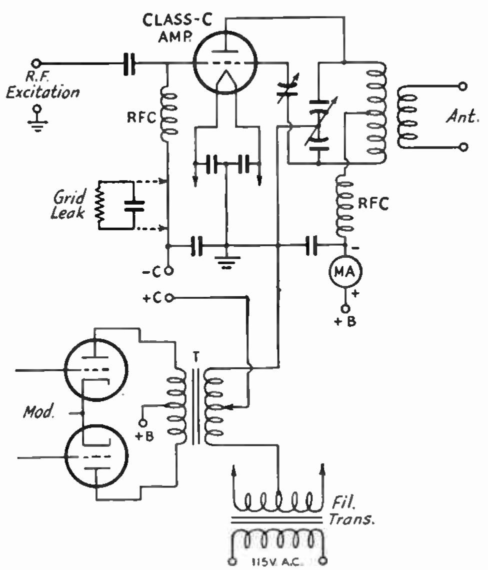


Fig. 10-17 — **Circuit arrangement for cathode modulation of a Class C r.f. amplifier. Values of by-pass condensers in the r.f. circuits should be the same as for other modulation methods.**

Because part of the modulation is by the control-grid method, the plate efficiency of the modulated amplifier must vary during modulation. The carrier efficiency therefore must be lower than the efficiency at the modulation peak. The required reduction in efficiency depends upon the proportion of grid modulation to plate modulation; the higher the percentage of plate modulation, the higher the permissible carrier efficiency, and vice versa. The audio power required from the modulator also varies with the percentage of plate modulation, being greater as this percentage is increased.

The way in which the various quantities vary is illustrated by the curves of Fig. 10-18. In these curves the performance of the cathode-modulated r.f. amplifier is plotted in terms of the tube ratings for plate-modulated telephony, with the percentage of plate modulation as a base.

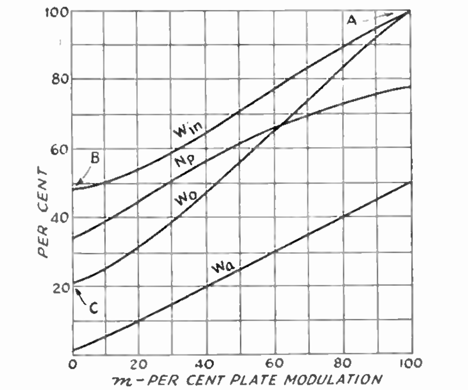


Fig. 10-18 — **Cathode-modulation performance curves, in terms of percentage of plate modulation plotted against percentage of Class C telephony tube ratings.**

W_{in} — D.c. plate input watts in terms of percentage of plate-modulation rating.
W_o — Carrier output watts in per cent of plate-modulation rating (based on plate efficiency of 77.5%).
W_a — Audio power in per cent of d.c. watts input.
N_p — Plate efficiency of the amplifier in percentage.

As the percentage of plate modulation is decreased, it is assumed that the grid modulation is increased to make the over-all modulation reach 100 per cent. The limiting condition, 100 per cent plate modulation and no grid modulation, is at the right (*A*); pure grid modulation is represented by the left-hand ordinate (*B* and *C*).

Example: Assume that the r.f. tube to be used has a 100% plate-modulation rating of 250 watts input and will give a carrier power output of 190 watts at that input. Cathode modulation with 40% plate modulation is to be used. From Fig. 10-18, the carrier efficiency will be 56% with 40% plate modulation, the permissible d.c. input will be 65% of the plate-modulation rating, and the r.f. output will be 48% of the plate-modulation rating. That is,

Power input = 250 × 0.65 = 162.5 watts
Power output = 190 × 0.48 = 91.2 watts

The required audio power, from the chart, is equal to 20% of the d.c. input to the modulated amplifier. Therefore

Audio power = 162.5 × 0.2 = 32.5 watts

The modulator should supply a small amount of extra power to take care of losses in the grid circuit. These should not exceed four or five watts.

Modulating Impedance

The modulating impedance of a cathode-modulated amplifier is approximately equal to

$$m\frac{E_b}{I_b}$$

where m = Percentage of plate modulation (expressed as a decimal)
E_b = D.c. plate voltage on modulated amplifier
I_b = D.c. plate current of modulated amplifier

Example: Assume that the modulated amplifier in the example above is to operate at a plate potential of 1250 volts. Then the d.c. plate current is

$$I = \frac{P}{E} = \frac{162.5}{1250} = 0.13 \text{ amp. (130 ma.)}$$

The modulating impedance is

$$m\frac{E_b}{I_b} = 0.4\frac{1250}{0.13} = 3846 \text{ ohms}$$

The modulating impedance is the load into which the modulator must work, just as in the case of pure plate modulation. This load must be matched to the load required by the modulator tubes by proper choice of the turns ratio of the modulation transformer, as described in the chapter on speech equipment.

Conditions for Linearity

R.f. excitation requirements for the cathode-modulated amplifier are midway between those for plate modulation and control-grid modulation. More excitation is required as the percentage of plate modulation is increased. Grid bias should be considerably beyond cut-off; fixed bias from a supply having good voltage regulation is preferred, especially when the percentage of plate modulation is small and the amplifier is operating more nearly like a grid-bias modulated stage. At the higher percentages of plate modulation a combination of fixed and grid-leak bias can be used, since the variation in rectified grid current is smaller. The grid leak should be by-passed for audio frequencies. The percentage of grid modulation may be regulated by choice of a suitable tap on the modulation-transformer secondary.

The cathode circuit of the modulated stage must be independent of other stages in the transmitter. When directly-heated tubes are modulated their filaments must be supplied from a separate transformer. The filament by-pass condensers should not be larger than about 0.002 μfd., to avoid by-passing the audio-frequency modulation.

Adjustment of Cathode-Modulated Amplifiers

In most respects, the adjustment procedure is similar to that for grid-bias modulation. The critical adjustments are antenna loading, grid bias, and excitation. The proportion of grid-bias to plate modulation will determine the operating conditions.

Adjustments should be made with the aid of an oscilloscope connected in the same way as for grid-bias modulation. With proper antenna loading and excitation, the normal wedge-shaped pattern will be obtained at 100 per cent modulation. As in the case of grid-bias modulation, too-light antenna loading will cause flattening of the upward peaks of modulation as also will too-high excitation. The cathode current will be practically constant with or without modulation when the proper operating conditions have been established.

Checking AM 'Phone Operation

● USING THE OSCILLOSCOPE

Proper adjustment of a 'phone transmitter is aided immeasurably by the oscilloscope. The 'scope will give more information, more accurately, than almost any collection of other instruments that might be named. Furthermore, an oscilloscope that is entirely satisfactory for the purpose is not necessarily an expensive instrument; the cathode-ray tube and its power supply are about all that are needed. Amplifiers and linear sweep circuits are by no means necessary.

In the simplest 'scope circuit, radio-frequency voltage from the modulated amplifier is applied directly to the vertical deflection plates of the tube, and audio-frequency voltage from the modulator is applied to the horizontal deflection plates. As the instantaneous amplitude of the audio signal varies, the r.f. output of the transmitter likewise varies, and this produces a wedge-shaped pattern or **trapezoid** on the screen. If the oscilloscope has a built-in horizontal sweep, the r.f. voltage is applied to the vertical plates as before (never through an amplifier) and the sweep will produce a pattern that follows the modulation envelope of the transmitter output, provided the sweep frequency is lower than the modulation frequency. This produces a **wave-envelope** modulation pattern.

The Wave-Envelope Pattern

The connections for the wave-envelope pattern are shown in Fig. 10-19A. The vertical deflection plates are coupled to the amplifier tank coil (or an antenna coil) through a twisted-pair line and pick-up coil. As shown in the alternative drawing, a resonant circuit tuned to the operating frequency may be connected to the vertical plates, using link coupling between it and the transmitter. This will eliminate r.f. harmonics, and the tuning control provides a convenient means for adjustment of the pattern height.

The position of the pick-up coil should be varied until an unmodulated carrier pattern, Fig. 10-20B, of suitable height is obtained. The horizontal sweep voltage should be adjusted to make the width of the pattern somewhat more than half the diameter of the screen. When voice modulation is applied, a rapidly-changing pattern of varying height will be obtained. When the maximum height of this pattern is just twice that of the carrier alone, the wave is being modulated 100 per cent. This is illustrated by Fig. 10-20D, where the point *X* represents the horizontal sweep line (reference line) alone, *YZ* is the carrier height, and *PQ* is the maximum height of the modulated wave.

If the height is greater than the distance *PQ*, as illustrated in *E*, the wave is overmodulated in the upward direction. Overmodulation in the downward direction is indicated by a gap in the pattern at the reference axis, where a single bright line appears on the screen. Overmodulation in either direction may take place even when the modulation in the other direction is less than 100 per cent.

The Trapezoidal Pattern

Connections for the trapezoid or wedge pattern as used for checking plate modulation are shown in Fig. 10-19B. The vertical plates of the c.r. tube are coupled to the transmitter tank through

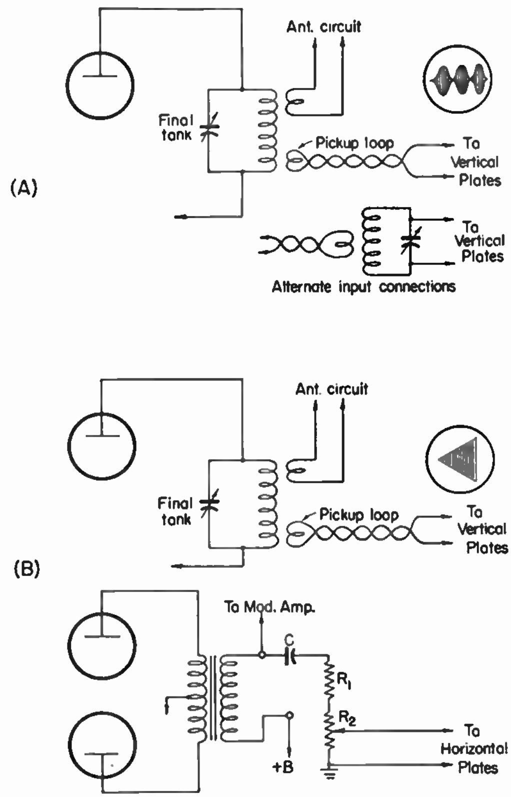


Fig. 10-19 — **Methods of connecting the oscilloscope for modulation checking. A — connections for wave-envelope pattern with any modulation method; B — connections for trapezoidal pattern with plate modulation. See Fig. 10-11 for 'scope connections for trapezoidal pattern with grid modulation.**

a pick-up loop, preferably using a tuned circuit, as shown in the upper drawing, adjustable to the operating frequency. Audio voltage from the modulator is applied to the horizontal plates through a voltage divider, R_1R_2. This voltage should be adjustable so a suitable pattern width can be obtained; a 0.25-megohm volume control can be used at R_2 for this purpose, with c.r. tubes up to the 3-inch size.

The resistance required at R_1 will depend on the d.c. plate voltage on the modulated amplifier. The total resistance of R_1 and R_2 in series should be about 0.25 megohm for each 100 volts of d.c. plate voltage. For example, if the modulated amplifier operates at 1500 volts, the total resistance should be 3.75 megohms, 0.25 megohm at R_2 and the remainder, 3.5 megohms, in R_1. R_1 should be composed of individual resistors not larger than 0.5 megohm each, in which case 1-watt resistors will be satisfactory.

For good low-frequency coupling the capacitance, in microfarads, of the blocking condenser, C, should at least equal $0.004/R$, where R is the total resistance ($R_1 + R_2$) in megohms. In the example above, where R is 3.75 megohms, the capacitance should be at least $0.004/3.75 = 0.001$ μfd., approximately. The voltage rating of the condenser should be at least twice the d.c. voltage applied to the modulated amplifier. The capacitance can be made up of two or more similar units in series, so long as the total capacitance is equal to that required, in case a single unit of sufficient voltage rating is not available. Two or more units may be used in parallel if condensers having adequate voltage rating but insufficient capacitance are available.

The corresponding 'scope connections for grid modulation were given in Fig. 10-11. This circuit will be satisfactory for checking screen-grid modulation (the audio connection of course being made to the screen grid rather than to the control grid) for d.c. screen voltages up to 200 volts or so, which will include most beam tetrodes. If the d.c. screen voltage, adjusted for proper modulation, exceeds 200 volts a voltage divider similar to that shown in Fig. 10-19 should be used, the values being calculated as described above using the screen voltage instead of the plate voltage.

Trapezoidal patterns for various conditions of modulation are shown in Fig. 10-20 at F to J, each alongside the corresponding wave-envelope pattern. With no signal, only the cathode-

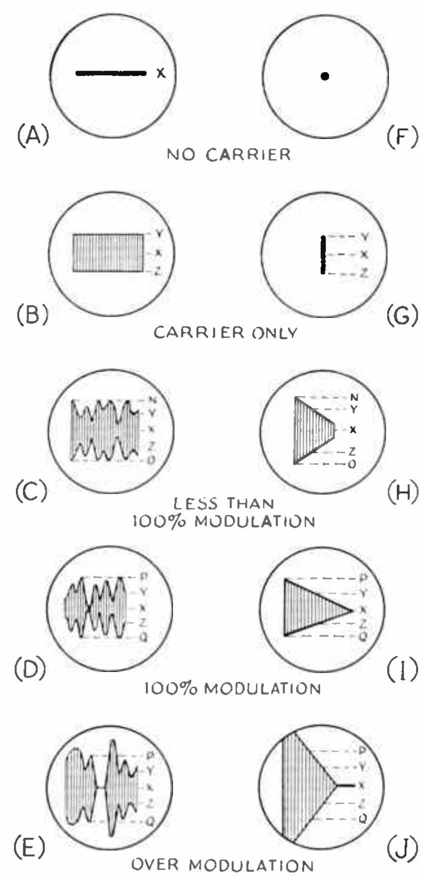


Fig. 10-20 — **Wave-envelope and trapezoidal patterns representing different conditions of modulation.**

ray spot appears on the screen. When the unmodulated carrier is applied, a vertical line appears; the length of the line should be adjusted, by means of the pick-up coil coupling, to a convenient value. When the carrier is modulated, the wedge-shaped pattern appears; the higher the modulation percentage, the wider and more pointed the wedge becomes. At 100 per cent modulation it just makes a point on the axis, X, at one end, and the height, PQ, at the other end is equal to twice the carrier height, YZ. Overmodulation in the upward direction is indicated by increased height over PQ, and in the downward direction by an extension along the axis X at the pointed end.

Checking Transmitter Performance

The trapezoidal pattern is far more useful than the wave-envelope pattern for checking the operation of a 'phone transmitter. The latter type of pattern is of use principally for checking modulation percentage, and even when the speech system is fed with a sine-wave tone for close examination of the pattern it is difficult to tell with sufficient accuracy whether the transmitter is operating linearly. Also, even when distortion is evident in the wave-envelope pattern there is no clue as to whether it is occurring in the modulated amplifier or is caused by a defect in the speech equipment.

On the other hand, the trapezoidal pattern is actually a graph of the modulation characteristic of the modulated amplifier. The sloping sides of the wedge show the r.f. amplitude for every value of instantaneous modulating voltage, exactly the type of curve plotted in Fig. 10-4. If these sides are perfectly straight lines, as drawn in Fig. 10-20 at H and I, the modulation characteristic is linear. If the sides show curvature, the characteristic is nonlinear to an extent that is shown by the degree to which the sides depart from perfect straightness. This is true regardless of the waveform of the modulating voltage.

If the speech system can be driven by a good audio sine-wave signal instead of a microphone, the trapezoidal pattern also will show the presence of even-harmonic distortion (the most common type, especially when the modulator is overloaded) in the speech amplifier or modulator. If there is no distortion in the audio system, the trapezoid will extend horizontally equal distances on each side of the vertical line representing the unmodulated carrier. If there is even-harmonic distortion the trapezoid will extend farther to one side of the unmodulated-carrier position than to the other. This is shown in Fig. 10-21. The probable cause is inadequate power output from the modulator, or incorrect load on the modulator.

An audio oscillator having reasonably good sine-wave output is highly desirable for testing both speech equipment and the 'phone transmitter as a whole. A very simple single-tone oscillator such as is shown in the chapter on measurements is quite adequate. With such an oscillator and the 'scope, the pattern is steady and can be studied closely to determine the effects of various operating adjustments.

The patterns shown in Figs. 10-21 and the top four groups of Fig. 10-22 show both correct and incorrect transmitter adjustments. The object of modulated-amplifier adjustment is to obtain a pattern closely resembling that in Fig. 10-22A, which shows excellent linearity (sides of wedge pattern quite straight) over the whole characteristic at 100 per cent modulation. Since no modulated amplifier is perfect, the sides will never be *perfectly* straight, but a close approach is possible. Different methods of modulation give different characteristic results. Fig. 10-22A is typical of correctly-operated plate modulation. With control-grid modulation the sides usually are somewhat concave, particularly near the point of the trapezoid, while screen modulation gives the characteristic pattern shown in Fig. 10-21. As mentioned earlier, it is necessary to drive the screen somewhat negative in order to reach complete plate-current cut-off and thus modulate 100 per cent downward.

Aside from overmodulation downward, Fig.

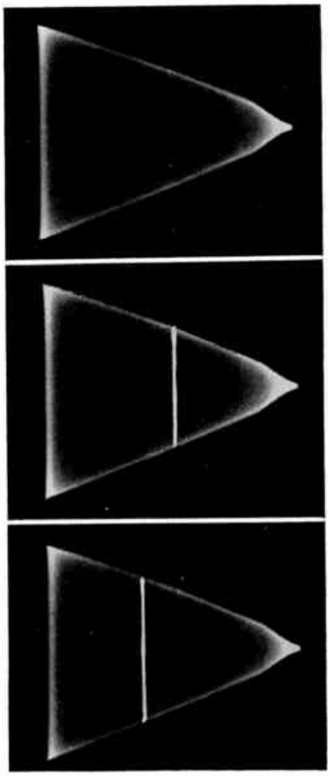
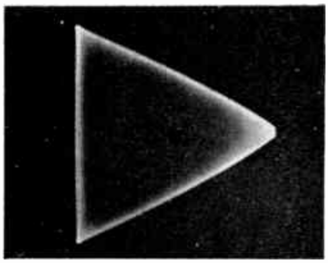
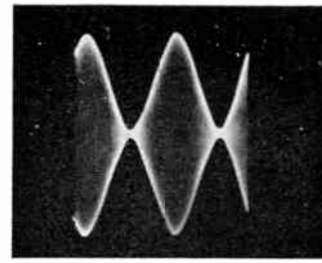


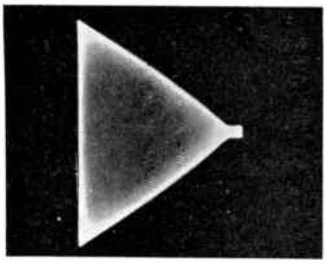
Fig. 10-21 — *Top* — a typical trapezoidal pattern obtained with screen modulation adjusted for optimum conditions. The sudden change in slope near the point of the wedge occurs when the screen voltage passes through zero. *Center* — If there is no audio distortion, the unmodulated carrier will have the height and position shown by the white line superimposed on the sine-wave modulation pattern. *Bottom* — Even-harmonic distortion in the audio system, when the audio signal applied to the speech amplifier is a sine wave, is indicated by the fact that the modulation pattern does not extend equal distances either side of the unmodulated carrier.



A

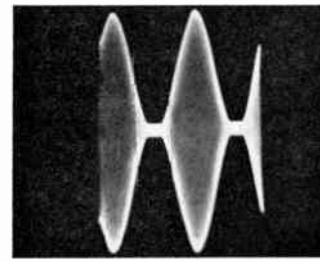
Properly-operated 'phone transmitter modulated 100 per cent.

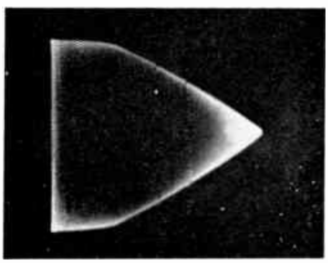




B

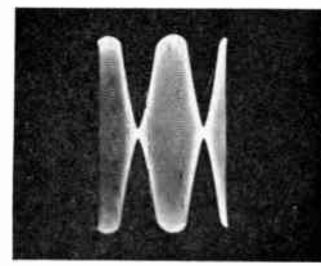
Overmodulation of a transmitter having high modulation capability. Distortion occurs only on the down-peaks.

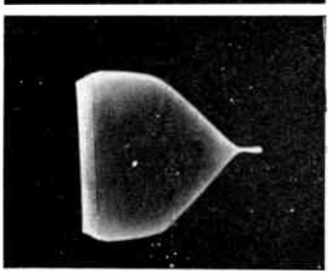




C

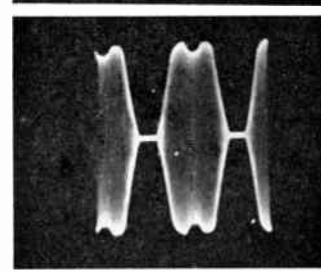
Nonlinearity in modulated r.f. stage, frequently caused by insufficient excitation of a plate-modulated amplifier or overexcitation of a grid-bias modulated amplifier. The amplifier modulates linearly in the downward direction but the up-peaks are flattened.

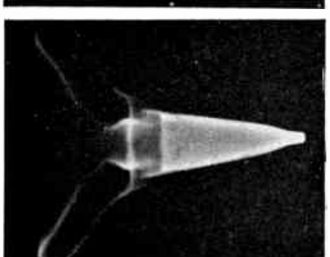




D

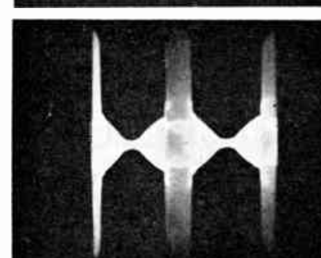
Overmodulation and nonlinear operation (insufficient modulation capability). These patterns are similar to those directly above, but with the modulation carried beyond 100 per cent in the downward direction.

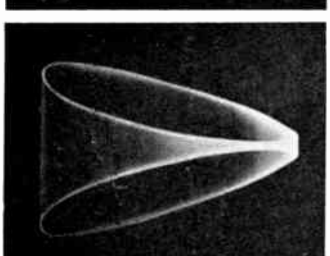




E

Overmodulation and parasitic oscillations in the modulated amplifier. The trapezoidal pattern also shows phase distortion caused by incorrect coupling between the oscilloscope and audio system.





F

Left — Phase distortion caused by incorrect coupling between audio system and oscilloscope. *Right* — Multiple pattern caused by incorrect setting of oscilloscope time-base control. In both cases the wave is modulated 100 per cent.

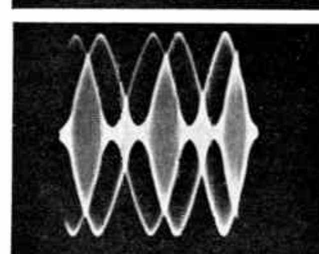


Fig. 10-22 — PHOTOGRAPHS OF TYPICAL OSCILLOSCOPE PATTERNS

These photographs show various conditions of modulation as displayed by the wedge or trapezoidal patterns in the left-hand column and the wave-envelope patterns in the right-hand column.

(Photographs reproduced through courtesy of the Allen B. DuMont Laboratories, Inc., Passaic, N. J.)

10-22B, which is easily cured by keeping the speech amplifier gain or speech intensity below the point that causes it, the most common type of improper operation is shown by the pattern of Fig. 10-22C. The flattening at the large end of the trapezoid results from the inability of the modulated amplifier to deliver sufficient power output on the modulation up-peak. With plate modulation the most likely cause is insufficient grid excitation or incorrect grid bias or both. With grid modulation this flattening is the result of attempting to operate the amplifier at too-high carrier efficiency. The remedy is to increase the loading on the output circuit and reduce the grid excitation, or both in combination, until the pattern sides are straight.

In this connection, it should be noted that while the trapezoidal pattern of Fig. 10-22C shows nonlinearity in the modulated amplifier, the corresponding wave-envelope pattern of the same figure could result *either* from this cause or from modulator overloading. With the trapezoidal pattern, modulator overloading will be evident by the fact that the position of the vertical line representing the unmodulated carrier will not be at the center of the pattern (when the modulating voltage is cut off) but modulator overloading will not affect the *shape* of the pattern. This assumes that the audio signal is a sine wave.

Curvature near the point of the trapezoid causing it to approach the horizontal axis more slowly than would occur with straight sides, indicates that the output power does not decrease rapidly enough in this region; it may be caused by r.f. leakage from the exciter through the final stage. This can be checked by removing the voltage from the modulated stage, when the carrier should disappear, leaving only the beam spot remaining on the screen (Fig. 10-20F). If a small vertical line remains, the amplifier should be carefully neutralized; if this does not eliminate the line, it is an indication that the 'scope is getting r.f. from lower-power stages, either by coupling through the final tank or via the pick-up loop.

Faulty Patterns

Figs. 10-20, 10-21, and 10-22A through D show what is normally to be expected in the way of pattern shapes when the oscilloscope is used to check modulation. If the actual patterns differ considerably from those shown, it may be that the pattern is faulty rather than the transmitter.

It is important that r.f. from the *modulated stage only* be coupled to the oscilloscope, and then only to the vertical plates. The effect of stray r.f. from other stages in the transmitter has been mentioned in the preceding section. If r.f. is present also on the horizontal plates, the pattern will lean to one side instead of being upright. If the oscilloscope cannot be moved to a position where the unwanted pick-up disappears, a small by-pass condenser (10 $\mu\mu$fd.) should be connected across the horizontal plates as close to the cathode-ray tube as possible. An r.f. choke (2.5 mh. or smaller) may also be connected in series with the ungrounded horizontal plate.

"Folded" trapezoidal patterns, and patterns in which the sides of the trapezoid are elliptical instead of straight, Fig. 10-22F (left), occur when the audio sweep voltage is taken from some point in the audio system other than that where the a.f. power is applied to the modulated stage. Such patterns are caused by a phase difference between the sweep voltage and the modulating voltage. The connections should always be as shown in Fig. 10-11 and 10-19B.

MODULATION CHECKING WITH THE PLATE METER

The plate milliammeter of the modulated amplifier provides a simple and fairly reliable means for checking the performance of a 'phone transmitter, although it does not give nearly as definite information as the oscilloscope does. If the modulated amplifier is perfectly linear, its plate current will not change when modulation is applied if

1) the upward modulation percentage does not exceed the modulation capability of the amplifier,

2) the downward modulation does not exceed 100 per cent, and

3) there is no change in the d.c. operating voltages on the transmitter when modulation is applied.

This is true of any of the methods of modulation discussed in this chapter, with the single exception of the controlled-carrier system. The plate meter cannot give a reliable check on the performance of the latter system because the plate current increases with the intensity of modulation. With this system the plate-current variations should be correlated with the transmitter performance as observed on an oscilloscope before the plate meter is used for checking modulation.

Plate Modulation

With plate modulation, a downward shift in plate current may indicate one or more of the following:

1) Insufficient excitation to the modulated r.f. amplifier.
2) Insufficient grid bias on the modulated stage.
3) The r.f. amplifier is not loaded properly to present the required value of modulating impedance to the modulator.
4) Insufficient output capacitance in the filter of the modulated-amplifier plate supply.
5) D.c. input to the r.f. amplifier, under carrier conditions, is in excess of the manufacturer's ratings for plate modulation. Alternatively, the filament emission of the amplifier tubes may be low.
6) In plate-and-screen modulation of tetrodes or pentodes, the screen is not being sufficiently modulated along with the plate. In systems in which the d.c. screen voltage is

obtained through a dropping resistor, a downward dip in plate current may occur if the screen by-pass condenser capacitance is large enough to by-pass audio frequencies.

7) Poor voltage regulation of the modulated-amplifier plate supply. This may be caused by voltage drop in the supply itself, when the modulated amplifier and a Class B amplifier are operated from the same supply, or may be caused by voltage drop in the primary supply from the power line when the modulator load is thrown on. It is readily checked by measuring the voltage with and without modulation. Poor line regulation will be shown by a drop in filament voltage with modulation.

Any of the following may cause an upward shift in plate current:

1) Overmodulation (excessive audio power, audio gain too great).
2) Incomplete neutralization of the modulated amplifier.
3) Parasitic oscillation in the modulated amplifier.

Grid Modulation

With any type of grid modulation, any of the following may cause a downward shift in modulated-amplifier plate current:

1) Too much r.f. excitation.
2) Insufficient grid bias, particularly with control-grid modulation. Grid bias is usually not critical with screen and suppressor modulation, the value of grid leak recommended for c.w. operation being satisfactory.
3) With control-grid modulation, excessive resistance in the bias supply.
4) Insufficient output capacitance in plate-supply filter.
5) Plate efficiency too high under carrier conditions; amplifier is not loaded heavily enough.

Because grid modulation is not perfectly linear (always less so than plate modulation) a properly-operating amplifier will show a small upward plate-current shift with modulation, 10 per cent or less with sine-wave modulation and amounting to an occasional upward flicker with voice. An upward plate current shift in excess of this may be caused by

1) Overmodulation (excessive modulating voltage).
2) Regeneration (incomplete neutralization).
3) With control-grid or suppressor modulation, bias too great.
4) With screen modulation, d.c. screen voltage too low.

In grid-modulation systems the modulator is not *necessarily* operating linearly if the plate current stays constant with or without modulation. It is readily possible to arrive at a set of operating conditions in which flattening of the up-peaks is just balanced by overmodulation downward, resulting in practically the same plate current as when the transmitter is unmodulated. The oscilloscope provides the only certain check on grid modulation. While the same type of improper operation is possible with plate modulation, it occurs only rarely.

COMMON TROUBLES IN THE 'PHONE TRANSMITTER

Noise and Hum on Carrier

Noise and hum may be detected by listening to the signal on a receiver, provided the receiver is far enough away from the transmitter to avoid overloading. The hum level should be low compared with the voice at 100 per cent modulation. Hum may come either from the speech amplifier and modulator or from the r.f. section of the transmitter. Hum from the r.f. section can be detected by completely shutting off the modulator; if hum remains when this is done, the power-supply filters for one or more of the r.f. stages have insufficient smoothing. With a hum-free carrier, hum introduced by the modulator can be checked by turning on the modulator but leaving the speech amplifier off; power-supply filtering is the likely source of such hum. If carrier and modulator are both clean, connect the speech amplifier and observe the increase in hum level. If the hum disappears with the gain control at minimum, the hum is being introduced in the stage or stages preceding the gain control. The microphone also may pick up hum, a condition that can be checked by removing the microphone from the circuit but leaving the first speech-amplifier grid circuit otherwise unchanged. A good ground (to a cold water pipe, for example) on the microphone and speech system usually is essential to hum-free operation.

Spurious Sidebands

A superheterodyne receiver having a crystal filter is needed for checking spurious sidebands outside the normal communication channel. The r.f. input to the receiver must be kept low enough by removing the antenna or by adequate separation from the transmitter, to avoid overloading and consequent spurious receiver responses. An "S"-meter reading of about half scale is satisfactory. With the crystal filter in its sharpest position tune through the region outside the normal channel limits (3 to 4 kilocycles each side of the carrier) while another person talks into the microphone. Spurious sidebands will be observed as intermittent "clicks" or crackles well away from the carrier frequency. Sidebands more than 3 to 4 kilocycles from the carrier should be of negligible strength, compared with the carrier in a properly-modulated 'phone transmitter. The causes are overmodulation or nonlinear operation.

With sine-wave modulation the relative intensity of sidebands can be observed if a tone of 1000 cycles or so is used, since the crystal filter readily can separate frequencies of this order. The "S"-meter will show how the spurious side frequencies (those spaced more than the modulating frequency from the carrier) compare with the carrier itself. Without an "S"-meter, the a.v.c.

should be turned off and the b.f.o. turned on; then the r.f. gain should be set to give a moderately strong beat note with the carrier. The intensity of side frequencies can be estimated from the relative strength of the beats as the receiver is tuned through the spectrum adjacent to the carrier.

R.F. in Speech Amplifier

A small amount of r.f. current in the speech amplifier — particularly in the first stage, which is most susceptible to such r.f. pick-up — will cause overloading and distortion in the low-level stages. Frequently also there is a regenerative effect which causes an audio-frequency oscillation or "howl" to be set up in the audio system. In such cases the gain control cannot be advanced very far before the howl builds up, even though the amplifier may be perfectly stable when the r.f. section of the transmitter is not turned on.

Complete shielding of the microphone, microphone cord, and speech amplifier is necessary to prevent r.f. pick-up, and a ground connection separate from that to which the transmitter is connected is advisable.

MODULATION MONITORING

It is always desirable to modulate as fully as possible, but 100 per cent modulation should not be exceeded — particularly in the downward direction — because harmonic distortion will be introduced and the channel width increased. This causes unnecessary interference to other stations. The oscilloscope is the best instrument for continuously checking the modulation. However, simpler indicators may be used for the purpose, once calibrated.

A convenient indicator, when a Class B modulator is used, is the plate milliammeter in the Class B stage, since plate current of the modulator fluctuates with the voice intensity. Using the oscilloscope, determine the gain-control setting and voice intensity that give 100 per cent modulation on voice peaks, and simultaneously observe the maximum Class B plate-milliammeter reading on the peaks. When this maximum reading is obtained, it will suffice to adjust the gain so that it is not exceeded.

A high resistance (1000-ohms-per-volt or more) rectifier-type voltmeter (copper-oxide or germanium type) also can be used for modulation monitoring. It should be connected across the output circuit of an audio driver stage where the power level is a few watts, and similarly calibrated against the oscilloscope to determine the reading that represents 100 per cent modulation.

The plate milliammeter of the modulated r.f. stage also is of value as an indicator of overmodulation. As explained earlier, the d.c. plate current stays constant if the amplifier is linear. When the amplifier is overmodulated, especially in the downward direction, the operation is no longer linear and the average plate current will change. A flicker of the pointer may therefore be taken as an indication of overmodulation or nonlinearity. However, since it is possible that under some operating conditions the plate current will remain constant even though the amplifier is considerably overmodulated, an indicator of this type is not wholly reliable unless it has been checked against an oscilloscope.

Overmodulation Indicators

Overmodulation on negative peaks is usually the worst type, as explained earlier in this chapter. The milliammeter in the negative-peak indicator of Fig. 10-23 will show a reading on each peak that carries the instantaneous voltage on a plate-modulated amplifier "below zero" — that is, negative. The rectifier, *V*, cannot conduct so long as the negative half-cycle of audio output voltage is less than the d.c. voltage applied to the r.f. tube.

The inverse-peak-voltage rating of the rectifier tube must be at least twice the d.c. plate voltage of the modulated amplifier. The filament transformer likewise must have insulation rated to withstand twice the d.c. plate voltage. Either mercury-vapor or high-vacuum rectifiers can be used. The 15-volt breakdown voltage of the former will introduce a slight error, since the plate voltage must go at least 15 volts negative before the rectifier will ionize, but the error is inconsequential at plate voltages above a few hundred volts.

The effectiveness of the monitor is improved if it indicates at somewhat less than 100 per cent modulation, as it will then warn of the danger of overmodulation before it actually occurs. It can be adjusted to indicate at any desired modulation percentage by making the meter return to a point on the power-supply bleeder as shown in the alternative diagram. The by-pass condenser, *C*, insures that the full audio voltage appears across the indicator circuit.

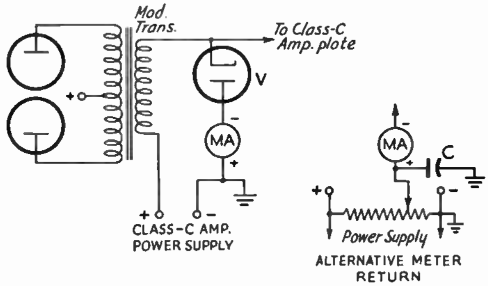


Fig. 10-23 — Negative-peak overmodulation indicator. The milliammeter *M.1* may be any low-range instrument (up to 0–50 ma. or so). The inverse-peak-voltage rating of the rectifier, *V*, must be at least twice the d.c. voltage applied to the plate of the r.f. amplifier. The alternative meter-return circuit can be used to indicate modulation in excess of any desired value below 100 per cent. The reactance of the by-pass condenser, *C*, at 100 cycles should be small compared with the resistance across which it is connected. An 8-μfd. electrolytic condenser will be satisfactory if the resistance it shunts is 1000 ohms or more.

Frequency and Phase Modulation

It is possible to convey intelligence by modulating any property of a carrier. These properties are amplitude, frequency and phase. Amplitude modulation (AM) is described in another chapter. When the frequency of the carrier is varied in accordance with the variations in a modulating signal, the result is **frequency modulation (FM).** Similarly, varying the phase of the carrier current is called **phase modulation (PM).**

Frequency and phase modulation are not independent, since the frequency cannot be varied without also varying the phase, and vice versa. The difference is largely a matter of definition.

The effectiveness of FM and PM for communication purposes depends almost entirely on the receiving methods. If the receiver will respond to frequency and phase changes but is insensitive to amplitude changes, it will discriminate against most forms of noise, particularly impulse noise such as is set up by ignition systems and other sparking devices. Special methods of detection are required to accomplish this result. Since most amateur receivers do not incorporate the proper circuits, the noise-reducing properties of FM or PM reception are seldom realized in amateur work.

Modulation methods for FM and PM are simple and require practically no audio power. There is also the advantage that, since there is no amplitude variation in the signal, interference to broadcast reception of the type resulting from rectification in the audio circuits of the b.c. receiver is substantially eliminated. These two points represent the principal reasons for the use of FM and PM in amateur work. Unfortunately, the user of FM or PM is unable to get the benefit of the inherent noise-reducing advantages of the system, and is furthermore at a considerable disadvantage with respect to AM of the same power, because most of his communication will be with amateurs using receivers designed specifically for AM.

Frequency Modulation

Fig. 11-1 is a representation of frequency modulation. When a modulating signal is applied, the carrier frequency is increased during one half-cycle of the modulating signal and decreased during the half-cycle of opposite polarity. This is indicated in the drawing by the fact that the r.f. cycles occupy less time (higher frequency) when the modulating signal is positive, and more time (lower frequency) when the modulating signal is negative. The change in the carrier frequency **(frequency deviation)** is proportional to the instantaneous amplitude of the modulating signal, so the deviation is small when the instantaneous amplitude of the modulating signal is small, and is greatest when the modulating signal reaches its peak, either positive or negative. That is, the frequency deviation follows the instantaneous changes in the amplitude of the modulating signal.

As shown by the drawing, the amplitude of the signal does not change during modulation.

Phase Modulation

To understand the difference between FM and PM it is necessary to appreciate that the frequency of an alternating current is determined by the *rate at which its phase changes.*

If the phase of the current in a circuit is changed there is an instantaneous frequency change during the time that the phase is being shifted. The amount of frequency change, or deviation, depends on how rapidly the phase shift is accomplished. It is also dependent upon the total amount of the phase shift. In a properly-operating PM system the amount of phase shift is proportional to the instantaneous amplitude of the modulating signal. The rapidity of the phase shift is directly proportional to the frequency of the modulating signal. Consequently, the frequency deviation in PM is proportional to both the amplitude and frequency of the modulating signal. The latter represents the outstanding difference between FM and PM, since in FM

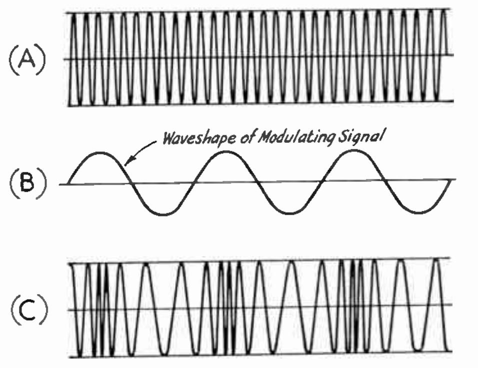


Fig. 11-1 — Graphical representation of frequency modulation. In the unmodulated carrier at A, each r.f. cycle occupies the same amount of time. When the modulating signal, B, is applied, the radio frequency is increased and decreased according to the amplitude and polarity of the modulating signal.

the frequency deviation is proportional only to the amplitude of the modulating signal.

Modulation Depth

Percentage of modulation in FM and PM has to be defined differently than for AM. Practically, "100 per cent modulation" is reached when the transmitted signal occupies a channel just equal to the bandwidth for which the *receiver* is designed. If the frequency deviation is greater than the receiver can accept, the receiver distorts the signal. However, on another receiver designed for a different bandwidth the same signal might be equivalent to only 25 per cent modulation.

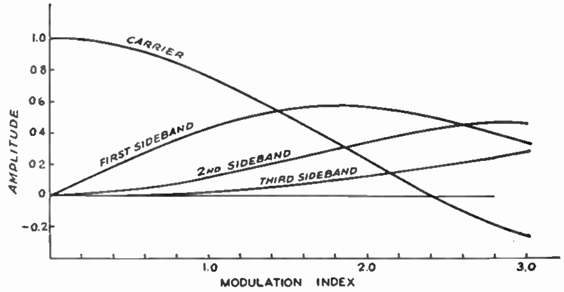


Fig. 11-2 — How the amplitude of the pairs of sidebands varies with the modulation index in an FM or PM signal. If the curves were extended for greater values of modulation index it would be seen that the carrier amplitude goes through zero at several points. The same statement also applies to the sidebands.

In amateur work "narrow-band" FM or PM (frequently abbreviated NFM) is defined as having the same channel width as a properly-modulated AM signal. That is, the channel width does not exceed twice the highest audio frequency in the modulating signal. NFM transmissions based on an upper audio limit of 3000 cycles therefore should occupy a channel no wider than 6 kc.

FM and PM Sidebands

The sidebands set up by FM and PM differ from those resulting from AM in that they occur at integral multiples of the modulating frequency on either side of the carrier rather than, as in AM, consisting of a single set of side frequencies for each modulating frequency. An FM or PM signal therefore inherently occupies a wider channel than AM.

The number of "extra" sidebands that occur in FM and PM depends on the relationship between the modulating frequency and the frequency deviation. The ratio between the frequency deviation, in cycles per second, and the modulating frequency, also in cycles per second, is called the **modulation index.** That is,

$$\text{Modulation index} = \frac{\text{Carrier frequency deviation}}{\text{Modulating frequency}}$$

Example: The maximum frequency deviation in an FM transmitter is 3000 cycles either side of the carrier frequency. The modulation index when the modulating frequency is 1000 cycles is

$$\text{Modulation index} = \frac{3000}{1000} = 3$$

At the same deviation with 3000-cycle modulation the index would be 1; at 100 cycles it would be 30, and so on.

In PM the modulation index is constant regardless of the modulating frequency; in FM it varies with the modulating frequency, as shown in the previous example. In an FM system the ratio of the *maximum* carrier-frequency deviation to the *highest* modulating frequency used is called the **deviation ratio.**

Fig. 11-2 shows how the amplitudes of the carrier and the various sidebands vary with the modulation index. This is for single-tone modulation; the first sideband (actually a pair, one above and one below the carrier) is displaced from the carrier by an amount equal to the modulating frequency, the second is twice the modulating frequency away from the carrier, and so on. For example, if the modulating frequency is 2000 cycles and the carrier frequency is 29,500 kc., the first sideband pair is at 29,498 kc. and 29,502 kc., the second pair is at 29,496 kc. and 29,504 kc., the third at 29,494 kc. and 29,506 kc., etc. The amplitudes of these sidebands depend on the modulation index, not on the frequency deviation. In AM, regardless of the percentage of modulation (so long as it does not exceed 100 per cent) the sidebands would appear *only* at 29,498 and 29,502 kc. under the same conditions.

Note that, as shown by Fig. 11-2, the carrier strength varies with the modulation index. (In amplitude modulation the carrier strength is constant; only the sideband amplitude varies.) At a modulation index of approximately 2.4 the carrier disappears entirely. It then becomes "negative" at a higher index, meaning that its phase is reversed as compared to the phase without modulation. In FM and PM the energy that goes into the sidebands is taken from the carrier, the *total* power remaining the same regardless of the modulation index.

Frequency Multiplication

Since there is no change in amplitude with modulation, an FM or PM signal can be amplified by an ordinary Class C amplifier without distortion. The modulation can take place in a very low-level stage and the signal can then be amplified by either frequency multipliers or straight amplifiers.

If the modulated signal is passed through one or more frequency multipliers, the modulation index is multiplied by the same factor that the carrier frequency is multiplied. For example, if modulation is applied on 3.5 Mc. and the final output is on 28 Mc. the total frequency multiplication is 8 times, so if the frequency deviation is 500 cycles at 3.5 Mc. it will be 4000 cycles at 28 Mc. Frequency multiplication offers a means for obtaining practically any desired amount of frequency deviation, whether or not the modulator itself is capable of giving that much deviation without distortion.

Narrow-Band FM and PM

"Narrow-band" FM or PM, the only type that is authorized for use on the lower frequencies where the 'phone bands are crowded, is defined as FM or PM that does not occupy a wider channel than an AM signal having the same audio modulating frequencies. Narrow-band operation requires using a relatively small modulation index.

If the modulation index (with single-tone modulation) does not exceed about 0.6 the most important extra sideband, the second, will be at least 20 db. below the unmodulated carrier level, and this should represent an effective channel width about equivalent to that of an AM signal. In the case of speech, a somewhat higher modulation index can be used. This is because the energy distribution in a complex wave is such that the modulation index for any one frequency component is reduced, as compared to the index with a sine wave having the same peak amplitude as the voice wave.

The chief advantage of narrow-band FM or PM for frequencies below 30 Mc. is that it eliminates or reduces certain types of interference to broadcast reception. Also, the modulating equipment is relatively simple and inexpensive. However, assuming the same unmodulated carrier power in all cases, narrow-band FM or PM is not as effective as AM with the methods of reception used by most amateurs. As shown by Fig. 11-2, at an index of 0.6 the amplitude of the first sideband is about 25 per cent of the unmodulated-carrier amplitude; this compares with a sideband amplitude of 50 per cent in the case of a 100 per cent modulated AM transmitter. That is, so far as effectiveness is concerned, a narrow-band FM or PM transmitter is about equivalent to a 100 per cent modulated AM transmitter operating at one-fourth the carrier power.

Comparison of FM and PM

Frequency modulation cannot be applied to an amplifier stage, but phase modulation can. PM is therefore readily adaptable to transmitters employing oscillators of high stability such as the crystal-controlled type. The amount of phase shift that can be obtained with good linearity is such that the maximum practicable modulation index is about 0.5 at the radio frequency at which the modulation takes place. Because the phase shift is proportional to the modulating frequency, this index can be used only at the highest frequency present in the modulating signal, assuming that all frequencies will at one time or another have equal amplitudes. Taking 3000 cycles as a suitable upper limit for voice work, and setting the modulation index at 0.5 for 3000 cycles, the frequency response of the speech-amplifier system above 3000 cycles must be sharply attenuated, to prevent sideband splatter. Also, if the "tinny" quality of PM as received on an FM receiver is to be avoided, the PM must be changed to FM, in which the modulation index decreases in inverse proportion to the modulating frequency. This requires shaping the speech-amplifier frequency-response curve in such a way that the output voltage is inversely proportional to frequency, at least over the voice range. When this is done the maximum modulation index can only be used at the lowest audio frequency, approximately 100 cycles in voice transmission, and must decrease in proportion to the increase in frequency. The result is that the maximum linear frequency deviation is only about 50 cycles, when PM is changed to FM. To increase the deviation to 3000 cycles requires a frequency multiplication of 3000/50, or 60 times.

In contrast, it is relatively easy to secure a fairly-large frequency deviation when a self-controlled oscillator is frequency-modulated directly. (True frequency modulation of a crystal-controlled oscillator results in only very small deviations and so requires a great deal of frequency multiplication.) The chief problem is to maintain a satisfactory degree of carrier stability, since the greater the inherent stability of the oscillator the more difficult it is to secure a wide frequency swing with linearity.

Methods of Frequency and Phase Modulation

● FREQUENCY MODULATION

The simplest and most satisfactory device for amateur FM is the reactance modulator. This is a vacuum tube connected to the r.f. tank circuit of an oscillator in such a way as to act as a variable inductance or capacitance.

Fig. 11-3 is a representative circuit. The control grid of the modulator tube, M, is connected across the oscillator tank circuit, C_1L_1, through resistor R_1 and blocking condenser C_2. C_8 represents the input capacitance of the modulator tube. The resistance of R_1 is made large compared to the reactance of C_8, so the r.f. current through R_1C_8 will be practically in phase with the r.f. voltage appearing at the terminals of the tank circuit. However, the voltage across C_8 will lag the current by 90 degrees. The r.f. current in the plate circuit of the modulator will be in phase with the grid voltage, and consequently is 90 degrees behind the current through C_8, or 90 degrees behind the r.f. tank voltage. This lagging current is drawn through the oscillator tank, giving the same effect as though an inductance were connected across the tank. The frequency increases in proportion to the amplitude of the lagging plate current of the modulator. The audio voltage, introduced through a radio-frequency choke, RFC, varies the transconductance of the tube and thereby varies the r.f. plate current.

The modulated oscillator usually is operated on a relatively low frequency, so that a high order of carrier stability can be secured. Frequency

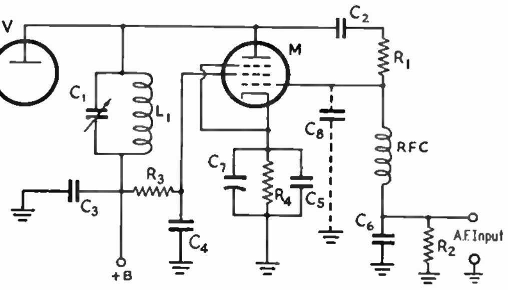


Fig. 11-3 — Reactance modulator using a high-transconductance pentode (6SG7, 6AG7, etc.).

C_1 — R.f. tank capacitance (see text).
C_2, C_3 — 0.001-μfd. mica.
C_4, C_5, C_6 — 0.0047-μfd. mica.
C_7 — 10-μfd. electrolytic.
C_8 — Tube input capacitance (see text).
R_1 — 47,000 ohms.
R_2 — 0.47 megohm.
R_3 — Screen dropping resistor; select to give proper screen voltage on type of modulator tube used.
R_4 — Cathode bias resistor; select as in case of R_3.
L_1 — R.f. tank inductance.
RFC — 2.5-mh. r.f. choke.

multipliers are used to raise the frequency to the final frequency desired. The frequency deviation increases with the number of times the initial frequency is multiplied: for instance, if the oscillator is operated on 6.5 Mc. and the output frequency is to be 52 Mc., an oscillator frequency deviation of 1000 cycles will be raised to 8000 cycles at the output frequency.

A reactance modulator can be connected to a crystal oscillator as well as to the self-controlled type. However, the resulting signal is more phase-modulated than it is frequency-modulated, for the reason that the frequency deviation that can be secured by varying the tuning of a crystal oscillator is quite small.

Design Considerations

The sensitivity of the modulator (frequency change per unit change in grid voltage) depends on the transconductance of the modulator tube. It increases when R_1 is made smaller in comparison with C_8. It also increases with an increase in L/C ratio in the oscillator tank circuit. Since the carrier stability of the oscillator depends on the L/C ratio, it is desirable to use the highest tank capacitance that will permit the desired deviation to be secured while keeping within the limits of linear operation.

A change in *any* of the voltages on the modulator tube will cause a change in r.f. plate current, and consequently a frequency change. Therefore it is advisable to use a regulated plate power supply for both modulator and oscillator. At the low voltages used (250 volts) the required stabilization can be secured by means of gaseous regulator tubes.

Speech Amplification

The speech amplifier preceding the modulator follows ordinary design, except that no power is required from it and the a.f. voltage taken by the modulator grid usually is small — not more than 10 or 15 volts, even with large modulator tubes. Because of these modest requirements, only a few speech stages are needed; a two-stage amplifier consisting of a pentode followed by a triode, both resistance-coupled, will more than suffice for crystal microphones.

PHASE MODULATION

The same type of reactance-tube circuit that is used to vary the tuning of the oscillator tank in FM can be used to vary the tuning of an amplifier tank and thus vary the phase of the tank current for PM. Hence the modulator circuit of Fig. 11-3 can be used for PM if the reactance tube works on an amplifier tank instead of directly on a self-controlled oscillator.

The phase shift that occurs when a circuit is detuned from resonance depends on the amount of detuning and the Q of the circuit. The higher the Q, the smaller the amount of detuning needed to secure a given number of degrees of phase shift. If the Q is at least 10, the relationship between phase shift and detuning (in kilocycles either side of the resonant frequency) will be substantially linear over a range of about 25 degrees. From the standpoint of modulator sensitivity, the Q of the tuned circuit on which the modulator operates should be as high as possible. On the other hand, the effective Q of the circuit will not be very high if the amplifier is delivering power to a load since the load resistance reduces the Q. There must therefore be a compromise between modulator sensitivity and r.f. power output from the modulated amplifier. An optimum figure for Q appears to be about 20; this allows reasonable loading of the modulated amplifier and the necessary tuning variation can be secured from a reactance modulator without difficulty. It is advisable to modulate at a very low power level — preferably in a transmitter stage where receiving-type tubes are used.

Reactance modulation of an amplifier stage usually also results in simultaneous amplitude modulation because the modulated stage is detuned from resonance as the phase is shifted. This must be eliminated by feeding the modulated signal through an amplitude limiter or one or more "saturating" stages — that is, amplifiers that are operated Class C and driven hard enough so that variations in the amplitude of the grid excitation produce no appreciable variations in the final output amplitude.

For the same type of reactance modulator, the speech-amplifier gain required is the same for PM as for FM. However, as pointed out earlier, the fact that the actual frequency deviation increases with the modulating audio frequency in PM makes it necessary to cut off the frequencies above about 3000 cycles before modulation takes place. If this is not done, unnecessary sidebands will be generated at frequencies considerably away from the carrier.

Reactance-Modulator Unit for Narrow-Band FM

The FM speech-amplifier and modulator unit shown in Figs. 11-4 and 11-5 uses a pentode reactance modulator in a circuit which is basically that of Fig. 11-3. It differs only in the detail that the audio signal is applied to the control grid in parallel with the r.f. voltage from the oscillator, instead of the series-feed arrangement shown in Fig. 11-3. Because of the parallel feed, resistor R_4 is incorporated in the circuit to prevent r.f. from appearing in the plate circuit of the speech-amplifier tube.

The unit uses miniature tubes for the sake of making a compact assembly that can be mounted in any convenient spot near the VFO tuned circuit. In Fig. 11-4 it is shown mounted on the outside of the VFO case. When this type of mounting is used the unit should be placed so that the lead between the VFO tuned circuit and the modulator is as short as possible. If there is space available, it is preferable to mount the unit inside the VFO cabinet.

The chassis for the unit is 4 inches long by 2 inches wide, and has a mounting lip 2 inches deep. As shown in the photographs, it is formed from a piece of aluminum with the edges turned over to stiffen it. The various components are easily accommodated underneath. The r.f. leads should be kept short and separated as much as possible from the audio and power-supply wiring.

Filament and plate power can usually be taken from the VFO supply, since the total plate current is only a few milliamperes. Filament current required is 0.6 amp. The microphone input is carried through a shielded lead to the unit, thus the microphone connector can be placed in any convenient location on the VFO unit itself. Once the proper setting of the gain control is found it need not be touched again, so screwdriver adjustment is quite adequate.

The adjustment of reactance modulators is discussed in a later section in this chapter.

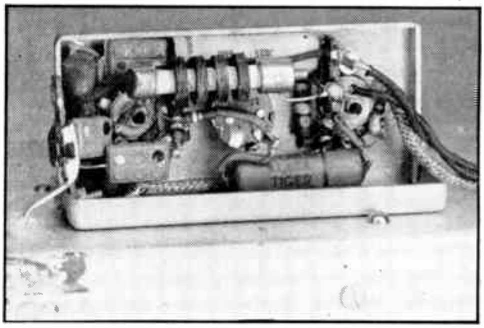


Fig. 11-5 — Underneath the modulator unit. The r.f. connection to the VFO goes through the feed-through bushing at the left.

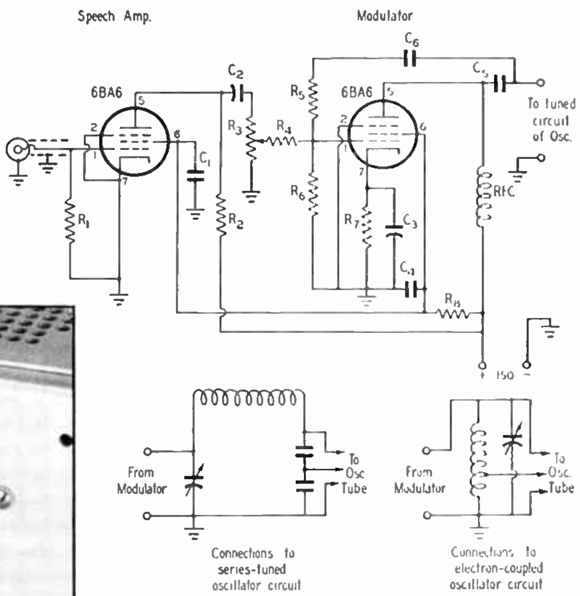


Fig. 11-6 — Circuit diagram of the narrow-band FM modulator unit.

C_1 — 680-μμfd. mica.
C_2, C_4 — 0.01-μfd. paper, 400 volts.
C_3 — 0.025-μfd. paper, 200 volts.
C_5, C_6 — 47-μμfd. mica.
R_1 — 1.2 megohms, ½ watt.
R_2, R_8 — 0.22 megohm, ½ watt.
R_3 — 0.5-megohm potentiometer.
R_4 — 0.1 megohm, ½ watt.
R_5 — 10,000 ohms, ½ watt.
R_6 — 0.47 megohm, ½ watt.
R_7 — 390 ohms, ½ watt.
RFC — 2.5-mh. r.f. choke.

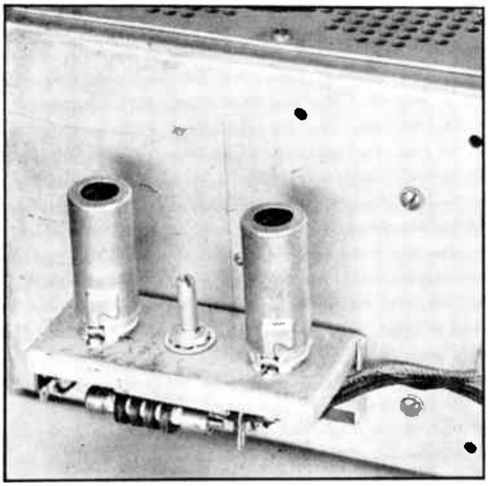


Fig. 11-4 — Miniature reactance modulator that can be used with any VFO. The shielded lead is for microphone input; the other two wires bring in filament and plate supply.

Checking FM and PM Transmitters

Accurate checking of the operation of an FM or PM transmitter requires different methods than the corresponding checks on an AM set. This is because the common forms of measuring devices either indicate amplitude variations only (a d.c. milliammeter, for example), or because their indications are most easily interpreted in terms of amplitude. There is no simple measuring instrument that indicates frequency deviation in a modulated signal directly.

However, there is one favorable feature in FM or PM checking. The modulation takes place at a very low level and the stages following the one that is modulated do not affect the linearity of modulation so long as they are properly tuned. Therefore the modulation may be checked *without putting the transmitter on the air*, or even on a dummy antenna. The power is simply cut off the amplifiers following the modulated stage. This not only avoids unnecessary interference to other stations during testing periods, but also keeps the signal at such a low level that it may be observed quite easily on the station receiver. A good receiver with a crystal filter is an essential part of the checking equipment of an FM or PM transmitter, particularly for narrow-band FM or PM.

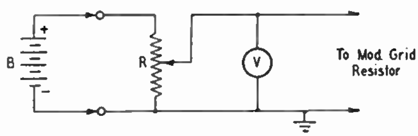


Fig. 11-7 — D.c. method of checking frequency deviation of a reactance-tube-modulated oscillator. A 500- or 1000-ohm potentiometer may be used at *R*.

The quantities to be checked in an FM or PM transmitter are the linearity and frequency deviation. Because of the essential difference between FM and PM the methods of checking differ in detail.

Reactance-Tube FM

It was explained earlier that in FM the frequency deviation is the same at any audio modulation frequency if the audio signal amplitude does not vary. Since this is true at *any* audio frequency it is true at zero frequency. Consequently it is possible to calibrate a reactance modulator by applying an adjustable d.c. voltage to the modulator grid and noting the change in oscillator frequency as the voltage is varied. A suitable circuit for applying the adjustable voltage is shown in Fig. 11-7. The battery, *B*, should have a voltage of 3 to 6 volts (two or more dry cells in series). The arrows indicate clip connections so that the battery polarity can be reversed.

The oscillator frequency deviation should be measured by using a receiver in conjunction with an accurately-calibrated frequency meter, or by any means that will permit accurate measurement of frequency differences of a few hundred cycles. One simple method is to tune in the oscillator on the receiver (disconnecting the receiving antenna, if necessary, to keep the signal strength well below the overload point) and then set the receiver b.f.o. to zero beat. Then increase the d.c. voltage applied to the modulator grid from zero in steps of about ½ volt and note the beat frequency at each change. Then reverse the battery terminals and repeat. The frequency of the beat note may be measured by comparison with a calibrated audio-frequency oscillator. Note that with the battery polarity positive with respect to ground the radio frequency will move in one direction when the voltage is increased, and in the other direction when the battery terminals are reversed. When several readings have been taken a curve may be plotted to demonstrate the relationship between grid voltage and frequency deviation.

A sample curve is shown in Fig. 11-8. The usable portion of the curve is the center part which is essentially a straight line. The bending at the ends indicates that the modulator is no longer linear; this departure from linearity will cause harmonic distortion and will broaden the channel occupied by the signal. In the example, the characteristic is linear 1.5 kc. on either side of the center or carrier frequency. This is the maximum deviation permissible at the frequency at which the measurement is made. At the final output frequency the deviation will be multiplied by the same number of times that the measurement frequency is multiplied. This must be kept in mind when the check is made at a frequency that differs from the output frequency.

A good modulation indicator is a "magic-eye" tube such as the 6E5. This should be connected across the grid resistor of the reactance modulator as shown in Fig. 11-9. Note its deflection (using the d.c. voltage method as in Fig. 11-7) at the maximum deviation to be used. This deflection represents "100 per cent

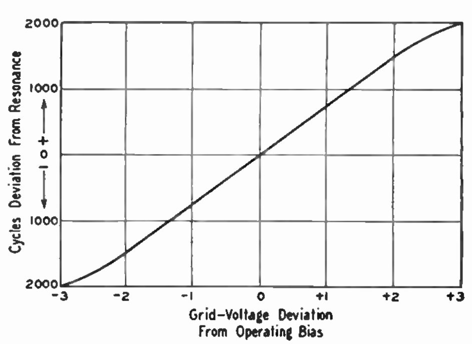


Fig. 11-8 — A typical curve of frequency deviation *vs.* modulator grid voltage.

modulation" and with speech input the gain should be kept at the point where it is just reached on voice peaks. If the transmitter is used on more than one band, the gain control should be marked at the proper setting for each band, because the signal amplitude that gives the correct deviation on one band will be either too great or too small on another. For narrow-band FM the proper deviation is approximately 2000 cycles (based on an upper a.f. limit of 3000 cycles and a deviation ratio of 0.7) at the final *output* frequency. If the output frequency is in the 29-Mc. band and the oscillator is on 7 Mc., the deviation at the *oscillator* frequency should not exceed 2000/4, or 500 cycles.

Checking with a Crystal-Filter Receiver

With PM the d.c. method of checking just described cannot be used, because the frequency deviation at zero frequency also is zero. For narrow-band PM it is necessary to check the actual width of the channel occupied by the transmission. (The same method also can be used to check FM.) For this purpose it is necessary to have a crystal-filter receiver and an a.f. oscillator that generates a 3000-cycle sine wave.

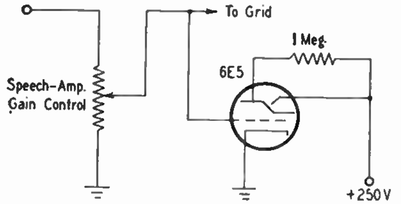


Fig. 11-9 — 6E5 modulation indicator for FM or PM modulators. To insure sufficient grid voltage for a good deflection, it may be necessary to connect the gain control in the modulator grid circuit rather than in an earlier speech-amplifier stage.

Keeping the signal intensity in the receiver at a medium level, tune in the carrier at the *output* frequency. Do not use the a.v.c. Switch on the beat oscillator, and set the crystal filter at its sharpest position. Peak the signal on the crystal and adjust the b.f.o. for any convenient beat note. Then apply the 3000-cycle tone to the speech amplifier (through an attenuator, if necessary, to avoid overloading; see chapter on audio amplifiers) and increase the audio gain until there is a small amount of modulation. Tuning the receiver near the carrier frequency will show the presence of sidebands 3 kc. from the carrier on both sides. With low audio input, these two should be the only sidebands detectable.

Now increase the audio gain and tune the receiver over a range of about 10 kc. on both sides of the carrier. When the gain becomes high enough, a second set of sidebands spaced 6 kc. on either side of the carrier will be detected. The signal amplitude at which these sidebands become detectable is the maximum speech amplitude that should be used. If the 6E5 modulation indicator is incorporated in the modulator, its deflection with the 3000-cycle tone will be the "100 per cent modulation" deflection for speech.

When this method of checking is used with a reactance-tube-modulated FM (not PM) transmitter, the linearity of the system can be checked by observing the *carrier* as the a.f. gain is slowly increased. The beat-note frequency will stay constant so long as the modulator is linear, but nonlinearity will be accompanied by a shift in the average carrier frequency that will cause the beat note to change in frequency. If such a shift occurs at the same time that the 6-kc. sidebands appear, the extra sidebands may be caused by modulator distortion rather than by an excessive modulation index. This means that the modulator is not capable of shifting the frequency over a wide-enough range. The 6-kc. sidebands should appear *before* there is any shift in the carrier frequency.

R.F. Amplifiers

The r.f. stages in the transmitter that follow the modulated stage may be designed and adjusted as in ordinary operation. In fact, there are no special requirements to be met except that all tank circuits should be carefully tuned to resonance (to prevent unwanted r.f. phase shifts that might interact with the modulation and thereby introduce hum, noise and distortion). In neutralized stages, the neutralization should be as exact as possible, also to minimize unwanted phase shifts. With FM and PM, all r.f. stages in the transmitter can be operated at the manufacturer's maximum c.w.-telegraphy ratings, since the average power input does not vary with modulation as it does in AM 'phone operation.

The output of the transmitter should be checked for amplitude modulation by observing the antenna current. It should not change from the unmodulated-carrier value when the transmitter is modulated. If there is no antenna ammeter in the transmitter, a flashlight lamp and loop can be coupled to the final tank coil to serve as a current indicator. If the carrier amplitude is constant, the lamp brilliance will not change with modulation.

Amplitude modulation accompanying FM or PM is just as much to be avoided as frequency or phase modulation that accompanies AM. A mixture of AM with either of the other two systems results in the generation of spurious sidebands and consequent widening of the channel. If the presence of AM is indicated by variation of antenna current with modulation, the cause is almost certain to be nonlinearity in the modulator. In very wide-band FM the selectivity of the transmitter tank circuits may cause the amplitude to decrease at high deviations, but this condition is not likely to occur on amateur frequencies at which wide-band FM would be used.

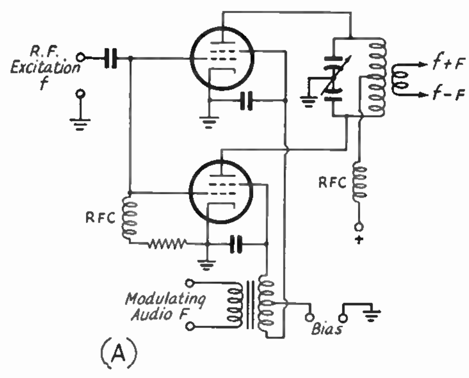
Reduced-Carrier and Single-Sideband

The most significant development in amateur radiotelephony in the past several years has been the increased use of single-sideband suppressed-carrier transmissions. This system has tremendous potentialities for increasing the effectiveness of 'phone transmission and for reducing interference. Because only one of the two sidebands normally produced in modulation is transmitted, the channel width is immediately cut in half. However, when only one sideband is transmitted the carrier — which is essential in double-sideband transmission — no longer is necessary; it can be supplied without too much difficulty at the receiver. With the carrier eliminated there is a great saving in power at the transmitter — or, from another viewpoint, a great increase in effective power output. Assuming that the same final-amplifier tube or tubes are used either for normal AM or for single-sideband, carrier suppressed, it can be shown that the use of SSB can give an effective gain of 9 db. over AM — equivalent to increasing the transmitter power 8 times. Eliminating the carrier also eliminates the heterodyne interference that wrecks so much communication in congested 'phone bands.

SUPPRESSING THE CARRIER

The carrier can be suppressed or nearly eliminated by an extremely sharp filter or by using a **balanced modulator.** The basic principle in any balanced modulator is to introduce the carrier in such a way that it does not appear in the output but so that the sidebands will. This requirement is satisfied by introducing the audio in push-pull and the r.f. drive in parallel, and connecting the output (plate circuit) of the tubes in push-pull, as shown in Fig. 12-1A. Balanced modulators can also be connected with the r.f. drive and audio inputs in push-pull and the output in parallel (Fig. 12-1B) with equal effectiveness. The choice of a balanced modulator circuit is generally determined by constructional considerations and the method of modulation preferred by the builder. Screen-grid modulation is shown in the examples in Fig. 12-1, but control-grid or plate modulation can be used equally as well. Balanced-modulator circuits using four rectifiers (germanium, copper oxide, or thermionic) in "bridge" or "ring" circuits are often used, particularly in commercial applications. Two-rectifier circuits are also available, and they are widely used in amateur SSB equipment. Examples of rectifier-type balanced modulators are shown in Fig. 12-2.

In any of the vacuum-tube circuits, there will be no output with no audio signal because the circuits are balanced. The signal from one tube is balanced or cancelled in the output circuit by the signal from the other tube. The circuits are thus balanced for any value of *parallel* audio signal. When push-pull audio is applied, the modulating voltages are of opposite polarity, and one tube will conduct more than the other. Since any modulation process is the same as "mixing" in receivers, sum and difference frequencies (sidebands) will be generated. The modulator is not



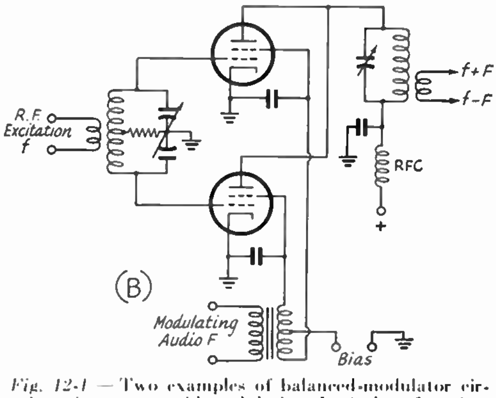


Fig. 12-1 — Two examples of balanced-modulator circuits using screen-grid modulation. In A the r.f. excitation is in parallel in both tubes, and the audio and output are in push-pull. In B the excitation and audio are in push-pull, the output is in parallel. In either case, the carrier frequency, *f*, does not appear in the output circuit — only the two sideband frequencies, $f + F$ and $f - F$, will appear. The bias fed to the screens is a practical requirement with all screen-grid tubes for proper linear operation, and is not a special requirement of balanced modulators.

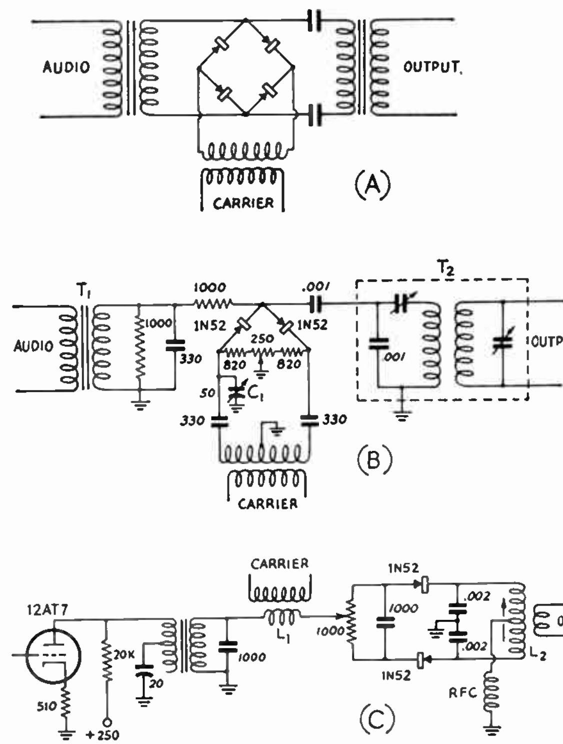


Fig. 12-2 — Typical rectifier-type balanced modulators.

The circuit at A is called a "bridge" balanced modulator and has been widely used in commercial work.

The balanced modulator at B is shown with constants suitable for operation at 450 kc. It is useful for working into a crystal bandpass filter. T_1 is a transformer designed to work from the audio source into a 600-ohm load, and T_2 is an ordinary i.f. transformer with the trimmer reconnected in series with a 0.001-μfd. condenser, for impedance-matching purposes from the modulator. The condenser C_1 is for carrier balance and may be found unnecessary in some instances — it should be tried connected on either side of the carrier input circuit and used where it is more effective. The 250-ohm potentiometer is normally all that is required for carrier balance. The carrier input should be sufficient to develop several volts across the resistor string.

The balanced modulator circuit at C is shown with constants suitable for operation at 3.9 Mc. T_3 is a small step-down output transformer (UTC R-38A), shunt-fed to eliminate d.c. from the windings. L_1 can be a small coupling coil wound on the "cold" end of the carrier-oscillator tank coil, with sufficient coupling to give two or three volts of r.f. across its output. L_2 is a slug-tuned coil that resonates to the carrier frequency with the effective 0.001 μfd. across it. The 1000-ohm potentiometer is for carrier balance.

balanced for the sidebands, and they will appear in the output.

The amount of carrier suppression is dependent upon the matching of the two tubes and their associated circuits. Normally two tubes of the same type will balance closely enough to give at least 15 or 20 db. carrier suppression without any adjustment. If further suppression is required, trimmer condensers to balance the grid-plate capacities and separate bias adjustments for setting the operating points can be used.

In the rectifier-type balanced modulators shown in Fig. 12-2, the diode rectifiers are connected in such a manner that, if they have equal forward resistances, no r.f. can pass from the carrier source to the output circuit via either of the two possible paths. The net effect is that no r.f. energy appears in the output. When audio is applied, it unbalances the circuit by biasing the diode (or diodes) in one path, depending upon the instantaneous polarity of the audio, and hence some r.f. will appear in the output. The r.f. in the output will appear as a double-sideband suppressed-carrier signal. (For a more complete description of diode-modulator operation, see "Diode Modulators," *QST*, April, 1953, page 39.)

In any diode modulator, the r.f. voltage should be at least 6 or 8 times the peak audio voltage, for minimum distortion. The usual operation involves a fraction of a volt of audio and several volts of r.f. The diodes should be matched as closely as possible — ohmmeter measurements of their forward resistances is the usual test.

(The circuit of Fig. 12-2B is described more fully in Weaver and Brown, "Crystal Lattice Filters for Transmitting and Receiving," *QST*, August, 1951. The circuit of Fig. 12-2C is suitable for use in a double-balanced-modulator circuit and is so described in "SSB, Jr.," *General Electric Ham News*, September, 1950.)

● REDUCED-CARRIER SIGNALS

Double-sideband reduced-carrier signals, obtained by unbalancing a balanced modulator sufficiently to allow some carrier to appear in the output, offer a number of advantages over conventional AM signals: considerably higher efficiency, where efficiency is defined as the ratio of sideband (useful) power output to total power input; high output with comparatively little audio power; and a considerable reduction in heterodyne interference. The signal can be received by ordinary methods, and merely sounds as though it had "a lot of modulation for the carrier."

In ordinary amplitude-modulated systems, the sideband amplitude can never exceed 0.5 the carrier amplitude without generating spurious side frequencies (when sine-wave modulation is used). Under these conditions, ⅔ of the total power is in the carrier and ⅓ is in the sidebands. However, with DSRC, generated by the unbalancing of a balanced modulator, it is possible to have *any* amplitude of sidebands without generating spurious side frequencies. In practical tests it has been found that a modulation factor of 4 is perfectly practical, and the distortion under normal demodulation is not enough to impair the communication value of the signal. Under these conditions, the sideband power is 2½ times as great as could be obtained with straight A3 transmission (grid-modulated) with the same tubes, or about ¾ of what could be obtained with the same tubes plate-modulated 100 per cent. Since the audio-power requirements can be kept low, and the no-modulation plate current may be only a little more than half of the full-signal plate current, the advantages of DSRC are obvious for work where the total power available is limited, as in mobile or portable work.

A DSRC signal can be generated at a low power level and amplified in a linear amplifier (discussed later in this chapter). Under these conditions, a relatively powerful signal can be obtained with a minimum of audio power and total power input to the entire transmitter.

(For further information on DSRC, see Grammer, "D.S.R.C. Radiotelephony," *QST*, May, 1951, and Grammer, "Practical D.S.R.C. Transmitter Design," *QST*, June, 1951.)

● SINGLE-SIDEBAND GENERATORS

Two basic systems for generating SSB signals are shown in Fig. 12-2. One involves the use of a bandpass filter having sufficient selectivity to pass one sideband and reject the other. Filters having such characteristics can only be constructed for relatively low frequencies, and most filters used by amateurs are designed to work somewhere between 10 and 20 kc. Good sideband filtering can be done at frequencies as high as 500 kc. by using multiple-crystal or electromechanical filters. The low-frequency oscillator output is combined with the audio output of a speech amplifier in a balanced modulator, and only the upper and lower sidebands appear in the output. One of the sidebands is passed by the filter and the other rejected, so that an SSB signal is fed to the mixer. The signal is there mixed with the output of a high-frequency r.f. oscillator to produce the desired output frequency. For additional amplification a linear r.f. amplifier (Class A or Class B) must be used. When the SSB signal is generated at 10 or 20 kc., it is generally first heterodyned to somewhere around 500 kc. and then to the operating frequency. This simplifies the problem of rejecting the "image" frequencies resulting from the heterodyne process. The problem of image frequencies in the frequency conversions of SSB signals differs from the problem in receivers because the beating-oscillator frequency becomes important. Either balanced modulators or sufficient selectivity must be used to attenuate these frequencies in the output and hence minimize the possibility of unwanted radiations.

The second system is based on the phase relationships between the carrier and sidebands in a modulated signal. As shown in the diagram, the audio signal is split into two components that are identical except for a phase difference of 90 degrees. The output of the r.f. oscillator (which may be at the operating frequency, if desired) is likewise split into two separate components having a 90-degree phase difference. One r.f. and one audio component are combined in each of two separate balanced modulators. The carrier is suppressed in the modulators, and the relative phases of the sidebands are such that one sideband is balanced out and the other is accentuated in the combined output. If the output from the balanced modulators is high enough, such an SSB exciter can work directly into the antenna, or the power level can be increased in a following amplifier.

Properly adjusted, either system is capable of good results. Arguments in favor of the filter system are that it is somewhat easier to adjust without an oscilloscope, since it requires only a receiver and a v.t.v.m. for alignment, and it is more likely to remain in adjustment over a long period of time. The chief argument against it, from the amateur viewpoint, is that it requires quite a few stages and at least one frequency conversion after modulation. The phasing system requires fewer stages and can be designed to require no frequency conversion, but its alignment and adjustment are often considered to be a little "trickier" than that of the filter system. This probably stemmed from lack of familiarity with the system rather than any actual difficulty, and now that commercially-available preadjusted audio-phasing networks are available, most of the alignment difficulty has been eliminated. In most cases the phasing system will cost less to apply to an existing transmitter.

Regardless of the method used to generate a SSB signal of 5 or 10 watts, the minimum cost will be found to be higher than for an AM transmitter of the same low power. However, as the power level is increased, the SSB transmitter becomes more economical than the AM rig, both initially and from an operating standpoint.

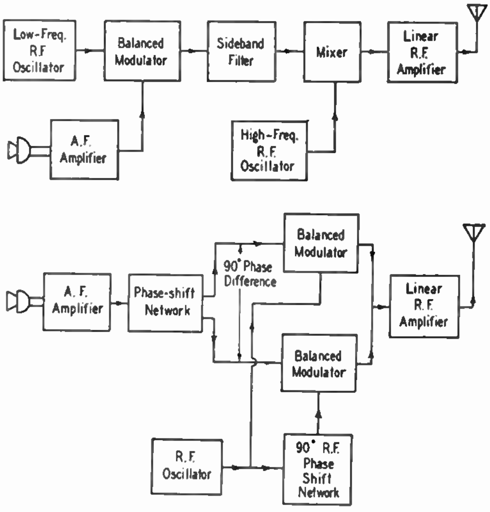


Fig. 12-3 — Two basic systems for generating single-sideband suppressed-carrier signals.

AMPLIFICATION OF SSB SIGNALS

When an SSB signal is generated at some frequency other than the operating frequency, it is necessary to change frequency by heterodyne methods. These are exactly the same as those used in receivers, and any of the normal mixer or converter circuits can be used. One exception to this is the case where the original signal and the heterodyning oscillator are not too different in frequency (as when heterodyning a 20-kc. signal to 500 kc.) and, in this case, a balanced mixer should be used, to eliminate the heterodyning oscillator frequency in the output.

To increase the power level of an SSB signal, a **linear amplifier** must be used. The simplest form of linear amplifier (r.f. or audio) is the Class A amplifier, which is used almost without exception throughout receivers and low-level speech equipment. While its linearity can be made relatively good, it is inefficient. The theoretical limit of efficiency is 50 per cent, and most practical amplifiers run 25–35 per cent efficient at full output. At low levels this is not worth worrying about, but when the 2- to 10-watt level is exceeded something else must be done to improve this efficiency and reduce tube, power-supply and operating costs.

Class AB_1 amplifiers make excellent linear amplifiers if suitable tubes are selected. Primary advantages of Class AB_1 amplifiers are that they give much greater output than straight Class A amplifiers using the same tubes, and they do not require any grid driving power (no grid current drawn at any time). Although triodes can be used for Class AB_1 operation, tetrodes or pentodes are usually to be preferred, since Class AB_1 operation requires high peak plate current without grid current, and this is easier to obtain in tetrodes and pentodes than in most triodes.

To obtain maximum output from tetrodes, pentodes and most triodes, it is necessary to operate them in Class AB_2. Although this produces maximum peak output, it increases the driving-power requirements and, what is more important, requires that the **driver regulation** (ability to maintain waveform under varying load) be good or excellent. The usual method to improve the driver regulation is to add fixed resistors across the grid circuit of the driven stage, to offer a load to the driver that is modified only slightly by the additional load of the tube when it is driven into the grid-current region. This increases the driver's output-power requirements. Further, it is desirable to make the grid circuit of the Class AB_2 stage a high-*C* circuit, to improve regulation and simplify coupling to the driver. A "stiff" bias source is also required, since it is important that the bias remain constant, whether or not grid current is drawn.

Class B amplifiers are theoretically capable of 78.5 per cent efficiency at full output, and practical amplifiers run at 60–70 per cent efficiency at full output. Tubes normally designed for Class B audio work can be used in r.f. linear amplifiers and will operate at the same power rating and efficiency provided, of course, that the tube is capable of operation at the radio frequency. The operating conditions for r.f. are substantially the same as for audio work — the only difference is that the input and output transformers are replaced by suitable r.f. tank circuits. Further, in r.f. circuits it is readily possible to operate only one tube if only half the power is wanted — push-pull is not a necessity in Class B r.f. work. However, the r.f. harmonics will be higher in the case of the single-ended amplifier, and this should be taken into consideration if TVI is a problem.

For proper operation of Class B amplifiers, and to reduce harmonics and facilitate coupling, the input and output circuits should not have a low *C*-to-*L* ratio. A good guide to the proper size of tuning condenser is Figs. 6-9 and 6-17 and, in case of any doubt, it is well to be on the high-capacity side. If zero-bias tubes are used in the Class B stage, it may not be necessary to add much "swamping" resistance across the grid circuit, because the grids of the tubes load the circuit at all times. However, with other tubes that require bias, the swamping resistor should be such that it dissipates from five to ten times the power required by the grids of the tubes. This will insure an almost constant load on the driver stage and good regulation of the grid voltage of the Class B stage.

Before going into detail on the adjustment and loading of the linear amplifier, a few general considerations should be kept in mind. If proper operation is expected, it is essential that the amplifier be so constructed, wired and neutralized that no trace of regeneration or parasitic instability remains. Needless to say, this also applies to the stages driving it.

The bias supply to the Class B linear amplifier should be quite stiff, such as batteries or some form of voltage regulator. If nonlinearity is noticed when testing the unit, the bias supply may be checked by means of a large electrolytic capacitor. Simply shunt the supply with 100 μfd. or so of capacity and see if the linearity improves. If so, rebuild the bias supply for better regulation. *Do not rely on a large condenser alone.*

Where tetrodes or pentodes are used, the screen supply should have good regulation and its voltage should remain constant under the varying current demands. If the maximum screen current does not exceed 30 or 35 ma., a string of VR tubes in series can be used to regulate the screen voltage. If the current demand is higher, it may be necessary to use an electronically-regulated power supply or a heavily-bled power supply with a current capacity of several times the current demand of the screen circuit.

Where VR tubes are used to regulate the screen supply, they should be selected to give a regulated voltage as close as possible to the tube's rated voltage, but it does not have to be exact. Minor differences in idling plate current can be made up by readjusting the grid bias.

From the standpoint of ease of adjustment and availability of proper operating voltages, a linear amplifier with Class AB_1 tetrodes or pentodes or one with zero-bias Class B triodes would be first choice. The Class B amplifier would require more driving power. (For examples of Class AB_1 tetrode amplifiers, see Russ, "The 'Little Firecracker' Linear Amplifier," *QST*, September, 1953, and Eckhardt, "The Single Side-Saddle Linear," *QST*, November, 1953.)

Table 12-I lists a few of the more popular tubes commonly used for SSB linear-amplifier operation. Except where otherwise noted, these ratings are those given by the manufacturer for audio work and and as such are based on a sine-wave signal. These ratings are adequate ones for use in SSB amplifier design, but they are conservative for such work and hence do not necessarily represent the maximum powers that can be obtained from the tubes in voice-signal SSB service. In no case should the *average* plate dissipation be exceeded for any considerable length of time, but the nature of a SSB signal is such that the average plate dissipation of the tube will run well below the peak plate dissipation. Hence in SSB operation the *peak* plate dissipation may exceed the average by several times.

Getting the most out of a linear amplifier is done by increasing the peak power without exceeding the average plate dissipation over any appreciable length of time. This can be done by raising the plate voltage or the peak current (or both), provided the tube can withstand the increase. For example, the 6146 is shown with 750 volts maximum on the plate, and it is quite likely that this can be increased to 900 or 1000 volts without any appreciable shortening of the life of the tube. However, the manufacturers have not released any data on such operation, and any extrapolation of the audio ratings is at the risk of the amateur. A 35- to 50-per cent increase above plate-voltage ratings should be perfectly safe in most cases. In a tetrode or pentode, the peak plate current can be boosted some by raising the screen voltage.

When running a linear amplifier at considerably higher than the audio ratings, the "two-tone test signal" (described below) should never be applied at full amplitude for more than a few seconds at any one time. The above statements about working tubes above ratings apply only when a voice signal is used — a prolonged whistle or two-tone test signal may damage the tube.

Adjustment of Amplifiers

The two critical adjustments for obtaining proper operation from the linear amplifier are the plate loading and the grid drive. Since these adjustments are preferably made with power on, it is a matter of practical convenience to have both controls readily available, at least during initial tune-up.

The 'scope can show misadjustment at a glance and will greatly facilitate all adjustments. In addition, it is the most reliable instrument for observing modulation amplitude and, once used, is likely to become the most nearly essential instrument in the shack.

With single sideband, 100 per cent modulation with a single tone is a pure r.f. output with no modulation envelope, and the point of amplifier overload is difficult to observe. However, if the input signal consists of two sine waves of different frequencies (for example, 1000 c.p.s. difference) but equal amplitudes, the output of the single-sideband transmitter should have the envelope shown in Fig. 12-4. This is called a "two-tone" test signal to distinguish it from other test signals. Its first advantage lies in the fact that any flattening of the positive peaks is readily discernible, which makes the adjustment of the linear-amplifier drive and output coupling as simple a procedure as that for AM systems. Flattening of the peaks (to be avoided) is illustrated in Fig. 12-5.

Those who use the filter method for obtaining single-sideband signals can obtain such a test signal by mixing the output of two audio oscillators of good waveform. The experimenters using the phasing method of single-side-band signal generation will recognize the pattern as that obtained when a single test tone is applied to *one* of the balanced modulators. For this latter group a two-tone test signal may be readily obtained by disabling one of the balanced modulators in the exciter and applying a single input tone.

Suppose that the linear amplifier has been coupled to a dummy load and the single-sideband exciter has been connected to its input. By observing the oscilloscope coupled to the amplifier output, it will be possible to adjust the drive and output coupling so that the peaks of the two-tone test signal waveform are on the verge of flattening. The peak input power may now be checked. This is readily possible, for with the two-tone test

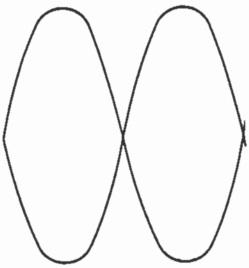


Fig. 12-4 — Oscillogram of a two-tone test signal through a linear amplifier.

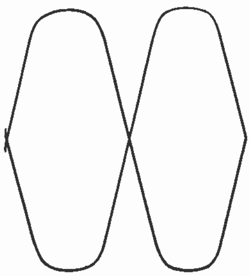


Fig. 12-5 — Flattening caused by overdrive or insufficient plate loading.

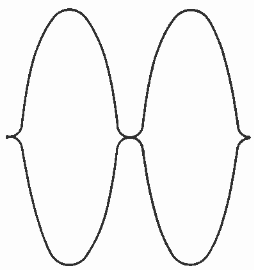


Fig. 12-6 — The distorted pattern obtained when the bias voltage is incorrect.

TABLE 12-I — LINEAR-AMPLIFIER TUBE-OPERATION DATA

Except where otherwise noted, ratings are manufacturers' for audio operation. Values given are for one tube. Driving powers represent tube losses only—circuit losses will increase the figures.

Tube	Class	Plate Voltage	Screen Voltage	D.C. Grid Voltage	Zero-Sig. D.C. Plate Current	Max.-Sig. D.C. Plate Current	Zero-Sig. D.C. Screen Current	Max.-Sig. D.C. Screen Current	Peak R.F. Grid Voltage	Max.-Sig Avg. Grid Current	Max.-Sig. Avg. Driving Power	Max.-Rated Screen Dissipation	Max.-Rated Grid Dissipation	Avg. Plate Dissipation	Max.-Sig. Useful Power Output
2E26	A_1	250	200	−14	35	42	7	10	14	—	0	2.5		10	5
	AB_2	400	125	−15	10	75		16	30		.2	2 5		10	21
		500	125	−15	11	75		16	30		.2	2.5		12.5	27
6146	AB_1	600	200	−50	26	120	.6	13	50	—	0	3		25	47
		750	200	−50	29	114	.5	14	50	—	0	3		25	60
	AB_2	600	185	−50	21	135	.5	15	57	.4	.02	3		25	58
		750	165	−45	18	120	.3	11	51	.4	.02	3		25	65
807 1625	AB_2	600	300	−30	30	100	.4	6	39		.1	3.5		25	40
		750	300	−32	26	120	.3	8	46		.1	3.5		30	60
811-A	B	1000	—	0	22	175	—	—	93		3.8	—		65	124
		1250	—	0	27	175	—	—	88	13	3.0	—		65	155
		1500	—	−4.5	16	157	—	—	85		2.2	—		65	170
4-65A	AB_1	1000	500	−85	15	85	0	12	85	—	0	10		45	40
		1500	500	−85	15	90	0	7	85	—	0	10		63	73
		1750	500	−90	10	85	0	9	90	—	0	10		62	88
	AB_2	1000	250	−30[1]	30	150	0	23	105		2.5	10	5	65	85
		1500	250	−35[1]	30	125	0	15	100		1.6	10	5	63	125
		1800	250	−35[1]	25	110	0	13	90		1.1	10	5	63	135
	B[2]	1500	300	−50[1]	33	200	0	35[3]	190	13	2.4	10	5	60	150
		2000	400	−75[1]	25	270	0	50[1]	270	17	4.6	10	5	65	300
		2500	500	−100[1]	20	230	0	35[3]	300	6	1.8	10	5	65	325
813	AB_2	2000	750	−90	20	158	.8	29	115		.1	22		100	228
		2250	750	−90	23	158	.8	29	115		.1	22		100	258
		2500	750	−95	18	180	.6	28	118		.2	22		125	325
4-125A	AB_1	1500	600	−90[1]	30	111	−.5	9	90	—	0	20		88	79
		2000	600	−94[1]	25	120	−.3	3	94	—	0	20		125	115
		2500	600	−96[1]	25	116	−.2	4	96	—	0	20		125	165
	AB_2	1500	350	−41[1]	44	200	0	17	141	9	1.25	20	5	125	175
		2000	350	−45[1]	36	150	0	3	105	7	.7	20	5	125	175
		2500	350	−43[1]	47	130	0	3	89	6	.5	20	5	122	200
4-250A	AB_1	1500	600	−95[1]	60	200	−.2	12	64	—	0	35		145	155
		2000	600	−104[1]	55	203	−.2	11	88	—	0	35		175	230
		2500	600	−110[1]	60	215	−.2	7	90	—	0	35		225	313
		3000	600	−116[1]	60	209	−.1	5	93	—	0	35		250	375
	AB_2	1500	300	−48[1]	50	243	0	17	96	11	1.1	35	10	150	214
		2000	300	−48[1]	60	255	0	13	99	12	1.2	35	10	185	325
		2500	300	−51[1]	60	250	0	12	100	11	1.1	35	10	205	420
		3000	300	−53[1]	63	237	0	17	99	10	1	35	10	190	520
304TL	AB_1	1500	—	−105	135	286	—	—	105	—	0	—		300	128
		2000	—	−160	100	273	—	—	160	—	0	—		300	245
		3000	—	−260	65	222	—	—	260	—	0	—		300	365

[1] Adjust to give stated zero-signal plate current.
[2] Single-sideband suppressed-carrier linear amplifier ratings, voice signal.
[3] Due to intermittent nature of voice, average dissipation is considerably less than max.-signal dissipation.

signal applied, the peak input power will be 1.57 times the d.c. power input to the linear amplifier. Should this be different from the design value for the particular linear amplifier, the drive and loading adjustments can be quickly changed in the proper direction (always adjusting the loading so that the peaks of the envelope are on the verge of flattening) and the proper value reached.

As a final check, before coupling the linear amplifier to the antenna, the single-sideband operator will do well to check the linearity of the system, since distortion in the linear amplifier probably will result in the generation of sidebands on the side that was suppressed in the exciter. Here again the two-tone test signal will be of great help, since distortion of the signal will be readily recognized. A check of the bias supply has already been recommended. The next most likely form of distortion will be caused by curvature of the tube characteristic near cut-off, and will be recognizable from a two-tone test pattern that looks like Fig. 12-6. A slight readjustment of bias (or applying a few volts of positive or negative bias, in the case of zero-bias tubes) will usually straighten out the kink that exists where the pattern crosses the zero axis. Make this adjustment with special care, however, because the dissipation of the tubes with no input signal will be very sensitive to this adjustment. There are a few tubes that will not permit this adjustment to be carried to the point where the kink is entirely eliminated without exceeding the rated plate dissipation.

The antenna may now be coupled to the linear amplifier until the plate input with the excitation as determined above is the same as that obtained with the dummy load. The system has now been adjusted for optimum performance, although it is well to monitor it with a 'scope.

(For further reading on linear amplifiers, see Long, "Sugar-Coated Linear-Amplifier Theory," *QST*, October, 1951, and Ehrlich, "How To Test and Align a Linear Amplifier," *QST*, May, 1952.)

VOICE-CONTROLLED BREAK-IN

Although it is possible for two SSB stations operating on widely different frequencies to work "duplex" if the carrier suppression is great enough (inadequate carrier suppression would be a violation of the FCC rules), most SSB operators prefer to use voice-controlled break-in and operate on the same frequency. This overcomes any possibility of violating the FCC rules and permits three or more stations to engage in a "round table."

Many various systems of voice-controlled break-in are in use, but they are all basically the same. Some of the audio from the speech amplifier is amplified and rectified, and the resultant d.c. signal is used to key an oscillator and one or more stages in the SSB transmitter and "blank" the receiver at the time that the transmitter is on. Thus the transmitter is on at any and all times that the operator is speaking but is off during the intervals between sentences. The voice-control circuit must have a small amount of "hold" built into it, so that it will hold in between words, but it should be made to turn on rapidly at the slightest voice signal coming through the speech amplifier. Both tube and relay keyers have been used with good success. Most voice-control systems require the use of headphones by the operator, but a loudspeaker can be used with the proper circuit. (See Nowak, "Voice-Controlled Break-In . . . and a Loudspeaker," *QST*, May, 1951, and Hunter, "Simplified Voice Control with a Loudspeaker," *QST*, October, 1953.)

A Phasing-Type SSB Exciter

The exciter shown in Figs. 12-7, 12-9 and 12-12 is an excellent unit for the amateur who might like to try single-sideband with a minimum of cost and effort. It requires r.f. driving power from one's present exciter and a power supply. It will deliver SSB output in the 3.9-Mc. 'phone band, either to an antenna for local work or to an r.f. amplifier adjusted for linear operation. The operating frequency can be varied over a wide range without seriously impairing the adjustment. Provision is made for transmitting either the upper or the lower sideband.

The schematic of the exciter is shown in Fig. 12-8. Four 6V6 tubes are used as balanced modulators. The plate circuit of the balanced modulators uses a push-pull-parallel arrangement. The grids of one pair of balanced modulators are fed through a phase-shift network consisting of a 300-ohm resistor and an inductance that is adjustable to 300 ohms reactance at the operating frequency. The grids of the second pair of balanced modulators are fed through a phase-shift network consisting of a 300-ohm resistor and a condenser which is adjustable to 300 ohms reactance at the operating frequency. The input impedance of the two phase-shift networks in parallel is 300 ohms.

Each balanced-modulator tube grid is fed through a blocking condenser and provided with grid-leak bias. The bias circuit of each balanced modulator is made adjustable for control of the carrier suppression. Provision is also made for the addition of fixed bias, in case the exciter is used in a voice-controlled circuit where the r.f. excitation is removed during listening periods.

Screen modulation is used, and the screen of each modulator tube is by-passed to ground for r.f. A transformer with a center-tapped secondary is used in the output of each audio amplifier to provide push-pull modulating voltages.

A reversing switch, S_1, allows switching to either the upper or lower sideband. If this switch has a center "off" position, it will facilitate using the "two-tone test" procedure mentioned earlier.

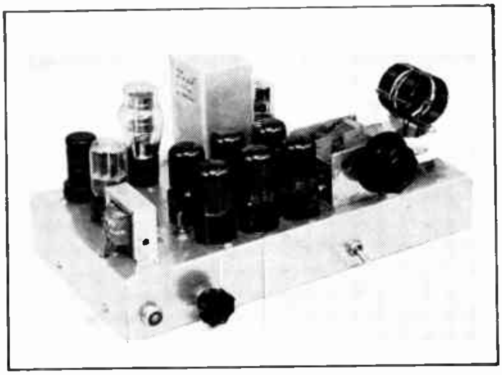


Fig. 12-7 — A small single-sideband exciter that includes voice-controlled break-in. Receiving-type tubes are used throughout.

Microphone input and audio gain control are at the left-hand side of the front — the switch selects the upper or lower sideband. (Revised version, W2UNJ, Aug., 1949, *QST*.)

A voltage divider is inserted between each output of the audio phase-shift network and the corresponding amplifier grid. One of these voltage dividers is made variable to provide for balancing of the two audio channels. The network constants are compensated for the load of these dividers.

Speech Amplifier and Voice Control

The speech amplifier is designed to attenuate both low and high frequencies, amplifying only the audio range required for good intelligibility. The wiring diagram is shown in Fig. 12-11. The output of the speech amplifier is coupled to the input of the audio phase-shift network through a transformer with a center-tapped secondary, to provide push-pull audio for the phase-shift network.

Part of the output of the speech amplifier is taken off through an adjustable voltage-divider circuit and blocking condenser to the voice-control circuit. There it is rectified by the diodes of the 6SQ7, and the resulting d.c. voltage is used to charge C_{14} negative. An audio choke prevents audio components from appearing across C_{14}. The triode section of the 6SQ7 is normally conducting and holding the relay closed, but when the negative voltage appears across C_{14} the 6SQ7 plate current is cut off and the relay opens. When the audio signal is removed, C_{14} discharges through R_{15} and the triode again conducts, closing the relay.

The Audio Phase-Shift Network

The audio phase-shift network requires close matching of resistance and capacity values and, to do this economically, advantage is taken of the fact that resistors and condensers in junk boxes and in stock at local dealers vary considerably from their nominal values.

Table 12-II is used in selecting the network components. The procedure is to collect as many resistors and condensers as possible with nominal values as indicated in the second column of the chart. Measure all of the condensers first, and select the six condensers whose measured values are closest to the "target values" in the third column. Enter the measured values of these condensers in the fourth column of the chart. Then calculate the "target values" for the resistors and select the six resistors whose measured values are closest to these target values.

A capacity bridge, of the type used by servicemen, and a good ohmmeter should give sufficient accuracy in selecting the network components. Absolute accuracy is not important, if the components are all in correct proportion to each other. A difference in percentage error between the resistance measurements and the capacitance measurements will merely shift the operating range of the network. The network components are mounted on a small sheet of insulating material to facilitate wiring.

Networks already adjusted are available through radio dealers — they can be used in this exciter to simplify the construction. The necessary circuit modifications for using a commercial network are shown in Fig. 12-10.

Construction

The exciter and its associated audio equipment are assembled on a 13 by 17 by 2-inch aluminum chassis. The four 6V6 balanced-modulator tubes are arranged in a square pattern toward the front center of the chassis, with the plate tuning condenser and coil off to one side and the 6K6 audio amplifier tubes on the other. The two modulation transformers are under the chassis directly below the plate tuning condenser. The speech amplifier is arranged along the left-hand side of the chassis, with the 6SJ7 at the rear and the output transformer on the top of the chassis at the front. The audio phase-shift network is below the output transformer.

The reactive components of the r.f. phasing network, L_1 and C_7, are mounted in a plug-in

TABLE 12-II
Phase-Shift Network Design Data

Part	*Nominal Value*	*Target Value*	*Measured Value*
C_1	0.001	0.00105	(C_{m1})
C_2	0.002	0.00210	(C_{m2})
C_3	0.006	0.00630	(C_{m3})
C_4	0.005	0.00475	(C_{m4})
C_5	0.01	0.00950	(C_{m5})
C_6	0.03	0.0285	(C_{m6})
R_1	100,000	$\frac{100}{C_{m1}} =$	
R_2	50,000	$\frac{105}{C_{m2}} =$	
R_3	15,000	$\frac{100}{C_{m3}} =$	
R_4	100,000	$\frac{453}{C_{m4}} =$	
R_5	50,000	$\frac{476}{C_{m5}} =$	
R_6	15,000	$\frac{453}{C_{m6}} =$	

All condensers mica, all resistors 1 watt.

shield can that mounts directly behind the balanced-modulator tubes. The shield can is grounded to the chassis through the spare pins of its plug. The voltage regulator tube is mounted to the left of the shield can, and the 6SQ7 voice-control tube is to the right. The components in the voice-control circuit are mounted under the chassis at the rear.

Associated Equipment

The r.f. input impedance of the exciter is 300 ohms, but a link line of lower characteristic impedance will operate satisfactorily for the short distance usually required. A means for adjusting the r.f. driving power is desirable. A surplus Command set transmitter (BC-696 or T-19/ARC-5), operating at low plate voltages, makes an ideal r.f. source, but any VFO or crystal oscillator with a few watts output will do.

The plate voltage for the speech amplifier must not be taken from the same point in the power supply that furnishes voltage for the 6K6 amplifiers, since interaction may occur that will upset the phase relationship at the output of the two 6K6s. If separate plate voltage sources are not available, an added filter section may be used to isolate the voltage to the speech amplifier.

The built-in voice-controlled relay can be used

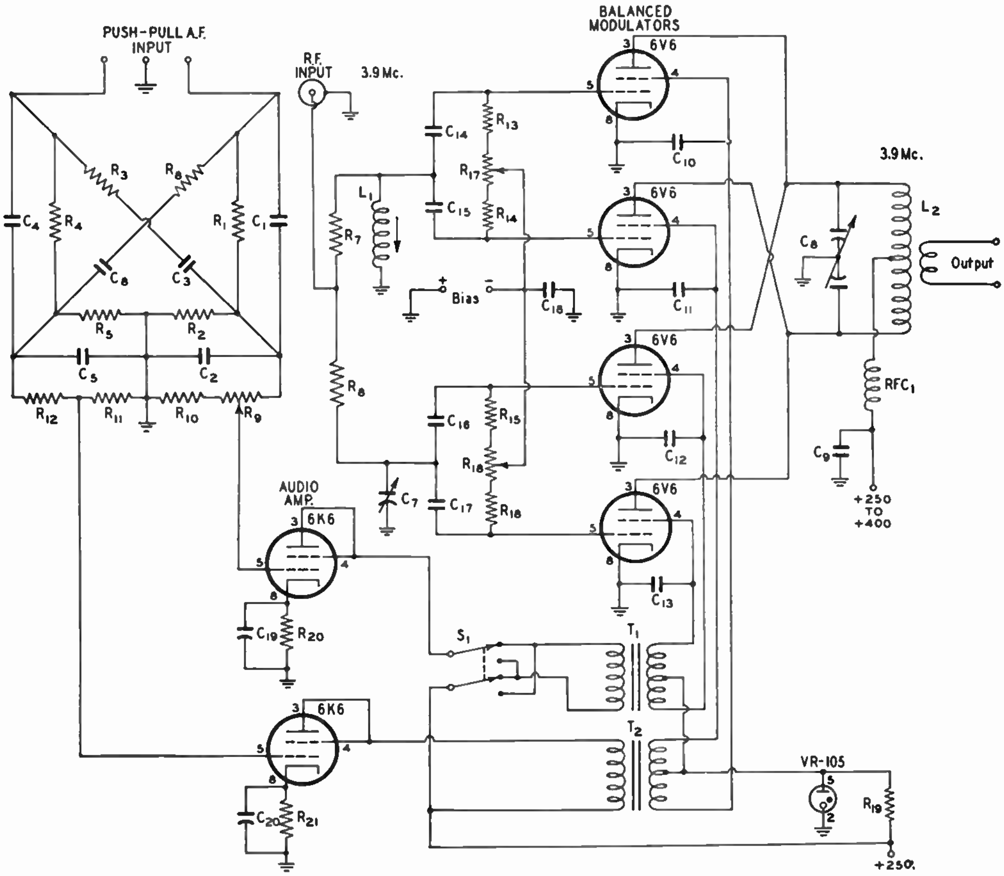


Fig. 12-8 — Circuit diagram of the single-sideband exciter.

C_1–C_6 — See Table 12-II.
C_7 — Air padder condenser. 3.9 Mc.: 150 $\mu\mu$fd.; 7 Mc.: 100 $\mu\mu$fd.; 14 Mc.: 35 $\mu\mu$fd.
C_8 — Approx. 400-$\mu\mu$fd. per section, b.c. receiver tuning condenser.
C_9 — 0.001-μfd. 1000-volt mica.
C_{10}–C_{18} — 0.001-μfd. 500-volt mica.
C_{19}, C_{20} — 4-μfd. 150-volt electrolytic.
R_1–R_6 — See Table 12-II.
R_7, R_8 — 300 ohms, 5 watts (5 1500-ohm 1-watt in parallel).
R_9 — 0.5-megohm linear volume control.
R_{10} — 0.47 megohm.
R_{11} — 0.75 megohm.
R_{12} — 0.24 megohm.
R_{13}–R_{16} — 10,000 ohms.
R_{17}, R_{18} — 15,000-ohm potentiometer, wirewound.
R_{19} — 7500 ohms, 10 watts.
R_{20}, R_{21} — 680 ohms, 2 watts.
All resistors 1-watt unless specified otherwise.
L_1 — 3.9 Mc.: 25 turns No. 28 enam.
7 Mc.: 18 turns No. 22 enam.
14 Mc.: 12 turns No. 20 enam.
All coils close-wound at mounting end of slot of National XR-50 slug-tuned form.
L_2 — 3.9 Mc.: 40-meter 75-watt tank coil with swinging link (Bud OLS-40).
7 Mc.: 20-meter 75-watt tank coil with swinging link (Bud OLS-20).
14 Mc.: 15-meter 75-watt tank coil with swinging link (Bud OLS-15).
RFC_1 — 2.5-mh. r.f. choke.
S_1 — D.p.d.t. toggle, preferably with center off. See text.
T_1, T_2 — 5-watt modulation transformer, 10,000 ohms c.t. to 4000 ohms (Stancor A-3812).

in a number of ways to provide the rapid voice break-in commonly used on 3.9-Mc. SSB 'phone. If a good c.w. break-in system is already in use at the station, the voice-control relay contacts may be substituted for the key, and no other changes are necessary.

If the local oscillator in the receiver will key in the plate voltage lead satisfactorily, then a simple voice break-in system may be obtained by using the relay contacts to shift the plate voltage from the receiver local oscillagor to the VFO. A drifting receiver oscillator must be avoided in this system, however.

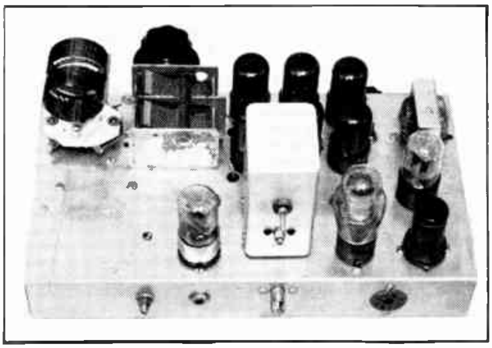


Fig. 12-9—**A rear view of the phasing-type exciter. The two r.f. phasing adjustments project from the shield can. The potentiometer shaft at the left sets the voice-control threshold level. The jack is for the keyed circuit, the r.f. connector takes the excitation cable, and the octal socket is for the power cable.**

Operating Conditions

If voice control is not used, and d.c. operating voltages are removed when excitation is removed for stand-by, then no fixed bias is required on the balanced modulators and a jumper can be placed across the bias terminals. When excitation is removed with d.c. voltages applied, as in voice-controlled operation, then 4½ volts of fixed bias should be used to limit the plate and screen currents on the balanced modulators.

With 400 volts applied to the balanced-modulator plates and 250 volts to all other plate supply inputs, the operating currents will be approximately as follows:

Total balanced-modulator plate current	85 ma.
VR tube supply current	20 ma.
Total 6K6 amplifier current	62 ma.
Total speech-amplifier current	12 ma.

The total balanced-modulator grid current, measured at the bias terminals, will vary with excitation, but it should be in the range 3 to 5 ma.

These currents will not change appreciably with varying audio input and, with the exception of the grid current, will not change appreciably when the excitation is removed, provided that 4½ volts of fixed bias is used on the balanced-modulator grids.

The exciter may be coupled directly to an antenna for use as a low-power transmitter, but most amateurs will wish to use it to drive a buffer or final amplifier. All stages following the exciter must be operated under Class A, AB, or B conditions. In general, the correct operating conditions for stages following the exciter may be found by referring to the audio operating conditions for the tube under consideration. Grid-bias and screen voltages should have very good regulation. For amateur voice operation, tubes may be operated considerably beyond the ratings given in the tube manuals, as discussed earlier. When the r.f. amplifier is operated Class AB_2, the grid tank circuit will require shunting by a resistor in order to provide better regulation of the exciting voltage. The value of this resistor is usually determined by experiment. It should be as small as possible consistent with good output and linear operation.

Adjustment

Adjustment of the exciter is best made under actual operating conditions. Connect the exciter to the transmitter, load the exciter with a dummy load, apply r.f. excitation, feed sine-wave audio into the speech amplifier, and tune the output tuning condenser in the conventional way for maximum output.

Reduce the audio input to zero, and adjust potentiometers R_{17} and R_{18} for minimum carrier output. Minimum carrier output may be determined by any sensitive r.f. indicator coupled to the final-amplifier plate circuit. A 0–1 milliammeter, in series with a crystal detector and a two-turn coupling loop, will make a satisfactory indicator. The meter should be by-passed with a 0.005-μfd. condenser. If a null indication cannot be obtained within the range of the potentiometers, the 6V6 tubes are not evenly matched. Exchanging the positions of the 6V6s may aid in obtaining the balance, or other tubes may have to be used.

After the carrier balance is obtained, tune in the r.f. source on the station receiver, and with the antenna terminals shorted, and the crystal selectivity in sharp position, adjust the crystal phasing to the point where only one sharply-peaked response is obtained as the receiver is tuned through the signal. Now apply sine-wave audio of about 1500-cycle frequency to the speech amplifier, and find the two sidebands on the receiver. Three distinct peak indications will be observed on the S-meter as the receiver is tuned. Set the receiver on the weaker of the two side-

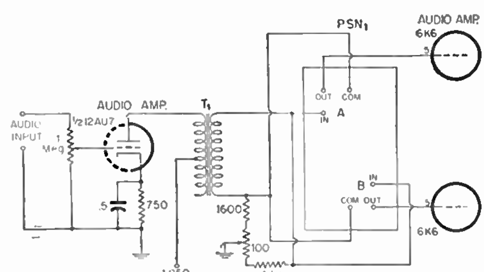


Fig. 12-10—**Circuit revisions for including a commercial phase-shift network in the SSB exciter of Fig. 12-8.**
***PSN_1*—Phase-shift network (Millen 75012 Phasing Unit).**
***T_1*—3:1 audio transformer, reversed (Stancor A-53-C).**

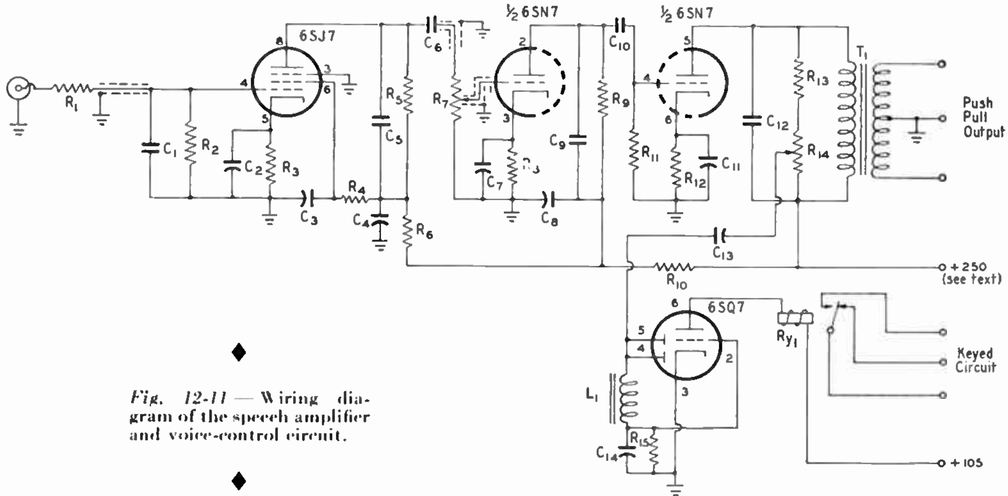


Fig. 12-11 — Wiring diagram of the speech amplifier and voice-control circuit.

C_1 — 100-μμfd. mica or ceramic.
C_2, C_7, C_{11} — 4-μfd. 150-volt electrolytic.
C_3 — 0.02-μfd. 400-volt paper.
C_4, C_8 — 8-μfd. 450-volt electrolytic.
C_5 — 270-μμfd. mica or ceramic.
C_6 — 0.001-μfd. mica or ceramic.
C_9 — 0.0033-μfd. mica or ceramic.
C_{10} — 0.002-μfd. mica or ceramic.
C_{12} — 0.005-μfd. ceramic or mica.
C_{13} — 0.01-μfd. 400-volt paper or ceramic.
C_{14} — 0.5-μfd. 200-volt paper.
R_1, R_9 — 0.1 megohm.
R_2 — 2.2 megohm.
R_3, R_{12} — 910 ohms.
R_4 — 1.0 megohm.
R_5 — 0.27 megohm.
R_6 — 27,000 ohms.
R_7 — 0.5-megohm volume control.
R_8 — 2700 ohms.
R_{10}, R_{13} — 10,000 ohms, 1 watt.
R_{11}, R_{15} — 0.47 megohm.
R_{14} — 15,000-ohm volume control.
All resistors ½-watt unless specified otherwise.
T_1 — 5-watt modulation transformer, 10,000 ohms c.t. to 4000 ohms (Stancor A-3812).
L_1 — Small filter or audio choke (Stancor C-1707).
Ry_1 — Sensitive 10,000-ohm relay.

bands and adjust L_1, C_7 and R_9 for minimum sideband strength. If suppression of the other sideband is desired, throw S_1 to its other position. A dip obtained with one set of adjustments is not necessarily the minimum. Other combinations should be tried. The final adjustment should give S-meter readings for the two sidebands which differ by at least 30 db. The bias voltage on all four balanced modulator tubes will be approximately equal.

After the adjustments have been completed, the r.f. drive to the exciter should be adjusted to the point where a decrease in drive will cause a decrease in output, but an increase in drive will not cause an increase in output. The complete adjustment procedure should then be rechecked. The rig is then ready for a microphone, an antenna, and an on-the-air test.

If an oscilloscope is available, a simpler and more reliable adjustment procedure may be used. Either linear or sine-wave horizontal sweep may be used on the oscilloscope. The vertical input should be coupled to the output of the transmitter in the same manner as is used for observing amplitude modulation. The sine-wave audio-frequency input to the speech amplifier should be any convenient multiple of the oscilloscope sweep frequency. A 60-cycle sweep frequency and a 600-cycle audio frequency are commonly used.

When the exciter is modulated with a single sine-wave audio frequency, the output should be a single radio frequency. Therefore, the oscilloscope should show a straight-edged band across the screen, the same indication as is given by an unmodulated carrier. This is illustrated in Fig. 12-13. If carrier output, or unwanted sideband output, is present, it will be indicated by "ripple" on the top and bottom edges of the oscilloscope picture. A small amount of ripple can be tolerated, but if the exciter is badly out of adjustment, the output will appear to be heavily modulated. Adjustment with the 'scope is accomplished by adjusting all controls to obtain the smallest possible amount of ripple. The oscil-

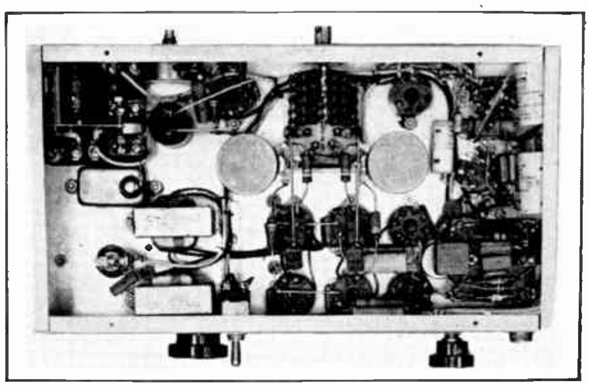


Fig. 12-12 — Underneath the chassis of the exciter. The two potentiometers are the bias balancing controls, R_{17} and R_{18}.

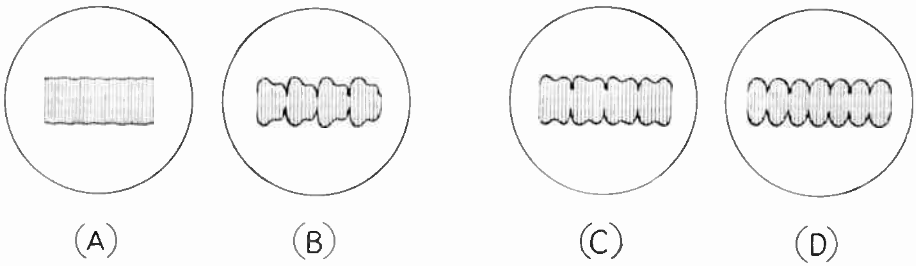


Fig. 12-13 — Sketches of the oscilloscope face showing different conditions of adjustment of the exciter unit. (A) shows the substantially clean carrier obtained when all adjustments are at optimum and a sine-wave signal is fed to the audio input. (B) shows improper r.f. phase and unbalance between the outputs of the two balanced modulators. (C) shows improper r.f. phasing but outputs of the two balanced modulators equal. (D) shows proper r.f. phasing but unbalance between outputs of two balanced modulators.

loscope may also be used for continuous monitoring during transmissions to avoid overloading of any stage of the transmitter. Overloading is indicated by a flattening of the modulation-peak patterns at the top and bottom. In observing these patterns, it is difficult to separate the effects of sideband and carrier suppression. However, considered separately, sideband or carrier suppression of 30 db. would give a 3 per cent ripple, 25 db. a ripple of 6 per cent, and 20 db. a 10 per cent ripple. Harmonics present in the audio modulating signal will modify the results and invalidate this test if they run more than 1 per cent, hence it is essential that the audio signal be as pure as possible.

The exciter is capable of driving any pair of beam tubes commonly used in amateur transmitters, or any pair of triodes in Class AB_1. A buffer stage will ordinarily be required to drive Class B triodes.

A Crystal-Filter SSB Exciter

The exciter uses a quartz crystal filter operating at 450 kc. (or vicinity). The filter allows a passband of 300 to 3000 cycles; the sideband rejection should run 35-40 db. over 300 to 3000 cycles. At no time within the reject range is the rejection less than 30 db.; at some places it approaches 60 db. Crystals suitable for use in the filter are available on the war surplus market for less than one dollar each. The most useful of these crystals are in the series that runs from 375 to 525 kc. in 1.388-kc. steps; this series is marked at 72 times the crystal frequency in a series of channels from 28.0 to 38.0 Mc. The holder pins have ½-inch spacing. The crystals may be socket-mounted or soldered directly into the filter at the builder's discretion.

The filter is of bridge design with complex entry and terminating sections. The complex sections are used to suppress the carrier and modify the response characteristics of the bridge. Fig. 12-14 shows the filter proper, set for rejection of the upper sideband. The transformers T_1 and T_2 are replacement-type 455-kc. i.f. transformers, mica-tuned and air-cored.

The original filter was designed to operate at a carrier frequency of 450 kc., although the filter will work at frequencies between 425 and 490 kc. without alteration of the circuit or transformers. Under the condition of design for 450-kc. carrier, crystal "B" is 2.78 kc. *higher* than 450 kc., or 2 channels higher in the crystal series. Crystal "C" is 1.39 kc. *lower* than 450 kc., or 1 channel lower. Crystal "D" is 450 kc. Crystal "A," also at 450 kc., is used in a crystal oscillator to generate the initial carrier. Channel markings on these crystals are as follows:

"A" — 32.4 Mc., Channel 324
"B" — 32.6 Mc., Channel 326
"C" — 32.3 Mc., Channel 323
"D" — 32.4 Mc., Channel 324

Any other group within the range of the i.f. transformers may be utilized; only the channel relationship need be retained.

A diagram of the exciter proper is shown in Fig. 12-15. The 6K8 hexode-triode serves as 450-kc. oscillator and audio mixer. Approximately 3 volts of audio is required at the signal grid of

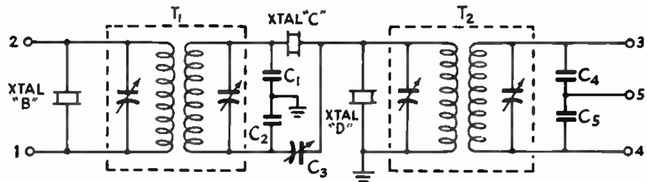


Fig. 12-14 — The 450-kc. crystal filter used for sideband and carrier rejection.
C_1, C_2, C_4, C_5 — 100-$\mu\mu$fd. mica or ceramic.
C_3 — 3- to 30-$\mu\mu$fd. ceramic trimmer.
T_1 — 455-kc. interstage i.f. transformer (Meissner 16-6659 or Miller 312-C2).
T_2 — 455-kc. diode i.f. transformer (Meissner 16-6660 or Miller 312-C4).
For a carrier frequency of 450 kc., the crystals are:

Crystal	*B*	*C*	*D*
High-freq. reject	452.8 kc.	448.6 kc.	450.0 kc.
Low-freq. reject	447.2 kc.	451.1 kc.	450.0 kc.

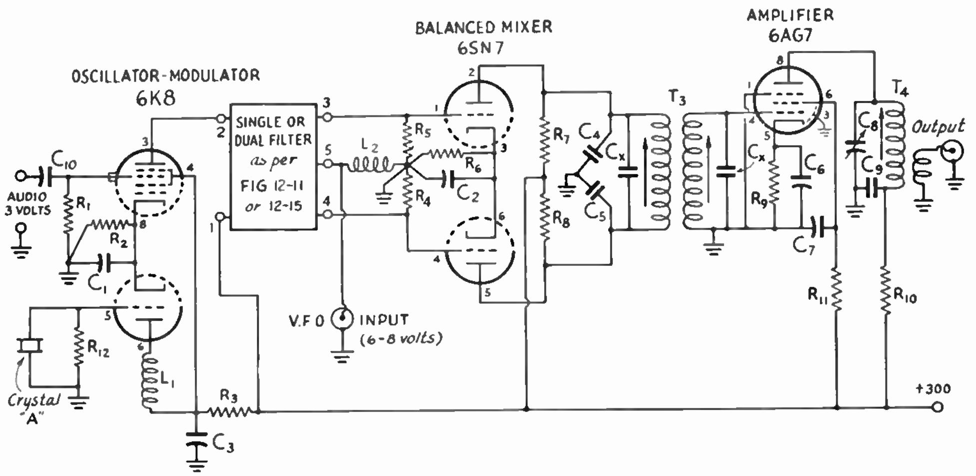


Fig. 12-15 — Complete diagram of the crystal-filter SSB exciter.

C_1, C_2, C_3, C_6, C_7 — 0.1-μfd. 400-volt paper.
C_4, C_5 — 39-μμfd. ceramic.
C_8 — 100-μμfd. variable air condenser, mounted in shield can with T_4.
C_9 — 0.02-μfd. 600-volt mica.
C_{10} — 0.01-μfd. 400-volt paper.
C_X — Trimmers in T_3.
R_1 — 0.47 megohm.
R_2 — 220 ohms.
R_3, R_{11} — 20,000 ohms, 1 watt.
R_4, R_5 — 0.1 megohm.
R_6, R_7, R_8 — 10,000 ohms.
R_9 — 150 ohms, 1 watt.
R_{10} — 1000 ohms.
R_{12} — 47,000 ohms.
All resistors ½ watt unless specified otherwise.
L_1 — 2.5-mh. r.f. choke.
L_2 — 0.5-mh. r.f. choke.
T_3 — 1.5-Mc. slug-tuned i.f. transformer (Miller No. 1166).
T_4 — 1.5-Mc. slug-tuned i.f. transformer. Secondary removed and 8-turn link wound over cold end of primary. All fixed capacitors removed.

the 6K8 for optimum results. The 6K8 delivers a carrier (450 kc.) and sidebands to the input of the filter. The filter rejects one sideband (depending upon the selection of crystals) and delivers single-sideband energy to the 6SN7 mixer. The filter also suppresses the carrier. The 6SN7 mixer combines the single-sideband energy (in the vicinity of 450 kc.) with the output of the VFO (3400 to 3550 kc.) and the sum products are recovered in the output (3850 to 4000 kc.). A VFO signal of about 6 to 8 volts is required. The output of the mixer is fed to the grid of a Class A 6AG7 tuned r.f. amplifier. The output is sufficient to drive a pair of 807s Class AB_2. Operation on 10 and 20 meters can be accomplished by heterodyning again to the desired band. Most VFOs in use cover or may be easily made to cover 3400 to 3550 kc. A single untuned 6AC7 amplifier following a BC-221 might be used as a VFO for this exciter.

Construction

The original transmitter was built for mobile operation and much hole drilling and experimentation has occurred on the chassis. Mounting the crystals on opposite sides of the transformers will keep stray capacity coupling at a minimum. No shielding other than that provided by the i.f. cans and the output tank can is required. It is important that capacity coupling *around* the crystal filter be minimized — no modulated signal must reach the 6SN7 mixer by any route except through the filter. If choice of sidebands is desired, a dual filter using 5 crystals will be required. This filter is shown in Fig. 12-17. A double-section wafer switch selects the upper or lower sideband. These wafer sections must be separated by approximately 3 inches to minimize stray coupling. It is recommended that the crystals be wrapped with several layers of adhesive tape and then strapped to the chassis with metal brackets; connections may then be made by soldering to the holder pins.

Alignment

Alignment of the filter is straightforward, and once aligned it will need little attention.

1) Crystal "A" is first removed from the circuit. This crystal is best provided with a socket mount so it can be removed during alignment.

Fig. 12-16 — The crystal-filter SSB exciter, as designed for mobile work, complete with receiver converter and VFO. The top dish is the exciter (with cover removed). The meter reads cathode current to a pair of 807s driven by the unit, and the two knobs handle carrier reinsertion and 6AG7 plate tuning. (W1JEO/9, Nov., 1950, *QST*.)

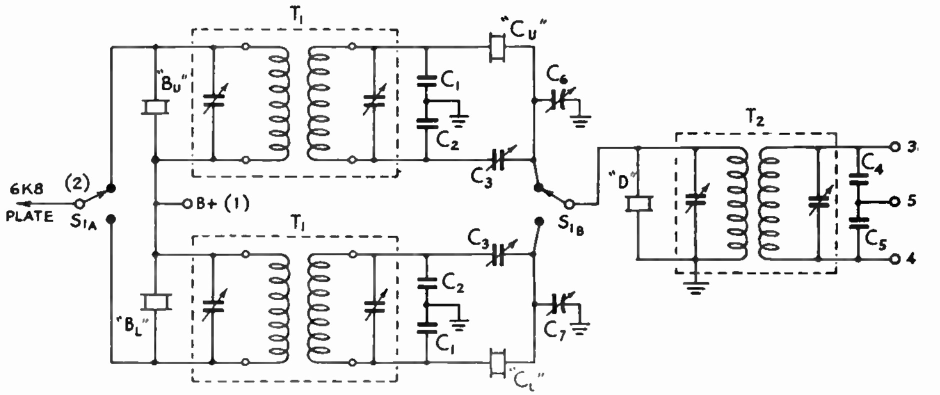


Fig. 12-17 — The double-channel crystal filter. All components are the same as in Fig. 12-14, except for the addition of the d.p.d.t. wafer switch, S_1, and the compensating condensers, C_6 and C_7 (3- to 30-μμfd. ceramic). The trimmer on the input side of T_2 is set at minimum and the alignment procedure is followed with C_6 or C_7 wherever the instructions call for adjusting the input condenser.

2) A calibrated signal generator covering the crystal range is connected to the grid of the triode section of the 6K8.

3) A vacuum tube voltmeter is connected from grid to ground of one of the 6SN7 grids.

4) Swing the signal generator through the crystal range until a maximum response is noted at the voltmeter. This will indicate the series-resonant frequency of crystal "C" and with the crystals described, based on a 450-kc. carrier, will be approximately 448.6 kc.

5) Align all transformer trimmers for maximum response on this frequency.

6) Next, adjust the signal generator *slowly* in the higher-frequency direction until a *null* is obtained. This will be the series-resonant frequency of crystal "D," 450 kc. with the crystals indicated.

7) Move the signal generator ½ kc. *lower* than this null and adjust the trimmer on the input side of T_2 for maximum response.

8) Return signal generator to null.

9) Move the signal generator approximately 1 to 1.2 kc. *higher* than the null and adjust C_3 for *minimum* response.

10) Move the signal generator *higher* until another null is found: this will be the series-resonant frequency of crystal "B," approximately 452.8 kc. with the crystals shown.

11) Continue approximately ½ kc. higher than this null and adjust the output trimmer on T_1 slightly for moderate null.

12) Repeat Steps 7 through 11 to compensate for interaction, and alignment is complete.

For alignment of the dual filter the procedure is identical but must be done once for each sideband. However, when adjusting the filter for rejecting the *lower* sideband and where Steps 1–12 mention "higher" insert "lower" and vice versa. The alignment chart, Fig. 12-18, will simplify the alignment procedure. For additional information, see Webb, "Aligning the Crystal-Filter S.S.B. Exciter," *QST*, August, 1952.

The slug-tuned i.f. transformer is peaked at 3930 kc. and then stagger-tuned slightly to provide coverage of the entire band. The 6AG7 plate tank capacitor is adjustable from the front panel and is touched up when shifting frequency. It is important that the 6AG7 amplifier stage be stable and show no tendency to oscillate. Good shielding of its grid and plate circuits is usually required, and occasionally a 47-ohm composition resistor from T_3 to Pin 4 of the 6AG7 may be necessary to avoid regeneration.

Many variations of this basic exciter circuit are possible. If a balanced modulator (using a pair of 6K8s) is used, the carrier suppression is readily obtained without close matching of crystals. Other filter circuits can be used, as those shown in Good, "Crystal Filter for 'Phone Reception," *QST*, October, 1951. For an advanced design for a crystal-filter SSB exciter, see Weaver & Brown, "Crystal Lattice Filters for Transmitting and Receiving," *QST*, August, 1951.

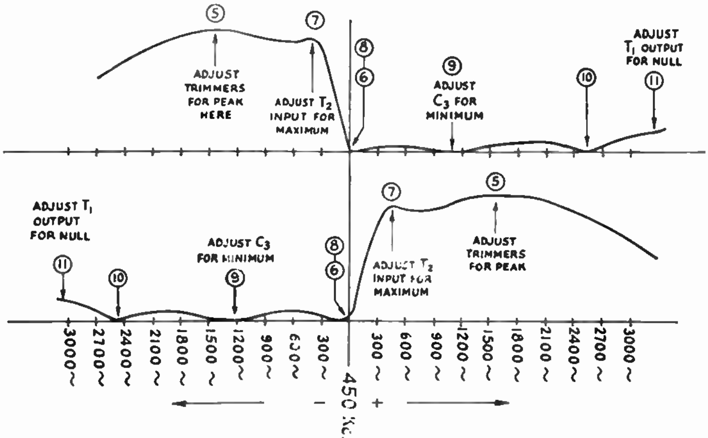


Fig. 12-18 — An alignment chart of the crystal filter. The numbers in the circles correspond to the steps outlined in the text.

A Two-Stage Linear Amplifier

The amplifier shown in Figs. 12-19 and 12-21 is designed to follow a low-powered SSB exciter. An 807 Class A driver is used to excite a pair of 811-As operating Class B. Only a few watts is required to drive the 807. The 811-As will deliver about 180 watts peak with 1000 volts on the plates and 250 watts peak at 1200 volts. Operation as a linear amplifier for SSB with 1500 volts on the plates is not recommended because the driver stage is likely to introduce too much distortion, although a small amount of fixed bias (3–4½ volts) on the grids of the 811-As will permit c.w. operation at this higher plate voltage.

The circuit is not unlike ordinary Class C practice, except for the bias voltages involved. The 807 stage uses cathode bias, and the 811-As run with zero bias (bias terminals short-circuited by a jumper wire). The 807 loading is adjusted by varying the position of the link coil in L_3 and the link to L_6 is controlled from the front panel.

A low-inductance by-pass condenser, C_2, made from a piece of coaxial line, helps to eliminate parasitics in the 807 stage, as does returning the screen by-pass condenser, C_3, to the cathode instead of to ground. Grid chokes, L_4 and L_5 and resistors R_3, R_4 and R_5 were found necessary to avoid high-frequency parasitic oscillations in the 811-A stage. All wiring other than r.f. was run in shield braid. Filament by-pass condensers were unnecessary.

Construction

The amplifier is built on a 13 by 17 by 3-inch aluminum chassis. The panel is an aluminum relay-rack panel, 15¾ inches high, held to the chassis by the shaft bearings and meters, and further braced by two strips of 1/16 by ½-inch brass.

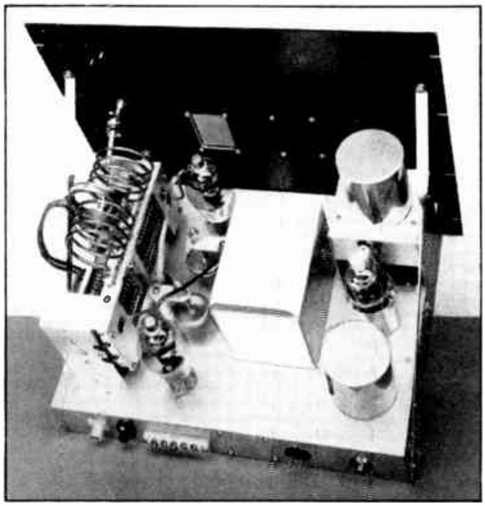


Fig. 12-19 — A rear view of the linear amplifier, showing the push-pull 811-A output amplifier and the 807 driver. The cover of the rectangular shield can slides off for access to the final grid coil. The round shield cans are for the 807 grid and plate coils.

The grid coil for the 807 plugs in to a socket mounted at the rear of the chassis and shielded by an ICA No. 1549 3-inch diameter aluminum shield can.

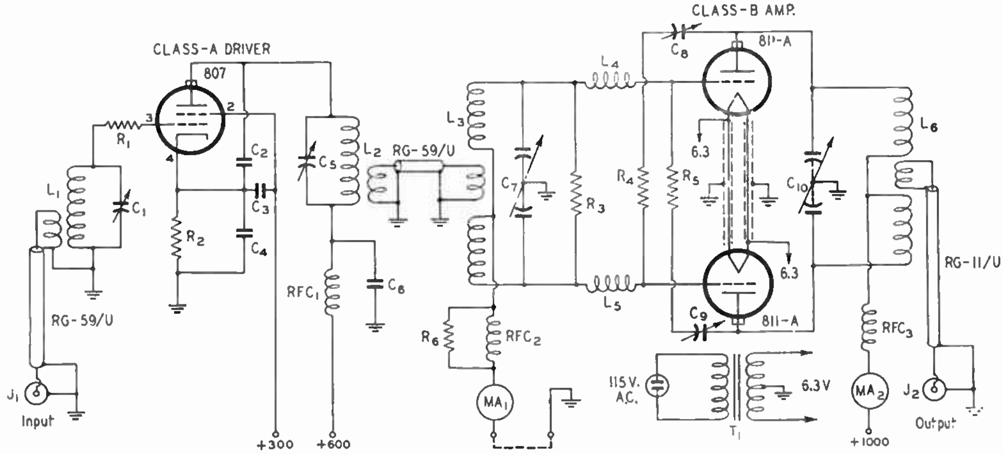


Fig. 12-20 — Wiring diagram of the linear amplifier.

C_1 — 140-μμfd. variable (Millen 19140).
C_2 — 13-μμfd. tubular, made of RG-58/U. Active length, 6 inches.
C_3, C_4 — 0.005-μfd. disc ceramic.
C_5 — 140-μμfd. variable (Millen 22140).
C_6 — 0.001-μfd. 1200-volt mica.
C_7 — Dual variable, 100-μμfd. per section (Millen 24100).
C_8, C_9 — Disc-type condenser (Bud NC-853).
C_{10} — Dual variable, 200-μμfd. per section, 0.077-inch spacing (National MC-200D).
R_1 — 100 ohms, ½ watt,
R_2 — 680 ohms, 2 watts.
R_3 — 2700 ohms, 4 watts (4 2700-ohm in series-parallel).
R_4, R_5 — 20 ohms, 2 watts.
R_6 — 1000 ohms, 1 watt.
All resistors are composition, not wirewound.
L_4, L_5 — 9 turns No. 12 enam., ½-inch diameter, 1¼ inches long.
MA_1 — 0–50 milliammeter.
MA_2 — 0–500 milliammeter.
RFC_1 — 2.5-mh. 125-ma. r.f. choke.
RFC_2 — 250-μh. 75-ma. r.f. choke (Millen 34300).
RFC_3 — 5-mh. 300-ma. r.f. choke (National R300S),

The plate coil plugs in to a socket mounted 4 inches above the chassis. The platform for the socket also shields the plate condenser, C_5. Another 3-inch diameter shield can protects the 807 plate coil. The plate by-pass condenser, C_6, is mounted under the chassis near the 807 socket, and the lead from C_5 and L_2 is brought down to it in shielded wire.

The grid coil for the 811-As is shielded by an ICA No. 29842 4 by 5 by 6 aluminum cabinet. To speed coil changing, the cabinet is fastened to the chassis and a friction-fit cover is made from a piece of sheet aluminum. The inside lips on the top of the cabinet should be bent down to allow more room for the hand that changes coils.

COIL TABLE FOR TWO-STAGE LINEAR AMPLIFIER

Band	*Turns*	*Wire No.*	*Diam.*	*Length*	*μh.*	*Link*	*Spacing*
L_1*							
3.9	22½	20 enam.	1	¾	10	4	1/16
14	10½	20 enam.	1	¾	2.5	3	1/16
L_2**							
3.9	25	20 enam.	1	⅞	11.2	4	1/16
14	11	20 enam.	1	¾	2.5	3	⅛
L_3***							
3.9	22	22 enam.	1¼	1¼	9.1	6	Adjustable
14	12	18 enam.	1¼	1⅛	3.3	4	Adjustable
L_6****							
3.9	22	16 enam.	2½	2¼	20	3	Adjustable
14	8	.15 tubing	2½	3¾	2.3	3	Adjustable

* Wound on Millen 15004 plug-in form.
** Wound on Millen 15005 plug-in form.
*** National AR-16-40S and AR-16-20S. 75-meter coil shunted by 150-μμfd. mica condenser.
**** B & W 80TVL with 18 turns removed, and B & W 15TVL.

The output tank condenser, C_{10}, is mounted on the chassis with aluminum brackets that also support the jack bar for the output coil, L_6. The variable link mounts on the jack bar.

Adjustment

With a signal from the exciter coupled through J_1, and plate and screen voltages on the 807, it should be quite possible to drive the 811-A grid current off scale (with no plate voltage on the 811-As). Back off the excitation to about 25 ma. grid current and neutralize the 811-A stage by adjusting C_8 and C_9. Use the "flick" in grid current as C_{10} is tuned through resonance, or a more sensitive indication such as a crystal diode and 0–1 milliammeter connected to J_2.

Couple a dummy load to J_2 and apply plate voltage to the 811-As. Couple an oscilloscope to the dummy load and apply a "two-tone" test signal, as described earlier in this chapter. The 811-A no-signal plate current should run around 40 or 50 ma., depending upon the plate voltage. Adjust the two-tone signal amplitude for 10 or 15 ma. grid current and resonate all circuits. Then increase the excitation until the two-tone pattern just begins to flatten on the peaks. When using 1000 volts on the 811-As, this flattening should not occur before MA_2 indicates 160 ma. or so — with 1200 volts the current should run up to 190 ma. without noticeable flattening. If flattening occurs sooner, it indicates that the 811-A stage should be coupled more tightly to its load, or that the 807 stage is not delivering enough drive to the 811-A grids.

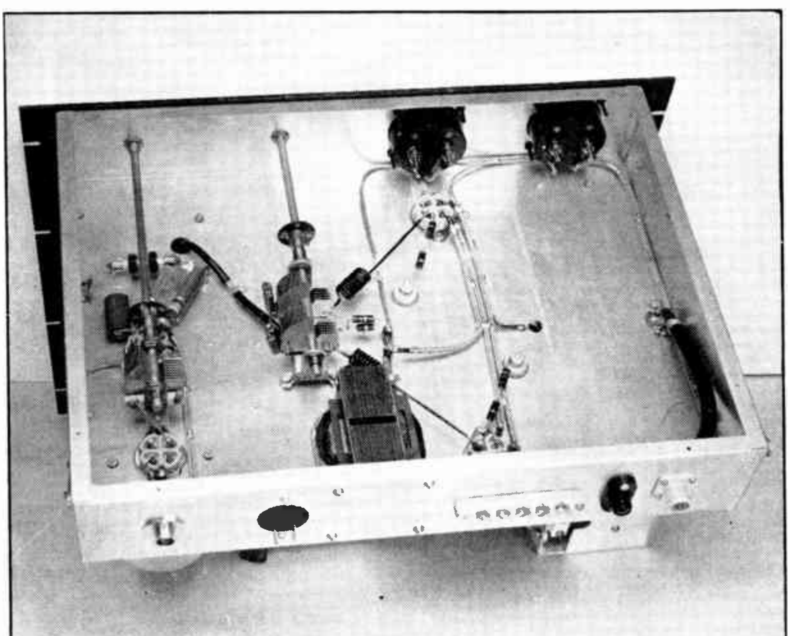


Fig. 12-21—Underneath the chassis, showing all but r.f. leads in shield braid. The coils in the leads from the split-stator grid condenser are parasitic chokes.

CHAPTER 13

Transmission Lines

The place where r.f. power is generated is very frequently not the place where it is to be utilized. A transmitter and its antenna are a good example: The antenna, to radiate well, should be high above the ground and should be kept clear of trees, buildings and other objects that might absorb energy, but the transmitter itself is most conveniently installed indoors where it is readily accessible. There are many other instances where power must be delivered from one point to another.

The means by which power is transported from point to point is the r.f. transmission line. At radio frequencies a line exhibits entirely different characteristics than it does at commercial power frequencies. This is because the speed at which electrical energy travels, while tremendously high as compared with mechanical motion, is not infinite. The peculiarities of r.f. transmission lines result from the fact that a time interval comparable with an r.f. cycle must elapse before energy leaving one point in the circuit can reach another just a short distance away.

Operating Principles

Suppose we have a battery and a pair of parallel wires extending to a very great distance. At the moment the battery is connected to the wires, electrons in the wire near the positive terminal will be attracted to the battery, and the same number of electrons in the wire near the negative battery terminal will be repelled outward along the wire.

Thus a current flows in each wire near the battery at the instant the battery is connected. However, a definite time interval will elapse before these currents are evident at a distance from the battery. The time interval may be very small. For example, one-millionth of a second (one microsecond) after the connection is made the currents in the wires will have traveled 300 meters, or nearly 1000 feet, from the battery terminals.

The current is in the nature of a charging current, flowing to charge the capacitance between the two wires. But unlike an ordinary condenser, the conductors of this "linear" condenser have appreciable inductance. In fact,

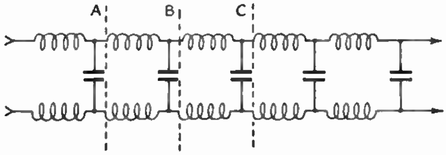


Fig. 13-1 — Equivalent of a transmission line in lumped circuit constants.

we may think of the line as being composed of a whole series of small inductances and capacitances connected as shown in Fig. 13-1, where each coil is the inductance of a very short section of one wire and each condenser is the capacitance between two such short sections.

Characteristic Impedance

An infinitely-long chain of coils and condensers connected as in Fig. 13-1, where each L is the same as all others and all the Cs have the same value, has an important property. To an electrical impulse applied at one end, the combination appears to have an impedance — called the **characteristic impedance** or **surge impedance** — that is approximately equal to $\sqrt{L/C}$, where L and C are the inductance and capacitance per unit length. This impedance is purely resistive.

In defining the characteristic impedance as $\sqrt{L/C}$, it is assumed that the conductors have no inherent resistance — that is, there is no I^2R loss in them — and that there is no power loss in the dielectric surrounding the conductors. In other words, it is assumed there is no power loss in or from the line no matter how great its length. This does not seem consistent with calling the characteristic impedance a pure resistance, which implies that the power supplied is all dissipated in the line. But in an infinitely-long line the effect, so far as the source of power is concerned, is exactly the same as though the power were dissipated in a resistance, because the power leaves the source and travels outward forever along the line.

The characteristic impedance determines the amount of current that can flow when a given voltage is applied to an infinitely-long line, in exactly the same way that a definite value of actual resistance limits current flow when a given voltage is applied.

The inductance and capacitance per unit length of line depend upon the size of the conductors and the spacing between them. The closer the two conductors and the greater their diameter, the higher the capacitance and the lower the inductance. A line with large conductors closely spaced will have low impedance, while one with small conductors widely spaced will have relatively high impedance.

"Matched" Lines

Actual transmission lines do not extend to infinity but have a definite length and are connected to, or **terminate in**, a load at the "output" end, or end to which the power is delivered. If the load is a pure resistance of a value equal to the characteristic impedance of the line, the current traveling along the line to the load does not find conditions changed in the least when it meets the load; in fact, the load just looks like still more transmission line of the same characteristic impedance. Consequently, connecting such a load to a short transmission line allows the current to travel in exactly the same fashion as it would on an infinitely-long line.

In other words, a short line terminated in a purely-resistive load equal to the characteristic impedance of the line acts just as though it were infinitely long. Such a line is said to be **matched.** In a matched transmission line, power travels outward along the line from the source until it reaches the load, where it is completely absorbed.

R.F. on Lines

The discussion above, although based on direct-current flow from a battery, also holds when an r.f. voltage is applied to the line. The difference is that the alternating voltage causes the amplitude of the current at the input terminals of the line to vary with the voltage, and the direction of current flow also periodically reverses when the polarity of the applied voltage reverses. In the time of one cycle the energy will travel a distance of one wavelength along the line wires. The current at a given instant at any point along the line is the result of a voltage that was applied at some *earlier* instant at the input terminals. Hence the instantaneous amplitude of the current is different at all points in a one-wavelength section of line; in fact, the current flows in opposite directions in the same wire in adjacent half-wavelength sections. However, at any given point along the line the current goes through similar variations with time that the current at the input terminals did.

The result of all this is that the current (and voltage) travels along the wire as a series of waves having a length equal to the velocity of travel divided by the frequency of the a.c. voltage. On an infinitely-long line, or one properly matched at the load, an ammeter inserted anywhere in the line will show the same current, since the ammeter averages out the variations in current during a cycle. It is only when the line is not properly matched that the wave motion becomes apparent. This is discussed in the next section.

STANDING WAVES

In the infinitely-long line (or its matched counterpart) the impedance is the same at any point on the line because the ratio of voltage to current is always the same. However, the impedance at the end of the line in Fig. 13-2 is zero — or at least extremely small — because the line is short-circuited at the end. The outgoing power, on meeting the short-circuit, reverses its direction of flow and goes back along the transmission line toward the input end. There is a large current in the short-circuit, but substantially no voltage across the line at this point. We now have a voltage and current representing the power going outward toward the short-circuit, and a second voltage and current representing the **reflected** power traveling back toward the source.

The reflected current travels at the same speed as the outgoing current, so its instantaneous value will be different at every point along the line, in the distance represented by the time of one cycle. At some points along the line the phase of the outgoing and reflected currents will be such that the currents cancel each other while at others the amplitude will be doubled. At in-between points the amplitude is between these two extremes. The points at which the currents are in and out of phase depend only on the *time* required for them to travel and so depend only on the *distance* along the line from the point of reflection.

In the short-circuit at the end of the line the two current components are in phase and the total current is large. At a distance of one-half wavelength back along the line from the short-circuit the outgoing and reflected components will again be in phase and the resultant current will again have its maximum value. This is also

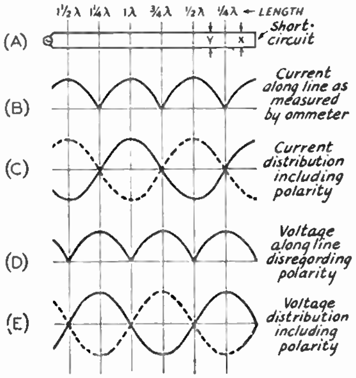


Fig. 13-2 — Standing waves of voltage and current along short-circuited transmission line.

true at any point that is a multiple of a half-wavelength from the short-circuited end of the line.

The outgoing and reflected currents will cancel at a point one-quarter wavelength, along the line, from the short-circuit. At this point, then, the current will be zero. It will also be zero at all points that are an *odd* multiple of one-quarter wavelength from the short-circuit.

If the current along the line is measured at successive points with an ammeter, it will be found to vary about as shown in Fig. 13-2B. The same result would be obtained by measuring the current in either wire, since the ammeter cannot measure phase. However, if the phase could be checked, it would be found that in each successive half-wavelength section of the line the currents at any given instant are flowing in opposite directions, as indicated by the solid line in Fig. 13-2C. Furthermore, the current in the second wire is flowing in the opposite direction to the current in the adjacent section of the first wire. This is indicated by the broken curve in Fig. 13-2C. The variations in current intensity along the transmission line are referred to as **standing waves.** The point of maximum line current is called a **current loop or current antinode** and the point of minimum line current a **current node.**

Voltage Relationships

Since the end of the line is short-circuited, the voltage at that point has to be zero. This can only be so if the voltage in the outgoing wave is met, at the end of the line, by a reflected voltage of equal amplitude and opposite polarity. In other words, the phase of the voltage wave is *reversed* when reflection takes place from the short-circuit. This reversal is equivalent to an extra half-cycle or half-wavelength of travel. As a result, the outgoing and returning voltages are in phase a quarter wavelength from the end of the line, and again out of phase a half-wavelength from the end. The standing waves of voltage, shown at D in Fig. 13-2, are therefore displaced by one-quarter wavelength from the standing waves of current. The drawing at E shows the voltages on both wires when phase is taken into account. The polarity of the voltage on each wire reverses in each half-wavelength section of transmission line. A voltage maximum is called a **voltage loop** or **antinode** and a voltage minimum is called a **voltage node.**

Open-Circuited Line

If the end of the line is open-circuited instead of short-circuited, there can be no current at the end of the line but a large voltage can exist. Again the outgoing power is reflected back toward the source. In this case, the outgoing and reflected components of *current* must be equal and opposite in phase in order for the total current at the end of the line to be zero. The outgoing and reflected components of voltage are in phase and add together. The result is that we again have standing waves, but the conditions are reversed as compared with a short-circuited line. Fig. 13-3 shows the open-circuited line case.

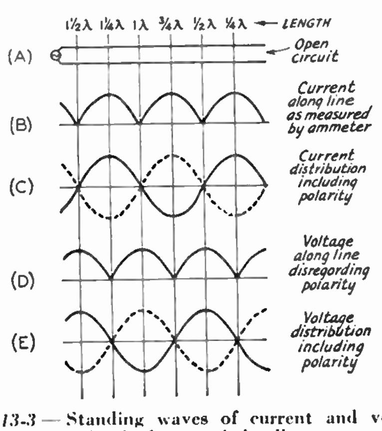


Fig. 13-3 — Standing waves of current and voltage along an open-circuited transmission line.

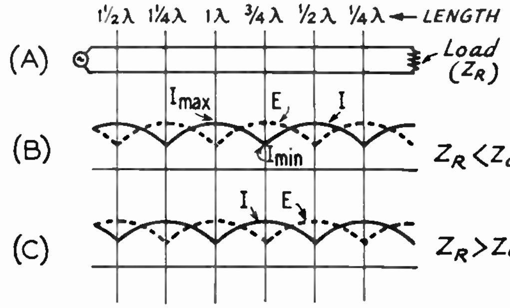


Fig. 13-4 — Standing waves on a transmission line terminated in a resistive load.

Lines Terminated in Resistive Load

Fig. 13-4 shows a line terminated in a resistive load. In this case at least part of the outgoing power is absorbed in the load, and so is not available to be reflected back toward the source. Because only part of the power is reflected, the reflected components of voltage and current do not have the same magnitude as the outgoing components. Therefore neither voltage nor current cancel completely at any point along the line. However, the *speed* at which the outgoing and reflected components travel is not affected by their amplitude, so the phase relationships are similar to those in open- or short-circuited lines.

It was pointed out earlier that if the load resistance, Z_r, is equal to the characteristic impedance, Z_0, of the line all the power is absorbed in the load. In such a case there is no reflected power and therefore no standing waves of current and voltage. This is a special case that represents the change-over point between "short-circuited" and "open-circuited" lines. If Z_r is less than Z_0, the current is largest at the load, while if Z_r is greater than Z_0 the voltage is largest at the load. The two conditions are shown at B and C, respectively, in Fig. 13-4.

The resistive termination is an important practical case. The termination is seldom an actual resistor, the most common terminations being resonant circuits or resonant antenna systems, both of which have essentially resistive impedances. If the load is reactive as well as resistive, the operation of the line resembles that shown in Fig. 13-4, but the presence of reactance in the load causes two modifications: The loops and nulls are shifted toward or away from the load; and the amount of power reflected back toward the source is increased, as compared with the amount reflected by a purely resistive load of the same total impedance. Both effects become more pronounced as the ratio of reactance to resistance in the load is made larger.

Standing-Wave Ratio

The ratio of maximum current to minimum current along a line, Fig. 13-5, is called the **standing-wave ratio.** The same ratio holds for maximum voltage and minimum voltage. It is a measure of the mismatch between the load and the line, and is equal to 1 when the line is perfectly matched. (In that case the "maximum" and "minimum" are the same, since the current and voltage do not vary along the line.) When the line is terminated in a purely-resistive load, the standing-wave ratio is

$$S.W.R. = \frac{Z_r}{Z_0} \text{ or } \frac{Z_0}{Z_r} \qquad \textbf{(13-A)}$$

Where $S.W.R.$ = Standing-wave ratio

Z_r = Impedance of load (must be pure resistance)

Z_0 = Characteristic impedance of line

Example: A line having a characteristic impedance of 300 ohms is terminated in a resistive load of 25 ohms. The s.w.r. is

$$S.W.R. = \frac{Z_0}{Z_r} = \frac{300}{25} = 12 \text{ to } 1$$

It is customary to put the larger of the two quantities, Z_r or Z_0, in the numerator of the fraction so that the s.w.r. will be expressed by a number larger than 1.

It is easier to measure the standing-wave ratio than some of the other quantities (such as the

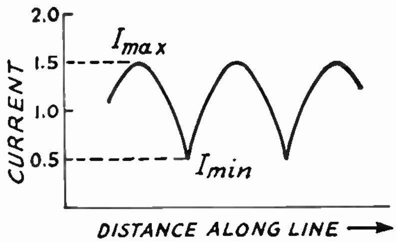


Fig. 13-5 — Measurement of standing-wave ratio. In this drawing, I_{max} is 1.5 and I_{min} is 0.5, so the s.w.r. $= I_{max}/I_{min} = 1.5/0.5 = 3$ to 1.

impedance of an antenna) that enter into transmission-line computations. Consequently, the s.w.r. is a convenient basis for work with lines. The higher the s.w.r., the greater the mismatch between line and load. In practical lines, the power loss in the line itself increases with the s.w.r.

INPUT IMPEDANCE

The input impedance of a transmission line is the impedance seen looking into the sending-end or input terminals; it is the impedance into which the source of power must work when the line is connected. If the load is perfectly matched to the line the line appears to be infinitely long, as stated earlier, and the input impedance is simply the characteristic impedance of the line itself. However, if there are standing waves this is no longer true; the input impedance may have a wide range of values.

This can be understood by referring to Figs. 13-2, 13-3, or 13-4. If the line length is such that standing waves cause the voltage at the input terminals to be high and the current low, then the input impedance is higher than the Z_0 of the line, since impedance is simply the ratio of voltage to current. Conversely, low voltage and high current at the input terminals mean that the input impedance is lower than the line Z_0. Comparison of the three drawings also shows that the range of input impedance values that may be encountered is greater when the far end of the line is open- or short-circuited than it is when the line has a resistive load. In other words, the higher the s.w.r. the greater the range of input impedance values when the line length is varied.

In addition to the variation in the absolute value of the input impedance with line length, the presence of standing waves also causes the input impedance to contain both reactance and resistance, even though the load itself may be a pure resistance. The only exceptions to this occur at the exact current loops or nodes, at which points the input impedance is a pure resistance. These are the only points at which the outgoing and reflected voltages and currents are exactly in phase: At all other distances along the line the current either leads or lags behind the voltage and the effect is exactly the same as though a capacitance or inductance were part of the input impedance of the line.

The input impedance can be represented by either a resistance and a capacitance, or by a resistance and an inductance, as shown in Fig. 13-6. Whether the impedance is inductive or capacitive depends on the characteristics of the load and the length of the line. It is possible to represent the equivalent circuit by resistance and reactance either in series or parallel, so long as the total impedance and phase angle are the same in either case. Meeting this last condition requires different values of resistance and reactance in the series case than in the parallel case.

The magnitude and character of the input impedance is quite important, since it determines the method by which the power source must be coupled to the line. The calculation of input impedance is rather complicated and its measurement is not feasible with ordinary equipment. Fortunately, in amateur work, it is unnecessary either to calculate or measure it. The proper coupling can be achieved by relatively simple methods described later in this chapter.

Unterminated Lines

The input impedance of a short-circuited or open-circuited line not an exact multiple of one-quarter wavelength long is practically a pure reactance. This is because there is very little power lost in the line. Such lines are frequently used as "linear" inductances and capacitances.

If a shorted line is less than a quarter wave long, as at X in Fig. 13-2, it will have inductive reactance. The reactance increases with the line length up to the quarter-wave point. Beyond that, as at Y, the reactance is capacitive, high near the quarter-wave point and becoming lower as the half-wave point is approached. It then alternates between inductive and capacitive in successive quarter-wave sections. Just the reverse is true of the open-circuited line.

At exact multiples of a quarter wavelength the impedance is purely resistive. It is apparent, from examination of B and D in Fig. 13-2, that at points that are a multiple of a half-wavelength — i.e., ½, 1, 1½ wavelengths, etc. — from the short-circuited end of the line the current and voltage

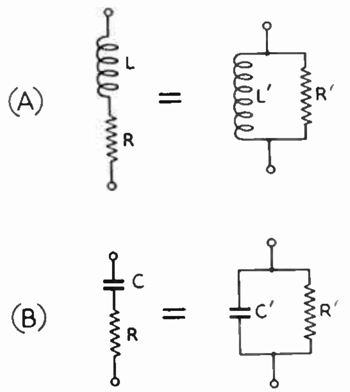


Fig. 13-6 — **Series and parallel equivalents of a line whose input impedance has both reactive and resistive components. The series and parallel equivalents do not have the same values; e.g., in A, *L* does not equal *L′* and *R* does not equal *R′*.**

have the same values that they do at the short-circuit. In other words, if the line were an exact multiple of a half-wavelength long the generator or source of power would "look into" a short-circuit. On the other hand, at points that are an odd multiple of a quarter wavelength — i.e., ¼, ¾, 1¼, etc. — from the short-circuit the voltage is maximum and the current is zero. Since $Z = E/I$, the impedance at these points is theoretically infinite. (Actually it is very high, but not infinite. This is because the current does not actually go to zero when there are losses in the line. Losses are always present, but usually are small.)

Impedance Transformation

The fact that the input impedance of a line depends on the s.w.r. and line length can be used to advantage when it is necessary to transform a given impedance into another value.

Study of Fig. 13-4 will show that, just as in the open- and short-circuited cases, if the line is one-half wavelength long the voltage and current are exactly the same at the input terminals as they are at the load. This is also true of lengths that are integral multiples of a half wavelength. It is also true for all values of s.w.r. Hence the input impedance of any line, no matter what its Z_0, that is a multiple of a half-wavelength long is exactly the same as the load impedance. Such a line can be used to transfer the impedance to a new location without changing its value.

When the line is a quarter wavelength long, or an odd multiple of a quarter wavelength, the load impedance is "inverted." That is, if the current is low and the voltage is high at the load, the input impedance will be such as to require high current and low voltage. The relationship between the load impedance and input impedance is given by:

$$Z_s = \frac{Z_0^2}{Z_r} \qquad \textbf{(13-B)}$$

where Z_s = Impedance looking into line (line length an odd multiple of one-quarter wavelength)
Z_r = Impedance of load (must be pure resistance)
Z_0 = Characteristic impedance of line

Example: A quarter-wavelength line having a characteristic impedance of 500 ohms is terminated in a resistive load of 75 ohms. The impedance looking into the input or sending end of the line is

$$Z_s = \frac{Z_0^2}{Z_r} = \frac{(500)^2}{75} = \frac{250{,}000}{75} = 3333 \text{ ohms}$$

If the formula above is rearranged, we have

$$Z_0 = \sqrt{Z_s Z_r} \qquad \textbf{(13-C)}$$

This means that if we have two values of impedance that we wish to "match," we can do so if we connect them together by a quarter-wave transmission line having a characteristic impedance equal to the square root of their product. A quarter-wave line, in other words, has the characteristics of a transformer.

Resonant and Nonresonant Lines

Because the input impedance of a line operating with a high s.w.r. is critically dependent on the line length, and furthermore is usually reactive as well as resistive, special tuning means are required for effective power transfer from the source to the line. Lines operated in this way are commonly called "tuned" or "resonant" lines. On the other hand, if the s.w.r. is low the input impedance is close to the Z_0 of the line and does not vary a great deal with the line length. Such lines are called "flat," or "untuned," or "nonresonant."

There is no sharp line of demarcation between tuned and untuned lines. If the s.w.r. is below 1.5 to 1 the line is essentially flat, since the same coupling method will work with all line lengths. If the s.w.r. is above 3 or 4 to 1 the type of coupling system, and its adjustment, will depend on the line length and such lines fall into the "tuned" category.

It is always advantageous to make the s.w.r. as low as possible. "Tuning the line" becomes necessary only when a considerable mismatch between the load and the line has to be tolerated. The most important practical example of this is when a single antenna is operated on several harmonically-related frequencies, in which case the antenna impedance will have widely-different values on different harmonics.

● RADIATION

Whenever a wire carries alternating current the electromagnetic fields travel away into space with the velocity of light. At power-line frequencies the field that "grows" when the current is increasing has plenty of time to return or "collapse" about the conductor when the current is decreasing, because the alternations are so slow. But at radio frequencies fields that travel only a relatively short distance do not have time to get back to the conductor before the next cycle commences. The consequence is that some of the electromagnetic energy is prevented from being restored to the conductor; in other words, energy is radiated into space in the form of electromagnetic waves.

The amount of energy radiated depends, among other things, on the length of the conductor in relation to the frequency or wavelength of the r.f. current. If the conductor is very short compared to the wavelength the energy radiated will be small. However, a transmission line used to feed power to an antenna is not short in this sense; in fact, it is almost always an appreciable fraction of a wavelength long and may have a length of several wavelengths.

The lines previously considered have consisted of two parallel conductors of the same diameter. Provided there is nothing in the system to destroy symmetry, at every point along the line the current in one conductor has the same intensity as the current in the other conductor at that point, but the currents flow in opposite directions. This was shown in Figs. 13-2C and 13-3C. It means that the fields set up about the two wires have the same intensity, but *opposite directions.* The consequence is that the total field set up about such a transmission line is zero; the two fields "cancel out." Hence no energy is radiated.

Actually, the fields do not completely cancel out because for them to do so the two conductors would have to occupy the same space, whereas they are slightly separated. However, the cancellation is substantially complete if the distance between the conductors is very small compared to the wavelength. Transmission line radiation will be negligible if the distance between the conductors is 0.01 wavelength or less, provided the currents in the two wires actually are balanced as described.

The amount of radiation also is proportional to the current flowing in the line. Because of the way in which the current varies along the line when there are standing waves, the effective current, for purposes of radiation, becomes greater as the s.w.r. is increased. For this reason the radiation is least when the line is flat. However, if the conductor spacing is small and the currents are balanced, the radiation from a line with even a high s.w.r. is inconsequential. A small unbalance in the line currents is far more serious.

Practical Line Characteristics

The foregoing discussion of transmission lines has been based on a line consisting of two parallel conductors. Actually, the **parallel-conductor** line is but one of two general types. The other is the **coaxial** or **concentric** line. The coaxial line consists of a round conductor placed in the center of a circular tube. The inside surface of the tube and the outside surface of the smaller inner conductor form the two conducting surfaces of the line.

In the coaxial line the fields are entirely inside the tube, because the tube acts as a shield to prevent them from appearing outside. This reduces radiation to the vanishing point. So far as the electrical behavior of coaxial lines is concerned, all that has previously been said about the operation of parallel-conductor lines applies. There are, however, practical differences in the construction and use of parallel and coaxial lines.

PARALLEL-CONDUCTOR LINES

A common type of parallel-conductor line used in amateur installations is one in which two wires (ordinarily No. 12 or No. 14) are supported a fixed distance apart by means of insulating rods called "spacers." The spacings used vary from two to six inches, the smaller spacings being necessary at frequencies of the order of 28 Mc. and higher so that radiation will be minimized. The construction is shown in Fig. 13-7. Such a line is said to be **air-insulated.** Typical spacers are shown in Fig. 13-8. The characteristic impedance of such "open-wire" lines is between 400 and 600 ohms, depending on the wire size and spacing.

Parallel-conductor lines also are sometimes constructed of metal tubing of a diameter of 1/4 to 1/2 inch. This reduces the characteristic impedance

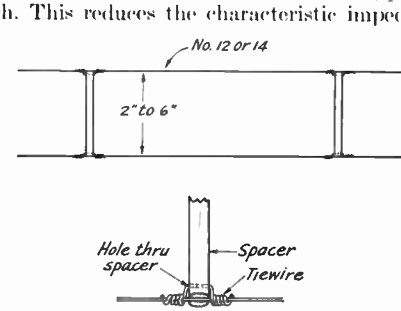


Fig. 13-7 — Typical construction of open-wire line. The line conductor fits in a groove in the end of the spacer, and is held in place by a tie-wire anchored in a hole near the groove.

of the line. Such lines are mostly used as quarter-wave transformers, when different values of impedance are to be matched.

Prefabricated parallel-conductor line with air insulation developed as a low-loss line for television reception can be used in transmitting applications. This line consists of two No. 18 conductors held at a spacing of one-half to one inch by molded-on spacers. The characteristic impedance is 375 to 450 ohms, depending on the spacing.

A convenient type of manufactured line is one in which the parallel conductors are imbedded in low-loss insulating material (polyethylene). It is commonly used as a TV lead-in and has a charac-

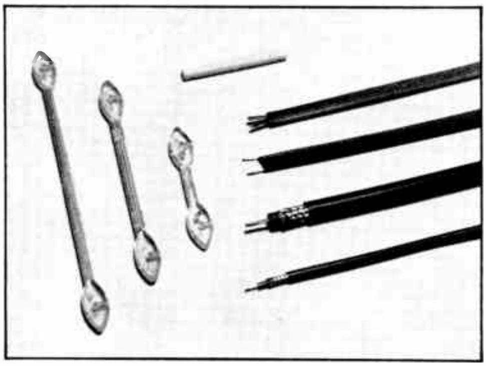


Fig. 13-8 — Typical manufactured transmission lines and spacers.

teristic impedance of 300 ohms. It is sold under various names, the most common of which is "Twin-Lead." This type of line has the advantages of light weight, close and uniform conductor spacing, flexibility and neat appearance. However, the losses in the solid dielectric are higher than in air, and dirt or moisture on the line tends to change the characteristic impedance. Moisture effects can be reduced by coating the line with silicone grease. A special form of 300-ohm Twin-Lead for transmitting uses a polyethylene tube with the conductors molded diametrically opposite; the longer dielectric path in such line reduces moisture troubles.

In addition to 300-ohm line, Twin-Lead is obtainable with a characteristic impedance of 75 ohms for transmitting purposes. Light-weight 75- and 150-ohm Twin-Lead also is available.

Characteristic Impedance

The characteristic impedance of an air-insulated parallel-conductor line is given by:

$$Z_0 = 276 \log \frac{b}{a} \qquad \text{(13-D)}$$

where Z_0 = Characteristic impedance
b = Center-to-center distance between conductors
a = Radius of conductor (in same units as b)

It does not matter what units are used for a and b so long as they are the *same* units. Both quantities may be measured in centimeters, inches, etc. Since it is necessary to have a table of common logarithms to solve practical problems, the solution is given in graphical form in Fig. 13-9 for a number of common conductor sizes.

In solid-dielectric parallel-conductor lines such as Twin-Lead the characteristic impedance cannot be calculated readily, because part of the electric field is in air as well as in the dielectric.

Unbalance in Parallel-Conductor Lines

When installing parallel-conductor lines care should be taken to avoid introducing electrical unbalance into the system. If for some reason the current in one conductor is higher than in the other, or if the currents in the two wires are not exactly out of phase with each other, the electromagnetic fields will not cancel completely and a considerable amount of power may be radiated by the line.

Maintaining good line balance requires, first of all, a balanced load at its end. For this reason the antenna should be fed, whenever possible, at a point where each conductor "sees" exactly the same thing. Usually this means that the antenna system should be fed at its electrical center. Even though the antenna appears to be symmetrical, physically, it can be unbalanced electrically if the part connected to one of the line conductors is inadvertently coupled to something (such as house wiring or a metal pole or roof) that is not duplicated on the other part of the antenna. Every effort should be made to keep the antenna as far as possible from other wiring or sizable

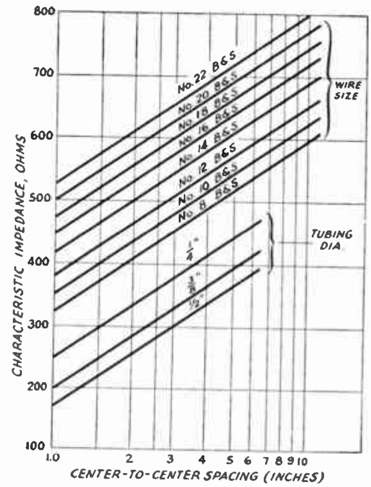


Fig. 13-9 — Chart showing the characteristic impedance of spaced-conductor parallel transmission lines with air dielectric. Tubing sizes given are for outside diameters.

metallic objects. The transmission line itself will cause some unbalance if it is not brought away from the antenna at right angles to it for a distance of at least a quarter wavelength.

In installing the line conductors take care to see that they are kept away from metal. The minimum separation between either conductor and all other wiring should be at least four or five times the conductor spacing. The shunt capacitance introduced by close proximity to metallic objects can drain off enough current (to ground) to unbalance the line currents, resulting in increased radiation. A shunt capacitance of this sort also constitutes a reactive load on the line, causing an impedance "bump" that will prevent making the line actually flat.

COAXIAL LINES

The most common form of coaxial line consists of either a solid or stranded-wire inner conductor surrounded by polyethylene dielectric. Copper braid is woven over the dielectric to form the outer conductor, and a waterproof vinyl covering is placed on top of the braid. This cable is made in a number of different diameters. It is moderately flexible, and so is convenient to install. Some different types are shown in Fig. 13-8. This solid coaxial cable is commonly available in impedances approximating 50 and 70 ohms.

Air-insulated coaxial lines have lower losses than the solid-dielectric type, but are less used in amateur work because they are expensive and difficult to install as compared with the flexible cable. The common type of air-insulated coaxial line uses a solid-wire conductor inside a copper tube, with the wire held in the center of the tube by means of insulating "beads" placed at regular intervals.

Characteristic Impedance

The characteristic impedance of an air-insulated coaxial line is given by the formula

$$Z_0 = 138 \log \frac{b}{a} \qquad \textbf{(13-E)}$$

where Z_0 = Characteristic impedance
b = Inside diameter of outer conductor
a = Outside diameter of inner conductor (in same units as b)

Curves for typical conductor sizes are given in Fig. 13-10.

The formula for coaxial lines is approximately correct for lines in which bead spacers are used, provided the beads are not too closely spaced. When the line is filled with a solid dielectric, the characteristic impedance as given by the chart should be multiplied by $1/\sqrt{K}$, where K is the dielectric constant of the material.

ELECTRICAL LENGTH

In the discussion of line operation earlier in this chapter it was assumed that currents traveled along the conductors at the speed of light. Actually, the velocity is somewhat less, the reason being that electromagnetic fields travel more slowly in material dielectrics than they do in free space. In air the velocity is practically the same as in empty space, but a practical line always has to be supported in some fashion by solid insulating materials. The result is that the fields are slowed down; the currents travel a shorter distance in the time of one cycle than they do in space, and so the wavelength along the line is less than the wavelength would be in free space at the same frequency.

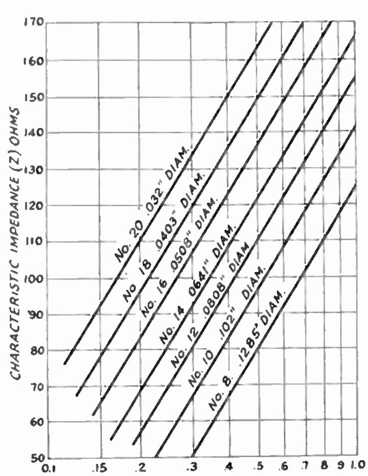


Fig. 13-10 — Chart showing characteristic impedance of various air-insulated concentric lines.

TABLE 13-I
Transmission-Line Data

Type	Description or Type Number	Characteristic Impedance	Velocity Factor	Capacitance per foot; μμfd.
Coaxial	Air-insulated	50–100	0.85[1]	
	RG-8/U	53	0.66	29.5
	RG-58/U	53	0.66	28.5
	RG-11/U	75	0.66	20.5
	RG-59/U	73	0.66	21.0
Parallel-Conductor	Air-insulated	200–600	0.975[2]	
	14-080[3]	75	0.68	19.0
	14-023[3]	75	0.71	20.0
	14-079[3]	150	0.77	10.0
	14-056[3]	300	0.82	5.8
	14-076[3]	300	0.84	3.9
	14-022[3]	300	0.85	3.0

[1] Average figure for small-diameter lines with ceramic beads.
[2] Average figure for lines insulated with ceramic spacers at intervals of a few feet.
[3] Amphenol type numbers and data. Line similar to 14-056 is made by several manufacturers, but rated loss may differ from that given in Fig. 13-11. Types 14-023, 14-076, and 14-022 are made for transmitting applications.

Whenever reference is made to a line as being so many wavelengths (such as a "half-wavelength" or "quarter wavelength") long, it is to be understood that the *electrical* length of the line is meant. Its actual physical length as measured by a tape always will be somewhat less. The physical length corresponding to an electrical wavelength is given by

$$\textit{Length in feet} = \frac{984}{f} \cdot V \qquad \textbf{(13-F)}$$

where f = Frequency in megacycles
V = Velocity factor

The **velocity factor** is the ratio of the actual velocity along the line to the velocity in free space. Values of V for several common types of lines are given in Table 13-I.

Example: A 75-foot length of 300-ohm Twin-Lead is used to carry power to an antenna at a frequency of 7150 kc. From Table 13-I, V is 0.82. At this frequency (7.15 Mc.) a wavelength is

$$\textit{Length (feet)} = \frac{984}{f} \cdot V = \frac{984}{7.15} \times 0.82$$

$$= 137.6 \times 0.82 = 112.8 \text{ ft.}$$

The line length is therefore 75/112.8 = 0.665 wavelength.

Because a quarter-wavelength line is frequently used as a linear transformer, it is con-

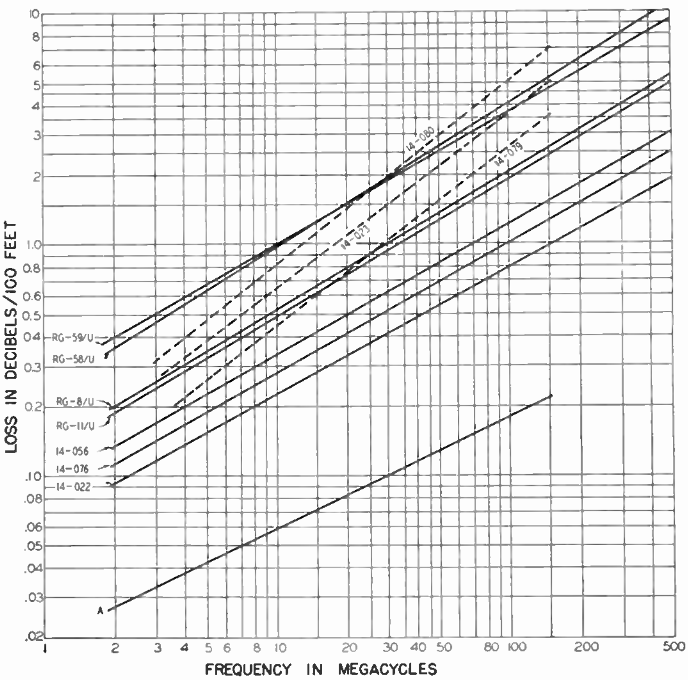


Fig. 13-11 — Attenuation data for common types of transmission lines. Curve A is the nominal attenuation of 600-ohm open-wire line with No. 12 conductors, not including dielectric loss in spacers nor possible radiation losses. Additional line data are given in Table 13-I.

venient to calculate the length of a quarter-wave line directly. The formula is

$$Length\ (feet) = \frac{246}{f} \cdot V \qquad \textbf{(13-G)}$$

where the symbols have the same meaning as above.

LOSSES IN TRANSMISSION LINES

There are three ways by which power may be lost in a transmission line: by radiation, by heating of the conductors (I^2R loss), and by heating of the dielectric, if any. There is no appreciable radiation loss from a coaxial line, but radiation from a parallel-conductor line may exceed the heat losses if the line is unbalanced. Since radiation losses cannot readily be estimated or measured, the following discussion is based only on conductor and dielectric losses.

Heat losses in both the conductor and the dielectric increase with frequency. Conductor losses also are greater the lower the characteristic impedance of the line, because a higher current flows in a low-impedance line for a given power input. The converse is true of dielectric losses because these increase with the voltage, which is greater on high-impedance lines. The dielectric loss in air-insulated lines is negligible (the only loss is in the insulating spacers) and such lines operate at high efficiency when radiation losses are low.

It is convenient to express the loss in a transmission line in decibels per unit length, since the loss in db. is directly proportional to the line length. Losses in various types of lines operated without standing waves (that is, terminated in a resistive load equal to the characteristic impedance of the line) are given in graphical form in Fig. 13-11. In these curves the radiation loss is assumed to be negligible.

When there are standing waves on the line the power loss increases as shown in Fig. 13-12. Whether or not the increase in loss is serious depends on what the original loss would have been if the line were perfectly matched. If the loss with perfect matching is very low, a large s.w.r. will not greatly affect the *efficiency* of the line — i.e.,

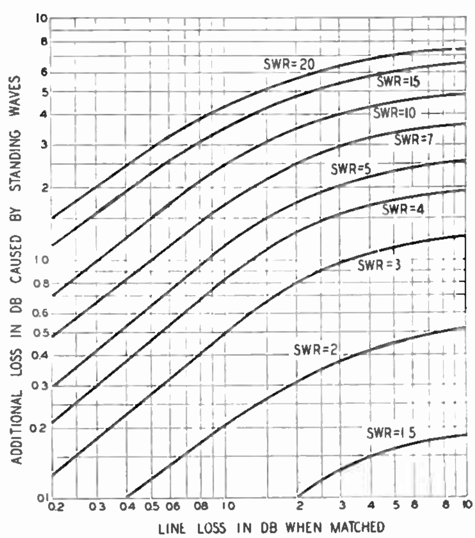


Fig. 13-12 — Effect of standing-wave ratio on line loss. The ordinates give the *additional* loss in decibels for the loss, under perfectly-matched conditions, shown on the horizontal scale.

the ratio of the power delivered to the load to the power put into the line.

> Example: A 150-foot length of RG-11/U cable is operating at 7 Mc. with a 5-to-1 s.w.r. If perfectly matched, the loss from Fig. 13-11 would be 1.5 × 0.4 = 0.6 db. From Fig. 13-12 the additional loss because of the s.w.r. is 0.73 db. The total loss is therefore 0.6 + 0.73 = 1.33 db.

An appreciable s.w.r. on a solid-dielectric line may result in excessive loss of power at the higher frequencies. Such lines, whether of the parallel-conductor or coaxial type, should be operated as nearly flat as possible, particularly when the line length is more than 50 feet or so. As shown by Fig. 13-12, the increase in line loss is not too serious so long as the s.w.r. is below 2 to 1, but increases rapidly when the s.w.r. rises above 3 to 1. Tuned transmission lines such as are used with multiband antennas always should be air-insulated, in the interests of highest efficiency.

Matching the Load to the Line

The load for a transmission line may be any device capable of dissipating r.f. power. When lines are used for transmitting applications the most common type of load is an antenna, but there are also practical cases where the grid circuit of a power amplifier may represent the load. When a transmission line is connected between an antenna and a receiver, the receiver input circuit (not the antenna) is the load, because the power taken from a passing wave is delivered to the receiver.

Whatever the application, the conditions existing at the load, and *only* the load, determine the standing-wave ratio on the line. If the load is purely resistive and equal in value to the characteristic impedance of the line, there will be no standing waves. If the load is not purely resistive, and/or is not equal to the line Z_0, there will be standing waves. No adjustments that can be made at the input end of the line can change the s.w.r., nor is it affected by changing the line length.

Only in a few special cases is the load inherently of the proper value to match a practicable transmission line. In all other cases it is necessary either to operate with a mismatch and accept the s.w.r. that results, or else to take steps to bring about a proper match between the line and load by means of transformers or similar devices. Impedance-matching transformers may take a variety of physical forms, depending on the circumstances.

Note that it is essential, if the s.w.r. is to be made as low as possible, that the load at the point of connection to the transmission line be purely resistive. In general, this requires that the load be tuned to resonance. If the load itself is not resonant at the operating frequency the tuning sometimes can be accomplished in the matching system.

THE ANTENNA AS A LOAD

Every antenna system, no matter what its physical form, will have a definite value of impedance at the point where the line is to be connected. The problem is to transform this **antenna input impedance** to the proper value to match the line. In this respect there is no one "best" type of line for a particular antenna system, because it is possible to transform impedances in any desired ratio. Consequently, any type of line may be used with any type of antenna. There are frequently reasons other than impedance matching that dictate the use of one type of line in preference to another, such as ease of installation, inherent loss in the line, and so on, but these are not considered in this section.

Although the input impedance of an antenna system is seldom known very accurately, it is often possible to make a reasonably close estimate of its value. The information in the chapter on antennas can be used as a guide.

Matching circuits may be constructed using ordinary coils and condensers, but are not used very extensively because they must be supported at the antenna and must be weatherproofed. The systems to be described use **linear transformers.**

The Quarter-Wave Transformer or "Q" Section

As described earlier in this chapter, a quarter-wave transmission line may be used as an impedance transformer. Knowing the antenna impedance and the characteristic impedance of the transmission line to be matched, the required characteristic impedance of a matching section such as is shown in Fig. 13-13 is

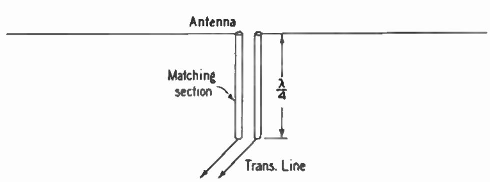


Fig. 13-13 — "Q" matching section, a quarter-wave impedance transformer.

$$Z = \sqrt{Z_1 Z_0}$$

where Z_1 is the antenna impedance and Z_0 is the characteristic impedance of the line to which it is to be matched.

> Example: To match a 600-ohm line to an antenna presenting a 72-ohm load, the quarter-wave matching section would require a characteristic impedance of $\sqrt{72 \times 600} = \sqrt{43{,}200} = 208$ ohms.

The spacings between conductors of various sizes of tubing and wire for different surge impedances are given in graphical form in Fig. 13-9. (With

½-inch tubing, the spacing in the example above should be 1.5 inches for an impedance of 208 ohms.)

The length of the quarter-wave matching section is given by Equation 13-G.

The antenna must be resonant at the operating frequency. Setting the antenna length by formula is amply accurate with single-wire antennas, but in other systems, particularly close-spaced arrays, the antenna should be adjusted to resonance before the matching section is connected.

When the antenna input impedance is not known accurately, it is advisable to construct the matching section so that the spacing between conductors can be changed. The spacing them may be adjusted to give the lowest possible s.w.r. on the transmission line.

Stub Matching

When a transmission line is not matched by the load, the impedance looking into the line toward the load varies with the distance from the load, as discussed earlier in this chapter. Considering the

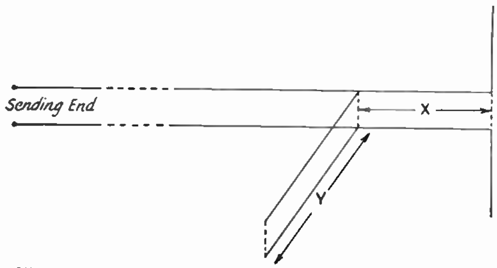


Fig. 13-14 — Matching the antenna to the line by means of a stub, Y. Curves for determining the lengths X and Y are given in Figs. 13-15 and 13-16, for the case where the line, section X and section Y all have the same characteristic impedance.

input impedance to be equivalent to a resistance in parallel with a reactance, at some distance along the line such as X in Fig. 13-14 the resistive part of the input impedance will be equal to the Z_0 of the line. If at this point a reactance equal to the reactive part of the input impedance, but of the opposite type, is connected across the line, the reactances will cancel and leave only the resistive component. From this point back to the transmitter or other source of energy the line will be matched.

The reactances used for matching in this way are usually **linear reactances** — sections of transmission line — called **stubs.** Stubs may be open or **closed,** depending on whether the free end is left open or is short-circuited, according to the type of reactance required in a particular case. The type and length of stub, as well as the point at which it should be attached to the line, can be found without any knowledge of the antenna input impedance, providing that the s.w.r. on the line can be measured before the stub is attached, and providing that the position of a current node (voltage loop) can be determined under the same conditions.

When the s.w.r. and the position of a current node are known Figs. 13-15 and 13-16 give the

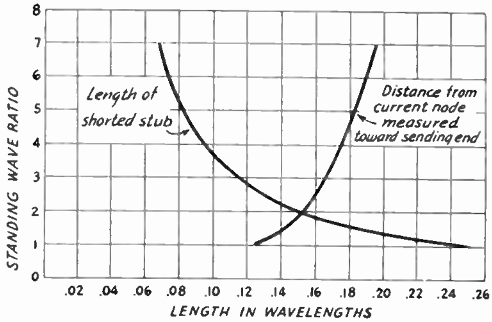


Fig. 13-15 — Graph for determining position and length of a *shorted* stub. Dimensions may be converted to linear units after values have been taken from the graph.

stub information necessary for impedance matching. Stub lengths are given in wavelengths, which may be converted to feet with the help of Equation 13-F. The data in Figs. 13-15 and 13-16 are based on the assumption that the line and stub both have the same Z_0.

With this system of matching it is not necessary that the antenna system be exactly resonant, since the match is based on the position of a current node along the line. The node nearest the antenna should be used for determining the position of the stub so that as much as possible of the transmission line will be operating with a low s.w.r.

Folded Dipoles

A half-wave antenna element can be made to match various line impedances if it is split into two or more parallel conductors with the transmission line attached at the center of only one of them. Various forms of such "folded dipoles" are shown in Fig. 13-17. Currents in all conductors are in phase in a folded dipole, and since the conductor spacing is small the folded dipole is equivalent in radiating properties to an ordinary single-conductor dipole. However, the current flowing into the input terminals of the antenna from the line is the current in one conductor only, and the entire power from the line is delivered at this value of current. This is equivalent to saying that the input impedance of the

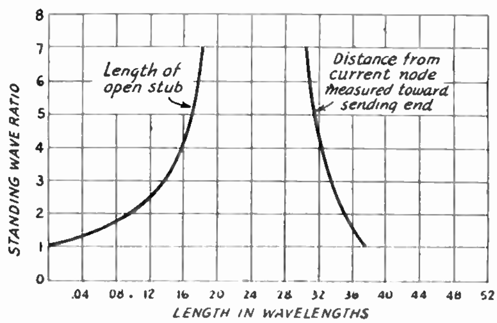


Fig. 13-16 — Graph for determining position and length of an *open* stub. Dimensions may be converted to linear units after values have been taken from the graph.

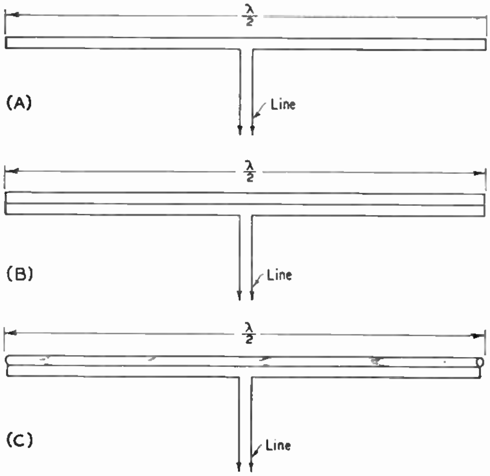


Fig. 13-17 — The folded dipole, a method for using the antenna element itself to provide an impedance transformation.

antenna has been raised by splitting it up into two or more conductors.

If the conductors of a folded dipole are all the same diameter and the spacing between them is small, the impedance at the input terminals is approximately equal to the input impedance of an ordinary dipole multiplied by the square of the number of conductors. A simple half-wave antenna has an impedance of about 70 ohms, so a two-conductor folded dipole will have an input impedance of 280 ohms, and a three-conductor dipole an impedance of 630 ohms. These values are sufficiently close for good matching to 300-ohm or 600-ohm line, respectively.

Other values of impedance ratio may be obtained by making one conductor larger in diameter than the other, as shown at C in Fig. 13-17. The required ratio of conductor radii (or diameters) for a desired impedance ratio using two conductors may be obtained from Fig. 13-18. Similar information for a 3-conductor dipole is given in Fig. 13-19. This graph applies where all three conductors are in the same plane and the two conductors not connected to the transmission line are equally spaced from the fed conductor, and have equal diameters (this diameter need not equal the diameter of the fed conductor). The unequal-conductor method has been found particularly useful in matching to low-impedance antennas such as directive arrays using close-spaced parasitic elements.

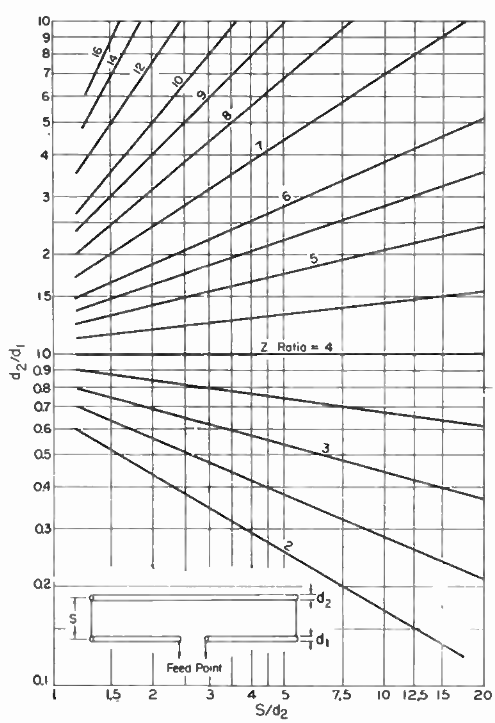


Fig. 13-18 — Impedance transformation ratio, two-conductor folded dipole. The dimensions d_1, d_2 and s are shown on the inset drawing. Curves show the ratio of the impedance (resistive) seen by the transmission line to the radiation resistance of the resonant antenna system.

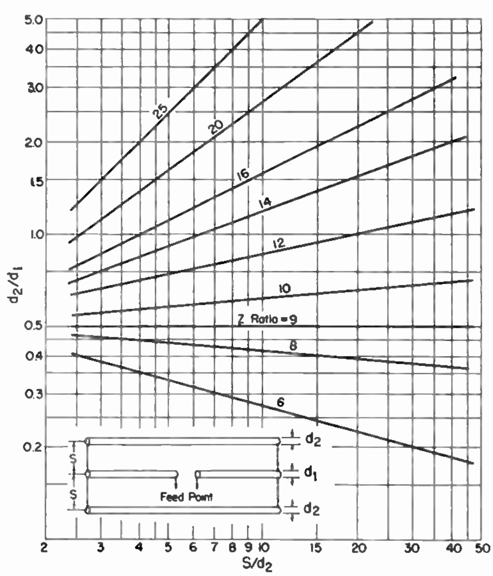


Fig. 13-19 — Impedance transformation ratio, three-conductor folded dipole. The dimensions d_1, d_2 and s are shown on the inset drawing. Curves show the ratio of the impedance (resistive) seen by the transmission line to the radiation resistance of the resonant antenna system.

The length of the antenna element should be such as to be approximately self-resonant at the median operating frequency. The length is usually not highly critical, because this method of matching tends to compensate for changes in antenna reactance with frequency and thus broadens the frequency-response curve of the antenna.

"T" and "Gamma" Matching Sections

The method of matching shown in Fig. 13-20A is based on the fact that the impedance between any two points along a resonant antenna is resistive, and has a value which depends on the spacing between the two points. It is therefore possible to choose a pair of points between which the impedance will have the right value to match a transmission line. In practice, the line cannot

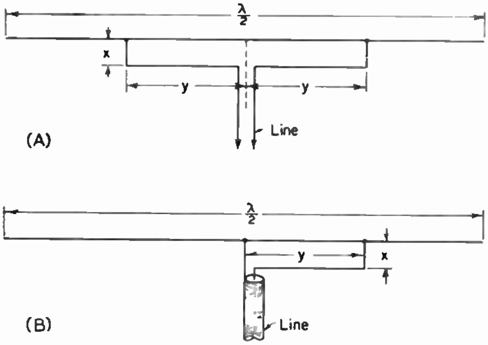


Fig. 13-20 — The "T" match and "gamma" match.

be connected directly at these points because the distance between them is much greater than the conductor spacing of a practicable transmission line. The "T" arrangement in Fig. 13-20A overcomes this difficulty by using a second conductor paralleling the antenna to form a matching section to which the line may be connected.

The "T" is particularly suited to use with a parallel-conductor line, in which case the two points along the antenna should be equidistant from the center so that electrical balance is maintained.

The operation of this system is somewhat complex. Each "T" conductor (y in the drawing) forms with the antenna conductor opposite it a short section of transmission line. Each of these transmission-line sections can be considered to be terminated in the impedance that exists at the point of connection to the antenna. Thus the part of the antenna between the two points carries a transmission-line current in addition to the normal antenna current. The two transmission-line matching sections are in series, as seen by the main transmission line.

If the antenna by itself is resonant at the operating frequency its impedance will be purely resistive, and in such case the matching-section lines are terminated in a resistive load. However, since these sections are shorter than a quarter wavelength their input impedance — i.e., the impedance seen by the main transmission line looking into the matching-section terminals — will be reactive as well as resistive. This prevents a perfect match to the main transmission line, since its load must be a pure resistance for perfect matching. The reactive component of the input impedance must be tuned out before a proper match can be secured.

One way to do this is to detune the antenna just enough, by changing its length, to cause reactance of the opposite kind to be reflected to the input terminals of the matching section, thus cancelling the reactance introduced by the latter. Another method, which is considerably easier to adjust, is to insert a variable condenser in series with the matching section where it connects to the transmission line, as shown in Fig. 13-21. A condenser having a maximum capacitance of 150 μμfd. or so will be about right in the average case, for 14 Mc. and higher. The condenser must be protected from the weather.

The method of adjustment commonly used is to cut the antenna for approximate resonance and then make the spacing x some value that is convenient constructionally. The distance y is then adjusted, while maintaining symmetry with respect to the center, until the s.w.r. on the transmission line is as low as possible. If the s.w.r. is not below 2 to 1 after this adjustment, the antenna length should be changed slightly and the matching-section taps adjusted again. This process may be continued until the s.w.r. is as close to 1 to 1 as possible.

When the series-condenser method of reactance compensation is used (Fig. 13-21) the antenna should be the proper length to be resonant at the operating frequency. Trial positions of the matching-section taps are taken, each time adjusting the condenser for minimum s.w.r., until the standing waves on the transmission line are brought down to the lowest possible value.

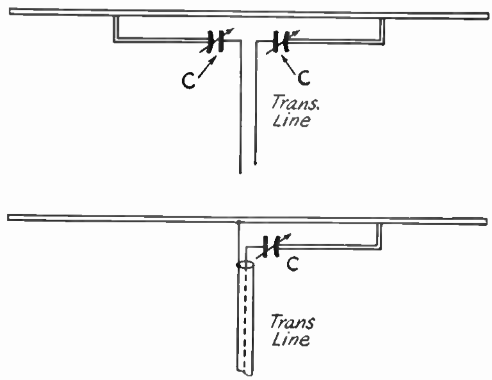


Fig. 13-21 — Using series condensers for tuning out reactance in the matching section with the "T" match and "gamma" match. The condenser C should have a maximum capacitance of approximately 150 μμfd. for 14 Mc. and may have proportionately lower capacitances for shorter wavelengths. Receiving-type condensers can be used for powers up to a few hundred watts.

The unbalanced ("gamma") arrangement in Fig. 13-20B is similar in principle to the "T," but is adapted for use with single coax line. The method of adjustment is the same.

The "Delta" Match

The matching system in Fig. 13-22 is based on the variation in impedance between two points symmetrically located with respect to the center of the antenna, as in the case of the "T" match, but uses a different matching section. If the two conductors of a transmission line are fanned out, the Z_0 of the line will increase with the increase in spacing. A fanned section of line can be used to match a given load impedance to the Z_0 of a uniformly-spaced transmission line, provided the line Z_0 is lower than the impedance of the load. Strictly, such a match can be made only if the conductor spacing in the fanned section of line increases at an exponential rate, but the "delta" arrangement in Fig. 13-22 is a rough approximation to this type of spacing.

Dimensions *a* and *b* in Fig. 13-22 depend on the antenna impedance (whether it is a simple half-wave antenna or the driven element of a multielement beam), the size of the conductors in the delta, and the Z_0 of the transmission line to be matched. Methods for calculation are not available, but dimensions for practical cases are given in the chapters on antennas.

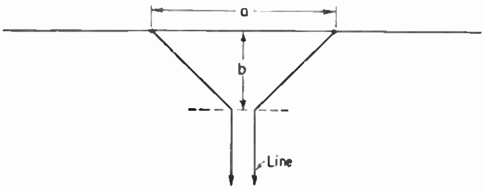


Fig. 13-22 — The "delta" matching section.

BALANCING DEVICES

An antenna with open ends, of which the half-wave type used as an illustration in this section is an example, is inherently a balanced radiator, having equal and opposite voltages at its ends and minimum voltage at the center. When opened at the center and fed with a parallel-conductor line this balance is maintained throughout the system, including the transmission line, so long as the causes of unbalance discussed earlier in this chapter are avoided.

If the antenna is fed at the center through a coaxial line, as indicated in Fig. 13-23A, this balance is upset because one side of the radiator is connected to the shield while the other is connected to the inner conductor. The antenna current on the side connected to the shield can flow down over the *outside* of the coaxial line, and the fields thus set up cannot be canceled by the fields from the inner conductor because the fields *inside* the line cannot escape through the shielding afforded by the outer conductor. Hence these "antenna" currents flowing on the outside of the line will be responsible for radiation. (In the gamma match of Fig. 13-20B such radiation is largely prevented because the radiator is continuous and the outer conductor is connected to its center, a point which is at ground potential.)

Linear Baluns

Line radiation can be prevented by a number of devices whose purpose is to detune or decouple the line for "antenna" currents and thus greatly reduce their amplitude. Such devices generally are known as **baluns** (a contraction for "balanced to unbalanced"). Fig. 13-23B shows one such arrangement, known as a **bazooka,** which uses a sleeve over the transmission line to form, with

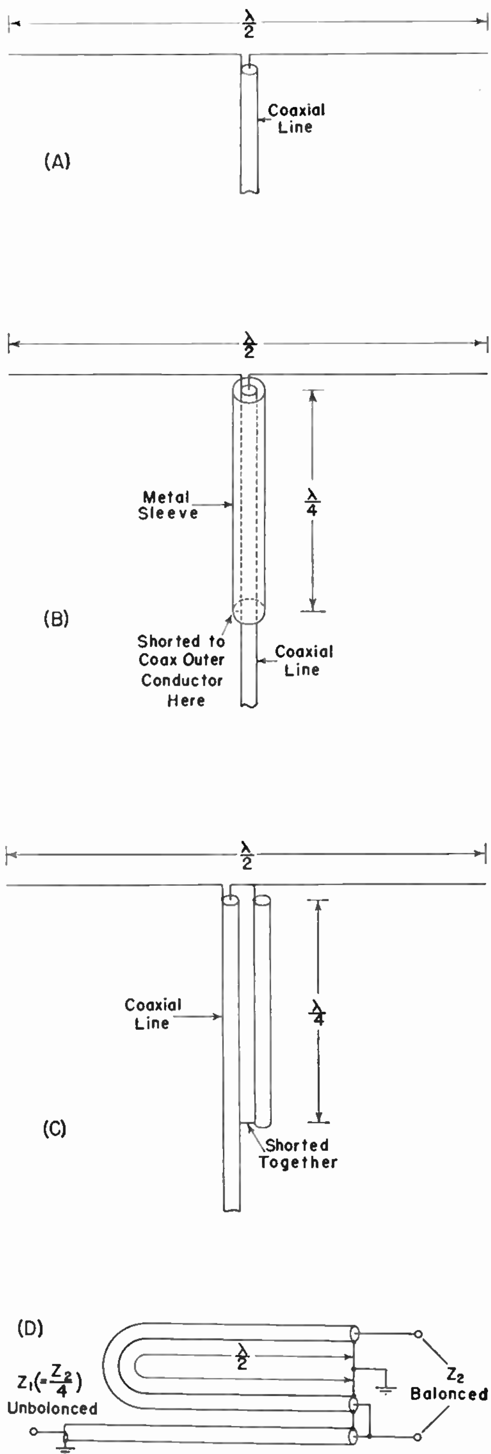


Fig. 13-23 — Radiator with coaxial feed (A) and methods of preventing unbalance currents from flowing on the outside of the transmission line (B and C). The half-wave phasing section shown at D is used for coupling between an unbalanced and a balanced circuit when a 4-to-1 impedance ratio is desired or can be accepted.

the outside of the outer line conductor, a shorted quarter-wave line section. As described earlier in this chapter, the impedance looking into the open end of such a section is very high, so that the end of the outer conductor of the coaxial line is effectively insulated from the part of the line below the sleeve. The length is an *electrical* quarter wave, and may be physically shorter if the insulation between the sleeve and the line is other than air. The bazooka has no effect on the impedance relationships between the antenna and the coaxial line.

Another method that gives an equivalent effect is shown at C. This uses a second conductor, generally of the same diameter as the coaxial line (a piece of the same type of line may be used, the inner conductor being disregarded) to form a parallel-conductor quarter-wave "insulator," thus isolating both halves of the antenna equally from the remainder of the line below the shorting connection.

Fig. 13-23D shows a third balun, in which equal and opposite voltages, balanced to ground, are taken from the inner conductors of the main transmission line and half-wave phasing section. Since the voltages at the balanced end are in series while the voltages at the unbalanced end are in parallel, there is a 4-to-1 step-down in impedance from the balanced to the unbalanced side. This arrangement is useful for coupling between a balanced 300-ohm line and a 75-ohm coaxial line, for example.

Coil Baluns

Another form of linear balun is shown in the upper drawing of Fig. 13-24. Two transmission lines of equal length having a characteristic impedance Z_0 are connected in series at one end and in parallel at the other. At the series-connected end the lines are balanced to ground and will match an impedance equal to $2Z_0$. At the parallel-connected end the lines will be matched by an impedance equal to $Z_0/2$. One side may be connected to ground at the parallel-connected end, provided the two lines have a length such that, considering each line as a single wire, the balanced end is effectively decoupled from the parallel-connected end. This requires a length that is an odd multiple of ¼ wavelength. The impedance transformation from the series-connected end to the parallel-connected end is 4 to 1.

A definite line length is required only for decoupling purposes, and so long as there is adequate decoupling the system will act as a 4-to-1 impedance transformer regardless of line length. If each line is wound into a coil, as in the lower drawing, the inductances so formed will act as choke coils and will tend to isolate the series-connected end from any ground connection that may be placed on the parallel-connected end. Balun coils made in this way will operate over a wide frequency range, since the choke inductance is not critical. The lower frequency limit is where the coils are no longer effective in isolating one line from the other; the length of line in each coil should be about equal to a quarter wavelength at the lowest frequency to be used.

The principal application of such coils is in going from a 300-ohm balanced line to a 75-ohm coaxial line. This requires that the Z_0 of the lines forming the coils be 150 ohms. Design data for winding the coils are not available; however, Equation 13-D can be used for determining the approximate wire spacing. Allowance should be made for the fact that the effective dielectric constant will be somewhat greater than 1 if the coil is wound on a form. The proximity effect between turns can be reduced by making the turn spacing somewhat larger than the conductor spacing. For operation at 3.5 Mc. and higher frequencies the length of each conductor should be about 60 feet. The conductor spacing can be adjusted to the proper value by terminating each line in a resistor equal to its characteristic impedance and adjusting the spacing until an s.w.r. bridge at the input end shows the line to be matched.

A balun of this type is simply a fixed-ratio transformer and does not make up for inaccurate

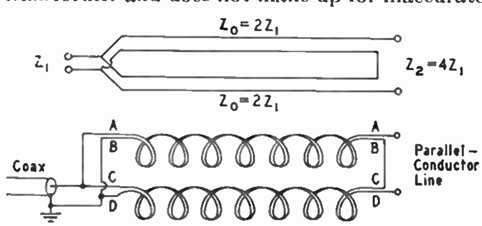


Fig. 13-24 — **Baluns for matching between push-pull and single-ended circuits. The impedance ratio is 4 to 1 from the push-pull side to the unbalanced side. Coiling the lines as shown in the lower drawing increases the frequency range over which satisfactory operation is obtained.**

matching elsewhere in the system. With a "300-ohm" line on the balanced end, for example, a 75-ohm coax cable will not be matched unless the 300-ohm line actually is terminated in a 300-ohm load.

NONRADIATING LOADS

Important practical cases of nonradiating loads for a transmission line are the grid circuit of a power amplifier (considered in the chapter on transmitters), the input circuit of a receiver, and another transmission line. This last case includes the "antenna tuner" — a misnomer because it is actually a device for coupling a transmission line to the transmitter. Because of its importance in amateur installations, the antenna coupler is considered separately in a later section of this chapter.

Coupling to a Receiver

A good match between an antenna and its transmission line does not guarantee a low standing-wave ratio on the line when the antenna system is used for receiving. The s.w.r. is determined wholly by what the line "sees" at the receiver's antenna-input terminals. For minimum s.w.r. the receiver input circuit must be matched to the line. The rated input impedance of a receiver is a nominal value that varies over a considerable range with frequency. Methods for bringing about a proper match are discussed in the chapter on receivers.

It should be noted that *if* the receiver is matched to the line, then it is desirable that the antenna and line also be matched, since this results in maximum signal transfer from the antenna to the line. If the receiver is *not* matched to the line, the input impedance of the line (at the terminals of the antenna itself) in turn cannot match the antenna impedance. In such a case the signal input to the receiver depends on the coupling system used between the line and the receiver. For greatest signal strength the coupling system has to be adjusted to the best compromise between receiver input impedance and load appearing at the input (antenna) end of the line. The proper adjustments must be determined by experiment.

A similar situation exists when the receiver input impedance inherently matches the line Z_0, but the line and antenna are mismatched. Under these conditions perfect matching at the receiver does not result in greatest signal strength; a deliberate mismatch has to be introduced so that the maximum power will be taken from the antenna.

The most desirable condition is that in which the receiver is matched to the line Z_0 and the line in turn is matched to the antenna. This transfers maximum power from the antenna to the receiver with the least loss in the transmission line.

Coupling the Transmitter to the Line

The type of coupling system that will be needed to transfer power adequately from the final r.f. amplifier to the transmission line depends almost entirely on the input impedance of the line. As shown earlier in this chapter, the input impedance is determined by the standing-wave ratio and the line length. The simplest case is that where the line is terminated in its characteristic impedance so that the s.w.r. is 1 to 1 and the input impedance is merely the Z_0 of the line, regardless of line length.

Coupling systems that will deliver power into a flat line are readily designed. For all practical purposes the line can be considered to be flat if the s.w.r. is no greater than about 1.5 to 1. That is, a coupling system designed to work into a pure resistance equal to the line Z_0 will have enough leeway to take care of the small variations in input impedance that will occur when the line length is changed, if the s.w.r. is higher than 1 to 1 but no greater than 1.5 to 1.

Coupling circuits suitable for coaxial lines are discussed in the chapter on transmitters. As stated in that chapter, an untuned "pick-up" or "link" coil connected directly to the transmission line should have an inductance such that the reactance at the operating frequency is approximately equal to the Z_0 of the line, to assure adequate coupling to a line that is actually flat. While this condition is sometimes met well enough at the higher frequencies, at least for coaxial lines, by manufactured link coils, it is definitely not met when a parallel-conductor line having a Z_0 of 300 ohms or more is used. The optimum pick-up coil for coupling to such lines will have about the same inductance as the plate tank coil itself.

Amateurs are frequently successful in coupling power into a line even though the pick-up coil is quite small and is loosely coupled to the amplifier tank coil. When such coupling is possible it is an indication that the line is operating at a fairly high s.w.r. and that the line length is such as to bring a current loop near the input end. It is customary to "prune" the line length in such cases until adequate coupling is secured — a practice that has given rise to the wholly fallacious belief, on the part of many, that pruning the line reduces the standing-wave ratio and that a flat line will load an amplifier with a small link and very loose coupling. Pruning the line accomplishes nothing if the line is actually flat because, as explained earlier in this chapter, the input impedance of a matched line is equal to its Z_0 regardless of the line length. If the line is not flat, pruning changes the input impedance and eventually results in a value such that the link or pick-up coil is actually tuned to the operating frequency by the line, a condition that will give maximum power transfer with minimum coupling. The higher the s.w.r. the more loose the coupling can be. Although there is nothing inherently wrong with this method of adjustment, it works only when the s.w.r. is fairly high and will not work with a line that actually is flat.

Tuned Coupling

A tuned coupling circuit has the same advantages, when used with properly-terminated parallel-conductor lines, that were outlined in the transmitting chapter in connection with coaxial lines. The principles are the same as well, but a resistance of 300 to 600 ohms is too high to be connected in series with a tuned circuit. Consequently, parallel-tuned circuits must be used with

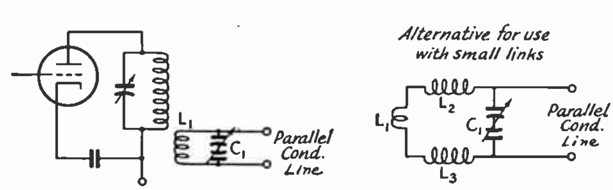


Fig. 13-25 — Tuned circuits for coupling to a flat parallel-conductor line. Values for C_1 are given in Table 13-II; L_1 is chosen to resonate with the value given at the operating frequency. In the alternative circuit the total inductance of L_1, L_2 and L_3 should equal L_1 in the circuit at the left.

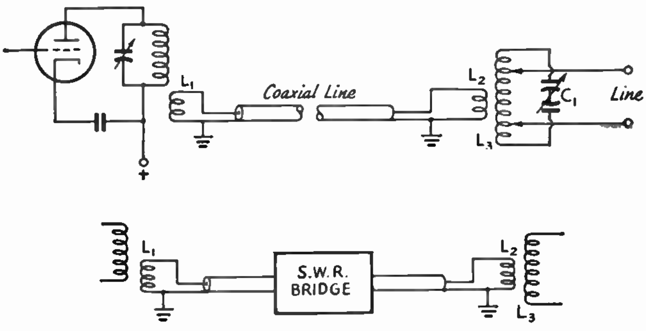


Fig. 13-26 — Matching circuits using a coaxial link, for use with parallel-conductor transmission lines. Adjustment set-up using an s.w.r. bridge is shown in the lower drawing. Design considerations and method of adjustment are discussed in the text.

these lines. Typical arrangements are shown in Fig. 13-25. The capacitance values given in Table 13-II are for a Q of 2 and are the *minimum* values that should be used. The Q may be increased, permitting full power transfer with looser coupling between the coils, by increasing the capacitance and decreasing the inductance correspondingly to maintain resonance.

The capacitance values given are the total capacitance required, so if a balanced condenser is used as indicated at C_1 in Fig. 13-25 each section of the condenser should have twice the capacitance given. A single-ended condenser may be used if care is taken to mount it far enough away from the chassis or any other grounded conductor so that the capacitance from stator and frame to ground is small. In such case the condenser should be tuned by an insulated extension shaft.

The series-tuned circuit shown in the transmitter chapter for coax line can be adapted to use with 75-ohm parallel-conductor line by using two variable condensers, one in each line conductor and each having twice the capacitance specified, and removing the ground connection. This is the best arrangement for maintaining balance to ground, but if reasonable care is taken to mount the condenser as described in the preceding paragraph, a single condenser may be used. In that case the only circuit difference is that neither side of the line should be grounded.

Link Coupling

The coupling arrangements for parallel-conductor line shown in Fig. 13-25 are not entirely satisfactory from a constructional standpoint. It is usually more convenient to build the coupling apparatus separate from the final amplifier, and this leads to greater operating flexibility as well. For lines operating at a low standing-wave ratio this is easily accomplished by connecting the amplifier and coupling circuits through a short length of transmission line or "link." When properly designed and adjusted, the tuning of both circuits will be completely independent of the length of the line connecting them. This method has the further advantage that, if the connecting line is coaxial cable, it offers an ideal spot for the insertion of low-pass filters for preventing harmonic interference to television and FM reception.

The circuit for coax-link coupling is given in Fig. 13-26. The constants of the tuned circuit C_1L_3 are not particularly critical; the principal requirement is that the circuit must be capable of being tuned to the operating frequency. Constants similar to those used in the plate tank circuit will be satisfactory. The construction of L_3 must be such that it can be tapped at least every turn. L_2 must be tightly coupled to L_3, and the inductance of L_2 should be approximately the value that gives a reactance equal to the Z_0 of the connecting line at the frequency in use. An average reactance of about 60 ohms will suffice for either 52- or 75-ohm coaxial line.

The coupling circuit at the amplifier end is merely designed and adjusted for working into a flat coaxial line, as described in the transmitter chapter. Hence the adjustment of coupling at the output end ($L_2L_3C_1$) is entirely independent of the adjustment at the input end (tank circuit and L_1).

When the system is properly designed and operated, the circuit formed by $L_2L_3C_1$ acts purely as a matching device to transform the input impedance of the main transmission line to a value equal to the Z_0 of the coaxial link.

The most satisfactory way to set up the system initially is to connect a coaxial s.w.r. bridge in the link as shown in Fig. 13-26. A simple resistance bridge such as is described in the chapter on measurements is perfectly adequate, requiring only that the transmitter output be reduced to a very low value so that the bridge will not be overloaded. Take a trial position of the line taps on L_3, keeping them equidistant from the center of the coil, and adjust C_1 for minimum s.w.r. as indicated by the bridge. If the s.w.r. is not close to 1 to 1, try new tap positions and adjust C_1 again, continuing this procedure until the s.w.r. is practically 1 to 1. The setting of C_1 and the tap positions may then be logged for future reference, since they will not change so long as the antenna system and frequency are not changed. At this point, check the link s.w.r. over the frequency

TABLE 13-II

Capacitance in μμfd. Required for Coupling to 300- and 600-Ohm Flat Lines with Tuned Coupling Circuit

Frequency Band Mc.	*Characteristic Impedance of Line* 300 ohms	600 ohms
1.8	600	300
3.5	300	150
7	150	75
14	75	40
28	40	20

Note: Inductance in circuit must be adjusted to resonate at operating frequency.

range normally used in that band, without changing the setting of C_1. No readjustment will be required if the s.w.r. does not exceed 1.5 to 1 over the range, but if it goes higher it is advisable to note as many settings of C_1 as may be necessary to keep the s.w.r. below 1.5 to 1 at any part of the band. Changes in the link s.w.r. are caused chiefly by changes in the s.w.r. on the main transmission line with frequency, and relatively little by the coupling circuit itself. A single setting of C_1 at midfrequency will suffice if the antenna itself is broad-tuning.

If it is impossible to get a 1-to-1 s.w.r. at any settings of the taps or C_1, the s.w.r. on the main transmission line is high and the line length is probably unfavorable. Ordinarily there should be no difficulty if the transmission-line s.w.r. is not more than about 3 to 1, but if the line s.w.r. is higher it may not be possible to bring the link s.w.r. down except by using the methods for reactance compensation described in a subsequent section.

The matching adjustment can be considerably facilitated by using a variable condenser in series with the matching-circuit coupling coil as shown in Fig. 13-27. The additional adjustment thus provided makes the tap settings on L_3 much less critical since varying C_2 has the effect of varying the coupling between the two circuits. For optimum control of coupling, L_2 should be somewhat larger than when C_2 is not used — perhaps twice the reactance recommended above — and the reactance of C_2 at maximum capacitance should be the same as that of L_2 at the operating frequency. L_3 and C_1 are the same as before. The method of adjustment is the same, except that for each trial tap position C_1 and C_2 are alternately adjusted, a little at a time, until the s.w.r. is brought to its lowest possible value. In general, the adjustment sought should be the one that keeps C_2 at the largest possible capacitance, since this broadens the frequency response. Also, the taps on L_3 should be kept as far apart as possible, while still permitting a match, since this also broadens the frequency response of the circuit.

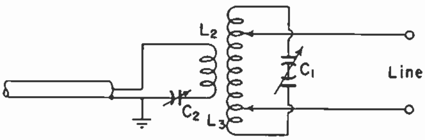


Fig. 13-27 — Using a series condenser for control of coupling between the link and line circuits with the coax-coupled matching circuit.

Once the matching circuit is properly adjusted, the s.w.r. bridge may be removed and full power applied to the transmitter. The input should be controlled by the coupling between L_3 and the amplifier tank coil, never by making any changes in the settings of the matching circuit. If the amplifier will not load properly, tuned coupling should be used into the coax link.

It is possible to use a circuit of this type without initially setting it up with the s.w.r. bridge. In such a case it is a matter of cut-and-try until adequate power transfer between the amplifier and main transmission line is secured. However, this method frequently results in a high s.w.r. in the link, with consequent power loss, "hot spots" in the coaxial cable, and tuning that is critical with frequency. The bridge method is simple and gives the optimum operating conditions quickly and with certainty.

"TUNED" LINES

If the s.w.r. on a transmission line is high enough to cause the input impedance to change appreciably as the applied frequency is varied, the coupling between the transmitter and the line must be changed accordingly to keep the amplifier loading constant. So far as the coupling apparatus is concerned, the principal difference between flat and tuned lines is that the system can be designed for relatively constant impedance for flat lines, but must be capable of coupling into a wide range of impedances if the line is "tuned."

As mentioned earlier, a simple coil can be used for coupling to a line having a high standing-wave ratio providing the line length is adjusted so there is a current loop near the point where it connects to the pick-up coil. The coupling will be maximum, for a given degree of separation between the pick-up coil and the amplifier tank coil, if the line is pruned to a length such that the input impedance is just sufficiently capacitive to cancel the inductive reactance of the pick-up coil. This can be done by cut-and-try. The higher the s.w.r. on the line the easier it becomes to load the amplifier with loose coupling between the two coils. Whether or not good loading can be obtained over a band of frequencies depends on the characteristics of the antenna system. The sharper the antenna and the higher the line s.w.r. the more difficult it becomes to operate over a band without progressively changing the line length.

Series and Parallel Tuning

Rather than adjusting the line length to fit a given coupling coil, it is more practical to adjust the coupling circuit to fit the conditions existing at the input end of the transmission line.

A high standing-wave ratio occurs principally on parallel-conductor lines, either because no attempt has been made at matching the antenna and the line or because the system is used for multiband operation, which precludes such matching. In the latter case, cutting the line length to a multiple of a quarter wavelength will bring either a current or voltage loop near the input terminals of the transmission line (assuming that the antenna itself is resonant) depending on the termination and the line length. If there is a current loop near the input end the impedance will be lower than the line Z_0; if a voltage loop, the input impedance will be higher than the line Z_0. In both cases the input impedances will be essentially resistive.

Under these conditions the circuit arrangements shown in Fig. 13-28 will work satisfactorily. Series tuning is used when a current loop occurs

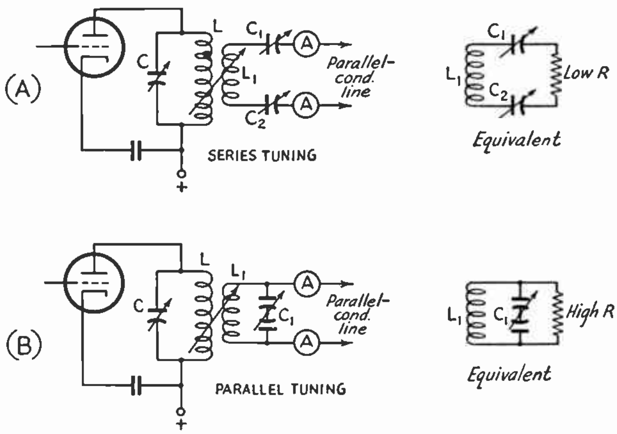


Fig. 13-28 — Series and parallel tuning. This method is useful with resonant lines when the length is such as to bring either a current or voltage loop near the input end. Design data and methods of adjustment are given in the text.

at the input end of the line; parallel tuning when there is a voltage loop at the input end. In the series case, the circuit formed by L_1, C_1 and C_2 with the line terminals short-circuited should tune to the operating frequency. C_1 and C_2 should be maintained at equal capacitance. In the parallel case, the circuit formed by L_1 and C_1 should tune to resonance with the line disconnected.

The L/C ratio in either circuit depends on the transmission line Z_0 and the standing-wave ratio. With series tuning, a high L/C ratio must be used if the s.w.r. is relatively low and the line Z_0 is high. With parallel tuning, a low L/C ratio must be used if the s.w.r. is relatively low and the transmission-line Z_0 also is low. With either series or parallel tuning the L/C ratio becomes less critical when the s.w.r. is high. As a first approximation, coil and condenser values of the same order as those used in the plate tank circuit may be tried.

To adjust the series-tuned circuit, first couple L_1 loosely to the amplifier tank coil and then vary C_1 and C_2, keeping their capacitances equal, until the setting is found that makes the amplifier plate current kick upward. Keep adjusting the amplifier tank condenser, C, for minimum plate current while this is being done. When the proper settings are found, increase the coupling between the two coils until the amplifier draws normal plate current with C adjusted for minimum. It is unnecessary to readjust C_1 and C_2 when the coupling is increased. Keep the coupling between the coils at the smallest value that will load the amplifier properly. If full loading cannot be obtained with the tightest possible coupling, use a coil of more inductance at L_1.

The same adjustment procedure is used with parallel tuning, except that there is only one condenser, C_1. If full loading cannot be secured, reduce the inductance of L_1 and increase C_1 correspondingly to maintain the same frequency, until the amplifier loads properly.

The r.f. ammeters shown in Fig. 13-28 are not strictly necessary, but are useful for indicating maximum output. They may be omitted if desired; in most cases the amplifier plate current is a good enough indication of output, providing the amplifier is operating at normal ratings and efficiency.

In case full loading cannot be obtained even when the L/C ratio is varied, the type of tuning in use probably is not suitable and should be changed; e.g., from series to parallel. If satisfactory loading still cannot be secured, the probability is that the s.w.r. is quite low and the coupling methods designed for flat lines, described earlier, should be used.

Two condensers are used in the series-tuned circuit in order to keep the line balanced to ground. This is because two identical condensers, both connected with either their stators or rotors to the line, will have the same capacitance to ground. A single condenser would be perfectly usable so far as the operation of the coupling circuit is concerned, but will slightly unbalance the circuit because the frame has more capacitance to ground than the stator. The unbalance is not especially serious unless the condenser is mounted near a large mass of metal, such as a chassis or shield assembly.

A balanced condenser is used in the parallel circuit, in preference to a single unit, for the same reason. An alternative scheme to maintain balance is to use two single-ended condensers in parallel, but with the frame of one connected to one side of the line and the frame of the other connected to the other side of the line. The same two condensers may be switched in series when series tuning is to be used.

Link Coupling

The circuits shown in Fig. 13-28 require a means for varying the coupling between two sizable coils, a thing that is somewhat inconvenient constructionally. It is easier to use separate fixed mountings for the final tank and antenna coils and couple them by means of a link. As explained in the chapter on circuit fundamentals, a *short* link is equivalent to providing mutual inductance between two tuned circuits. Typical arrangements for series and parallel tuning are shown in Fig. 13-29. Although these drawings show variable coupling at both ends of the link, a fixed link coil can be used at either end so long as variable coupling is available at the other.

There is no essential difference between the tuning procedures with these circuits and those of Fig. 13-28. The only change is that the coupling is adjusted by means of a link instead of by varying the spacing between L and L_1.

In cases where the link will be more than a few inches long, or when coaxial cable is to be used for the link, it is much better to consider the link as a transmission line that should be properly matched. The circuit of Fig. 13-26 is recommended in that case, except that either a series- or parallel-tuned circuit is substituted for C_1L_3 in that figure. The same considerations apply with respect to the sizes of the link coils, and the best adjustment procedure is that using an s.w.r. bridge.

Lines of Random Length

Series or parallel tuning will always work satisfactorily with lines having a high standing-wave ratio so long as the electrical length of the line is approximately a multiple of a quarter wavelength. However, it is not always possible to couple satisfactorily when intermediate line lengths are used. This is because at some lengths the input impedance of the line has a considerable reactive component, and because the resistive component is too large to be connected in series with a tuned circuit and too low to be connected in parallel.

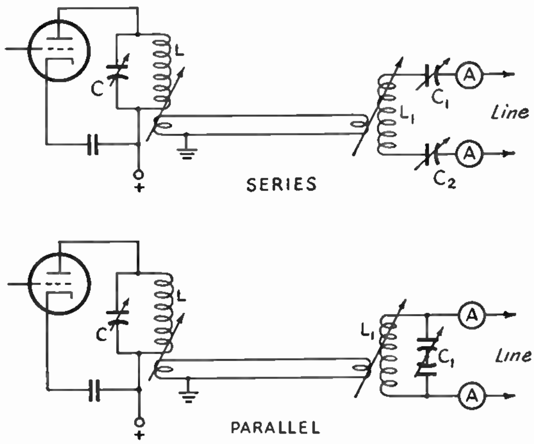


Fig. 13-29 — Link-coupled series and parallel tuning.

The coupling system shown in Fig. 13-26 is capable of handling the resistive component of the input impedance of the transmission lines used in most amateur installations, regardless of the standing-wave ratio on the line. Consequently, it can generally be used wherever either series or parallel tuning would normally be called for, simply by setting the taps properly on the coil. (A possible exception is where the s.w.r. is considerably higher than 10 to 1 and the line length is such as to bring a current loop at the input end. In such a case the resistance may be only a few ohms, which is difficult to match by means of taps on a coil.)

Within limits, the same circuit is capable of being adjusted to compensate for the reactive component of the input impedance; this merely means that a 1-to-1 s.w.r. in the link will be obtained at a different setting of C_1 (Fig. 13-26) than would be the case if the line "looked like" a pure resistance. Sometimes, however, C_1 does not have enough range available to give complete compensation, particularly when (as is the case with some line lengths when the s.w.r. is high) the input impedance is principally reactive.

Under such conditions it is necessary, if the line length cannot be changed to a more satisfactory value, to provide additional means for compensating for or "canceling out" the reactive component of the input impedance. As described earlier in this chapter (Fig. 13-6) the input impedance can be considered to be equivalent to a circuit consisting either of resistance and inductance or resistance and capacitance. It is generally more convenient to consider these elements as a parallel combination, so if the line "looks like" $L'R'$ at A in Fig. 13-6, it is apparent that if we connect a capacitance of the right value across L' the circuit will become resonant and will appear to be a pure resistance of the value R'. Similarly, connecting an inductance of the right value across C'' in Fig. 13-6B will resonate the circuit and the impedance will be equal to R'. The resistive impedance that remains can easily be matched to the coax link by means of the circuit of Fig. 13-26.

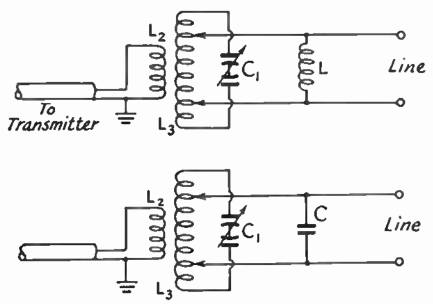


Fig. 13-30 — Reactance cancellation on random-length lines having a high standing-wave ratio.

The practical application of this principle is shown in Fig. 13-30, where L and C are the reactances required to cancel out the line reactance, L for cases where the line is capacitive, C for lines having inductive reactance. The amount of either inductance or capacitance required is easily determined by trial. Using the s.w.r. bridge in the coax link, first disconnect the main transmission line and connect a noninductive resistor to the line terminals. A ½- or 1-watt carbon resistor of about the same resistance as the line Z_0 will do. Adjust the coil taps and C_1 for a 1-to-1 standing-wave ratio in the link, as described earlier. This determines the proper setting of C_1 for a purely resistive load. Then take off the resistor and connect the line, again adjusting the taps and C_1 for minimum s.w.r. If a 1-to-1 ratio can be obtained further compensation is not needed, but if not, make the s.w.r. as low as possible and compare the new setting of C_1 with the original setting. If the capacitance has increased, the line reactance is inductive and a condenser must be connected at C in Fig. 13-30. The amount of capacitance needed to bring the proper setting of C_1 near the original setting can be determined by trial. On the other hand, if the capacitance of C_1 is less than the original, an inductance must be connected at L. Trial values will show when the proper tuning conditions have been reached. It is not necessary that C_1 be at exactly the original setting after the compensating reactance has been adjusted; it is sufficient that it be somewhere in the same vicinity.

Using this procedure practically any length of line can be coupled properly to the transmitter, even when the line s.w.r. is quite high. Unfortunately, no specific values can be suggested for L and C, since they vary widely with line length and s.w.r. Their values usually are comparable with the values used in the regular coupling circuits at the same frequency.

Coupler or Matching-Circuit Construction

The design of matching or "antenna coupler" circuits has been covered in the preceding section, and the adjustment procedure also has been outlined. Since circuits of this type are most frequently used for transferring power from the transmitter to a parallel-conductor transmission line, a principal point requiring attention is that of maintaining good balance to ground. If the coupler circuit is appreciably unbalanced the currents in the two wires of the transmission line will also be unbalanced, resulting in radiation from the line.

In most cases the matching circuit will be built on a metal chassis, following common practice in the construction of transmitting units. The chassis, because of its relatively large area, will tend to establish a "ground" — even though not actually grounded — particularly if it is assembled with other units of the transmitter in a rack or cabinet. The components used in the coupler, therefore, should be placed so that they are electrically symmetrical with respect to the chassis and to each other.

In general, the construction of a coupler circuit should physically resemble the tank layouts used with push-pull amplifiers. In parallel-tuned circuits a split-stator condenser should be used. The condenser frame should be insulated from the chassis because, depending on line length and other factors, harmonic reduction and line balance may be improved in some cases by grounding and in others by not grounding. It is therefore advisable to adopt construction that permits either. Provision also should be made for grounding the center of the coil, for the same reason. The coil in a parallel-tuned circuit should be mounted so that its hot ends are symmetrically placed with respect to the chassis and other components. This equalizes stray capacitances and helps maintain good balance.

When the coupler is of the type that can be shifted to series or parallel tuning as required, two separate single-ended condensers will be satisfactory. As described earlier, they should be connected so that both frames go to the same side of the circuit — i.e., either to the coil or to the line — for series tuning, and when used in parallel for parallel tuning should be connected frame-to-stator.

A coupler designed and adjusted so that the connecting link acts as a matched transmission line may be placed in any convenient location. Some amateurs prefer to install the coupler at the point where the main transmission line enters the station. This helps maintain a neat station layout when an air-insulated parallel-conductor transmission line is used. With solid-dielectric lines, which lend themselves well to neat installation indoors, it is probably more desirable to install the coupler where it can be reached easily for adjustment and band-changing. The use of coaxial line between the transmitter and coupler is strongly recommended if the link line is more than a few inches long, for the reasons outlined in the preceding section.

● COAX-COUPLED MATCHING CIRCUIT

The matching unit shown in Fig. 13-31 is constructed according to the design principles outlined earlier in this chapter. It uses a parallel-tuned circuit with taps for matching a parallel-conductor line through a link coil to a coaxial line to the transmitter. It will handle about 500

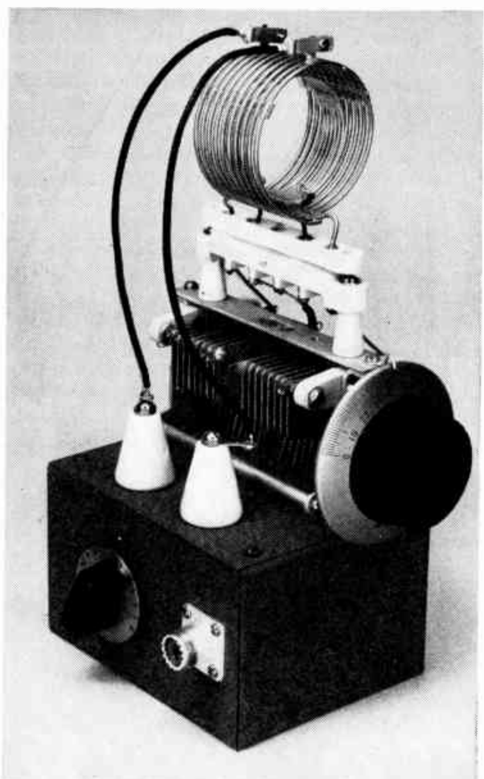


Fig. 13-31 — A coax-coupled matching circuit of simple construction. The entire circuit is mounted on a 3 by 4 by 5 box. C_1 is inside; C_2 and the plug-in coil assembly are mounted on top.

watts of r.f. power and will work, without modification, into lines having an s.w.r. below 3 or 4 to 1. If the s.w.r. is high, it may be necessary to compensate for the reactive part of the input impedance of the line, at certain line lengths, by using an additional coil or condenser as discussed earlier. The necessity for such compensation can be avoided, on lines having a high s.w.r., by making the electrical length of the line a multiple of a quarter wavelength.

As shown by the circuit diagram, Fig. 13-32, the link circuit is adjusted by means of a variable condenser, C_1, to facilitate matching the main transmission line to the coax link. The coils are constructed from commercially-available coil material, and the link inductances are chosen to provide adequate coupling for flat lines. The link coil, of smaller diameter than the tank coil, is mounted inside the latter at the center. Duco cement is used to hold the coils together at their bottom tie strips. The coils are mounted on Millen type 40305 plugs and require no other support than the stiffness of the short lengths of wire going into the end prongs of the plug from the tank coil. Short lengths of spaghetti tubing are slipped over the leads to the link coil where they go between the tank coil turns to reach the plug.

Taps on the tank coil for connection to a parallel-conductor transmission line are made by bending ordinary soldering lugs around the wire and soldering them in place. The clips are Johnson type 235-860, adjusted so that they fit snugly over the taps when pushed on sidewise. Used this way, the clips provide an easy and rapid method of connecting and disconnecting the line. The proper positions for the taps may be determined by first using the clips in the normal fashion.

The maximum length of coil that can be mounted satisfactorily on the plugs is about 4 inches, and a coil of this size cannot be tuned to the 3.5-Mc. band with the 100-μμfd.-per-section split-stator condenser used in this unit. To cover the 3.5-Mc. band it is necessary to shunt the coil with an additional capacitance of about 75 μμfd.

The matching circuit should be adjusted with the aid of an s.w.r. bridge, as described earlier in this chapter. In general, the tuning will be less critical, and the circuit will work over a wider frequency range without readjustment, if the taps are kept as far toward the ends of the coil as possible and C_1 is set at the largest capacitance that will permit bringing the s.w.r. in the coax link down to 1 to 1.

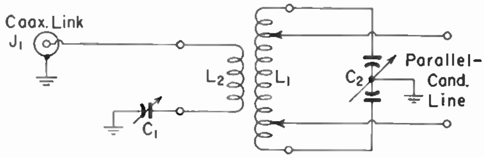


Fig. 13-32 — Circuit diagram of the coax-coupled matching circuit.

C_1 — 300-μμfd. variable, approximately 0.024″ spacing.
C_2 — 100 μμfd. per section, 1500 volts (National TMK-100D).
J_1 — Chassis-type coax connector.

Coil Data

Band	*L_1, turns*	*L_2, turns*
3.5 Mc.*	24 (17 μh.)	10 (5 μh.)
7 Mc.	18 (12 μh.)	6 (2.5 μh.)
14 Mc.	10 (5 μh.)	3 (1 μh.)
21–28 Mc.	6 (2.5 μh.)	2

* Add 75 μμfd. in parallel with C_2.

L_1 — No. 12 tinned wire, 2½ inches dia., 6 turns per inch (B & W 3905-1).
L_2 — No. 16 wire, 2 inches dia., 10 turns per inch (B & W 3907 or 3907-1).

A "UNIVERSAL" MATCHING CIRCUIT

The matching circuit shown in Fig. 13-33 offers considerable flexibility in that it can be used as a tapped-coil matching network of the same type as that just described, and also can be used as either a series- or parallel-tuned "antenna coupler." It can also be adapted to other types of coupling by simple changes in the plug-connection arrangement of the coils.

Two condensers are used in the tank circuit. Their rotors are insulated from each other but are turned simultaneously by a right-angle drive unit. When used either for parallel tuning or the tapped-coil method of matching, the rotors are connected together to form a split-stator condenser having a maximum capacitance of 150

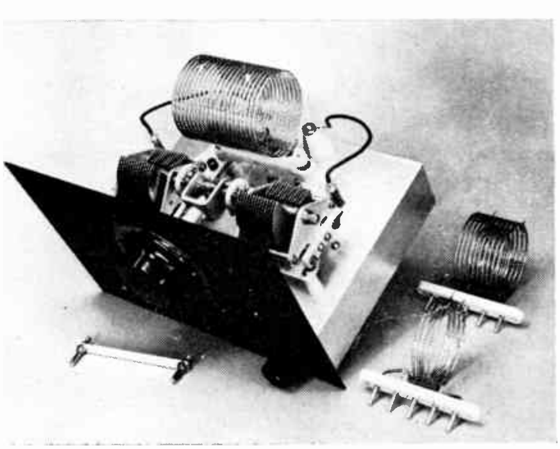


Fig. 13-33 — A coupler or matching network that can also be used for series or parallel tuning of tuned lines.

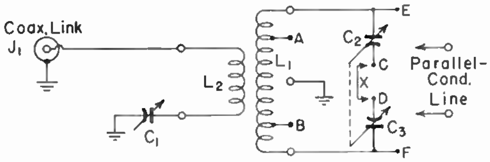


Fig. 13-34 — Circuit diagram of the "universal" coax-coupled matching network. For use as a tapped matching circuit, connect the line to taps on L_1, as at *A-B*, and connect the jumper, X, to *A-B*; the jumper is also used for parallel tuning but with the line connected to *E-F*. For series tuning, remove the jumper and connect the line to *C-D*. The ground connection to the middle prong of the coil socket is provided for cases where it is desirable to ground the center of L_1.

C_1 — 300-$\mu\mu$fd. variable, approximately 0.024″ spacing.
C_2, C_3 — 300-$\mu\mu$fd. variable, 1000 volts (National TMS-300).
J_1 — Chassis-type coax connector.

Coil Data

Band	L_1, *turns*	L_2, *turns*
3.5–7 Mc.	20 (14 μh.)	10 (5 μh.)
7–14 Mc.	10 (5 μh.)	6 (2.5 μh.)
14–28 Mc.	4 (1.5 μh.)	2

L_1 — No. 12 tinned wire, 2½ inches dia., 6 turns per inch (B & W 3905-1).
L_2 — No. 16 wire, 2 inches dia., 10 turns per inch (B & W 3907 or 3907-1).

$\mu\mu$fd. When used for series tuning the condenser frames connect to the parallel-conductor transmission line, the jumper that connects the rotors together being removed.

The unit is built on a 7 by 9 by 2 aluminum chassis and has a 7 by 10 panel. The tank condensers are mounted on small aluminum plates supported on ¾-inch stand-off insulators, to insulate the frames from the chassis; this method is preferable to mounting the condensers directly on the insulators as it lessens the mechanical strain on the latter. The soldering lugs projecting from the condensers provide means for connecting the line clips for series and parallel tuning. The jumper for connecting the rotors together is in the foreground; it uses banana plugs that fit into jacks mounted on the condenser mounting plates. The link condenser is located underneath the chassis.

The coils shown are designed primarily for use in the tapped matching circuit or for parallel tuning, but will also be satisfactory for series tuning if the transmission line length is such as to bring a current loop near the input end. Coil taps are made in the same way as in the coupler previously described. Soldering lugs are also used as taps on C_2 and C_3 to make the necessary connections for series or parallel tuning. Because of the fairly large value of maximum capacitance available when the tank condensers, C_2 and C_3, are used together as a split-stator condenser, it is possible to cover a 2-to-1 frequency range. Consequently, only three coil assemblies are needed to cover the 3.5- to 30-Mc. range, and each one can be used for two (in the case of the smallest coil, three) adjacent amateur bands.

As a tapped matching circuit, adjustment is the same as for the unit just described. When using either series or parallel tuning, the s.w.r. bridge should be used as before, adjusting C_1 and C_2-C_3 for minimum s.w.r. in the coax link.

Antennas

An *antenna system* can be considered to include the antenna proper (the portion that radiates the r.f. energy), the feedline, and any coupling devices used for transferring power from the transmitter to the line and from the line to the antenna. Some simple systems may omit the transmission line or one or both of the coupling devices. This chapter will describe the antenna proper, and in many cases will show popular types of lines, as well as line-to-antenna couplings where they are required. However, it should be kept in mind that *any* antenna proper can be used with *any* type of feedline if a suitable coupling is used between the antenna and the line. Changing the line does not change the type of antenna.

Selecting an Antenna

In selecting the type of antenna to use, the majority of amateurs are somewhat limited through space and structural limitations to simple antenna systems, except for v.h.f. operation where the small space requirements make the use of multielement beams readily possible. This chapter will consider antennas for frequencies as high as 30 Mc. — a later chapter will describe the popular types of v.h.f. antennas. However, even though the available space may be limited, it is well to consider the propagation characteristics of the frequency band or bands to be used, to insure that best possible use is made of the available facilities. The propagation characteristics of the various bands, up to 30 Mc., are described in Chapter Four. In general, antenna construction and location become more critical and important on the higher frequencies. On the lower frequencies (3.5 and 7 Mc.) the vertical angle of radiation and the plane of polarization may be of relatively little importance; at 28 Mc. they may be all-important. On a given frequency, the particular type of antenna best suited for long-distance communication may not be as good for shorter-range work as would a different type.

Definitions

The important properties of an antenna proper are its polarization, vertical and horizontal angles of maximum radiation, impedance, gain and bandwidth.

The **polarization** of a straight-wire antenna is determined by its position with respect to the earth. Thus a vertical antenna radiates vertically-polarized waves, while a horizontal antenna radiates horizontally-polarized waves in a direction broadside to the wire and vertically-polarized waves at high vertical angles off the ends of the wire. The wave from an antenna in a slanting position, or from the horizontal antenna in directions other than mentioned above, contains both horizontal and vertical components.

The vertical angle of maximum radiation of an antenna is determined by the free-space pattern of the antenna, its height above ground, and the nature of the ground. The angle is measured in a vertical plane with respect to a tangent to the earth at that point, and it will usually vary with the horizontal angle, except in the case of a simple vertical antenna. The **horizontal angle of maximum radiation** of an antenna is determined by the free-space pattern of the antenna.

The **impedance** of the antenna at any point is the ratio of the voltage to the current at that point. It is important in connection with feeding power to the antenna, since it constitutes the load to the line offered by the antenna. It can be either resistive or complex, depending upon whether or not the antenna is resonant.

The **field strength** produced by an antenna is proportional to the current flowing in it. When there are standing waves on an antenna, the parts of the wire carrying the higher current have the greater radiating effect. All resonant antennas have standing waves — only terminated types, like the terminated rhombic and terminated "V," have substantially uniform current along their lengths.

The ratio of power required to produce a given field strength, with a "comparison" antenna, to the power required to produce the same field strength with a specified type of antenna is called the **power gain** of the latter antenna. The field is measured in the optimum direction of the antenna under test. In amateur work, the comparison antenna is generally a half-wave antenna at the same height and having the same polarization as the antenna under consideration. Power gain usually is expressed in decibels.

In unidirectional beams (antenna systems with maximum radiation in only one direction) the **front-to-back** ratio is the ratio of power radiated in the maximum direction to power radiated in the opposite direction. It is also a measure of the reduction in received signal when the beam direction is changed from that for maximum response to the opposite direction. Front-to-back ratio is usually expressed in decibels.

The **bandwidth** of an antenna generally refers to the frequency range over which the gain and impedance are substantially constant. It is of importance primarily in connection with multielement beams fed by a "flat" transmission line.

Ground Effects

The radiation pattern of any antenna that is many wavelengths distant from the ground and all other objects is called the **free-space pattern** of that antenna. The free-space pattern of an antenna is almost impossible to obtain in practice, except in the v.h.f. and u.h.f. ranges. Below 30 Mc., the location of the antenna with respect to ground plays an important part in determining the actual radiation pattern of the antenna.

When any antenna is near the ground the free-space pattern is modified by reflection of radiated waves from the ground, so that the actual pattern is the resultant of the free-space pattern and ground reflections. This resultant is dependent upon the height of the antenna, its position or orientation with respect to the surface of the ground, and the electrical characteristics of the ground. The effect of a perfectly-reflecting ground is such that the original free-space field strength may be multiplied by a factor which has a maximum value of 2, for complete reinforcement, and having all intermediate values to zero, for complete cancellation. These reflections only affect the radiation pattern in the vertical plane — that is, in directions upward from the earth's surface — and not in the horizontal plane, or the usual geographical directions.

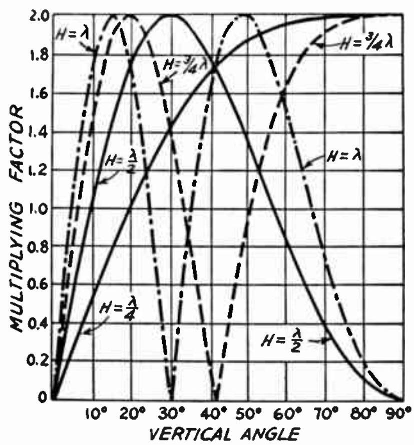


Fig. 14-1 — Effect of ground on radiation of horizontal antennas at vertical angles for four antenna heights. This chart is based on perfectly-conducting ground.

Fig. 14-1 shows how the multiplying factor varies with the vertical angle for several representative heights for horizontal antennas. As the height is increased the angle at which complete reinforcement takes place is lowered, until for a height equal to one wavelength it occurs at a vertical angle of 15 degrees. At still greater heights, not shown on the chart, the first maximum will occur at still smaller angles.

Radiation Angle

The vertical angle of maximum radiation, is of primary importance, especially at the higher frequencies. It is advantageous, therefore, to erect the antenna at a height that will take advantage of ground reflection in such a way as to reinforce the space radiation at the most desirable angle. Since low angles usually are most effective, this generally means that the antenna should be high — at least one-half wavelength at 14 Mc., and preferably three-quarters or one wavelength, and at least one wavelength, and preferably higher, at 28 Mc. The physical height required for a given height in wavelengths decreases as the frequency is increased, so that good heights are not impracticable; a half-wavelength at 14 Mc. is only 35 feet, approximately, while the same height represents a full wavelength at 28 Mc. At 7 Mc. and lower frequencies the higher radiation angles are effective, so that again a useful antenna height is not difficult of attainment. Heights between 35 and 70 feet are suitable for all bands, the higher figures being preferable.

Imperfect Ground

Fig. 14-1 is based on ground having perfect conductivity, whereas the actual earth is not a perfect conductor. The principal effect of actual ground is to make the curves inaccurate at the lowest angles; appreciable high-frequency radiation at angles smaller than a few degrees is practically impossible to obtain over horizontal ground. Above 15 degrees, however, the curves are accurate enough for all practical purposes, and may be taken as indicative of the result to be expected at angles between 5 and 15 degrees.

The effective ground plane — that is, the plane from which ground reflections can be considered to take place — seldom is the actual surface of the ground but is a few feet below it, depending upon the character of the soil.

Impedance

Waves that are reflected directly upward from the ground induce a current in the antenna in passing, and, depending on the antenna height, the phase relationship of this induced current to the original current may be such as either to increase or decrease the total current in the antenna. For the same power input to the antenna, an increase in current is equivalent to a decrease in impedance, and vice versa. Hence, the impedance of the antenna varies with height. The theoretical curve of variation of radiation resistance for an antenna above perfectly-reflecting ground is shown in Fig. 14-2. The impedance approaches the free-space value as the height becomes large, but at low heights may differ considerably from it.

Choice of Polarization

Polarization of the transmitting antenna is generally unimportant on frequencies between 3.5 and 30 Mc. However, the question of whether the antenna should be installed in a horizontal or vertical position deserves consideration for other reasons. A vertical half-wave or quarter-wave antenna will radiate equally well in all *horizontal* directions, so that it is substantially nondirectional, in the usual sense of the word. If installed horizontally, however, the antenna will tend to show directional effects, and will radiate best in the direction at right angles, or broadside, to the wire. The radiation in such a case will be least in the direction toward which the wire points.

The vertical angle of radiation also will be affected by the position of the antenna. If it were not for ground losses at high frequencies, the vertical half-wave antenna would be preferred because it would concentrate the radiation horizontally.

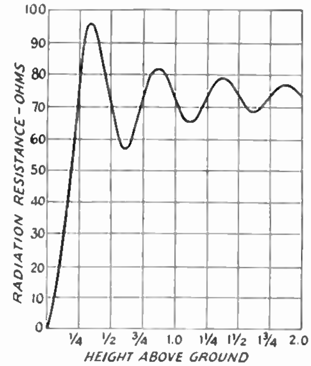


Fig. 14-2 — Theoretical curve of variation of radiation resistance for a half-wave horizontal antenna, as a function of height in wavelength above perfectly-reflecting ground.

The Half-Wave Antenna

The fundamental form of antenna is a single wire whose length is approximately equal to half the transmitting wavelength. It is the unit from which many more-complex forms of antennas are constructed. It is variously known as a **half-wave dipole, half-wave doublet,** or **Hertz** antenna.

The length of a half-wavelength in space is:

$$\textit{Length} \text{ (feet)} = \frac{492}{\textit{Freq.} \text{ (Mc.)}} \qquad \text{(14-A)}$$

The actual length of a half-wave antenna will not be exactly equal to the half-wave in space, but depends upon the thickness of the conductor in relation to the wavelength as shown in Fig. 14-3, where K is a factor that must be multiplied by the half-wavelength in free space to obtain the resonant antenna length. An additional shortening effect occurs with wire antennas supported by insulators at the ends because of the capacitance added to the system by the insulators **(end effect).** The following formula is sufficiently accurate for wire antennas at frequencies up to 30 Mc.:

$$\textit{Length of half-wave antenna} \text{ (feet)} = \frac{492 \times 0.95}{\textit{Freq.} \text{ (Mc.)}} = \frac{468}{\textit{Freq.} \text{ (Mc.)}} \qquad \text{(14-B)}$$

Example: A half-wave antenna for 7150 kc. (7.15 Mc.) is $\frac{468}{7.15}$ = 65.45 feet, or 65 feet 5 inches.

Above 30 Mc. the following formulas should be used, particularly for antennas constructed from rod or tubing. K is taken from Fig. 14-3.

$$\textit{Length of half-wave antenna} \text{ (feet)} = \frac{492 \times K}{\textit{Freq.} \text{ (Mc.)}} \qquad \text{(14-C)}$$

$$\textit{or length} \text{ (inches)} = \frac{5905 \times K}{\textit{Freq.} \text{ (Mc.)}} \qquad \text{(14-D)}$$

Example: Find the length of a half-wavelength antenna at 29 Mc., if the antenna is made of 2-inch diameter tubing. At 29 Mc., a half-wavelength in space is $\frac{492}{29}$ = 16.97 feet, from Eq. 14-A. Ratio of half-wavelength to conductor diameter (changing wavelength to inches) is $\frac{16.97 \times 12}{2}$ = 101.8. From Fig. 14-3, K = 0.963 for this ratio. The length of the antenna, from Eq. 14-C, is $\frac{492 \times 0.963}{29}$ = 16.34 feet, or 16 feet 4 inches. The answer is obtained directly in inches by substitution in Eq. 14-D: $\frac{5905 \times 0.963}{29}$ = 196 inches.

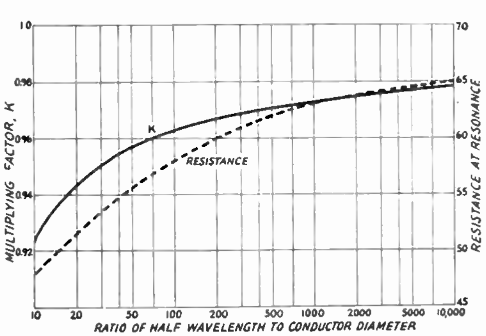


Fig. 14-3 — Effect of antenna diameter on length for half-wave resonance, shown as a multiplying factor, K, to be applied to the free-space half-wavelength (Equation 14-A). The effect of conductor diameter on the impedance measured at the center also is shown.

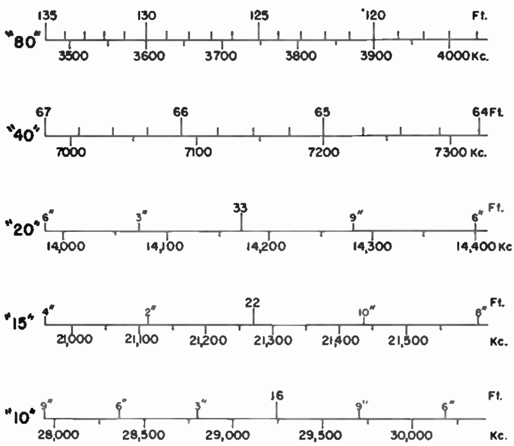


Fig. 14-4 — The above scales, based on Eq. 14-B, can be used to determine the length of a half-wave antenna of wire.

Current and Voltage Distribution

When power is fed to a half-wave antenna, the current and voltage vary along its length. The current is maximum at the center and nearly zero at the ends, while the opposite is true of the r.f. voltage. The current does not actually reach zero at the current nodes, because of the end effect; similarly, the voltage is not zero at its node because of the resistance of the antenna, which consists of both the r.f. resistance

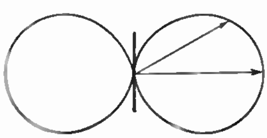


Fig. 14-5 — The free-space radiation pattern of a half-wave antenna. The antenna is shown in the vertical position. This is a cross-section of the solid pattern described by the figure when rotated on its vertical axis. The "doughnut" form of the solid pattern can be more easily visualized by imagining the drawing glued to a piece of cardboard, with a short length of wire fastened on it to represent the antenna. Twirling the wire will give a visual representation of the solid radiation pattern.

of the wire (*ohmic resistance*) and the radiation resistance. The radiation resistance is an *equivalent* resistance, a convenient conception to indicate the radiation properties of an antenna. The radiation resistance is the equivalent resistance that would dissipate the power the antenna radiates, with a current flowing in it equal to the antenna current at a current loop (maximum). The ohmic resistance of a half-wavelength antenna is ordinarily small enough, in comparison with the radiation resistance, to be neglected for all practical purposes.

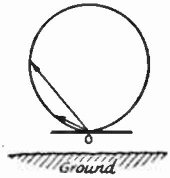


Fig. 14-6 — Illustrating the importance of vertical angle of radiation in determining antenna directional effects. Off the end, the radiation is greater at higher angles. Ground reflection is neglected in this drawing of the free-space pattern of a horizontal antenna.

Impedance

The radiation resistance of an infinitely-thin half-wave antenna in free space — that is, sufficiently removed from surrounding objects so that they do not affect the antenna's characteristics — is 73 ohms, approximately. The value under practical conditions is commonly taken to be in the neighborhood of 70 ohms. It is pure resistance, and is measured at the center of the antenna. The impedance is minimum at the center, where it is equal to the radiation resistance, and increases toward the ends. The actual value at the ends will depend on a number of factors, such as the height, the physical construction, the insulators at the ends, and the position with respect to ground.

Conductor Size

The impedance of the antenna also depends upon the diameter of the conductor in relation to the wavelength, as shown in Fig. 14-3. If the diameter of the conductor is made large, the capacitance per unit length increases and the inductance per unit length decreases. Since the radiation resistance is affected relatively little, the decreased L/C ratio causes the Q of the antenna to decrease, so that the resonance curve becomes less sharp. Hence, the antenna is capable of working over a wide frequency range. This effect is greater as the diameter is increased, and is a property of some importance at the very-high frequencies where the wavelength is small.

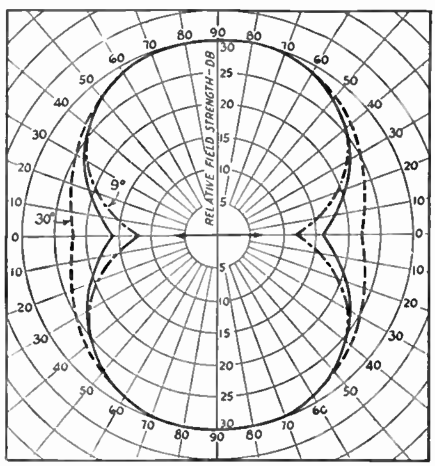


Fig. 14-7 — Horizontal pattern of a horizontal half-wave antenna at three vertical radiation angles. The solid line is relative radiation at 15 degrees. Dotted lines show deviation from the 15-degree pattern for angles of 9 and 30 degrees. The patterns are useful for shape only, since the amplitude will depend upon the height of the antenna above ground and the vertical angle considered. The patterns for all three angles have been proportioned to the same scale, but this does not mean that the maximum amplitudes necessarily will be the same. The arrow indicates the direction of the horizontal antenna wire.

Radiation Characteristics

The radiation from a half-wave antenna is not uniform in all directions but varies with the angle with respect to the axis of the wire. It is most intense in directions perpendicular to the wire and zero along the direction of the wire, with intermediate values at intermediate angles. This is shown by the sketch of Fig. 14-5, which represents the radiation pattern in free space. The relative intensity of radiation is proportional to the length of a line drawn from the center of the figure to the perimeter. If the antenna is vertical, as shown in the figure, then the field strength will be uniform in all horizontal directions; if the antenna is horizontal, the relative field strength will depend upon the direction of the receiving point with respect to the direction of the antenna wire. The variation in radiation at various vertical angles from a half-wavelength horizontal antenna is indicated in Figs. 14-6 and 14-7.

FEEDING THE HALF-WAVE ANTENNA

Direct Feed

If possible, it is advisable to locate the antenna at least a half-wavelength from the transmitter and use a transmission line to carry the power from the transmitter to the antenna. However, in many cases this is impossible, particularly on the lower frequencies, and direct feed must be used. Three examples of direct feed are shown in Fig. 14-8. In the method shown at A, C_1 and C_2 should be about 150 $\mu\mu$fd. each for the 3.5-Mc. band, 75 $\mu\mu$fd. each at 7 Mc., and proportionately smaller at the higher frequencies. The antenna coil connected between them should resonate to 3.5 Mc. with about 60 or 70 $\mu\mu$fd., for the 80-meter band, for 40 meters it should resonate with 30 or 35 $\mu\mu$fd., and so on. The circuit is adjusted by using loose coupling between the antenna coil and the transmitter tank coil and adjusting C_1 and C_2 until resonance is indicated by an increase in plate current. The coupling between the coils should then be increased until proper plate current is drawn. It may be necessary to reresonate the transmitter tank circuit as the coupling is increased, but the change should be small.

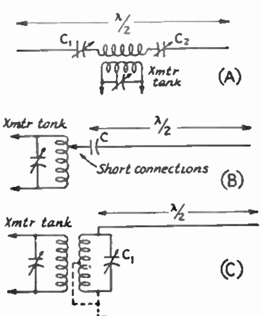


Fig. 14-8 — Methods of directly exciting the half-wave antenna. A, current feed, series tuning; B, voltage feed, capacitive coupling; C, voltage feed, with inductively-coupled antenna tank. In A, the coupling circuit is not included in the effective electrical length of the antenna system proper.

The circuits in Fig. 14-8B and C are used when only one end of the antenna is accessible. In B, the coupling is adjusted by moving the tap toward the "hot" or plate end of the tank coil — the condenser C may be of any convenient value that will stand the voltage, and it doesn't have to be variable. In the circuit at C, the antenna tuned circuit (C_1 and the antenna coil) should be similar to the transmitter tank circuit. The antenna tuned circuit is adjusted to resonance with the antenna connected but with loose coupling to the transmitter. Heavier loading of the tube is then obtained by tightening the coupling between the antenna coil and the transmitter tank coil.

Of the three systems, that at A is preferable because it is a symmetrical system and generally results in less r.f. power "floating" around the shack. The system of B is undesirable because it provides practically no protection against the radiation of harmonics, and it should only be used in emergencies.

Transmission-Line Feed for Half-Wave Antennas

Since the impedance at the center of a half-wavelength antenna is in the vicinity of 75 ohms, it offers a good match for 75-ohm two-wire transmission lines. Several types are available on the market, with different power-handling capabilities. They can be connected in the center of the antenna, across a small strain insulator to provide a convenient connection point. Coaxial line of 75 ohms impedance can also be used, but it is heavier and thus not as convenient. In either case, the transmission line should be run away at right angles to the antenna for at least one-quarter wavelength, if possible, to avoid current unbalance in the line caused by pick-up from the antenna. The antenna length is calculated from Equation 14-B, for a half-wavelength antenna. When No. 12 or No. 14 enameled wire is used for the antenna, as is generally the case, the length of the wire is the over-all length measured from the loop through the insulator at each end. This is illustrated in Fig. 14-9.

The use of 75-ohm line results in a "flat" line over most of any amateur band. However, by making the half-wave antenna in a special manner, called the **two-wire** or **folded dipole,** a good match is offered for a 300-ohm line. Such an antenna is shown in Fig. 14-10. The open-wire line shown in Fig. 14-10 is made of

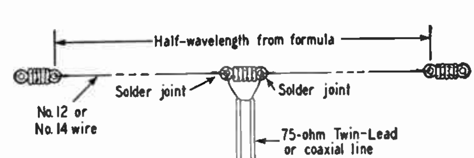


Fig. 14-9 — Construction of a half-wave doublet fed with 75-ohm line. The length of the antenna is calculated from Equation 14-B or Fig. 14-4.

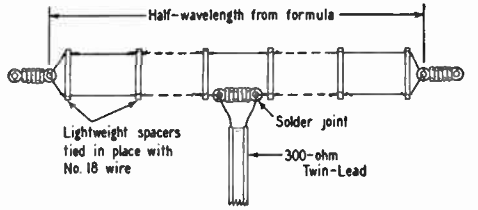


Fig. 14-10 — The construction of an open-wire folded doublet fed with 300-ohm line. The length of the antenna is calculated from Equation 14-B or Fig. 14-4.

No. 12 or No. 14 enameled wire, separated by lightweight spacers of Lucite or other material (it doesn't have to be a *low-loss* insulating material), and the spacing can be on the order of from 4 to 8 inches, depending upon what is convenient and what the operating frequency is. At 14 Mc., 4-inch separation is satisfactory, and 8-inch or even greater spacing can be used at 3.5 Mc.

The half-wavelength antenna can also be made from the proper length of 300-ohm line, opened on one side in the center and connected to the feedline. After the wires have been soldered together, the joint can be strengthened by molding some of the excess insulating material (polyethylene) around the joint with a hot iron, or a suitable lightweight clamp of two pieces of Lucite can be devised.

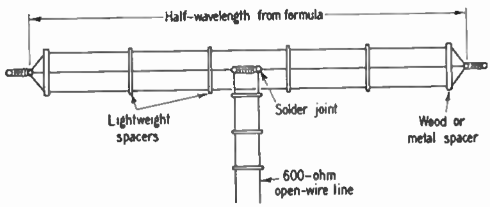


Fig. 14-11 — The construction of a 3-wire folded dipole is similar to that of the 2-wire folded dipole. The end spacers may have to be slightly stronger than the others because of the greater compression force on them. The length of the antenna is obtained from Equation 14-B or Fig. 14-4. A suitable line can be made from No. 14 wire spaced 4½ to 5 inches, or from No. 12 wire spaced 6 inches.

Similar in some respects to the two-wire folded dipole, the three-wire folded dipole of Fig. 14-11 offers a good match for a 600-ohm line. It is favored by amateurs who prefer to use an open-wire transmission line instead of the 300-ohm insulated line. The three wires of the antenna proper should all be of the same diameter.

Another method for offering a match to a 600-ohm open-wire line with a half-wavelength antenna is shown in Fig. 14-12. The system is called a **delta match**. The line is "fanned" as it approaches the antenna, to have a gradually-increasing impedance that equals the antenna impedance at the point of connection. The dimensions are fairly critical, but careful measurement before installing the antenna and matching section is generally all that is necessary. The length of the antenna, L, is calculated from Equation **14-B** or Fig. 14-4. The length of section C is computed from:

$$C \text{ (feet)} = \frac{118}{Freq. \text{ (Mc.)}} \qquad \textbf{(14-E)}$$

The feeder clearance, E, is found from

$$E \text{ (feet)} = \frac{148}{Freq. \text{ (Mc.)}} \qquad \textbf{(14-F)}$$

Example: For a frequency of 7.1 Mc., the length

$$L = \frac{468}{7.1} = 65.91 \text{ feet, or 65 feet 11 inches.}$$

$$C = \frac{118}{7.1} = 16.62 \text{ feet, or 16 feet 7 inches.}$$

$$E = \frac{148}{7.1} = 20.84 \text{ feet, or 20 feet 10 inches.}$$

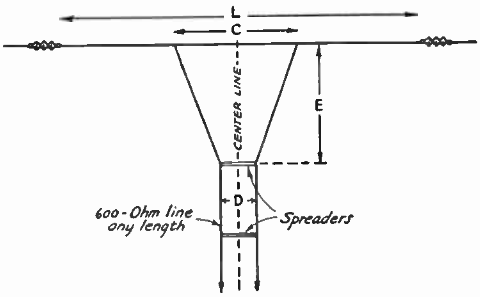


Fig. 14-12 — Delta-matched antenna system. The dimensions C, D, and E are found by formulas given in the text. It is important that the matching section, E, come straight away from the antenna without any bends.

Since the equations hold only for 600-ohm line, it is important that the line be close to this value. This requires 4¾-inch spaced No. 14 wire, 6-inch spaced No. 12 wire, or 3¾-inch spaced No. 16 wire.

If a half-wavelength antenna is fed at the center with other than 75-ohm line, or if a two-wire dipole is fed with other than 300-ohm line, standing waves will appear on the line and coupling to the transmitter may become awkward for some line lengths, as described in the preceding chapter. However, in many cases it is not convenient to feed the half-wave antenna with the correct line (as is the case where multiband operation of the same antenna is desired), and sometimes it is not convenient to feed the antenna at the center. Where multiband operation is desired (to be discussed later) or when the antenna must be

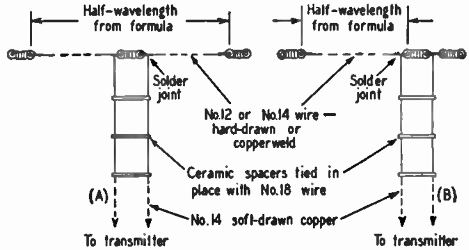


Fig. 14-13 — The half-wave antenna can be fed at the center or at the end with an open-wire line. The antenna length is obtained from Equation 14-B or Fig. 14-4.

fed at one end by a transmission line, an open-wire line of from 450 to 600 ohms impedance is generally used. The impedance at the end of a half-wavelength antenna is in the vicinity of several thousand ohms, and hence a standing-wave ratio of 4 or 5 is not unusual when the line is connected to the end of the antenna. It is advisable, therefore, to keep the losses in the line as low as possible. This requires the use of ceramic or Micalex feeder spacers, if any appreciable power is used. For low-power installations in dry climates, dry wood spacers that have been boiled in paraffin are satisfactory. Mechanical details of half-wavelength antennas fed with open-wire lines are given in Fig. 14-13. If the power level is low, below 100 watts or so, 300-ohm Twin-Lead can be used in place of the open line.

Long-Wire Antennas

An antenna will be resonant so long as an integral number of standing waves of current and voltage can exist along its length; in other words, so long as its length is some integral multiple of a half-wavelength. When the antenna is more than a half-wave long it usually is called a long-wire antenna, or a harmonic antenna.

Current and Voltage Distribution

Fig. 14-14 shows the current and voltage distribution along a wire operating at its fundamental frequency (where its length is equal to a half-wavelength) and at its second, third and fourth harmonics. For example, if the fundamental frequency of the antenna is 7 Mc., the current and voltage distribution will be as shown at A. The same antenna excited at 14 Mc. would have current and voltage distribution as shown at B. At 21 Mc., the third harmonic of 7 Mc., the current and voltage distribution would be as in C; and at 28 Mc., the fourth harmonic, as in D. The number of the harmonic is the number of half-waves contained in the antenna at the particular operating frequency.

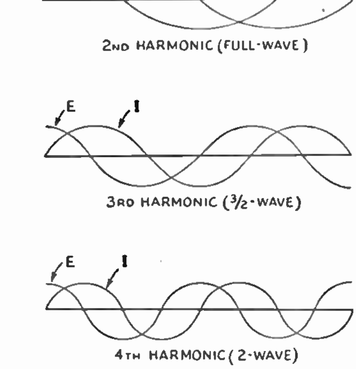


Fig. 14-14 — **Standing-wave current and voltage distribution along an antenna when it is operated at various harmonics of its fundamental resonant frequency.**

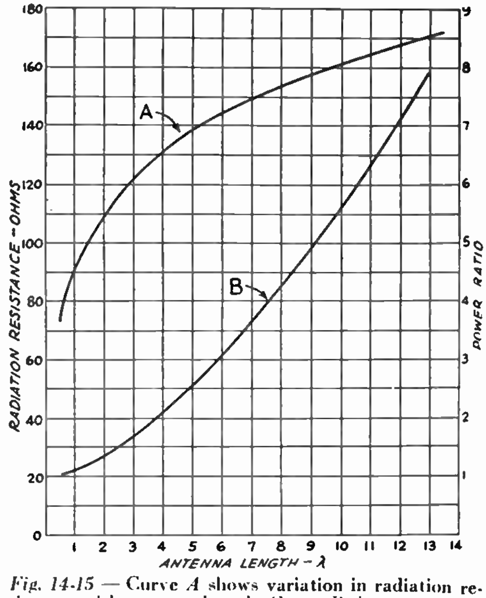


Fig. 14-15 — **Curve *A* shows variation in radiation resistance with antenna length. Curve *B* shows power in lobes of maximum radiation for long-wire antennas as a ratio to the maximum radiation for a half-wave antenna.**

The polarity of current or voltage in each standing wave is opposite to that in the adjacent standing waves. This is shown in the figure by drawing the current and voltage curves successively above and below the antenna (taken as a zero reference line), to indicate that the polarity reverses when the current or voltage goes through zero. Currents flowing in the same direction are *in phase;* in opposite directions, *out of phase.*

It is evident that one antenna may be used for harmonically-related frequencies, such as the various amateur bands. The long-wire or harmonic antenna is the basis of multiband operation with one antenna.

Physical Lengths

The length of a long-wire antenna is not an exact multiple of that of a half-wave antenna because the end effects operate only on the end sections of the antenna; in other parts of the wire these effects are absent, and the wire length is approximately that of an equivalent

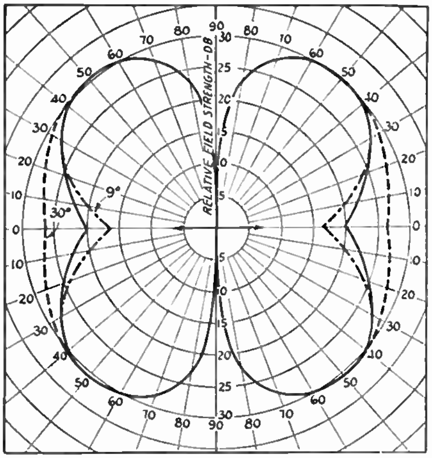


Fig. 14-16 — Horizontal patterns of radiation from a *full-wave* antenna. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. All three patterns are drawn to the same relative scale: actual amplitudes will depend upon the height of the antenna.

portion of the wave in space. The formula for the length of a long-wire antenna, therefore, is

$$\textit{Length}\ (\text{feet}) = \frac{492\ (N-0.05)}{\textit{Freq.}\ (\text{Mc.})} \qquad \textbf{14-G}$$

where N is the number of *half*-waves on the antenna.

Example: An antenna 4 half-waves long at 14.2 Mc. would be $\frac{492\ (4 - 0.05)}{14.2} = \frac{492 \times 3.95}{14.2}$ = 136.7 feet, or 136 feet 8 inches.

It is apparent that an antenna cut as a half-wave for a given frequency will be slightly off resonance at exactly twice that frequency (the second harmonic), because of the decreased influence of the end effects when the antenna is more than one-half wavelength long. The effect is not very important, except for a possible unbalance in the feeder system and consequent radiation from the feedline. If the antenna is fed in the exact center, no unbalance will occur at any frequency, but end-fed systems will show an unbalance in all but one frequency, the frequency for which the antenna is cut.

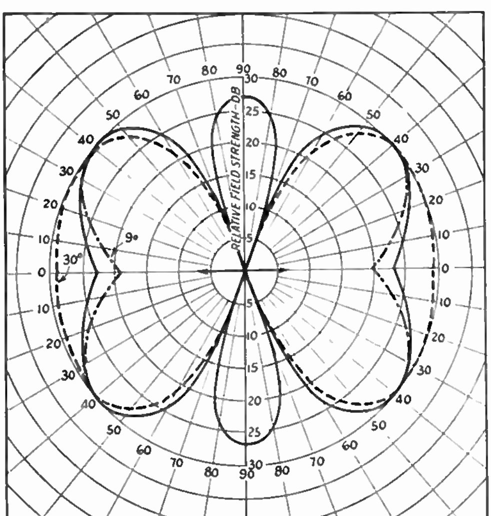


Fig. 14-17 — Horizontal patterns of radiation from an antenna *three half-waves* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. Minor lobes coincide for all three angles.

Impedance and Power Gain

The radiation resistance as measured at a current loop becomes larger as the antenna length is increased. Also, a long-wire antenna radiates more power in its most favorable direction than does a half-wave antenna in its most favorable direction. This power gain is secured at the expense of radiation in other directions. Fig. 14-15 shows how the radiation resistance and the power in the lobe of maximum radiation vary with the antenna length.

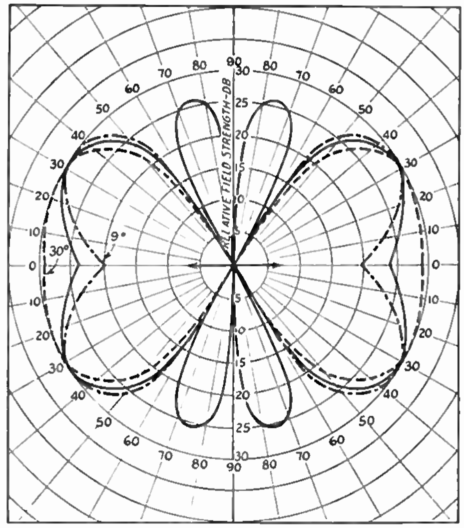


Fig. 14-18 — Horizontal patterns of radiation from an antenna *two wavelengths* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

Directional Characteristics

As the wire is made longer in terms of the number of half-wavelengths, the directional effects change. Instead of the "doughnut" pattern of the half-wave antenna, the directional characteristic splits up into "lobes" which make various angles with the wire. In general, as the length of the wire is increased the direction in which maximum radiation occurs tends to approach the line of the antenna itself.

Directional characteristics for antennas one wavelength, three half-wavelengths, and two wavelengths long are given in Figs. 14-16, 14-17 and 14-18, for three vertical angles of radiation. Note that, as the wire length increases, the radiation along the line of the antenna becomes more pronounced. Still longer antennas can be considered to have practically "end-on" directional characteristics, even at the lower radiation angles.

Methods of Feeding

In a long-wire antenna, the currents in adjacent half-wave sections must be out of phase, as shown in Fig. 14-14. The feeder system must not upset this phase relationship. This requirement is met by feeding the antenna at either end or at any current *loop*. A two-wire feeder cannot be inserted at a current *node*, however, because this invariably brings the currents in two adjacent half-wave sections in phase; if the phase in one section could be reversed, then the currents in the feeders necessarily would have to be in phase and the feeder radiation would not be canceled out.

No point on a long-wire antenna offers a reasonable impedance for a direct match to any of the common types of transmission lines. The most common practice is to feed the antenna at one end or at a current loop with a low-loss open-wire line and accept the resulting standing-wave ratio of 4 or 5. When a better match is required, "stubs" are generally used (described in the preceding chapter).

Multiband Antennas

As suggested in the preceding section, the same antenna may be used for several bands by operating it on harmonics. When this is done it is necessary to use resonant feeders, since the impedance matching for nonresonant feeder operation can be accomplished only at one frequency unless means are provided for changing the length of a matching section and shifting the point at which the feeder is attached to it.

Furthermore, the current loops shift to a new position on the antenna when it is operated on harmonics, further complicating the feed situation. It is for this reason that a half-wave antenna that is center-fed by a solid-dielectric line is practically useless for harmonic operation; on all even harmonics there is a voltage maximum occurring right at the feed point, and the resultant impedance mismatch is so bad that there is a large standing-wave ratio and consequently high losses arise in the solid dielectric. It is wise not to attempt to use on its harmonics a half-wave antenna center-fed with coaxial cable. High-impedance solid-dielectric lines such as 300-ohm Twin-Lead may be used, however, provided the power does not exceed a few hundred watts.

When the same antenna is used for work in several bands, it must be realized that the directional characteristic will vary with the band in use.

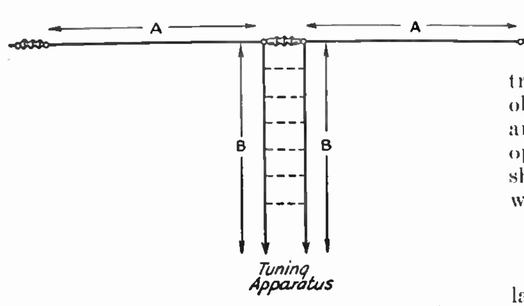


Fig. 14-19 — **Practical arrangement of a shortened antenna. The total length, $A + B + B + A$, should be a half-wavelength for the lowest-frequency band, usually 3.5 Mc. See Table 14-1 for lengths and tuning data.**

Simple Systems

The most practical simple multiband antenna is one that is a half-wavelength long at the lowest frequency and is fed either at the center or one end with an open-wire line. Although the standing-wave ratio on the feedline will not approach 1.0 on any band, if the losses in the line are low the system will be efficient. From the standpoint of reduced feedline radiation, a center-fed system is superior to one that is end-fed, but the end-fed arrangement is often more convenient and should not be ignored as a possibility. The center-fed antenna will not have the same radiation pattern as an end-fed one of the same length, except on frequencies where the over-all length of the antenna is a half-wavelength or less. The end-fed antenna acts like a long-wire antenna on all bands (for which it is longer than a half-wavelength), but the center-fed one acts like two antennas of half that length fed in phase. For example, if a full-wavelength antenna is fed at one end, it will have a radiation pattern as shown in Fig. 14-16, but if it is fed in the center the pattern will be somewhat similar to Fig. 14-7, with the maximum radiation broadside to the wire. Either antenna is a good radiator, but if the radiation pattern is a factor, the point of feed must be considered.

Since multiband operation of an antenna does not permit matching of the feedline, some attention must be paid to the length of the feedline if convenient transmitter-coupling arrangements are to be obtained. Table 14-I gives some suggested antenna and feeder lengths for multiband operation. In general, the length of the feedline should be some integral multiple of a quarter wavelength at the lowest frequency.

Antennas for Restricted Space

If the space available for the antenna is not large enough to accommodate the length necessary for a half-wave at the lowest frequency to be used, quite satisfactory operation can be secured by using a shorter antenna and making

TABLE 14-I

Multiband Resonant-Line Fed Antennas

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
With end feed:			
120	60	14-Mc. 'phone	series
136	67	3.5-Mc. c.w. 7 through 28-Mc.	series parallel
67	33	7 and 21 Mc. 14 and 28 Mc.	series parallel
With center feed:			
137	67	3.5 through 28 Mc.	parallel
68	66	7 and 21 Mc. 14 and 28 Mc.	series parallel
68	34	7 through 28 Mc.	parallel

The antenna lengths given represent compromises for harmonic operation because of different end effects on different bands. The 136-foot end-fed antenna is slightly long for 3.5 Mc. but will work well in the region (3500–3600 kc.) that quadruples into the 14-Mc. band. Bands not listed are not recommended for the particular antenna. The center-fed systems are less critical as to length. Tuning connections are for open-wire line and may differ for 300-ohm Twin-Lead.

The end-fed and center-fed antennas will have the same directional characteristics only on the lowest frequency, as explained in the text.

up the missing length in the feeder system. The antenna itself may be as short as a quarter wavelength and still radiate fairly well, although of course it will not be as effective as one a half-wave long. Nevertheless, such a system is useful where operation on the desired band otherwise would be impossible.

Resonant feeders are a practical necessity with such an antenna system, and a center-fed antenna will give best all-around performance. With end feed the feeder currents become badly unbalanced.

With center feed practically any convenient length of antenna can be used, if the feeder length is adjusted to accommodate at least one half-wave around the whole system.

A practical antenna of this type can be made as shown in Fig. 14-19. Table 14-II gives a few recommended lengths. However, the antenna can be made any convenient length, provided the total length of wire is a half-wavelength at the lowest frequency, or an integral multiple of a half-wavelength.

In using the tables, it should be held in mind that the "type of tuning" will vary from that listed if the feed-line lengths are not as shown or if solid-dielectric line (Twin-Lead) is used. This should not be interpreted as a fault in the antenna, and any tuning system (series or parallel) that works well without any trace of heating is quite satisfactory.

Bent Antennas

Since the field strength at a distance is proportional to the current in the antenna, the high-current part of a half-wave antenna (the center quarter wave, approximately) does most of the radiating. Advantage can be taken of this fact when the space available does not permit building an antenna a half-wave long. In this case the ends may be bent, either horizontally or vertically, so that the

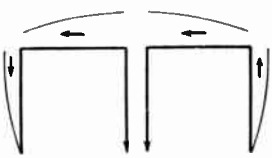


Fig. 14-20 — Folded arrangement for shortened antennas. The total length is a half-wave, not including the feeders. The horizontal part is made as long as convenient and the ends dropped down to make up the required length. The ends may be bent back on themselves like feeders to cancel radiation partially. The horizontal section should be at least a quarter wave long.

total length equals a half-wave, even though the straightaway horizontal length may be as short as a quarter wave. The operation is illustrated in Fig. 14-20. Such an antenna will be a somewhat better radiator than a quarter-wavelength antenna on the lowest frequency, but is not so desirable for multiband operation because the ends play an increasingly important part as the frequency is raised. The performance of the system in such a case is difficult to predict, especially if the ends are vertical (the most convenient arrangement) because of the complex combination of horizontal and vertical polarization which results as well as the dissimilar directional characteristics. However, the fact that the radiation pattern is incapable of prediction does not detract from the general usefulness of the antenna. For one-band operation, end-loading with coils (5 feet or so in from each end) is practical and efficient.

TABLE 14-II

Antennas and Feeder Lengths for Short Multiband Antennas, Center-Fed

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
100	83	3.5 Mc. 7, 14, 21 Mc. 28 Mc.	parallel series series or parallel
68	34	3.5 Mc. 7, 14, 21 and 28 Mc.	series parallel
50	43	7, 14, 21 and 28 Mc.	parallel
33	51	7, 14, 21 and 28 Mc.	parallel
33	31	7 and 21 Mc. 14 and 28 Mc.	series parallel

Vertical Antennas

A vertical quarter-wavelength antenna is often used in the low-frequency amateur bands to obtain low-angle radiation. Four typical examples and suggested methods for feeding are shown in Fig. 14-21. The antenna may be wire or tubing supported by wood or insulated guy wires. When

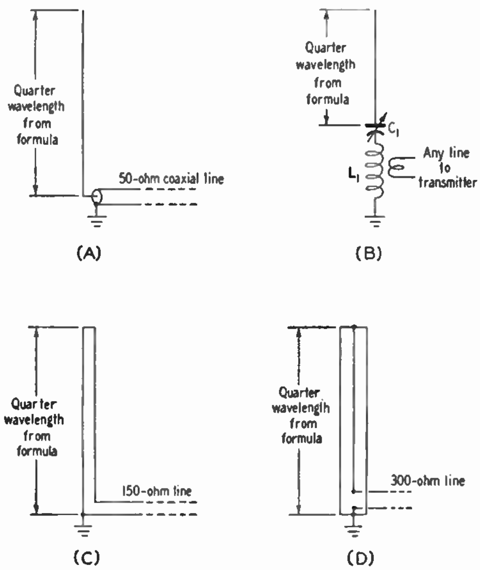


Fig. 14-21 — **A quarter-wavelength antenna can be fed directly with 50-ohm coaxial line (A) with a low standing-wave ratio, or a coupling network can be used (B) that will permit a line of any impedance to be used. In (B), L_1 and C_1 should resonate to the operating frequency, and L_1 should be larger than is normally used in a plate tank circuit at the same frequency.**

By using multiwire antennas, the quarter-wave vertical can be fed with (C) 150- or (D) 300-ohm line.

tubing is used for the antenna, or when guy wires (broken up by insulators) are used to reinforce the structure, the length given by the formula is likely to be long by a few per cent. A check of the standing-wave ratio on the line will indicate the frequency at which the s.w.r. is minimum, and the length of the antenna can be adjusted accordingly.

The examples shown in Fig. 14-21 all require an antenna insulated from the ground, to provide for the feed point. A *grounded* tower or pipe can be used as a radiator by employing "shunt feed," which consists of tapping the inner conductor of the coaxial-line feed up on the tower until the best match is obtained, in much the same manner as the "gamma match" (described later) is used on a horizontal element. If the antenna is not an electrical quarter-wavelength long, it is necessary to tune out the reactance by adding capacity or inductance between the coaxial line and the shunting conductor. A metal tower supporting a TV antenna or rotary beam can be shunt-fed only if all of the wires and leads from the supported antenna run down the center of the tower and underground away from the tower, since otherwise they would become part of the low-frequency antenna system.

● THE GROUND-PLANE ANTENNA

A ground-plane antenna is a vertical quarter-wavelength antenna using an artificial metallic ground, usually consisting of four rods or wires perpendicular to the antenna and extending radially from its base Unlike the quarter-wavelength vertical antennas without an artificial ground, the ground-plane antenna will give low-angle radiation regardless of the height above actual ground. It is a useful antenna for DX work in any of the amateur bands below 30 Mc.

The vertical portion of the ground-plane antenna can be made of self-supported aluminum tubing, or a top-supported wire, depending upon the necessary length and the available supports. The radials are also made of tubing or heavy wire, depending upon the available supports and necessary lengths.

The radiation resistance of a ground-plane antenna varies with the diameter of the vertical element, as shown in Fig. 14-22. Since the radiation resistance is usually in the vicinity of 30 to 32 ohms, the antenna can be fed with 75-ohm coaxial line if a quarter-wavelength matching section of 50-ohm coaxial line is used between the line and the antenna. (See Chapter Thirteen, "Quarter-Wave Transformers.")

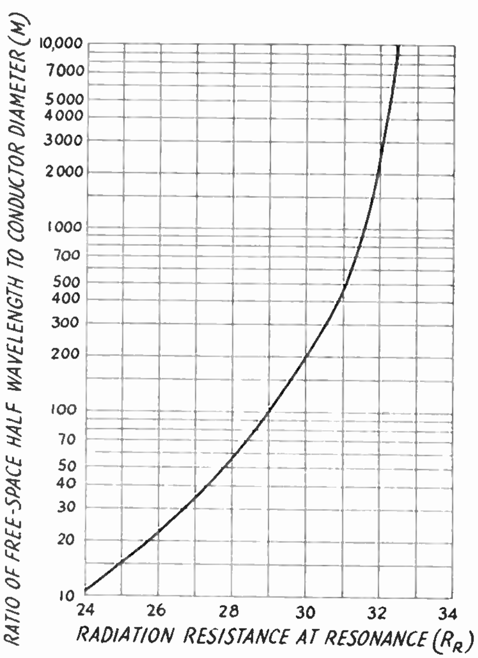


Fig. 14-22 — **Radiation resistance of a quarter-wave antenna (with ground plane or grounded) as a function of *M*. The values apply only when the antenna is of the resonant length.**

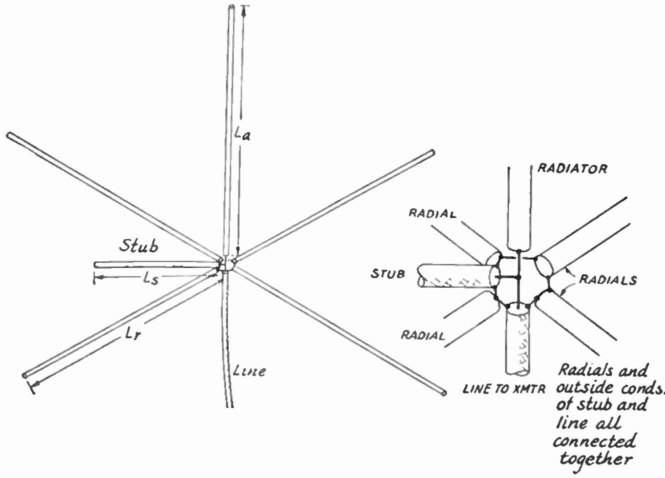


Fig. 14-23 — The ground plane antenna with shunt matching. The antenna length, L_a, matching stub length, L_s, and radial length, L_r, are determined as described in the text, for matching a transmission line of given characteristic impedance. As shown in the insert, the radials and the outside conductors of the stub and line are all connected together.

It is also possible to feed the ground-plane antenna with coaxial line and a "shunt" matching section, as shown in Fig. 14-23. The various values required for proper matching will depend on the particular type of line used, as well as on the radiation resistance, resonant length, and reactance per unit length of the antenna. These antenna characteristics are dependent on the length/diameter ratio — that is, the ratio of a half wavelength in free space to the diameter of the antenna element — and allowance must be made for this factor. The necessary information for design purposes is given in Figs. 14-22, 14-24 and 14-25.

Determining the antenna dimensions can be reduced to a series of steps, as follows:

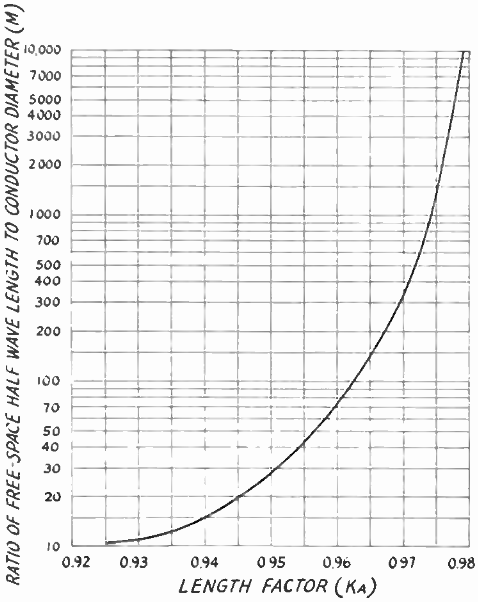


Fig. 14-24 — The antenna-length factor as a function of the ratio of a free-space half wavelength to the conductor diameter. The length factor multiplied by a free-space quarter wavelength is the length of a quarter-wave radiator resonant at the selected frequency.

First determine M, the ratio of a free-space half wavelength to the conductor diameter. The following formula may be used:

$$M = \frac{5906}{FD},$$

where F = frequency in megacycles,
D = conductor diameter in inches.

Using the value of M so found, read the length factor (K_a) from Fig. 14-24, the reactance change per 1 per cent change in length (K_x) from Fig. 14-25, and the radiation resistance (R_r) from Fig. 14-22.

Since the antenna is to be shortened, these values must be modified appropriately. The actual radiation resistance, after the antenna is properly shortened, will be

$$R_o = R_r - \frac{Z_1}{4R_r} \text{ ohms},$$

where R_o = radiation resistance after shortening,
Z_1 = characteristic impedance of transmission line to be matched.

The proper value of capacitive reactance in the shortened antenna is given by

$$X_a = SR_o \text{ ohms},$$

where X_a = capacitive reactance of antenna, and

$$S = \sqrt{\frac{Z_1}{R_o} - 1}.$$

The antenna length that gives the proper capacitive reactance is

$$L_a = \frac{2953K_aK_b}{F} \text{ inches},$$

where L_a = required antenna length, and

$$K_b = 1 - \frac{X_a}{100K_x}.$$

The only remaining steps are to find the dimensions of the inductive stub and the length of the radial ground-plane rods.

The required stub reactance is given by

$$X_s = \frac{Z_1}{S} \text{ ohms},$$

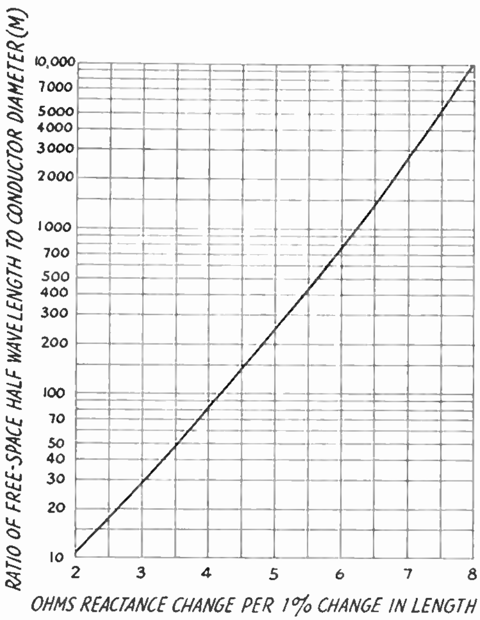


Fig. 14-25 — Reactance change with antenna length as a function of M, for quarter-wave ground-plane (or grounded) antennas. If the antenna is longer than the resonant length the reactance is inductive; if shorter, the reactance is capacitive. The curve is accurate for lengths within 10 per cent of the resonant length. Multiply reactance values by 2 for half-wave antennas.

where X_s = inductive reactance of stub.

The length of the shorted stub is

$$L_s = \frac{32.8VL}{F} \text{ inches,}$$

where L_s = stub length,
V = velocity factor of line used in stub,
L = length of stub in electrical degrees having required X_s.

L is equal to the angle whose tangent is X_s/Z_s, where Z_s is the characteristic impedance of the stub.

The length of each radial is given by

$$L_r = \frac{2953K_a}{F} \text{ inches,}$$

the length being measured from the center line of the radiator to the tip of the radial.

If the radials have a different diameter than the radiator (a common practice) the M and K_a for radials and antenna must be considered separately.

Example: Assume a ground-plane antenna to be constructed with a vertical radiator of 2-inch diameter tubing and radials of No. 10 (0.10-inch diam.) wire, for a frequency of 7.1 Mc. and to be matched to 72-ohm RG-8/U coaxial line by using a stub of the same material.

$$F = 7.1 \text{ Mc.}, D = 2 \text{ inches}, Z_l = Z_s = 72 \text{ ohms},$$
$$V = 0.66, M = 5906 \div (7.1 \times 2) = 416.$$

From Figs. 14-24, 14-25 and 14-22, it is found that

$$K_a = 0.971, K_x = 5.5, R_r = 30.9.$$

From the formula,

$$R_o = 30.9 - \frac{72}{4 \times 30.9} = 30.3 \text{ ohms}$$

and the factor

$$S = \sqrt{\frac{72}{30.3} - 1} = 1.09.$$

Hence $X_a = 1.09 \times 30.3 = 33$ ohms.

Also, $K_b = 1 - \frac{33}{100 \times 5.5} = 0.94.$

Thus the antenna length,

$$L_a = \frac{2953 \times 0.971 \times .94 \text{ 8 inches.}}{7.1} = 380 \text{ inches} = 31 \text{ feet.}$$

To find the stub dimensions,

$$X_s = \frac{72}{1.09} = 66.$$

L is the angle whose tangent is $66 \div 72 = 0.918$, and from a table of tangents is found to be 42.6 degrees.

Then $L_s = \frac{32.8 \times 0.66 \times 42.6}{7.1} = 130$ inches = 10 feet 10 inches

For the radials,

$$M = 5906 \div (7.1 \times 0.1) = 8340, K_a = 0.978.$$

Hence $L_r = \frac{2953 \times 0.978}{7.1} = 407$ inches = 33 feet 11 inches.

Antennas for 160 Meters

Results on 1.8 Mc. will depend to a large extent on the antenna system and the time of day or night. Almost any random long wire that can be tuned to resonance will work during the night but it will generally be found very ineffective during the day. A vertical antenna — or rather an antenna from which the radiation is predominantly vertically polarized — is probably the best for 1.8-Mc. operation. A horizontal antenna (horizontally-polarized radiation) will give better results during the night than the day because daytime absorption in the ionosphere is so high at this frequency that the reflected wave is too weak to be useful. At night the performance improves because nighttime ionosphere conditions generally permit the reflected wave to return to earth without too much attenuation. The vertically-polarized radiator gives a strong ground wave that is effective day or night, and it is to be preferred on 1.8 Mc.

There is another reason why a vertical antenna is better than a horizontal for 160-meter operation. The low-angle radiation from a horizontal antenna ⅛ or ¼ wavelength above ground is almost insignificant. Any reasonable height is small in terms of wavelength, so that a horizontal antenna on 160 meters is a poor radiator at angles useful for long distances ("long," that is, for this band). Its chief usefulness is over relatively short distances at night.

Bent Antennas

Since ideal vertical antennas are generally out of the question for practical amateur work, the best compromise is to bend the antenna in such a way that the high-current

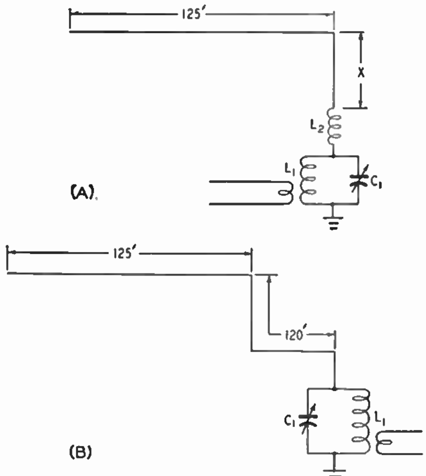


Fig. 14-26 — **Bent antenna for the 160-meter band. In the system at A, the vertical portion (length X) should be made as long as possible. In either antenna system, L_1C_1 should resonate at 1900 kc., roughly. To adjust L_2 in antenna A, resonate L_1C_1 alone to the operating frequency, then connect it to the antenna system and adjust L_2 for maximum loading. Further loading can be obtained by increasing the coupling between L_1 and the link.**

portions of the antenna run vertically. It is, of course, advisable to place the antenna so that the highest currents in the antenna occur at the highest points above actual ground. Two antenna systems designed along these lines are shown in Fig. 14-26. The antenna at A uses a loading coil, L_2, to increase the electrical length of the antenna to a half wavelength, so that the antenna can be fed at its high-voltage point through the coupling circuit L_1C_1. The antenna of Fig. 14-26B uses a full half-wavelength of wire but is bent so that the high-current portion runs vertically. The horizontal portion running to L_1C_1 should run 8 or 10 feet above ground.

Grounds

A good ground connection is generally important on 160 meters. The ideal system is a number of wire radials buried a foot or two underground and extending 50 to 100 feet from the central connection point. As many radials as possible should be used.

If the soil is good (not rocky or sandy) and generally moist, a low-resistance connection to the cold-water pipe system in the house will often serve as an adequate ground system. The connection should be made close to where the pipe enters the ground, and the surface of the pipe should be scraped clean before tightening the ground clamp around the pipe.

A 6- or 8-foot length of 1-inch water pipe, driven into the soil at a point where there is considerable natural moisture, can be used for the ground connection. Three or four pipes driven into the ground 8 or 10 feet apart and all joined together at the top with heavy wire are more effective than the single pipe.

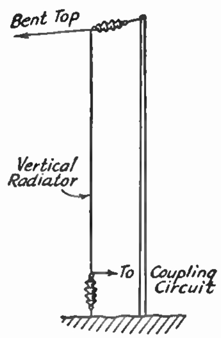


Fig. 14-27 — An arrangement for keeping the main radiating portion of the antenna vertical.

The use of a counterpoise is recommended where a buried system is not practicable or where a pipe ground cannot be made to have low resistance because of poor soil conditions. A counterpoise consists of a number of wires supported from 6 to 10 feet above the surface of the ground. Generally the wires are spaced 10 to 15 feet apart and located to form a square or polygonal configuration under the vertical portion of the antenna.

Long-Wire Directive Arrays

● THE "V" ANTENNA

It has been emphasized that, as the antenna length is increased, the lobe of maximum radiation makes a more acute angle with the

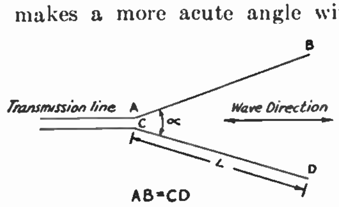


Fig. 14-28 — **The basic "V" antenna, made by combining two long wires.**

wire. Two such wires may be combined in the form of a horizontal "V" so that the main lobes from each wire will reinforce along a line bisecting the angle between the wires. This increases both gain and directivity, since the lobes in directions other than along the bisector cancel to a greater or lesser extent. The horizontal "V" antenna therefore transmits best in either direction (is bidirectional) along a line bisecting the "V" made by the two wires. The power gain depends upon the length of the wires. Provided the necessary space is available, the "V" is a simple antenna to build and operate. It can also be used on harmonics, so that it is suitable for multiband work. A top view of the "V" antenna is shown in Fig. 14-28.

Fig. 14-29 shows the dimensions that should be followed for an optimum design to obtain maximum power gain for different-sized "V" antennas. The longer systems give good performance in multiband operation. Angle ∝ is approximately equal to twice the

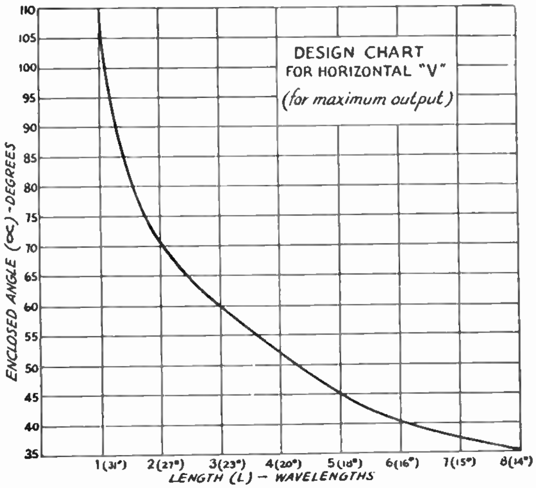


Fig. 14-29 — Design chart for horizontal "V" antennas, giving the enclosed angle between sides *vs.* the length of the wires. Values in parentheses represent approximate wave angle for height of one-half wavelength.

angle of maximum radiation for a single wire equal in length to one side of the "V."

The wave angle referred to in Fig. 14-29 is the vertical angle of maximum radiation. Tilting the whole horizontal plane of the "V" will tend to increase the low-angle radiation off the low end and decrease it off the high end.

The gain increases with the length of the wires, but is not exactly twice the gain for a single long wire as given in Fig. 14-15. In the longer lengths the gain will be somewhat increased, because of mutual coupling between the wires. A "V" eight wavelengths on a leg, for instance, will have a gain of about 12 db. over a half-wave antenna, whereas twice the gain of a single eight-wavelength wire would be only approximately 9 db.

The two wires of the "V" must be fed out of phase, for correct operation. A resonant line may simply be attached to the ends, as shown in Fig. 14-28. Alternatively, a quarter-wave matching section may be employed and the antenna fed through a nonresonant line. If the antenna wires are made multiples of a half-wave in length (use Equation 14-G for computing the length), the matching section will be closed at the free end. A stub can be connected across the resonant line to provide a match, as described in the preceding chapter.

THE RHOMBIC ANTENNA

The horizontal rhombic or "diamond" antenna is shown in Fig. 14-30. Like the "V," it requires a great deal of space for erection, but it is capable of giving excellent gain and directivity. It also can be used for multiband operation. In the terminated form shown in Fig. 14-30, it operates like a nonresonant transmission line, without standing waves, and is unidirectional. It may also be used without the terminating resistor, in which case there are standing waves on the wires and the antenna is bidirectional.

The important quantities influencing the design of the rhombic antenna are shown in Fig. 14-30. While several design methods may be used, the one most applicable to the conditions existing in amateur work is the so-called "compromise" method. The chart of Fig. 14-31 gives design information based on a given length and wave angle to determine the remaining optimum dimensions for best operation. Curves for value of length of two, three and four wavelengths are shown, and any intermediate values may be interpolated.

With all other dimensions correct, an increase in length causes an increase in power gain and a slight reduction in wave angle. An increase in height also causes a reduction in wave angle and an increase in power gain, but not to the same extent as a proportionate increase in length. For multiband work, it is satisfactory to design the rhombic antenna on the basis of 14-Mc. operation, which will permit work from the 7- to 28-Mc. bands as well.

A value of 800 ohms is correct for the terminating resistor for any properly-constructed rhombic, and the system behaves as a pure resistive load under this condition. The terminating resistor must be capable of safely dissipating one-half the power output (to eliminate the rear pattern), and should be noninductive. Such a resistor may be made up from a carbon or graphite rod or from a long 800-ohm transmission line using resistance wire. If the carbon rod or a similar form of lumped resistance is used, the device should be suitably protected from weather effects; i.e., it should be covered with a good asphaltic compound and sealed in a small lightweight box or fiber tube. Suitable nonreactive terminating resistors are also available commercially.

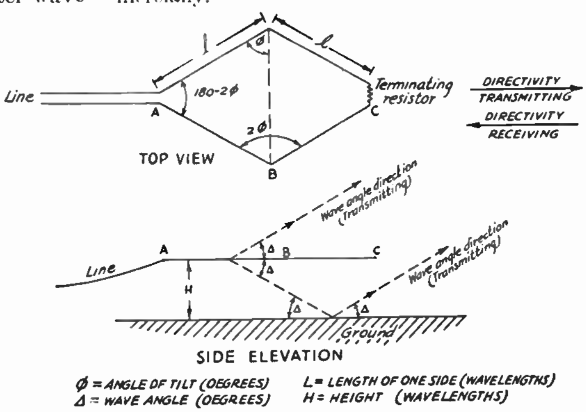


Fig. 14-30 — The horizontal rhombic or diamond antenna, terminated. Important design dimensions are indicated; details in text.

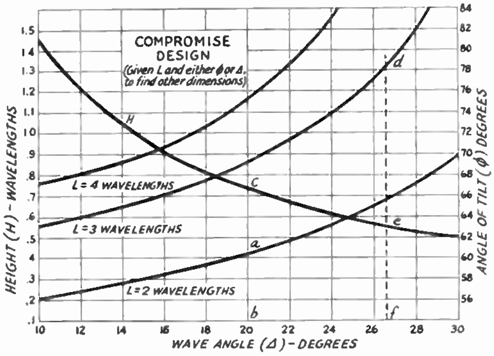


Fig. 14-31 — **Compromise-method design chart for rhombic antennas of various leg lengths and wave angles. The examples at the right illustrate the use of the chart:**

»

(1) Given:

Length (L) = 2 wavelengths
Desired wave angle (Δ) = 20°.

To Find: H, Φ.

Method:

Draw vertical line through point a (L = 2 wavelengths) and point b on abscissa (Δ = 20°). Read angle of tilt (Φ) for point a and height (H) from intersection of line ab at point c on curve H.

Result:

Φ = 60.5°.
H = 0.73 wavelength.

(2) Given:

Length (L) = 3 wavelengths.
Angle of tilt (Φ) = 78°.

To Find: H, Δ.

Method:

Draw a vertical line from point d on curve L = 3 wavelengths at Φ = 78°. Read intersection of this line on curve H (point e) for height, and intersection at point f on the abscissa for Δ.

Result:

H = 0.56 wavelength.
Δ = 26.6°.

For feeding the antenna, the antenna impedance will be matched by an 800-ohm line, which may be constructed from No. 16 wire spaced 20 inches or from No. 18 wire spaced 16 inches. The 800-ohm line is somewhat ungainly to install, however, and may be replaced by an ordinary 600-ohm line with only a negligible mismatch. Alternatively, a matching section may be installed between the antenna terminals and a low-impedance line. However, when such an arrangement is used, it will be necessary to change the matching-section constants for each different band on which operation is contemplated.

The same design details apply to the unterminated rhombic as to the terminated type. When used without a terminating resistor, the system is bidirectional. Resonant feeders are generally used with the unterminated rhombic. A nonresonant line may be used by incorporating a matching section at the antenna, but is not readily adaptable to satisfactory multiband work.

Rhombic antennas will give a power gain of 8 to 12 db. or more for leg lengths of two to four wavelengths, when constructed according to the charts given. In general, the larger the antenna, the greater the power gain.

Directive Arrays with Driven Elements

By combining individual half-wave antennas into an **array** with suitable spacing between the antennas (called **elements**) and feeding power to them simultaneously, it is possible to make the radiated fields from the individual elements add in a favored direction, thus increasing the field strength in that direction as compared to that produced by one antenna element alone. In other directions the fields will more or less oppose each other, giving a reduction in field strength. Thus a power gain in the desired direction is secured at the expense of a power reduction in other directions.

Besides the spacing between elements, the instantaneous direction of current flow (*phase*) in individual elements determines the directivity and power gain. There are several methods of arranging the elements. If they are strung end to end, so that all lie on the same straight line, the elements are said to be **collinear.** If they are parallel and all lying in the same plane, the elements are said to be **broadside** when the phase of the current is the same in all, and **end-fire** when the currents are not in phase. Elements that receive power from the transmitter through the transmission line are called **driven elements.**

The power gain of a directive system increases with the number of elements. The proportionality between gain and number of elements is not simple, however. The gain depends upon the effect that the spacing and phasing has upon the radiation resistance of the elements, as well as upon their number.

Collinear Arrays

Simple forms of collinear arrays, with the current distribution, are shown in Fig. 14-32. The two-element array at A is popularly known as "two half-waves in phase." It will be recognized as simply a center-fed antenna operated at its second harmonic. The way in which the number of elements may be extended for increased directivity and gain is shown in Fig. 14-28B. Note that quarter-wave phasing sections are used between elements; these give the reversal in phase necessary to make the currents in individual antenna elements all flow in the same direction at the same instant. Any phase-reversing section may be used as a quarter-wave matching section for attaching a nonresonant feeder, or a resonant transmission line may be substituted for any of the quarter-wave sections. Also, the antenna may be end-fed

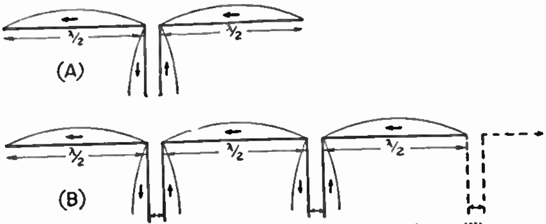


Fig. 14-32 — Collinear half-wave antennas in phase. The system at A is generally known as "two half-waves in phase." B is an extension of the system; in theory the number of elements may be carried on indefinitely, but practical considerations usually limit the elements to four.

by any of the systems previously described, or any element may be center-fed. It is best to feed at the center of the array, so that the energy will be distributed uniformly among the elements.

The gain and directivity depend upon the number of elements and their spacing, center-to-center. This is shown by Table 14-III. Although three-quarter wave spacing gives greater gain, it is difficult to construct a suitable phase-reversing system when the ends of the antenna elements are widely separated. For this reason, the half-wave spacing is most generally used in actual practice.

Collinear arrays may be mounted either horizontally or vertically. Horizontal mounting gives increased horizontal directivity, while the vertical directivity remains the same as for

TABLE 14-III

Theoretical Gain of Collinear Half-Wave Antennas

Spacing between centers of adjacent half-waves	*Number of half-waves in array vs. gain in db.*				
	2	3	4	5	6
½ wave	1.8	3.3	4.5	5.3	6.2
¾ wave	3.2	4.8	6.0	7.0	7.8

a single element at the same height. Vertical mounting gives the same horizontal pattern as a single element, but concentrates the radiation at low angles. It is seldom practicable to use more than two elements vertically at frequencies below 14 Mc. because of the excessive height required.

Broadside Arrays

Parallel antenna elements with currents in phase may be combined as shown in Fig. 14-33 to form a **broadside** array, so named because the direction of maximum radiation is broadside to the plane containing the antennas. Again the gain and directivity depend upon the number of elements and the spacing, the gain for different spacings being shown in Fig. 14-34. Half-wave spacing generally is used, since it simplifies the problem of feeding the system when the array has more than two elements. Table 14-IV gives theoretical gain as a function of the number of elements with half-wave spacing.

Broadside arrays may be suspended either with the elements all vertical or with them horizontal and one above the other **(stacked)**. In the former case the horizontal pattern becomes quite sharp, while the vertical pattern is the same as that of one element alone. If the array is suspended horizontally, the horizontal pattern is equivalent to that of one element while the vertical pattern is sharpened, giving low-angle radiation.

Broadside arrays may be fed either by resonant transmission lines or through quarter-wave matching sections and nonresonant lines. In Fig. 14-33, note the "crossing over" of the feeders, which is necessary to bring the elements into proper phase relationship.

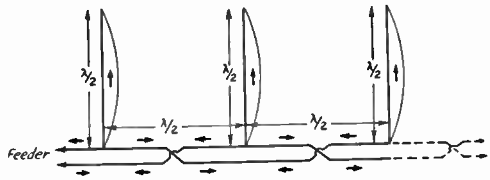


Fig. 14-33 — Broadside array using parallel half-wave elements. Arrows indicate the direction of current flow. Transposition of the feeders is necessary to bring the antenna currents in phase. Any reasonable number of elements may be used. The array is bidirectional, with maximum radiation "broadside" or perpendicular to the antenna plane (perpendicularly through this page).

Combined Broadside and Collinear Arrays

Broadside and collinear arrays may be combined to give both horizontal and vertical directivity, as well as additional gain. The general plan of constructing such antennas is shown in Fig. 14-35. The lower angle of radiation resulting from stacking elements in the vertical plane is desirable at the higher frequencies. In general, doubling the number of elements in an array by stacking will raise the gain from 2 to 4 db., depending upon whether vertical or horizontal elements are used — that is, whether the stacked elements are of the broadside or collinear type.

The arrays in Fig. 14-35 are shown fed from one end, but this is not especially desirable in

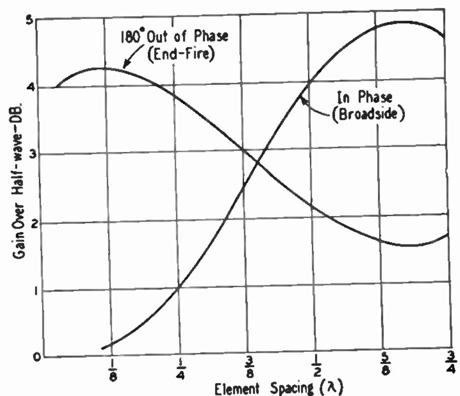


Fig. 14-34 — Gain *vs.* spacing for two parallel half-wave elements combined as either broadside or end-fire arrays.

TABLE 14-IV Theoretical Gain *vs.* Number of Broadside Elements (Half-Wave Spacing)	
No. of elements	*Gain*
2	4 db.
3	5.5
4	7
5	8
6	9

the case of large arrays. Better distribution of energy between elements, and hence better over-all performance, will result when the feeders are attached as nearly as possible to the center of the array. Thus, in the eight-element array at A, the feeders could be introduced at the middle of the transmission line between the second and third set of elements, in which case the connecting line would not be transposed between the second and third set of elements. Alternatively, the antenna could be constructed with the transpositions as shown and the feeder connected between the adjacent ends of either the second or third pair of collinear elements.

A four-element array of the general type shown in Fig. 14-35B, known as the "lazy-H"

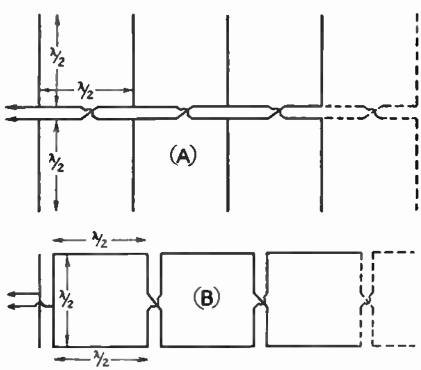


Fig. 14-35 — Combination broadside and collinear arrays. A, with vertical elements; B, with horizontal elements. Both arrays give low-angle radiation. Two or more sections may be used. The gain in db. will be equal, approximately, to the sum of the gain for one set of broadside elements (Table 14-IV) plus the gain of one set of collinear elements (Table 14-III). For example, in A each broadside set has four elements (gain 7 db.) and each collinear set two elements (gain 1.8 db.), giving a total gain of 8.8 db. In B, each broadside set has two elements (gain 4 db.) and each collinear set three elements (gain 3.3 db.), making the total gain 7.3 db. The result is not strictly accurate, because of mutual coupling between the elements, but is good enough for practical purposes.

antenna, has been quite frequently used. This arrangement is shown, with the feed point indicated, in Fig. 14-36.

End-Fire Arrays

Fig. 14-37 shows a pair of parallel half-wave elements with currents out of phase. This is known as an end-fire array because it radiates best along the plane of the antennas, as shown.

The end-fire array may be used either vertically or horizontally (elements at the same height), and is well adapted to amateur work because it gives maximum gain with relatively close element spacing. Fig. 14-34 shows how the gain varies with spacing. End-fire elements may be combined with additional collinear and broadside elements to give a further increase in gain and directivity.

Either resonant or nonresonant lines may be used with this type of array. Nonresonant lines

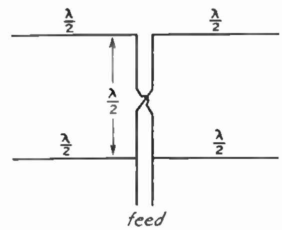


Fig. 14-36 — A four-element combination broadside-collinear array, popularly known as the "lazy-H" antenna. A closed quarter-wave stub may be used at the feed point to match into a 600-ohm transmission line, or resonant feeders may be attached at the point indicated. The gain over a half-wave antenna is 5 to 6 db.

preferably are matched to the antenna through a quarter-wave matching section or phasing stub.

Phasing

Figs. 14-35 and 14-37 illustrate a point in connection with feeding a phased antenna system which sometimes is confusing. In Fig. 14-37, when the transmission line is connected as at A there is no crossover in the line connecting the two antennas, but when the transmission line is connected to the center of the connecting line the crossover becomes necessary (B). This is because in B the two halves of the connecting line are simply *branches* of the same line. In other words, even though the connecting line in B is a half-wave in length, it is not actually a half-wave line but *two quarter-wave lines in parallel.* The same thing is true of the untransposed line of Fig. 14-35B. Note that, under these conditions, the antenna elements are in phase when the line is not transposed, and out of phase when the transposition is made. The opposite is the case when the half-wave line simply joins two an-

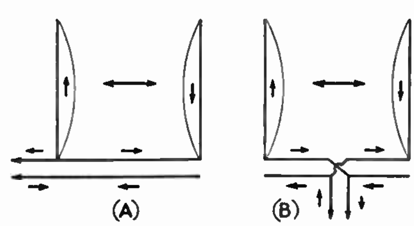


Fig. 14-37 — End-fire arrays using parallel half-wave elements. The elements are shown with half-wave spacing to illustrate feeder connections. In practice, closer spacings are desirable, as shown by Fig. 14-34. Direction of maximum radiation is shown by the large arrows.

tenna elements and does not have the feed line connected to its center, as in Fig. 14-33.

Adjustment of Arrays

With arrays of the types just described, using half-wave spacing between elements, it will usually suffice to make the length of each element that given by Equations 14-B or 14-C. The half-wave phasing lines between the parallel elements should be of open-wire construction, and their length can be calculated from:

$$\text{Length of half-wave line (feet)} = \frac{480}{\text{Freq. (Mc.)}} \qquad \text{(14-H)}$$

Example: A half-wavelength phasing line for 28.8 Mc. would be $\frac{480}{28.8} = 16.66$ feet = 16 feet 8 inches.

The spacing between elements can be made equal to the length of the phasing line. No special adjustments of line or element length or spacing are needed, provided the formulas are followed closely.

With collinear arrays of the type shown in Fig. 14-32B, the same formula may be used for the element length, while the length of the quarter-wave phasing section can be found from the following formula:

$$\text{Length of quarter-wave line (feet)} = \frac{240}{\text{Freq. (Mc.)}} \qquad \text{(14-I)}$$

Example: A quarter-wavelength phasing line for 14.25 Mc. would be $\frac{240}{14.25} = 16.84$ feet = 16 feet 10 inches.

If the array is fed in the center it should not be necessary to make any particular adjustments, although, if desired, the whole system can be resonated by connecting an r.f. ammeter in the shorting link of each phasing section and moving the link back and forth to find the maximum-current position. This refinement is hardly necessary in practice, however, so long as all elements are the same length and the system is symmetrical.

The phasing sections can be made of 300-ohm Twin-Lead, if low power is used. However, the lengths of the phasing sections must then be only 84 per cent of the length obtained in the two formulas above.

Example: The half-wavelength line for 28.8 Mc. would become 0.84 × 16.66 = 13.99 feet = 14 feet 0 inches.

Using Twin-Lead for the phasing sections is most useful in arrays such as that of Fig. 14-32B, or any other system in which the element spacing is not controlled by the length of the phasing section.

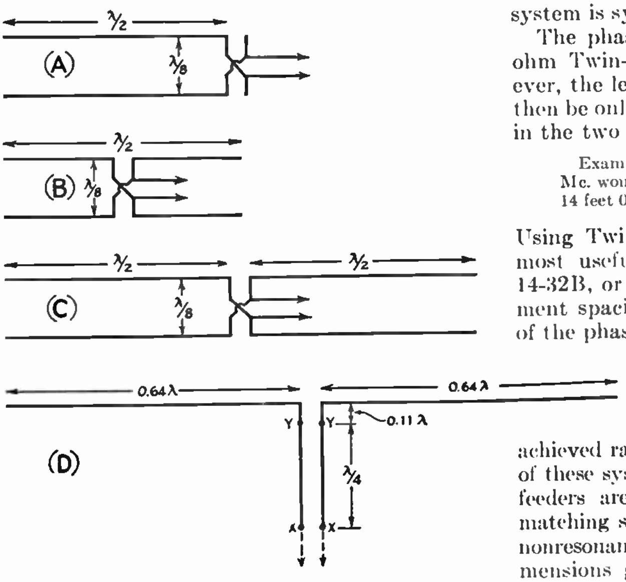


Fig. 14-38 — Simple directive-antenna systems. A is a two-element end-fire array; B is the same array with center feed, which permits use of the array on the second harmonic, where it becomes a four-element array with quarter-wave spacing. C is a four-element end-fire array with ⅛-wave spacing. D is a simple two-element broadside array using extended in-phase antennas ("extended double-Zepp"). The gain of A and B is slightly over 4 db. On the second harmonic, B will give about 5-db. gain. With C, the gain is approximately 6 db., and with D, approximately 3 db. In A, B and C, the phasing line contributes about ⅛ wavelength to the transmission line; when B is used on the second harmonic, this contribution is ⅛ wavelength. Alternatively, the antenna ends may be bent to meet the transmission line, in which case each feeder is simply connected to one antenna. In D, points Y-Y indicate a quarter-wave point (high current) and X-X a half-wave point (high voltage). The line may be extended in multiples of quarter waves if resonant feeders are to be used. A, B and C may be suspended on wooden spreaders. The plane containing the wires should be parallel to the ground.

Simple Arrays

Several simple directive-antenna systems using driven elements have achieved rather wide use among amateurs. Four of these systems are shown in Fig. 14-38. Tuned feeders are assumed in all cases; however, a matching section readily can be substituted if a nonresonant transmission line is preferred. Dimensions given are in terms of wavelength; actual lengths can be calculated from the equations for the antenna and from the equation above for the resonant transmission line or matching section. In cases where the transmission line proper connects to the midpoint of a phasing line, only *half* the length of the latter should be added to the line to find the quarter-wave point.

At A and B are two-element end-fire arrangements using close spacing. They are electrically equivalent; the only difference is in the method of connecting the feeders. B may also be used on the second harmonic, although the spacing is not optimum (Fig. 14-34) for such operation.

A close-spaced four-element array is shown at C. It will give about 2 db. more gain than the two-element array.

The antenna at D, commonly known as the "extended double-Zepp," is designed to take advantage of the greater gain possible with collinear antennas having greater than half-wave center-to-center spacing, but without introducing feed complications. The elements are made longer than a half-wave in order to bring this about. The gain is 3 db. over a single half-wave antenna, and the broadside directivity is fairly sharp.

The antennas of A and B may be mounted either horizontally or vertically; horizontal suspension (with the elements in a plane parallel to the ground) is recommended, since this tends to give low-angle radiation without an unduly sharp horizontal pattern. Thus these systems are useful for coverage over a wide horizontal angle. The system at C, when mounted horizontally, will have a sharper horizontal pattern than the two-element arrays because of the effect of the collinear arrangement. The vertical pattern, however, will be the same as that of the antennas in A and B.

Directive Arrays with Parasitic Elements

Parasitic Excitation

The antenna arrays previously described are bidirectional; that is, they will radiate in directions both to the "front" and to the "back" of the antenna system. If radiation is wanted in only one direction, it is necessary to use different element arrangements. In most of these arrangements the additional elements receive power by induction or radiation from the driven element, generally called the "antenna," and reradiate it in the proper phase relationship to achieve the desired effect. These elements are called *parasitic* elements, as contrasted to the driven elements which receive power directly from the transmitter through the transmission line.

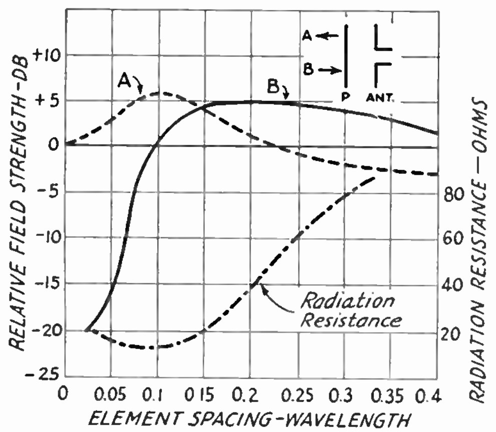


Fig. 14-39 — Gain *vs.* element spacing for an antenna and one parasitic element. The reference point, 0 db., is the field strength from a half-wave antenna alone. The greatest gain is in direction *A* at spacings of less than 0.14 wavelength, and in direction *B* at greater spacings. The front-to-back ratio is the difference in db. between curves *A* and *B*. Variation in radiation resistance of the driven element also is shown. These curves are for a self-resonant parasitic element. At most spacings the gain as a reflector can be increased by slight lengthening of the parasitic element; the gain as a director can be increased by shortening. This also improves the front-to-back ratio.

The parasitic element is called a **director** when it reinforces radiation on a line pointing to it from the antenna, and a **reflector** when the reverse is the case. Whether the parasitic element is a director or reflector depends upon the parasitic-element tuning, which usually is adjusted by changing its length.

Gain vs. Spacing

The gain of an antenna with parasitic elements varies with the spacing and tuning of the elements, and thus for any given spacing there is a tuning condition that will give maximum gain at this spacing. The maximum front-to-back ratio seldom, if ever, occurs at the same condition that gives maximum forward gain. The impedance of the driven element also varies with the tuning and spacing, and thus the antenna system must be tuned to its final condition before the match between the line and the antenna can be completed. However, the tuning and matching may interlock to some extent, and it is usually necessary to run through the adjustments several times to insure that the best possible tuning has been obtained.

Two-Element Beams

A 2-element beam is useful where space or other considerations prevent the use of the larger structure required for a 3-element beam. The general practice is to tune the parasitic element as a reflector and space it about 0.15 wavelength from the driven element, although some successful antennas have been built with 0.1-wavelength spacing and director tuning. Gain *vs.* element spacing for a 2-element antenna is given in Fig. 14-39, for the special case where the parasitic element is resonant, but it is indicative of the performance to be expected under maximum-gain tuning conditions.

Three-Element Beams

Where room is available for an over-all length greater than 0.2 wavelength, a 3-element beam is preferable to one with only 2 elements. Once the over-all length has been decided upon, the curves of Fig. 14-40 can be used to determine the proper spacing of director and reflector. If, for example, the distance between director and reflector can be made 0.4 wavelength, Fig. 14-40 shows that a spacing of 0.2D-0.2R gives a gain of 7.9 db., and a spacing of 0.25D-0.15R gives a gain of 8.2 db. Obviously the latter is the better choice, although the practical difference might be difficult to measure, and practical (mechanical) considera-

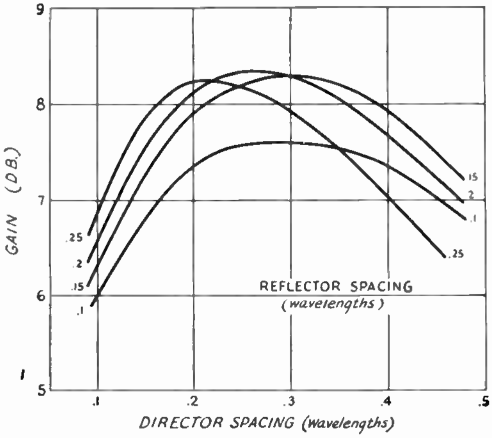


Fig. 14-40 — Gain *vs.* element spacing for 3-element beams using a driven element and a director and a reflector. The 0-db. reference level is the field strength from a half-wavelength antenna alone. These curves are for the system tuned for maximum forward gain. The element spacing shown is the fraction of a wavelength determined by $\frac{984}{f\ (\text{Mc.})}$. Thus a wavelength at 14.2 Mc. = 984/14.2 = 69.3 feet. A spacing of 0.15 wavelength at 14.2 Mc. would be 0.15 × 69.3 = 10.4 feet = 10 feet 5 inches.

tions might call for using the more balanced 0.2D-0.2R construction.

When the over-all length has been decided upon, and the element spacing has been determined, the element lengths can be found by referring to Fig. 14-41. It must be remembered that the lengths determined by these charts will vary slightly in actual practice with the element diameter and the method of supporting the elements, and the tuning of a beam should always be checked after installation. However, the lengths obtained by the use of the charts will be close to correct in practically all cases, and they can be used without checking if the beam is difficult of access.

The preferable method for checking the beam is by means of a field-strength meter or the S-meter of a communications receiver, used in conjunction with a half-wave dipole antenna located at least 10 wavelengths away and as high as or higher than the beam that is being checked. A few watts of power fed into the antenna will give a useful signal at the observation point, and the power input to the transmitter (and hence the antenna) should be held constant for all of the readings. Beams tuned on the ground and then lifted into place are subject to tuning errors and cannot be depended upon. The impedance of the driven element will vary with the height above ground, and good practice dictates that all final matching between antenna and line be done with the antenna in place at its normal height above ground.

Simple Systems: the Rotary Beam

Two- and 3-element systems are popular for rotary-beam antennas, where the entire antenna system is rotated, to permit its gain and directivity to be utilized for any compass direction. They may be mounted either horizontally (with the plane containing the elements parallel to the earth) or vertically.

A 4-element beam will give still more gain than a 3-element one, provided the support is

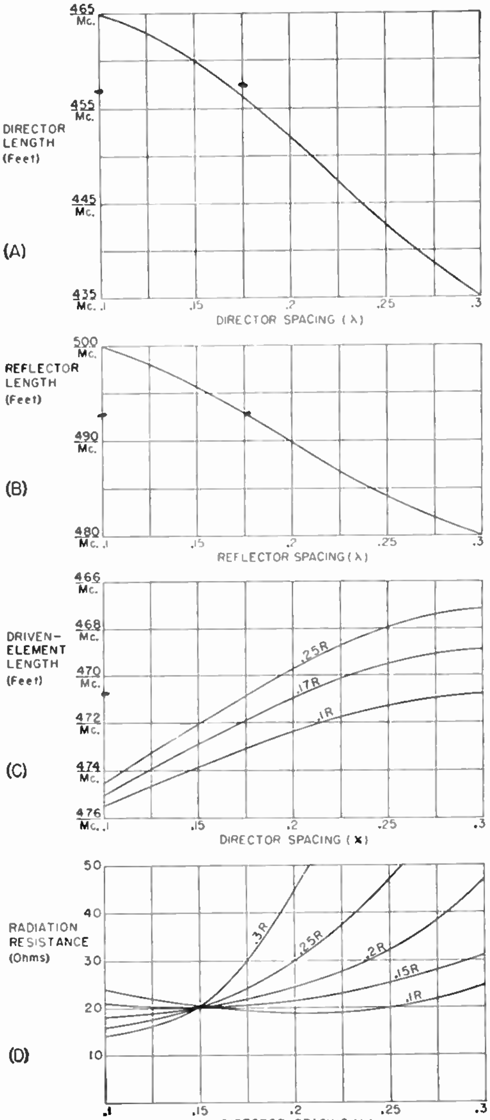


Fig. 14-41 — Element lengths for a 3-element beam. These lengths will hold closely for tubing elements supported at or near the center. The radiation resistance (D) is useful information in planning for a matching system, but it is subject to variation with height above ground and must be considered only as an approximation.

The driven-element length (C) may require modification for tuning out reactance if a T- or gamma-match feed system is used, as mentioned in the text.

A 0.2D-0.2R beam cut for 28.6 Mc. would have a director length of 452/28.6 = 15.8 = 15 feet 10 inches, a reflector length of 490/28.6 = 17.1 = 17 feet 1 inch, and a driven-element length of 470.5/28.6 = 16.45 = 16 feet 5 inches.

sufficient for at least 0.2-wavelength spacing between elements. The tuning for maximum gain involves many variables, and complete gain and tuning data are not available.

The elements in close-spaced (less than one-quarter wavelength element spacing) arrays preferably should be made of tubing of one-half to one-inch diameter. A conductor of large diameter not only has less ohmic resistance but also has lower *Q;* both these factors are important in close-spaced arrays because the impedance of the driven element usually is quite low compared to that of a single half-wave dipole. With 3- and 4-element arrays the radiation resistance of the driven element may be so low that ohmic losses in the conductor can consume an appreciable fraction of the power. Low radiation resistance means that the antenna will work over only a small frequency range without retuning unless large-diameter conductors are used.

Feeding Close-Spaced Arrays

Any of the usual methods of feed may be applied to the driven element of a parasitic array. The preferred methods are shown in Fig. 14-42. Resonant feeders are not recommended for lengths greater than a half-wavelength unless open-wire lines of copper-tubing conductors are used.

Three versions of the popular "T"-match are shown, for two-wire lines of Twin-Lead at A, for single coaxial line at B, and for double coaxial line at C. The match is adjusted by moving the shorting bars, keeping them equidistant from the center, until the minimum s.w.r. is obtained on the line. If the s.w.r. minimum is not 1.5 or less, the transmitter frequency should be shifted to find the frequency where the minimum s.w.r. occurs. If it is higher than the original test frequency, increase the antenna element length slightly. The parasitic element lengths taken from Fig. 14-41 should not require much adjustment unless considerably different spacing is used, but it may be necessary to change the position of the shorting bars and the length of the antenna element once or twice before the s.w.r. at the test frequency is acceptable. The matching section may be made of the same type of conductor as the element and spaced a few inches from it. The length of the matching section will be greater with higher-impedance lines and with wider element spacing. A good starting point for a 28-Mc. wide-spaced (0.2D-0.15R) beam fed with 300-ohm Twin-Lead is 28 inches each side of center. A similar antenna and line on 14 Mc. might require about 56 inches each side.

The gamma match, shown in Fig. 14-42D, can be considered as one-half a "T"-match, and the same principles hold. However, when the length of the element is changed, in an effort to minimize the s.w.r., only the side to which the movable bar is connected should be changed — the other side should remain at one-half the length obtained from Fig. 14-41.

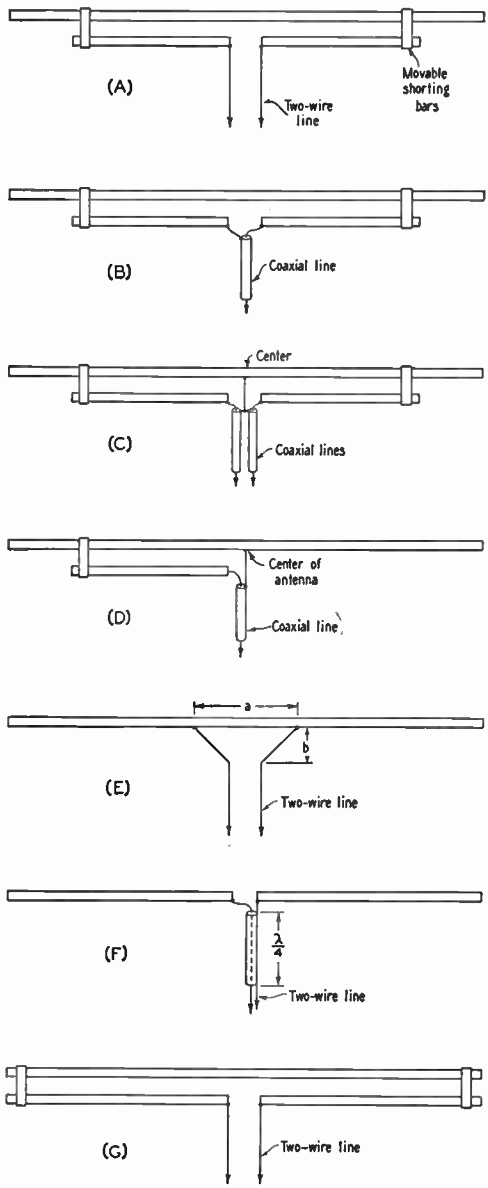


Fig. 14-42 — Recommended methods of feeding the driven antenna element in close-spaced parasitic arrays. The parasitic elements are not shown. A, B, C, "T"-match; D, "gamma" match; E, delta matching transformer; F, coaxial-line quarter-wave matching section; G, folded dipole. Adjustment details are discussed in the text.

With 52-ohm coaxial line feed, the length of the matching element may run around 15 to 20 inches in a 28-Mc. beam, and twice this value in a 14-Mc. array.

An alternative to adjusting the element length for tuning out the residual reactance is to use a small variable condenser in series at the junction of the coaxial-cable inner conductor and the matching section of the gamma match. A small 140-$\mu\mu$fd. receiving-type variable is adequate at powers of a few hundred watts, and it can be weatherproofed by mounting it in a small plastic cup or other housing. The T-match of Figs. 14-42 A, B or C would require two condensers, one in each side.

The delta matching transformer shown at E is probably easier to install, mechanically, than any of the others. The positions of the taps (dimension *a*) must be determined experimentally, along with the length, *b*, by checking the standing-wave ratio on the line as adjustments are made. Dimension *b* should be about 15 per cent longer than *a*.

The coaxial-line matching section at F will work with fair accuracy into a close-spaced parasitic array of 2, 3 or 4 elements without necessity for adjustment. The line is used as a quarter-wavelength transformer, and, if its characteristic impedance is 70 ohms (RG-11/U), it will give a good match to a 600-ohm line when the resistance at the termination is about 8.5 ohms. Over a range of 5 to 15 ohms the mismatch, and therefore the standing-wave ratio, will be less than 2-to-1. The length of the quarter-wave section may be calculated from

$$Length\ (\text{feet}) = \frac{246V}{f} \qquad \text{(14–J)}$$

where V = Velocity factor
f = Frequency in Mc.

Example: A quarter-wave transformer of RG-11/U is to be used at 28.7 Mc. From the table in Chapter Thirteen, $V = 0.66$.

$$\text{Length} = \frac{246 \times 0.66}{28.7} = 5.67 \text{ feet}$$
$$= 5 \text{ feet } 8 \text{ inches}$$

The folded-dipole antenna, Fig. 14-42G, presents a good match for the line when properly designed. Details are given in Chapter Thirteen. Different impedance step-up ratios can be obtained by varying the number of conductors or their diameter ratio.

Sharpness of Resonance

Peak performance of a multielement parasitic array depends upon proper phasing or tuning of the elements, which can be exact for one frequency only. In the case of close-spaced arrays, which because of the low radiation resistance usually are quite sharp-tuning, the frequency range over which optimum results can be secured is only of the order of 1 or 2 per cent of the resonant frequency, or up to about 500 kc. at 28 Mc. However, the antenna can be made to work satisfactorily over a wider frequency range by adjusting the director or directors to give maximum gain at the *highest* frequency to be covered, and by adjusting the reflector to give optimum gain at the *lowest* frequency. This sacrifices some gain at all frequencies, but maintains more uniform gain over a wider frequency range.

As mentioned in the preceding paragraphs, the use of large-diameter conductors will broaden the response curve of an array because the larger diameter lowers the Q. This causes the reactances of the elements to change rather slowly with frequency, with the result that the tuning stays near the optimum over a considerably wider frequency range than is the case with wire conductors.

Combination Arrays

It is possible to combine parasitic elements with driven elements to form arrays composed of collinear driven and parasitic elements and combination broadside-collinear-parasitic elements. Thus two or more collinear elements might be provided with a collinear reflector or director set, one parasitic element to each driven element. Or both directors and reflectors might be used. A broadside-collinear array could be treated in the same fashion.

When combination arrays are built up, a rough approximation of the gain to be expected may be obtained by adding the gains for each type of combination. Thus the gain of two broadside sets of four collinear arrays with a set of reflectors, one behind each element, at quarter-wave spacing for the parasitic elements, would be estimated as follows: From Table 14-III, the gain of four collinear elements is 4.5 db. with half-wave spacing; from Fig. 14-34 or Table 14-IV, the gain of two broadside elements at half-wave spacing is 4.0 db.; from Fig. 14-40, the gain of a parasitic reflector at quarter-wave spacing is 4.5 db. The total gain is then the sum, or 13 db. for the sixteen elements. Note that it makes no difference in the final result if the array is considered as a grouping of several sets of antennas plus reflectors or as an array of antennas plus an array of reflectors. The actual gain of the combination array will depend, in practice, upon the way in which the power is distributed between the various elements and upon the effect which mutual coupling between elements has upon the radiation resistance of the array, and may be somewhat higher or lower than the estimate.

A great many directive-antenna combinations can be worked out by combining elements according to these principles.

● RECEIVING ANTENNAS

Nearly all of the properties possessed by an antenna as a radiator also apply when it is used for reception. Current and voltage distribution, impedance, resistance and directional characteristics are the same in a receiving antenna as if it were used as a transmitting antenna. This reciprocal behavior makes possible the design of a receiving antenna of optimum performance based on the same considerations that have been discussed for transmitting antennas.

The simplest receiving antenna is a wire of random length. The longer and higher the wire, the more energy it abstracts from the wave. Because of the high sensitivity of modern receivers, sometimes only a short length of wire strung

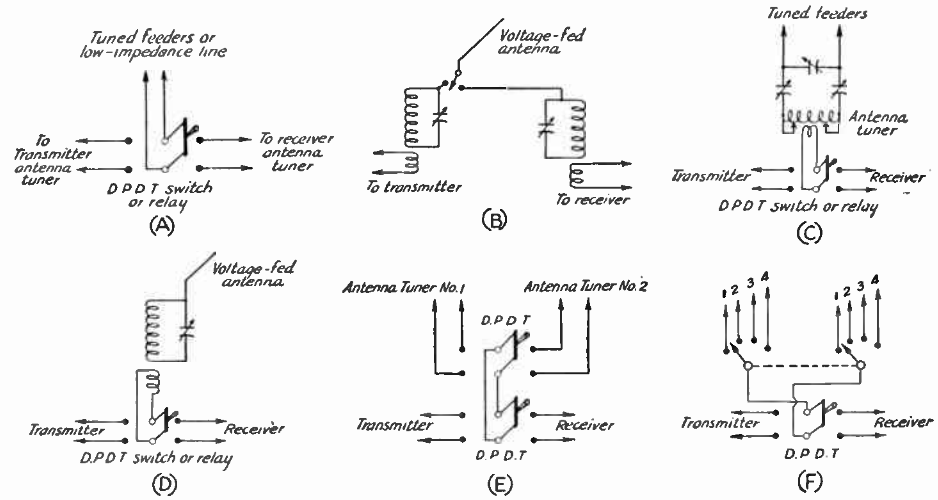


Fig. 14-43 — Antenna-switching arrangements for various types of antennas and coupling systems. A — For tuned lines with separate antenna tuners or low-impedance lines. B — For a voltage-fed antenna. C — For a tuned line with a single antenna tuner. D — For a voltage-fed antenna with a single tuner. E — For two tuned-line antennas with a tuner for each antenna or for two low-impedance lines. F — For combinations of several two-wire lines.

around the room is used for a receiving antenna, but such an antenna cannot be expected to give good performance, although it is adequate for loud signals on the 3.5- and 7-Mc. bands. It will serve in emergencies, but a longer wire outdoors is always better.

The use of a tuned antenna improves the operation of the receiver, however, because the signal strength is raised more in proportion to the stray noises picked up than is the case with wires of random length. Since the transmitting antenna usually is given the best location, it can also be expected to serve best for receiving. This is especially true when a directive antenna is used, since the directional effects and power gain of directive transmitting antennas are the same for receiving as for transmitting.

In selecting a directional receiving antenna it is preferable to choose a type that gives very little response in all but the desired direction (small minor lobes). This is even more important than high gain in the desired direction, because the cumulative response to noise and unwanted-signal interference in the smaller lobes may offset the advantage of increased desired-signal gain. The feed line from the antenna should be balanced so that it will not pick up signals and thus greatly reduce the directivity effects.

Antenna Switching

Switching of the antenna from receiver to transmitter is commonly done with a change-over relay, connected in the antenna leads or the coupling link from the antenna tuner. If the relay is one with a 115-volt a.c. coil, the switch or relay that controls the transmitter plate power will also control the antenna relay. If the convenience of a relay is not desired, porcelain knife switches can be used and thrown by hand.

Typical arrangements are shown in Fig. 14-43. If coaxial line is used, the use of a coaxial relay is recommended, although on the lower-frequency bands a regular switch or change-over relay will work almost as well.

Antenna Construction

The use of good materials in the antenna system is important, since the antenna is exposed to wind and weather. To keep electrical losses low, the wires in the antenna and feeder system must have good conductivity and the insulators must have low dielectric loss and surface leakage, particularly when wet.

For short antennas, No. 14 gauge hard-drawn enameled copper wire is a satisfactory conductor. For long antennas and directive arrays, No. 14 or No. 12 enameled copper-clad steel wire should be used. It is best to make feeders and matching stubs of ordinary soft-drawn No. 14 or No. 12 enameled copper wire, since hard-drawn or copper-clad steel wire is difficult to handle unless it is under considerable tension at all times. The wires should be all in one piece; where a joint cannot be avoided, it should be carefully soldered.

In building a two-wire open line, the spacer insulation should be of as good quality as in the antenna insulators proper. For this reason, good ceramic spacers are advisable. Wooden dowels boiled in paraffin may be used with untuned lines, but their use is not recommended for tuned lines. The wooden dowels can be attached to the feeder wires by drilling small holes and binding them to the feeders with wire.

At points of maximum voltage, insulation is most important, and Pyrex glass, Isolantite or Steatite insulators with long leakage paths are recommended for the antenna. Glazed porcelain also is satisfactory. Insulators should be

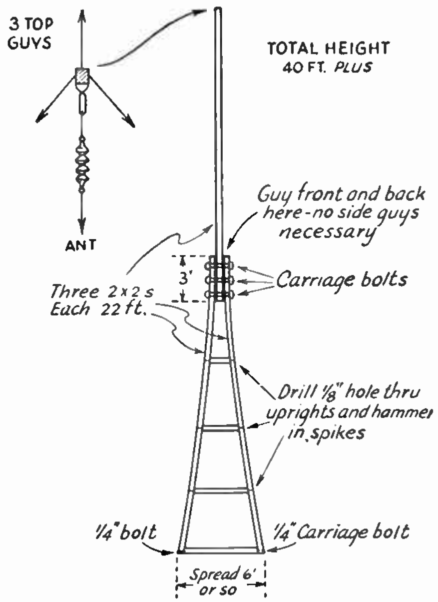


Fig. 14-44 — **Details of a simple 40-foot "A"-frame mast suitable for erection in locations where space is limited.**

cleaned once or twice a year, especially if they are subjected to much smoke and soot.

In most cases poles or masts are desirable to lift the antenna clear of surrounding buildings, although in some locations the antenna will be sufficiently in the clear when strung from one chimney to another or from a housetop to a tree. Small trees usually are not satisfactory as points of suspension for the antenna because of their movement in windy weather. If the antenna is strung from a point near the center of the trunk of a large tree, this difficulty is not so serious. Where the antenna wire must be strung from one of the smaller branches, it is best to tie a pulley firmly to the branch and run a rope through the pulley to the antenna, with the other end of the rope attached to a counterweight near the ground. The counterweight will keep the tension on the antenna wire reasonably constant even when the branches sway or the rope tightens and stretches with varying climatic conditions.

Telephone poles, if they can be purchased and installed economically, make excellent supports because they do not ordinarily require guying in heights up to 40 feet or so. Many low-cost television-antenna supports are now available, and they should not be overlooked as possible antenna aids.

"A"-FRAME MAST

The simple and inexpensive mast shown in Fig. 14-44 is satisfactory for heights up to 35 or 40 feet. Clear, sound lumber should be selected. The completed mast may be protected by two or three coats of house paint.

If the mast is to be erected on the ground, a couple of stakes should be driven to keep the bottom from slipping and it may then be "walked up" by a pair of helpers. If it is to go on a roof, first stand it up against the side of the building and then hoist it from the roof, keeping it vertical. The whole assembly is light enough for two men to perform the complete operation — lifting the mast, carrying it to its permanent berth, and fastening the guys — with the mast vertical all the while. It is entirely practicable, therefore, to erect this type of mast on any small, flat area of roof.

By using 2 × 3s or 2 × 4s, the height may be extended up to about 50 feet. The 2 × 2 is too flexible to be satisfactory at such heights.

SIMPLE 40-FOOT MAST

The mast shown in Fig. 14-45 is relatively strong, easy to construct, readily dismantled, and costs very little. Like the "A"-frame, it is suitable for heights of the order of 40 feet.

The top section is a single 2 × 3, bolted at the bottom between a pair of 2 × 3s with an overlap of about two feet. The lower section thus has two legs spaced the width of the narrow side of a 2 × 3. At the bottom the two

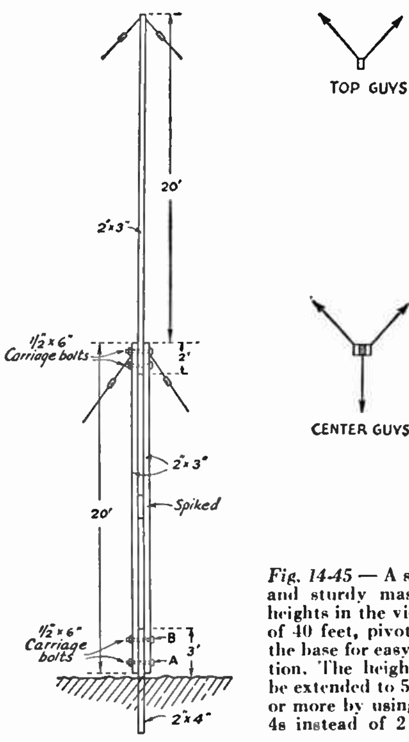


Fig. 14-45 — **A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 × 4s instead of 2 × 3s.**

legs are bolted to a length of 2 × 4 which is set in the ground. A short length of 2 × 3 is placed between the two legs about halfway up the bottom section, to maintain the spacing.

The two back guys at the top pull against the antenna, while the three lower guys prevent buckling at the center of the pole.

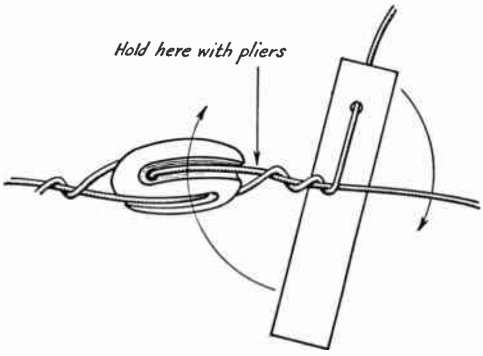


Fig. 14-46 — Using a lever for twisting heavy guy wires.

The 2 × 4 section should be set in the ground so that it faces the proper direction, and then made vertical by lining it up with a plumb bob. The holes for the bolts should be drilled beforehand. With the lower section laid on the ground, bolt *A* should be slipped in place through the three pieces of wood and tightened just enough so that the section can turn freely on the bolt. Then the top section may be bolted in place and the mast pushed up, using a ladder or another 20-foot 2 × 3 for the job. As the mast goes up, the slack in the guys can be taken up so that the whole structure is in some measure continually supported. When the mast is vertical, bolt *B* should be slipped in place and both *A* and *B* tightened. The lower guys can then be given a final tightening, leaving those at the top a little slack until the antenna is pulled up, when they should be adjusted to pull the top section into line.

GUYS AND GUY ANCHORS

For masts or poles up to about 50 feet, No. 12 iron wire is a satisfactory guy-wire material. Heavier wire or stranded cable may be used for taller poles or poles installed in locations where the wind velocity is high.

More than three guy wires in any one set usually are unnecessary. If a horizontal antenna is to be supported, two guy wires in the top set will be sufficient in most cases. These should run to the rear of the mast about 100 degrees apart to offset the pull of the antenna. Intermediate guys should be used in sets of three, one running in a direction opposite to that of the antenna, while the other two are spaced 120 degrees either side. This leaves a clear space under the antenna. The guy wires should be adjusted to pull the pole slightly back from vertical before the antenna is hoisted so that when the antenna is pulled up tight the mast will be straight.

When raising a mast that is big enough to tax the facilities available, it is some advantage to know nearly exactly the length of the guys. Those on the side on which the pole is lying can then be fastened temporarily to the anchors beforehand, which assures that when the pole is raised, those holding opposite guys will be able to pull it into nearly-vertical position with no danger of its getting out of control. The guy lengths can be figured by the right-angled-triangle rule that "the sum of the squares of the two sides is equal to the square of the hypotenuse." In other words, the distance from the base of the pole to the anchor should be measured and squared. To this should be added the square of the pole length to the point where the guy is fastened. The square root of this sum will be the length of the guy.

Guy wires should be broken up by strain insulators, to avoid the possibility of resonance at the transmitting frequency. Common practice is to insert an insulator near the top of each guy, within a few feet of the pole, and then cut each section of wire between the insulators to a length which will not be resonant either on the fundamental or harmonics. An insulator every 25 feet will be satisfactory for frequencies up to 30 Mc. The insulators should be of the "egg" type with the insulating material under compression, so that the guy will not part if the insulator breaks.

Twisting guy wires onto "egg" insulators may be a tedious job if the guy wires are long and of large gauge. The simple time- and finger-saving device shown in Fig. 14-46 can be made from a piece of heavy iron or steel by drilling a hole about twice the diameter of the guy wire about a half inch from one end of the piece. The wire is passed through the insulator, given a single turn by hand, and then held with a pair of pliers at the point shown in the sketch. By passing the

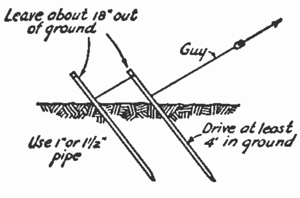


Fig. 14-47 — Pipe guy anchors. One pipe is sufficient for small masts, but two installed as shown will provide the additional strength required for the larger poles.

wire through the hole in the iron and rotating the iron as shown, the wire may be quickly and neatly twisted.

Guy wires may be anchored to a tree or building when they happen to be in convenient spots. For small poles, a 6-foot length of 1-inch pipe driven into the ground at an angle will suffice. Additional bracing will be provided by using two pipes, as shown in Fig. 14-47.

HALYARDS AND PULLEYS

Halyards or ropes and pulleys are important items in the antenna-supporting system. Particular attention should be directed toward the choice of a pulley and halyards for a high mast since replacement, once the mast is in position, may be a major undertaking if not entirely impossible.

Galvanized-iron pulleys will have a life of only a year or so. Especially for coastal-area installations, marine-type pulleys with hardwood blocks and bronze wheels and bearings should be used.

For short antennas and temporary installations, heavy clothesline or window-sash cord may be used. However, for more permanent jobs, 3/8-inch or 1/2-inch waterproof hemp rope should be used. Even this should be replaced about once a year to insure against breakage.

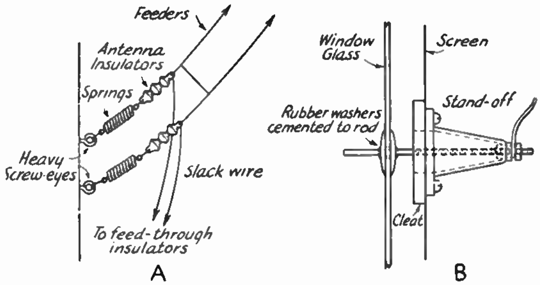


Fig. 14-49 — A — Anchoring feeders takes the strain from feed-through insulators or window glass. B — Going through a full-length screen, a cleat is fastened to the frame of the screen on the inside. Clearance holes are cut in the cleat and also in the screen.

Nylon rope, used during the war as glider tow rope, is, of course, one of the best materials for halyards, since it is weatherproof and has extremely long life.

It is advisable to carry the pulley rope back up to the top in "endless" fashion in the manner of a flag hoist so that if the antenna breaks close to the pole, there will be a means for pulling the hoisting rope back down.

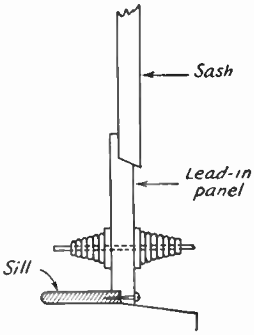


Fig. 14-48 — An antenna lead-in panel may be placed over the top sash or under the lower sash of a window. Substituting a smaller height sash in half the window will simplify the weatherproofing problem where the sash overlap.

BRINGING THE ANTENNA OR FEED LINE INTO THE STATION

The antenna or transmission line should be anchored to the outside wall of the building, as shown in Fig. 14-49, to remove strain from the lead-in insulators. Holes cut through the walls of the building and fitted with feed-through insulators are undoubtedly the best means of bringing the line into the station. The holes should have plenty of air clearance about the conducting rod, especially when using tuned lines that develop high voltages. Probably the best place to go through the walls is the trimming board at the top or bottom of a window frame which provides flat surfaces for lead-in insulators. Either cement or rubber gaskets may be used to waterproof the exposed joints.

Where such a procedure is not permissible, the window itself usually offers the best opportunity. One satisfactory method is to drill holes in the glass near the top of the upper sash. If the glass is replaced by plate glass, a stronger job will result. Plate glass may be obtained from automobile junk yards and drilled before placing in the frame. The glass itself provides insulation and the transmission line may be fastened to bolts fitting the holes. Rubber gaskets will render the holes waterproof. The lower sash should be provided with stops to prevent damage when it is raised. If the window has a full-length screen, the scheme shown in Fig. 14-49B may be used.

As a less permanent method, the window may be raised from the bottom or lowered from the top to permit insertion of a board which carries the feed-through insulators. This lead-in arrangement can be made weatherproof by making an overlapping joint between the board and window sash, as shown in Fig. 14-48.

LIGHTNING PROTECTION

An ungrounded radio antenna, particularly if large and well elevated, is a lightning hazard.

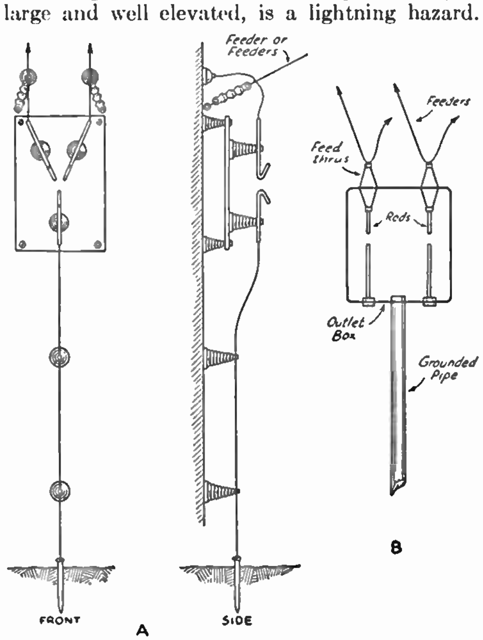


Fig. 14-50 — Low-loss lightning arresters for transmitting-antenna installations.

When grounded, it provides a measure of protection. Therefore, grounding switches or lightning arresters should be provided. Examples of construction of low-loss arresters are shown in Fig. 14-50. At A, the arrester electrodes are mounted by means of stand-off insulators on a fireproof asbestos board. At B, the electrodes are enclosed in a standard steel outlet box. The gaps should be made as small as possible without danger of breakdown during operation. Lightning-arrester systems require the best ground connection obtainable.

The most positive protection is to ground the antenna system when it is not in use; grounded flexible wires provided with clips for connection to the feeder wires may be used. The ground lead should be of short length and run, if possible, directly to a driven pipe or water pipe where it enters the ground outside the building.

Rotary-Beam Construction

It is a distinct advantage to be able to shift the direction of a beam antenna at will, thus securing the benefits of power gain and directivity in any desired compass direction. A favorite method of doing this is to construct the antenna so that it can be rotated in the horizontal plane. The use of such rotatable antennas is usually limited to the higher frequencies — 14 Mc. and above — and to the simpler antenna-element combinations if the structure size is to be kept within practicable bounds. For the 14-, 21- and 28-Mc. bands such antennas usually consist of two to four elements and are of the parasitic-array type described earlier in this chapter. At 50 Mc. and higher it becomes possible to use more elaborate arrays because of the shorter wavelength and thus obtain still higher gain. Antennas for these bands are described in another chapter.

The problems in rotary-beam construction are those of providing a suitable mechanical support for the antenna elements, furnishing a means of rotation, and attaching the transmission line so that it does not interfere with the rotation of the system.

Elements

The antenna elements usually are made of metal tubing so that they will be at least partially self-supporting, thus simplifying the supporting structure. The large diameter of the conductor is beneficial also in reducing resistance, which becomes an important consideration when close-spaced elements are used.

Dural tubes often are used for the elements, and thin-walled corrugated steel tubes with copper coating also are available for this purpose. The elements frequently are constructed of sections of telescoping tubing making length adjustments for tuning quite easy. Electrician's thin-walled conduit also is suitable for rotary-beam elements.

If steel elements are used, special precautions should be taken to prevent rusting. Even copper-coated steel does not stand up indefinitely, since the coating usually is too thin. The elements should be coated both inside and out with slow-drying aluminum paint. For coating the inside, a spray gun may be used, or the paint may be poured in one end while rotating the tubing. The excess paint may be caught as it comes out the bottom end and poured through again until it is certain that the entire inside wall has been covered. The ends should then be plugged up with corks sealed with glyptal varnish.

Supports

The supporting framework for a rotary beam usually is made of wood or metal, using as lightweight construction as is consistent with the required strength. Generally, the frame is not required to hold much weight, but it must be extensive enough so that the antenna elements can be supported without excessive sag, and it must have sufficient strength to stand up under the maximum wind in the locality. The design of the frame will depend on the size and strength of the elements, whether they are mounted horizontally or vertically, and the method used to rotate the antenna.

The general preference is for horizontal polarization, primarily because less height is required to clear surrounding obstructions when all the antenna elements are in the horizontal plane. This is important at 14 and 21 Mc. where the elements are fairly long.

The support may be coupled to the pole by any convenient means which permits rotation or, alternatively, it may be firmly fastened to the pole and the latter rotated in bearings affixed to the side of the house.

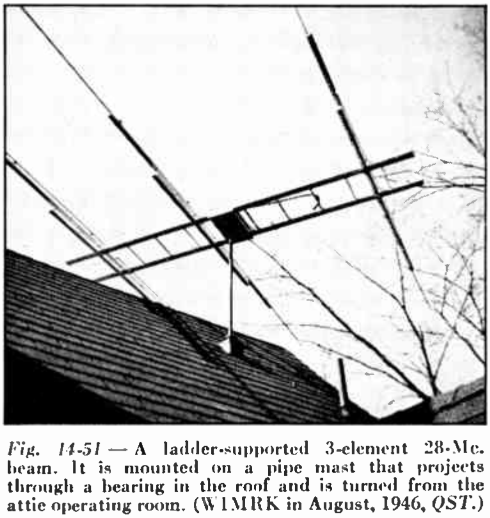


Fig. 14-51 — A ladder-supported 3-element 28-Mc. beam. It is mounted on a pipe mast that projects through a bearing in the roof and is turned from the attic operating room. (W1MRK in August, 1946, *QST*.)

One type of construction is shown in Fig. 14-51. It uses a section of ordinary ladder as the main support, with crosspieces to hold the tubing antenna elements.

Metal can be used to support the elements of the rotary beam. For 28 Mc., a piece of 2-inch diameter duraluminum tubing makes a good "boom" for supporting the elements. The elements can be made to slide through suitable holes in the boom, or special clamps and brackets can be fashioned to support the elements. Fittings for TV antennas can often be used on 28-Mc. beams.

With all-metal construction, delta, "gamma" or "T"-match are the only practical matching methods to use to the line, since anything else requires opening the driven element at the center, and this complicates the support problem for that element.

A Wooden Boom for 14 Mc.

Many amateurs prefer to build their beam booms from standard pieces of lumber, and the beam shown in Figs. 14-52 and 14-53 is an example of excellent design in wooden-boom construction. The boom members are two 20-foot 2 × 4s fastened to the 4 × 12 × 24-inch center block with six lag screws. The two center screws serve as the axis for tilting — the other four lock the boom in position after final assembly and adjustment have been completed. The blocks midway from each end are 2 × 4s spaced about six inches apart, with a long bolt between them. When this bolt is drawn tight, a very sturdy box brace is formed. The crossarms are 3 × 3s twelve feet long, bolted to the boom with carriage bolts.

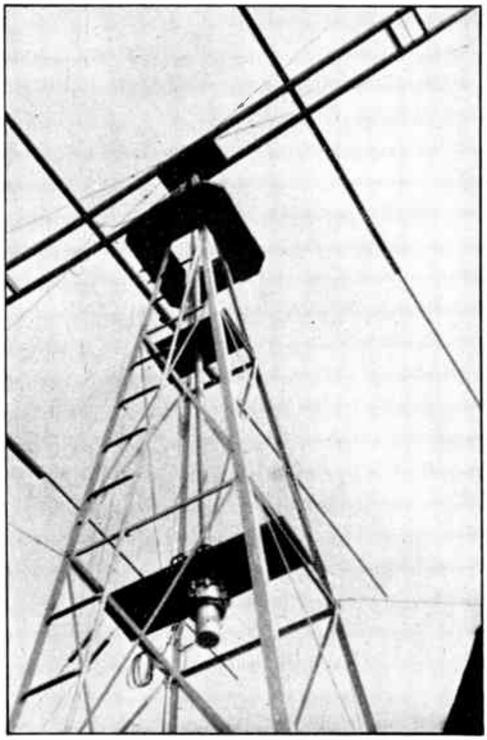


Fig. 14-52 — A wooden boom for a 4-element 14-Mc. boom can be made quite strong by judicious use of guy wires. This installation is made on a windmill tower, and the drive motor is mounted halfway down on the tower. (W6MJB, Nov., 1947, *QST.*)

The umbrella guys should have turnbuckles in them, and the guys are fastened to the center support after the beam has been permanently locked in its horizontal position. With the turnbuckles properly adjusted, there will be no sag in the boom and the elements will be neat.

The elements are 1⅜- and 1½-inch diameter duraluminum tubing, supported by 1½-inch stand-off insulators. Hose clamps are used to hold the elements on the insulators. Final adjustment of element lengths is possible through "hairpin" loops. The tower for the beam shown in Fig. 14-52 was a Sears-Roebuck windmill tower. The driving motor for the beam was located halfway down the tower, the torque being transmitted through a length of 1½-inch drive shaft. A pipe flange is welded to the drive shaft and bolted to the center block. A cone bearing is obtained by turning both the flange and a sleeve of 2-inch pipe to match, as shown in Fig. 14-53.

One method of matching the line to the antenna is to use a quarter wavelength of 75-ohm Twin-Lead between the radiator and the slip-ring contacts, to match a 600-ohm line from the slip rings to the transmitter.

A 600-ohm open-wire line is run to a point about halfway up on the tower, then up the side of the tower to the slip rings. The slip rings are mounted on the top of the tower, directly under the center block. A quarter-wavelength matching section of transmitting-type 75-ohm Amphenol Twin-Lead hangs in a loop between the driven element and the slip-ring contacts.

"Plumber's-Delight" Construction

The lightest beam to build is the so-called "plumber's delight" — an array constructed entirely of metal, with no insulating members between the elements and the supporting structure. Suggested constructional details are shown in Figs. 14-54, 14-55, 14-56, 14-57 and 14-58.

The boom can be built of two lengths of 3-inch diameter 61S-T6 dural tubing of 0.072-inch wall thickness, as shown in Fig. 14-54. The two sections are spliced together with a three-foot length of 6 × 6 oak, turned down at each end to fit inside the tubing. The center of the block is left square to provide a flat surface to attach to the vertical rotating pipe. At each extremity of this boom is cut a hole the exact diameter of the parasitic elements. A two-foot length of ¾-inch pipe, complete with flange mounting plate, is bolted to the top surface of the oak block, and a single guy wire is run to each end of the boom. An egg insulator and a turnbuckle are placed in each guy. The turnbuckles should be tightened until there is no sag in the boom when it is supported at the center, and then safety-wired.

Finally the center block should be given a good coat of paint or varnish.

The elements can be made of three 12-foot lengths of dural tubing, the two outside lengths telescoping inside the center section. The ends of the center section should be slotted for a distance of about 4 inches with a hack saw, but it is advisable to do the slotting after the center sections have been assembled on the boom. The parasitic-element center sections are fastened to the boom with ¼-inch bolts, as shown in Fig. 14-55, while the driven element is secured in a cradle made of half sections of iron pipe welded together, as shown in Fig. 14-56. The cradle is bolted to the boom with three ¼-inch bolts, and the driven element is held fast with two bolts or with adjustable aircraft-tubing clamps.

The feed line for the antenna can be any balanced line, of from 200 to 600 ohms impedance, and it is most conveniently coupled through a "T"-match. This "T"-match assembly can be made from two 4-foot lengths of dural tubing joined together by a piece of broomstick, as shown in Fig. 14-58. The "T" is connected to the antenna by two clamps fashioned of 1-inch-wide brass strip.

A convenient method for supporting the boom atop the pipe used to rotate the beam is shown in Fig. 14-57. A "U"-channel into which the boom will fit is welded to the end of the pipe. Holes are drilled in the side of the channel corresponding to holes in the boom. The boom is hoisted up and positioned between the two flanges and a bolt run through the flanges and the boom. The boom can then be swung into a horizontal position and the second bolt put in place.

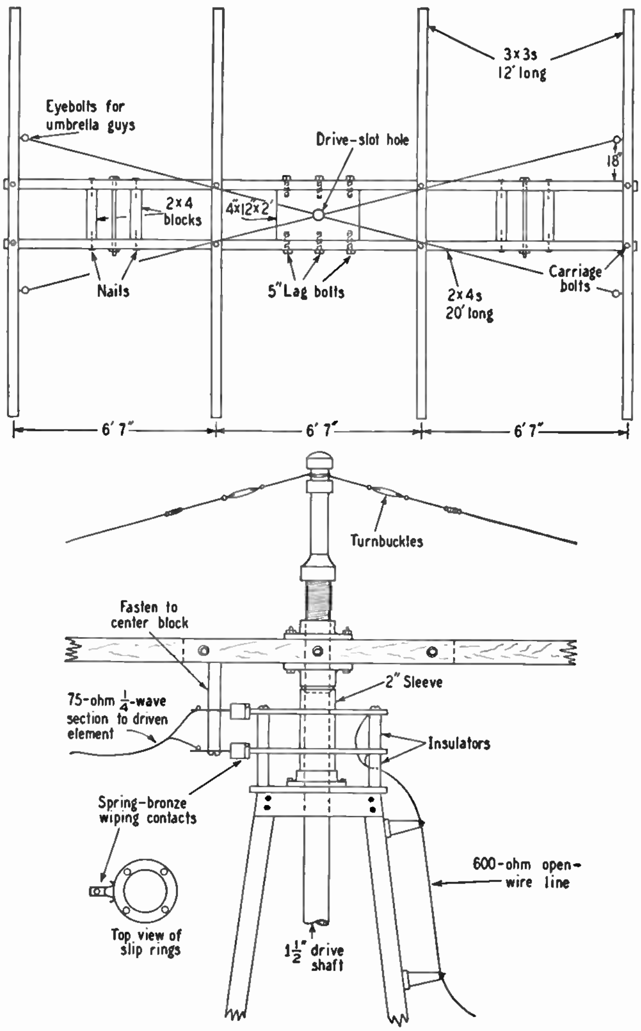


Fig. 14-53 — Details of the wooden boom, its method of support and the construction of the slip rings.

Feeder Connections

For beams that rotate only 360 degrees, it is common to bring off feeders by making a short section of the feeder, just where it leaves the rotating member, of flexible wire. Enough slack should be left so that there is no danger of breaking or twisting. Stops should be placed on the rotating shaft of the antenna so that it will be impossible for the feeders to "wind up."

For continuous rotation, the sliding contact is simple and, when properly built, quite practicable. The chief points to keep in mind are that the contact surfaces should be wide enough to take care of wobble in the rotating shaft, and that the contact surfaces should be kept clean. Spring contacts are essential, and an "umbrella" or other scheme for keeping rain off the contacts is a desirable addition. Sliding contacts preferably should be used with nonresonant open-wire lines, so that the line current is low.

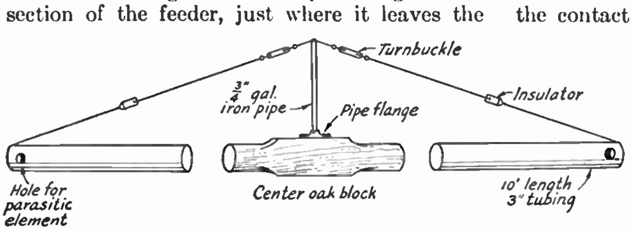


Fig. 14-54 — The boom is made of two 10-foot lengths of dural tubing slipped over a 3-foot oak block and held in place with 2-inch wood screws. Guy wires from the center add strength to the boom structure.

The possibility of poor connec-

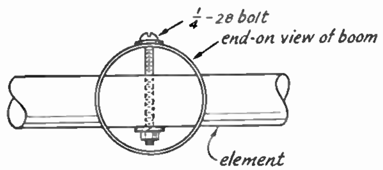


Fig. 14-55 — The center element section is held in the boom with a ¼-28 machine screw, nut and lock washer. The guy wire attaches to the head of the bolt.

tions in sliding contacts can be avoided by using inductive coupling at the antenna, with one coil rotating on the antenna and the other fixed in position, the two coils being arranged so that the coupling does not change when the antenna is rotated. A quarter-wave feeder system is connected to a tuned pick-up circuit whose inductance is coupled to a link. The link coil connects to a twisted-pair transmission line, but any type of line such as flexible coaxial cable can be used. The circuit would be adjusted in the same way as any link-coupled circuit, and the number of turns in the link should be varied to give proper loading on the transmitter. The rotating coupling circuit tunes to the transmitting frequency. The system is equivalent to a link-coupled antenna tuner mounted on the pole, using a parallel-tuned tank at the end of a quarter-wave line to center-

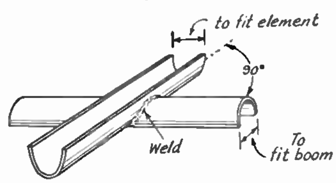


Fig. 14-56 — The clamp for the driven element is made by splitting 1-foot lengths of iron pipe and welding them as shown.

feed the antenna. For constant coupling, the two coils should be rigid and the pole should rotate without wobble. The two coils might be made a part of the upper bearing assembly holding the rotating pole in position.

There are other variations of the inductive-coupled system. The tuned circuit might, for instance, be placed at the end of a 600-ohm line, and a one-turn link used to couple directly to the center of the antenna, if the construction of the rotary member permits. In this case the coupling can be varied by changing the *L/C* ratio in the tuned circuit. For mechanical strength the coupling coils preferably should be made of ¼-inch copper tubing, well braced with insulating strips to keep them rigid.

Rotation

It is convenient but not essential to use a motor to rotate the beam. If a rope-and-pulley arrangement can be brought into the operating room or if the pole can be mounted near a window in the operating room, hand rotation of the beam will work.

If the use of a rope and pulleys is impracticable, motor drive is about the only alternative. There are several complete motor driven rotators on the market, and they are easy to mount, convenient to use, and require little or no maintenance. Generally speaking, light-weight units are better because they reduce the load on the mast or tower.

The speed of rotation should not be too great — one or two r.p.m. is about right. This

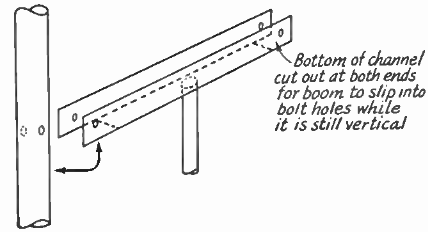


Fig. 14-57 — The mounting plate is made from a length of "U"-channel iron cut and drilled as shown. The boom is raised vertically until one set of bolt holes is in line and a bolt is slipped through. The boom is then swung into its horizontal position and the other bolt is put in place.

requires a considerable gear reduction from the usual 1750-r.p.m. speed of small induction motors; a large reduction is advantageous because the gear train will prevent the beam from turning in weather-vane fashion in a wind. The usual beam does not require a great deal of power for rotation at slow speed, and a ⅛-hp. motor will be ample. A reversible motor should be used so that it will not be necessary to go through nearly 360 degrees to bring the beam back to a direction only slightly different, but in the opposite direction of rotation, to the direction to which it may be pointed at the moment. In cases where the pole is stationary and only the supporting framework rotates, it will be necessary to mount the motor and gear train in a housing on or near the top of the pole. If the pole rotates, the motor can be installed in a more accessible location.

Driving motors and gear housings will stand the weather better if given a coat of aluminum paint followed by two coats of enamel and a coat of glyptal varnish. Even commercial units will last longer if treated with glyptal varnish. Be sure that the surfaces are clean and free from grease before painting. Grease can be removed by brushing with kerosene and then squirting the surface with a solid stream of water. The work can then be wiped dry with a rag.

The power and control leads to the rotator should be run in electrical conduit or in lead covering, and the metal should be grounded.

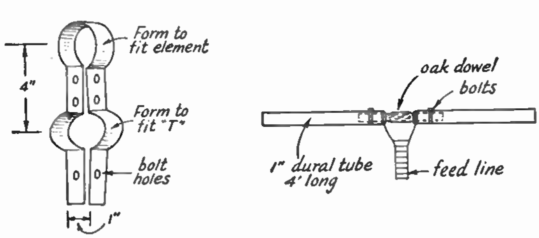


Fig. 14-58 — Details of the "T"-match assembly.

About V.H.F.

While it is possible to use the frequencies above 30 Mc. without knowing anything about wave propagation, the amateur who understands something of the means by which his signals reach distant points will be able to do a better job of it. Because much of the pleasure and satisfaction to be derived from v.h.f. work lie in making the best possible use of propagation vagaries associated with natural phenomena, a working knowledge of the basic principles of wave propagation is a most useful tool for the v.h.f. operator.

Characteristics of the Bands Above 50 Mc.

The assignments from 50 Mc. up are superior to our lower bands in one outstanding respect: their ability to provide interference-free communication consistently within a limited service area. Lower frequencies are more subject to varying conditions that impair their effectiveness for work over a radius of 100 miles or less at least part of the time, and the heavy occupancy they support creates a continuing interference problem. Our v.h.f. bands, on the other hand, are seldom crowded, and their characteristics for local work are more stable. Because of these attributes the 50- and 144-Mc. bands, particularly, enjoy considerable popularity in areas where there are dense concentrations of population.

In addition, it has been found that there are several media by which v.h.f. signals are propagated beyond the local range, and operation on the v.h.f. bands has been taken up by many operators who must depend almost entirely on "DX" for their contacts. The latter group, particularly, will benefit from a familiarity with common propagation phenomena. The material to follow is intended to supplement the more detailed information in Chapter Four, dealing with wave propagation as it affects the world above 50 Mc.

50 to 54 Mc.

This band is borderline territory between the frequencies regularly used for long-distance communication and those normally employed for local work. Thus just about every form of wave propagation to be found throughout the radio spectrum will appear, on occasion, in the 50-Mc. region. This diversity has contributed greatly to the growing popularity of the 50-Mc. band in the amateur picture.

During the peak years of the sunspot cycle it is occasionally possible to work 50-Mc. DX of worldwide proportions, by reflection of signals from the F_2 layer. Sporadic-E skip provides opportunities for work over distances from 400 to 2500 miles or so during the early summer months, regardless of the solar cycle. Reflection from the aurora regions accounts for communication over 100 to 600-mile paths during pronounced ionospheric disturbances. The ever-changing weather pattern offers frequent opportunities for extension of the normal coverage to as much as 300 miles. This tropospheric condition develops most often during the warmer months, but may occur at any season. In the absence of any favorable propagation, the average well-equipped 50-Mc. station should be able to work regularly over a radius of 75 to 100 miles or more, depending on local terrain.

144 to 148 Mc.

Ionospheric effects are greatly reduced at 144 Mc. It is doubtful whether F_2-layer reflection ever occurs at this frequency, and sporadic-E skip is rare. Aurora DX is fairly common, but the signals are generally weaker than on 50 Mc. Tropospheric effects are more pronounced than on 50 Mc., and distances covered during favorable weather conditions are greater than on lower bands. Air-mass boundary bending has been responsible for communication on 144 Mc. over distances in excess of 1100 miles, and 500-mile work is fairly common in the warmer months. The reliable working range under normal conditions is slightly less than on 50 Mc., when comparable equipment and antennas are used.

220 Mc. and Higher

Ionospheric propagation is unlikely at 220 Mc. and up, but tropospheric bending is more prevalent than on lower bands. Amateur experience on 220 and 420 Mc. is showing that they can be as useful as 144 Mc., when comparable equipment is used. Under minimum conditions the range may be slightly shorter, but when signals are good on 144 Mc., they may be better on 220 or 420. Even at 1215 Mc. and higher there is evidence that paths well beyond line of sight may be covered successfully.

Propagation Phenomena

The various known means by which v.h.f. signals may be propagated over unusual distances are discussed below.

F_2-Layer Reflection

The "normal" contacts made on 28 Mc. and lower frequencies are the result of reflection of the transmitted wave by the F_2 layer, the ionization density of which varies with solar activity, the highest frequencies being reflected at the peak of the 11-year solar cycle. The maximum usable frequency (m.u.f.) for F_2 reflection also rises and falls with other well-defined cycles, including daily, monthly, and seasonal variations, all related to conditions on the sun and its position with respect to the earth.

At the low point of the 11-year cycle, such as the period we encountered in the early '50s, the m.u.f. may reach 28 Mc. only during a short period each spring and fall, whereas it may go to 60 Mc. or higher at the peak of the cycle. The fall of 1946 saw the first authentic instances of long-distance work on 50 Mc. by F_2-layer reflection, and as late as 1950 contacts were still being made in the more favorable areas of the world by this medium. In the northern latitudes there are peaks of m.u.f. each spring and fall, with a low period during the summer and a slight dropping-off during the midwinter months. At or near the Equator conditions are more or less constant at all seasons.

Fortunately the F_2 m.u.f. is quite readily determined by observation, and means are available whereby it may be estimated quite accurately for any path at any time. It is predictable for months in advance,[1] enabling the v.h.f. worker to arrange test schedules with distant stations at propitious times. As there are numerous signals, both harmonics and fundamental transmissions, on the air in the range between 28 and 50 Mc., it is possible for an observer to determine the approximate m.u.f. by careful listening in this range. A series of daily observations will serve to show if the m.u.f. is rising or falling from day to day, and once the peak for a given month is determined it can be assumed that the peak for the following month will occur about 27 days later, this cycle coinciding with the turning of the sun on its axis. The working range, via F_2 skip, is roughly comparable to that on 28 Mc., though the *minimum* distance is somewhat longer. Two-way work on 50 Mc. by reflection from the F_2 layer has been accomplished over distances ranging from 2200 to 10,500 miles. The maximum frequency for F_2 reflection is believed to be in the vicinity of 70 Mc. F_2 DX on 50 Mc. is unlikely again before 1956.

Sporadic-E Skip

Patchy concentrations of ionization in the E-layer region are often responsible for re-

[1] *Basic Radio Propagation Predictions*, issued monthly, three months in advance, by the Central Radio Propagation Laboratory of the National Bureau of Standards. Order from the Supt. of Documents, Washington 25, D. C.; $1.00 per year.

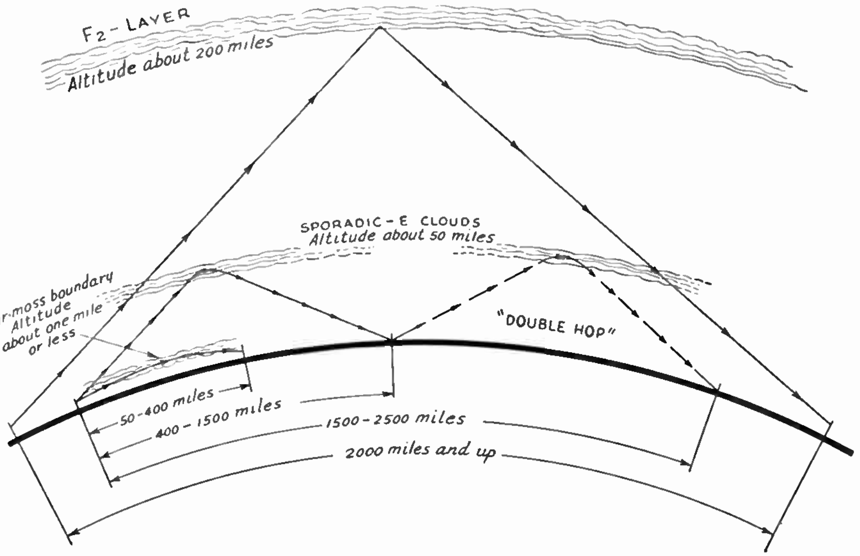


Fig. 15-1 — The principal means by which v.h.f. signals may be returned to earth. The F_2 layer, highest of the known reflecting regions of the ionosphere, is capable of reflecting 50-Mc. signals during the peak period of the 11-year solar cycle. Such communication may be world-wide in scope. Sporadic ionization of the E layer produces the familiar "short skip" contacts over medium distances at 28 and 50 Mc. On these bands it is a fairly frequent occurrence regardless of the solar cycle. It is most common in May through August. Refraction of v.h.f. waves also takes place at air-mass boundaries in the lower atmosphere, making possible communication over distances of several hundred miles, usually without a skip zone, on all v.h.f. bands.

flection of signals on 28 and 50 Mc. This is the popular "short skip" that provides fine contacts on both bands in the range between 400 and 1300 miles. It is most common in May, June and July, during morning and early evening hours, but it may occur at any time or season. Since it is largely unpredictable, at our present state of knowledge, sporadic-E skip is of high "surprise value." Multiple-hop effects may appear, when ionization develops simultaneously over large areas, making possible work over distances of more than 2500 miles.

The upper limit of frequency for sporadic-E skip is not positively known, but scattered instances of 144-Mc. propagation over distances in excess of 1000 miles indicate that E-layer reflection, possibly aided by tropospheric effects, may be responsible.

Aurora Effect

Low-frequency communication is occasionally wiped out by absorption of these frequencies in the ionosphere, when ionospheric storms, associated with variations in the earth's magnetic field, occur. During such disturbances, however, v.h.f. signals may be reflected back to earth, making communication possible over distances not normally workable in the v.h.f. range. Magnetic storms may be accompanied by an aurora-borealis display, if the disturbance occurs at night and visibility is good. When the aurora is confined to the northern sky, aiming a directional array at the auroral curtain will bring in signals strongest, regardless of the true direction to the transmitting station. When the display is widespread there may be only a slight improvement noted when the directional array is aimed north. The latter condition is often noticed during the period around the peak of the 11-year cycle, when solar activity is spread well over the sun's surface, instead of being concentrated in the region near the solar equator.

Aurora-reflected signals are characterized by a rapid flutter, which lends a "dribbling" sound to 28-Mc. carriers and may render modulation on 50- and 144-Mc. signals completely unreadable. The only satisfactory means of communication then becomes straight c.w. The effect may be noticeable on signals from any distance other than purely local, and stations up to about 800 miles in any direction may be worked at the peak of the disturbance. Unlike the two methods of propagation previously described, aurora effect exhibits no skip zone. It is observed frequently on 50 Mc., and pronounced disturbances affect the 144-Mc. band similarly. The highest frequency for aurora reflection is not yet known.

Scatter

When long-distance communication is possible on 50 Mc., stations within the skip zone may be heard with a wavery quality indicative of multipath reception. Such signals have traversed a normal ionospheric path, via either the F_2 or E layer, and a small amount of energy has returned to the receiver by reflection from a distant point on the earth's surface. The process is similar to that of a radar echo, except that an ionospheric route is followed.

The effect is most marked with high-gain directional arrays and high transmitter power. The direction from which scatter signals are observed indicates the region of most intense ionization, and adaptations of radar methods make it possible to "sound" the ionosphere to determine what distances and directions may be covered on a given frequency.

Reflections from Meteor Trails

Probably the least-known means of v.h.f. wave propagation is that resulting from the passage of meteors across the signal path. Reflections from the ionized meteor trails may be noted as a Doppler-effect whistle on the carrier of a signal already being received, or they may cause bursts of reception from stations not normally receivable. Sudden large increases in strength of normally-weak signals are another manifestation of this effect. Ordinarily such reflections are of little value in extending communication ranges, since the increases in signal strength are of short duration, but meteor showers of considerable magnitude and duration may provide fluttery v.h.f.

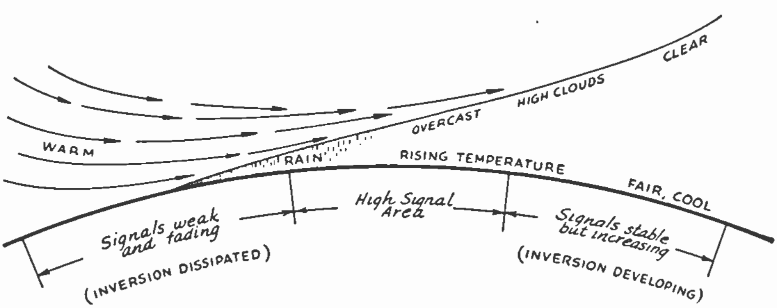


Fig. 15-2 — Illustrating a typical weather sequence, with associated variations in v.h.f. propagation. At the right is a cold air mass (fair weather, high or rising barometer, moderate summer temperatures). Approaching this from the left is a warm moist air mass, which overruns the cold air at the point of contact, creating a temperature inversion and considerable bending of v.h.f. waves. At the left, in the storm area, the inversion is dissipated and signals are weak and subject to fading. Barometer is low or falling at this point.

signals from distances up to 1000 miles or more. Signals so reflected have a combination of the characteristics of aurora and sporadic-*E* skip.

Tropospheric Bending

Refraction of radio waves takes place whenever a change in refractive index is encountered. This may occur at one of the ionized layers of the ionosphere, as mentioned above, or it may exist at the boundary area between two different types of air masses, in the region close to the earth's surface. A warm, moist air mass from over the Gulf of Mexico, for instance, may overrun a cold, dry air mass which may have had its origin in northern Canada. Each tends to retain its original characteristics for considerable periods of time, and there may be a well-defined boundary between the two for as much as several days. When such air-mass boundaries exist along the path between two v.h.f. stations separated by 50 to 300 miles or more, a considerable degree of refraction takes place, and signals run high above the average value. Under ideal conditions there may be almost no attenuation, and signals from far beyond the visual horizon will come through with strength comparable to that of local stations.

Many factors other than air-mass movement of a continental character may provide increased v.h.f. operating range. The convection that takes place along our coastal areas in warm weather is a good example. The rapid cooling of the earth after a hot day in summer, with the air aloft cooling more slowly, is another, producing a rise in signal strength in the period around sundown. The early-morning hours, when the sun heats the air aloft, before the temperature of the earth's surface begins its daily rise, may frequently be the best hours of the day for extended v.h.f. range, particularly in clear, calm weather, when the barometer is high and the humidity low.

Any weather condition that produces a pronounced boundary between air masses of different temperature and humidity characteristics provides the medium by which v.h.f. signals cover abnormal distances. The ambitious v.h.f. enthusiast soon learns to correlate various weather manifestations with radio-propagation phenomena. By watching temperature, barometric pressure, changing cloud formations, wind direction, visibility, and other easily-observed weather signs, he is able to tell with a reasonable degree of accuracy what is in prospect on the v.h.f. bands.

The responsiveness of radio waves to varying weather conditions increases with frequency. Our 50-Mc. band is considerably more sensitive to weather variations than is the 28-Mc. band, and the 144-Mc. band may show strong signals from far beyond visual distances when the lower frequencies are relatively inactive. The maximum distance over which tropospheric propagation is frequently observed on 50 Mc. is in the neighborhood of 300 miles. On 144 Mc. distances of 500 miles are not uncommon. It is probable that this tendency continues on up through the microwave range, and that our assignments in the u.h.f. and s.h.f. portions of the frequency spectrum may someday support communication over distances far in excess of the optical range. Already 144-Mc. tropospheric communication by amateurs has passed the 1100-mile mark, and even greater distances are believed possible on this and higher frequencies.

STATION LOCATIONS

In line with our early notions of v.h.f. wave propagation, it was once thought that only highly-elevated v.h.f. stations had any chance of working beyond a few miles. Almost all the work was done by portable stations operating from mountain tops, and only hilltop home sites were considered suitable for fixed-station work. It is still true that the fortunate amateur who lives at the top of a hill enjoys a certain advantage over his fellows on the v.h.f. bands, but high elevation is not the all-important factor it was once thought to be.

Improvements in equipment, the wide use of high-gain antenna systems, and an awareness of the opportunities afforded by weather phenomena have enabled countless v.h.f. workers to achieve excellent results from seemingly poor locations. In 50-Mc. DX work particularly, elevation has ceased to be an important factor, though it may help in extending the range of operation somewhat under normal conditions. A high elevation is somewhat more helpful on 144 Mc. and higher frequencies, particularly when no unusual propagation factors are present, as during the winter months. Other factors, such as close proximity to large bodies of water, may more than compensate for lack of elevation during the other seasons of the year, however.

Stations situated in sea-level locations along our coasts have been consistent in their ability to work long distances on 144 Mc.; weather variations provide interesting propagation effects over our Middle Western plain areas; and even the worker situated in mountainous country need not necessarily feel that he is prevented by the nature of his horizon from doing interesting work. Contacts have been made on 50 and 144 Mc. over distances in excess of 100 miles in all kinds of terrain.

The consistently-reliable nature of 50 and 144 Mc. for work over such a radius and more, regardless of weather, time or season, and the occasional opportunities these frequencies afford for exciting DX, have caused an increasing number of amateurs to migrate to the v.h.f. bands for extended-local communication, once thought possible only on the lower frequencies.

V.H.F. Receivers

Even more than in work on lower frequencies, receiver performance is all-important in the v.h.f. station. High sensitivity and good signal-to-noise ratio, necessary attributes in a receiving system for 50 Mc. and higher bands, are best attained through the use of a converter, working in conjunction with a communications receiver designed for lower frequencies. Though receivers and converters for 50, 144, and even 220 Mc. are available on the amateur market, it is possible for the v.h.f. worker to build his own with fully as good results, and at a considerable saving in cost.

In its basic principles, modern receiving equipment for these bands differs little from that employed on lower frequencies, and the same order of selectivity may be used in amateur work up to at least 450 Mc. The greatest practical selectivity should be used in v.h.f. work, as well as on the frequencies below 30 Mc., as it not only permits more stations to operate in a given band, but is an important factor in improving the signal-to-noise ratio. The effective sensitivity of a receiver having "communication" selectivity can be made considerably better than is possible with broadband systems. First on 56 Mc., more than a decade ago, then more recently on 144 Mc., and currently on 220 and 420 Mc., the change to selective superheterodyne receivers marked the beginning of real extensions of the operating range.

The superregenerative receiver, once very popular for v.h.f. work, is now used principally for portable operation, or for other applications where maximum sensitivity and selectivity are not of prime importance. It is still capable of surprising performance, for a given number of tubes and components, but its lack of selectivity, its poor signal-to-noise ratio, and its tendency to radiate a strong interfering signal rule out the superregenerator as a fixed-station receiver in areas where there is appreciable v.h.f. activity.

● R.F. AMPLIFIER DESIGN

The amount of noise generated within the receiver itself is an important factor in the effectiveness of v.h.f. receiving gear. At lower frequencies the external noise is a limiting factor, but at 50 Mc. and higher the receiver noise figure, gain and selectivity determine the ability of the system to respond to weak signals. Proper selection of r.f. amplifier tubes and appropriate circuit design aimed at low noise figure are of more importance in the v.h.f. receiver "front end" than mere gain.

Certain triode or triode-connected pentode tubes have been found superior in this respect, their superiority becoming more pronounced as we go higher in frequency. At 144 Mc., for instance, a triode r.f. stage may give substantially the same gain as a pentode, but with a much lower noise figure. With the exception of the simplest unit, the equipment described in the following pages incorporates low-noise r.f. amplifier technique.

When triodes are used as r.f. amplifiers some form of neutralization of the grid-plate capacitance is required. This can be capacitive, as is commonly used in transmitting applications,

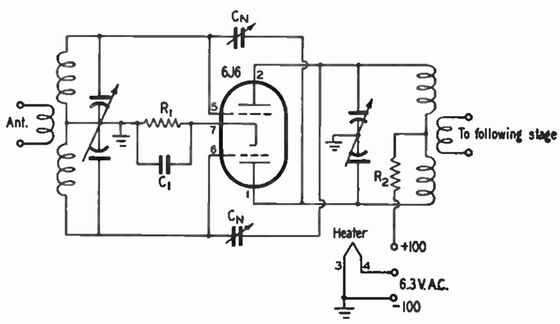


Fig. 16-1 — **Schematic diagram of a push-pull r.f. amplifier for v.h.f. receiver use. This circuit is well suited to use with antenna systems fed by balanced lines. Coil and condenser sizes will be governed by the band for which the amplifier is to be used.**

C_1 — 0.005-μfd. disc ceramic.

C_N — Neutralizing capacitance, about 2 $\mu\mu$fd. May be made from lengths of 75-ohm Twin-Lead about 1½ inches long.

R_1 — 150 ohms, ½-watt carbon.

R_2 — 1000 ohms, ½-watt carbon.

or inductive. The alternative to neutralization is the use of grounded-grid technique. Circuits for v.h.f. triode r.f. amplifier stages are given in Figs. 16-1 through 16-4.

A dual triode operated as a neutralized push-pull amplifier is shown at 16-1. This arrangement is well adapted to v.h.f. preamplifier applications, or as the first stage in a converter, particularly when a balanced transmission line such as the popular 300-ohm Twin-Lead is used. It is relatively selective

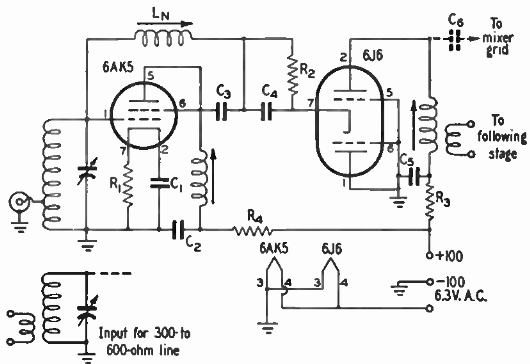


Fig. 16-2 — Circuit of the cascode r.f. amplifier. Preferred antenna coupling methods for coaxial or balanced lines are shown. The first r.f. grid coil, and the neutralizing coil, L_N, should be a high-Q design. Other coils are not critical as to Q.

C_1, C_2, C_4, C_5 — 0.005-μfd. disc ceramic.
C_3, C_6 — 50-$\mu\mu$fd. ceramic.
R_1, R_2 — 100 ohms, ½-watt carbon.
R_3, R_4 — 1000 ohms, ½-watt carbon.
L_N — Should resonate at signal frequency with 6AK5 grid-plate capacitance.

and may require resistive loading of the plate circuit, when used as a preamplifier. The loading effect of the following circuit may be sufficient to give the required bandwidth, when the push-pull stage is inductively coupled to the mixer.

A two-stage triode amplifier having excellent noise figure and broadband characteristics is shown in Fig. 16-2. Commonly called the cascode, it uses a triode or triode-connected pentode followed by a triode grounded-grid stage. This circuit is extremely stable and uncritical in adjustment. At 50 Mc. and higher its over-all gain is at least equal to the best single-stage pentode amplifier and its noise figure is far lower.

Neutralization is accomplished by the coil L_N, whose value is such that it resonates at the signal frequency with the grid-plate capacitance of the tube. Its inductance is not critical; it may be omitted from the circuit without the stage going into oscillation, but neutralization results in a lower noise figure than is possible without it. Any of several v.h.f. tubes may be used in the cascode circuit, the most popular arrangement being the 6AK5-6J6 combination, Fig. 16-2.

A simplified version of the cascode, using a dual triode tube designed especially for this application, is shown in Fig. 16-3. By reducing stray capacitance through direct coupling between the two triode sections, this circuit makes for improved performance at the frequencies above 100 Mc. The two sections of the tube are in series, as far as plate voltage is concerned, so it requires higher voltage than the other circuits shown.

The neutralization process for the cascode and neutralized-triode amplifiers is somewhat similar. With the circuit operating normally the neutralizing adjustments (capacitance of C_N in Fig. 16-1; inductance of L_N in Figs. 16-2 and 16-3) can be set for best signal-to-noise ratio. The middle of the range over which no oscillation occurs is approximately the proper setting. Finer adjustment can be made by disconnecting one heater lead from the r.f. amplifier tube socket and adjusting the neutralizing for *minimum* signal. The best results are obtained using a noise generator, adjusting for lowest noise figure, but the two methods described above will provide a fair approximation. Noise generators and their use in v.h.f. receiver adjustment are treated in July, 1953, *QST*, page 10.

Grounded-grid r.f. amplifier technique is illustrated in Fig. 16-4. Here the input circuit is connected in the cathode lead, with the grid of the tube grounded, to act as a shield between cathode and plate. The grounded-grid circuit is stable and easily adjusted, and is well adapted to broadband applications. The gain per stage is low, so that two or more stages are ordinarily required.

Choice of tubes is fairly limited, the best for the job being the 6J4, 6AN4, 6AJ4 and 6AM4, triodes especially designed for grounded-grid service. The 6J6 is used occasionally, as in Fig. 16-2. Disc-seal tubes such as the "lighthouse" and "pencil tube" types are often used as r.f. amplifiers above 500 Mc., where most miniature tubes become ineffective because of excessive lead inductance.

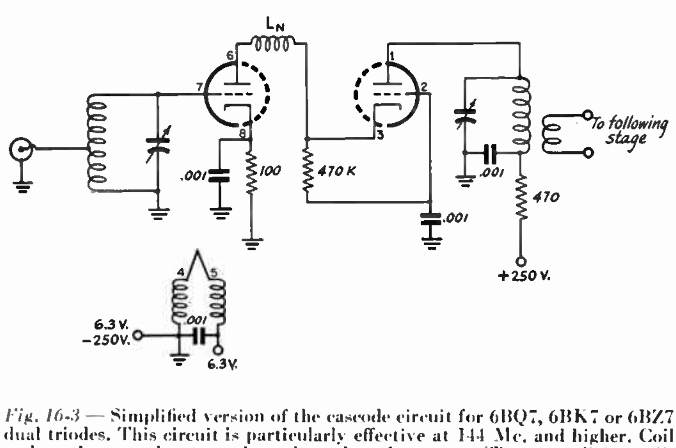


Fig. 16-3 — Simplified version of the cascode circuit for 6BQ7, 6BK7 or 6BZ7 dual triodes. This circuit is particularly effective at 144 Mc. and higher. Coil and condenser values not given depend on frequency. The neutralizing coil, L_N, should resonate at the signal frequency. R.f. chokes in the heater circuit should be resonant with the plate-to-ground capacitance of the first triode section, at the highest frequency to be covered. They are bifilar wound.

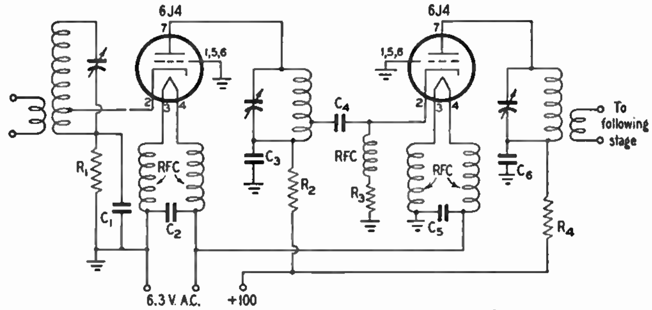


Fig. 16-4 — Grounded-grid r.f. amplifier. Position of cathode taps on coils should be adjusted for lowest noise figure.

C_1, C_2, C_3, C_5, C_6 — 0.005-μfd. disc ceramic.
C_4 — 50-μμfd. ceramic.
R_1, R_3 — 220 ohms, ½-watt carbon.
R_2, R_4 — 470 ohms, ½-watt carbon.

MIXER CIRCUITS

Triode tubes are favored for v.h.f. applications, as they are less critical as to operating conditions and the highest frequency at which they will operate satisfactorily is well above that of most pentodes. When used in converters having no r.f. amplifier stage triodes are usually quieter in operation as well.

A simple triode mixer circuit is shown in Fig. 16-5A. The grid circuit is tuned to the signal frequency, the plate circuit to the intermediate frequency. A dual-triode version is given at B. The latter is particularly suitable for use at the higher frequencies. Frequently a dual triode is used as a combination mixer-oscillator, using the circuits of Figs. 16-5A and 16-6A. The amount of oscillator injection is usually not critical, but in the interest of stability it should be kept as low as practical. In dual triodes having separate cathodes (7F8, 12AT7, 2C51, etc.) some external coupling may be required, but the common cathode of the 6J6 will provide sufficient injection in most cases. If the injection is more than necessary it can be reduced by dropping the oscillator plate voltage, either directly or by increasing the value of the dropping resistor, R_1.

A pentode mixer may be less subject to oscillator pulling than a triode, and it will probably require less injection voltage. If a pentode mixer is used, its plate current should be held to the lowest usable value, to reduce tube noise. This may be controlled by varying the mixer screen voltage. The principal use of pentode mixers in v.h.f. work is in the interest of simplicity of circuit layout, as in multiband converters employing bandswitching.

Occasionally oscillation near the signal frequency may be encountered in v.h.f. mixers. This usually results from stray lead inductance in the mixer plate circuit, and is most common with triode mixers. It may be corrected by connecting a small capacitance from plate to cathode, *directly* at the tube socket. Ten to 25 μμfd. will be sufficient, depending on the signal frequency.

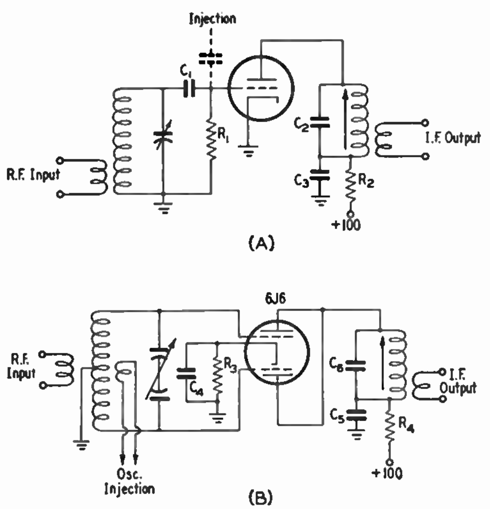


Fig. 16-5 — Two types of triode mixers suitable for v.h.f. receivers. A single-ended triode circuit is shown at A. The tube may be half of a dual triode, with the other portion used as the oscillator, or separate tubes may be used. The dual-triode version, B, is particularly useful for 144 Mc. and higher bands.

C_1 — 50-μμfd. ceramic or mica.
C_2, C_6 — 30- to 50-μμfd. ceramic or mica.
C_3, C_4, C_5 — 0.005-μfd. disc ceramic.
R_1 — 1 megohm, ½ watt.
R_2, R_4 — 1000 ohms, ½ watt.
R_3 — 150 ohms, ½ watt.

OSCILLATOR STABILITY

When a high-selectivity i.f. system is employed in v.h.f. reception, the stability of the oscillator is extremely important. Slight variations in oscillator frequency that would not be noticed when a broadband i.f. amplifier is used become intolerable when the passband is reduced to crystal-filter proportions.

One satisfactory solution to this problem is the use of a crystal-controlled oscillator, with frequency multipliers if needed, to supply the injection voltage. Such a converter usually employs one or more broadband r.f. amplifier stages, and tuning is done by varying the intermediate frequency to cover the desired frequency range.

When a tunable oscillator and a fixed intermediate frequency are used, special attention must be paid to the oscillator design, to be sure that it is mechanically and electrically stable. The tuning condenser should be solidly built, preferably of the double-bearing type. Split-stator condensers specifically designed for v.h.f. service, usually having ball-bearing end plates and special construction to insure short leads, are well worth their extra cost. Leads should be made with stiff wire, to reduce vibration effects. Mechanical stability of air-wound coils can be improved by tying the turns together with narrow strips of household cement at several points.

Recommended oscillator circuits for v.h.f. work are shown in Fig. 16-6. The single-ended oscillator may be used for 50 or 144 Mc. with good results. The push-pull version is recommended for higher frequencies and may also be used on the two lower bands, as well. Circuit A works well with almost any small triode, the 6AB4, 6AF4 or one half of a 6J6 or 12AT7 being most commonly used. The 6J6 is well suited to push-pull applications, as shown in circuit 16-6B.

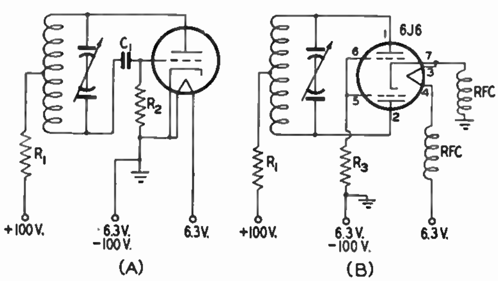


Fig. 16-6 — **Recommended circuits for v.h.f. oscillators. The push-pull arrangement at B is recommended for 220 and 420 Mc., particularly.**

C_1 — 50 μμfd.
R_1 — Any small carbon resistor, 1000 ohms or less.
R_2 — 10,000 ohms, ½ watt.
R_3 — 3000 to 5000 ohms, ½ watt.

THE I.F. AMPLIFIER

Superheterodyne receivers for 50 Mc. and up should have fairly high intermediate frequencies, to reduce both oscillator pulling and image response. Approximately 10 per cent of the signal frequency is commonly used, with 10.7 Mc. being set up as the standard i.f. for commercially-built FM receivers. This particular frequency has a disadvantage for 50-Mc. work, in that it makes the receiver subject to image response from 28-Mc. signals, if the oscillator is on the low side of the signal frequency. A spot around 7 Mc. is favored for amateur converter service, as practically all communications receivers are capable of tuning this range.

For selectivity with a reasonable number of i.f. stages, double conversion is usually employed in complete receivers for the v.h.f. range. A 7-Mc. intermediate frequency, for instance, is changed to 455 kc., by the addition of a second mixer-oscillator. This procedure is, of course, inherent in the use of a v.h.f. converter ahead of a communications receiver.

If the receiver so used is lacking in sensitivity, the over-all gain of the converter-receiver combination may be inadequate. This can be corrected by building an i.f. amplifier stage into the converter itself. Such a stage is useful even when the gain of the system is adequate without it, as the gain control can be used to permit operation of the converter with receivers of widely-different performance. If the receiver has an S-meter, its adjustment may be left in the position used for lower frequencies, and the converter gain set so as to make the meter read normally on v.h.f. signals.

Where reception of wide-band FM or unstable signals of modulated oscillators is desired, a converter may be used ahead of an FM broadcast receiver. A superregenerative detector operating at the intermediate frequency, with or without additional i.f. amplifier stages, also may serve as an i.f. and detector system for reception of wide-band signals. By using a high i.f. (10 to 30 Mc. or so) and by resistive loading of the i.f. transformers, almost any desired degree of bandwidth can be secured, providing good voice quality on all but the most unstable signals. Any of these methods may be used for reception in the microwave region, where stabilized transmission is extremely difficult at the current state of the art.

THE SUPERREGENERATIVE RECEIVER

The simplest type of v.h.f. receiver is the superregenerator. It affords fair sensitivity with few tubes and elementary circuits, but its weaknesses, listed earlier, have relegated it to applications where small size and low power consumption are important considerations.

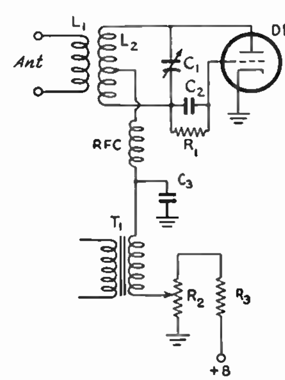


Fig. 16-7 — **Superregenerative detector circuit using a self-quenched detector. L_2C_1 tunes to the signal frequency. Typical values for other components are given below.**

C_2 — 47 μμfd.
C_3 — 0.001 to 0.005 μfd.
R_1 — 2 to 10 megohms.
R_2 — 50,000-ohm potentiometer.
R_3 — 47,000 ohms, 1 watt.
RFC — Single-layer r.f. choke, for frequency involved.
T_1 — Interstage audio transformer.

Its sensitivity results from the use of an alternating quenching voltage, usually in the range between 20 and 200 kc., to interrupt the normal oscillation of a regenerative detector. The regeneration can thus be increased far beyond the amount usable in a straight regenerative circuit. The detector itself can be made to furnish the quenching voltage, or a separate oscillator tube can be used. Regeneration is usually controlled by varying the plate voltage in triode detectors, or the screen voltage in the case of pentodes. A typical circuit is shown in Fig. 16-7.

Crystal-Controlled Converters for 50, 144 and 220 Mc.

The family of converters shown in Figs. 16-8 through 16-16 was designed to provide optimum reception on all v.h.f. bands. Crystal-controlled injection is used to insure stability, and the r.f. circuit design provides the lowest practical noise figure for each frequency. Special attention has been paid to the reduction of spurious responses, often a troublesome point in broadband converter design. A separate converter section for each band connects to a common i.f. amplifier and power supply by means of a single plug and cable. This carries the mixer output, and plate and filament voltages.

The R.F. Circuits

A pentode r.f. amplifier (6CB6) is used in the 50-Mc. converter in the interest of simplicity. With proper design, such a stage can be made to deliver a satisfactory noise figure at 50 Mc. Its performance is quite adequate; it will be found that outside noise picked up by the antenna will be the limiting factor in weak-signal reception, even in a quiet receiving location.

The 144- and 220-Mc. converters have modified cascode circuits with dual triodes (6BQ7A, 6BK7 or 6BZ7) in the first stages. The 220-Mc. converter has an additional pentode stage, to build up the gain and improve the ability of the converter to reject unwanted frequencies. It will be noted that the converters differ somewhat as to circuitry in other respects, but this was done primarily to show examples of various circuit techniques, rather than because of any superiority of one approach over another. This applies particularly to the methods of coupling between stages.

When a fixed injection frequency is used with a variable intermediate frequency, the r.f. and i.f. circuits of the converter must be made broadband, to avoid the need for readjusting them as the receiver with which the converter is used is tuned across the i.f. range. Spurious responses, both at the i.f. range and at frequencies adjacent to the desired signal frequencies, pose a special problem. Bandpass characteristics are attained through the use of overcoupled double-tuned circuits in the converter r.f. circuits. These circuits present a high impedance at the signal frequency, but they look like a short circuit to signals in the i.f. range that are picked up by the antenna.

Spurious responses that might develop as the result of the injection of unwanted frequencies at the mixer grid are reduced by the use of a separate tube for the mixer, and coupling the injection voltage from the multiplier stage through a link. Isolation of the mixer and multiplier stages is further increased in the 144- and 220-Mc. converters by the installation of a shield partition along the middle of the base plate.

Crystal Oscillator Details

Crystal frequencies were selected so that all bands would start at the same spot on the communications receiver dial; in this case 7000 kc. Crystal frequencies, multiplier details and i.f. tuning ranges are shown in Table 16-I. Other i.f. tuning ranges that may be better suited to some communications receivers may be employed by suitable alteration of the crystal and multiplier frequencies.

A fairly high oscillator frequency is desirable, to reduce the possibility of oscillator harmonics appearing in the tuning range, as well as to keep down the number of multiplier stages. Each con-

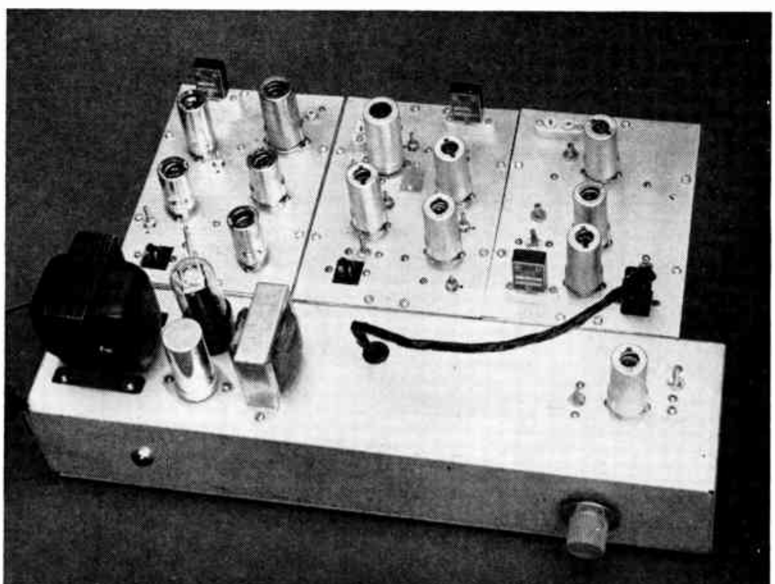


Fig. 16-8 — Crystal-controlled converters for 220, 144 and 50 Mc. (*l. to r.*) with their common i.f. amplifier and power supply. All chassis are standard sizes, requiring a minimum of metal work.

verter in this series uses a readily-obtainable crystal operating on its third overtone. This may result in a frequency of oscillation that is not exactly three times that marked on the crystal, but it is close enough for ordinary calibration purposes. Overtone crystals of the desired frequency may be obtained on order, at somewhat higher prices than for fundamental-type crystals. Conventional operation of crystals in the 7-Mc. range, making up the multiplication with additional stages, is not recommended because of the difficulty in avoiding birdies from crystal harmonics. In the overtone circuit, no frequency lower than the overtone at which the crystal oscillates is heard.

Layout

Each converter is built on a single 5 × 7-inch aluminum plate, and mounted on a standard chassis that serves as shielding and case. The three 5 × 7 × 3-inch chassis are bolted to the back of the i.f. unit, to be described later. In this way each converter is a separate entity, permitting the constructor to build any one of them, omitting those bands in which he may not be interested. The shape of the i.f. unit is not important, and it could very readily be built in more compact fashion if less than the three converters are planned. The method of construction shown requires a minimum of metal work, and a converter can be rebuilt or replaced without affecting the operation of the others.

As only three tubes are used in the 50-Mc. converter they are arranged in a single line down the middle of the base plate. The other models have the oscillator-multiplier and amplifier-mixer sections separated by a vertical shield partition.

● THE 50-MC. CONVERTER

The simplest of the three converters is the 50-Mc. unit, shown in Figs. 16-9 and 16-10. The r.f. and mixer stages use 6CB6 pentodes and a 6J6 serves as crystal oscillator and multiplier. A somewhat lower noise figure could have been obtained with a triode r.f. amplifier, but the design shown has a noise figure under 5 db. With the considerable external noise picked up by the antenna at 50 Mc., even in a quiet location, there is little to be gained in weak-signal reception by going lower than this figure.

The bottom view of the converter, Fig. 16-9, shows the r.f. amplifier socket and components at the left side. A small shield across the socket isolates the grid and plate circuits. The r.f. plate tuning condenser, C_2, is near the center. The plate coil, L_3, is the lower of the two coils in the middle of the photograph, with the mixer grid coil, L_4, just above it. An enameled-wire link may be seen running from this coil to the doubler plate coil, L_{10}, at the lower right. The oscillator inductance, L_9, is at the upper right corner.

Two methods of antenna coupling are shown in the schematic, Fig. 16-10, but the constructor need install only the one that is suited to the type of transmission line he intends to use to feed his antenna system. If coax is used, connection is made directly to the r.f. amplifier grid coil, L_2. This same type of connection may be used with a balun for balanced lines, or the coupling winding, L_1, may be added. In some instances it may be desirable to connect a trimmer between J_1 and L_2, as shown in the 220-Mc. converter, if spurious signals are a problem.

TABLE 16-I
Crystal-Controlled Converter Data

Band (Mc.)	*Injection (Mc.)*	*I.F. (Mc.)*	*Crystal (kc.)*	*Overtone & Multiplication*
50	43	7–11	7166	3rd × 2
144	137	7–11	7611	3rd × 3 × 2
220	213	7–12	7100	3rd × 5 × 2
*420**	382	50–54*	7071	3rd × 3 × 3 × 3 × 2
*420**	406	26–30*	7518	same

* For covering 432 to 436 Mc. only. To tune the rest of the band additional crystal frequencies or a wider i.f. tuning range must be used.

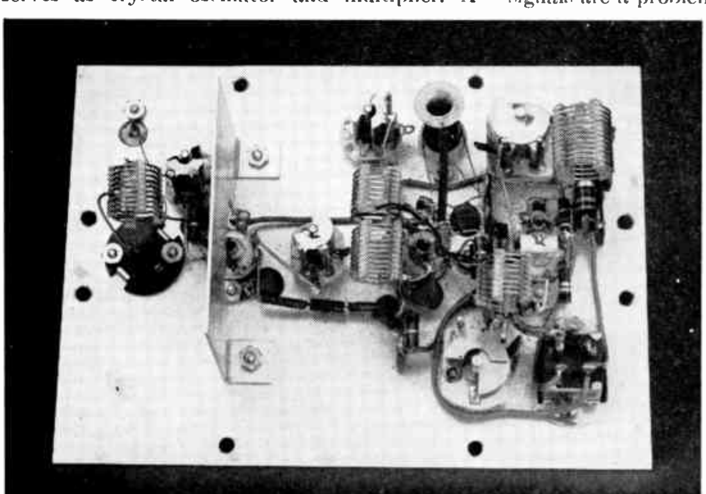


Fig. 16-9 — Bottom view of the 50-Mc. converter. The r.f. amplifier socket, divided by a shield partition, is at the left. Crystal oscillator and multiplier components are at the right, with the mixer in the middle.

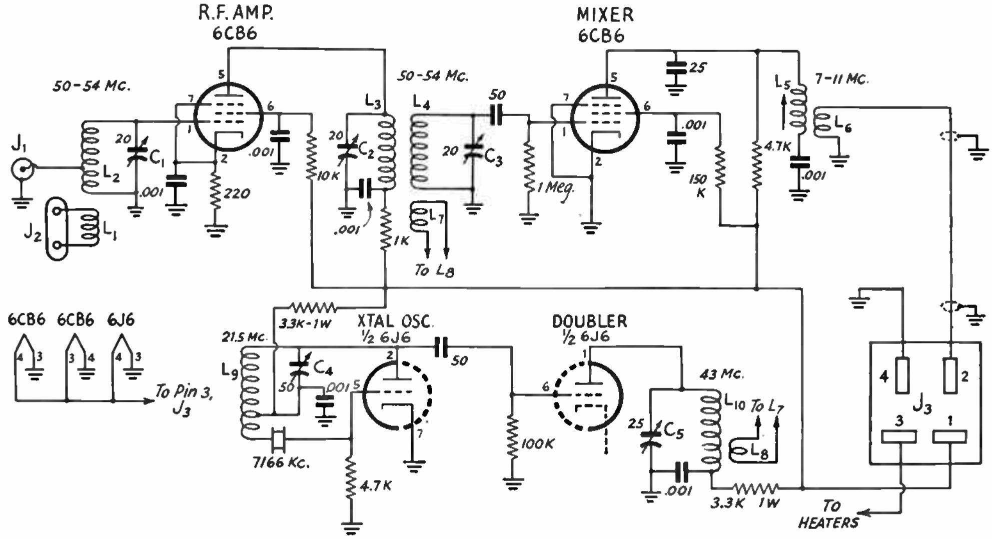


Fig. 16-10 — Schematic diagram of the 50-Mc. crystal-controlled converter.

C_1, C_2, C_3 — 20-μμf. min. variable (Johnson 20M11).
C_4 — 50-μμf. min. padder (Hammarlund MAPC 50).
C_5 — 25-μμf. min. padder (Hammarlund MAPC 25).
L_1 — 3 turns fine ins. wire wound over cold end of L_2.
L_2, L_4 — 9 turns No. 20 tinned, ½-inch diam., 11/16 inch long (B & W Miniductor No. 3003).
L_3 — 10½ turns similar to L_3. These coils are mounted in line with their cold ends ⅛ inch apart.
L_5 — No. 28 enameled wire close-wound one inch on ⅜-inch slug-tuned form (National XR-91). Lacquer and dry before winding L_6. Wind on upper portion of form.
L_6 — 10 turns same wound over cold end of L_5.
L_7, L_8 — Loop of No. 22 enameled wire inserted in cold ends of L_4 and L_{10}, connected by link of same material. Fasten in place with cement.
L_9 — 13 turns No. 20 tinned, ⅝-inch diam., ¾ inch long, tapped at 3½ turns from crystal end (B & W No. 3007).
L_{10} — 8 turns similar to L_2.
J_1 — Coaxial fitting.
J_2 — Crystal socket for antenna terminal.
J_3 — 4-pin male chassis fitting (Jones P-304-AB).

Adjustment of the converter is very simple. First the oscillator and multiplier are tuned up, with the r.f. and mixer tubes out of their sockets, or with their plate voltage removed. Proper adjustment of the overtone oscillator follows practice outlined in the introductory portion of Chapter Seventeen, and the doubler portion need only be resonated for maximum output initially. This can be checked with a 60-ma. pilot lamp connected across a one-turn loop coupled to the cold end of L_{10}. The frequency of the output should be checked to be sure that the right overtone and harmonic are being used, and the oscillator tested to see that it is controlled by the crystal.

Now a signal source will be helpful. This can be a signal generator, an amateur signal, or the harmonic of a receiver or transmitter oscillator of known frequency. If the signal is derived locally it should be possible to hear it with only the mixer and oscillator-multiplier stages running, and with no pick-up antenna. If a weak signal is used it may be necessary to put a temporary coupling winding (similar to L_1) on the mixer grid coil, L_4. Peak this circuit and the slug in the mixer plate circuit for maximum response. The plate voltage should be removed from the r.f. stage during this period, but the tube should be left in the socket with the heater voltage on.

Next feed the signal into the r.f. stage, by either of the coupling methods shown, and peak L_2 and L_3 for maximum response. There should be a considerable rise in noise as the adjustments are made, so the noise level can be used as an indication of resonance in the absence of a test signal.

The converter is now ready for final adjustment, for best signal-to-noise ratio and uniform response across the band. The first can best be done with a noise generator, though a test signal can be used. Noise figure will be affected principally by the tuning of the first stage, and by the adjustment of the antenna coupling. Watch for improvements in the margin of signal over noise, rather than maximum gain, as these two characteristics may not occur coincidentally. The coupling between L_3 and L_4 affects the passband of the system and the tuning of these circuits and the slug in the mixer plate winding can be staggered to provide uniform response across the band. Peaking of the input circuit may be necessary as the receiver is tuned across the entire band, though a setting can be made for the middle of the range most used and this will hold for at least a megacycle either way. Receiver noise can be used as a check on the uniformity of response, in the absence of signals.

The amount of injection from the multiplier should be set at the least that will provide satisfactory performance. This will not be at all critical, but more injection than needed will increase the tendency to spurious response. It is controlled by the size and position of the coupling loops, L_7 and L_8. In the original model they are about two-thirds the diameter of the windings in which they

CHAPTER 16

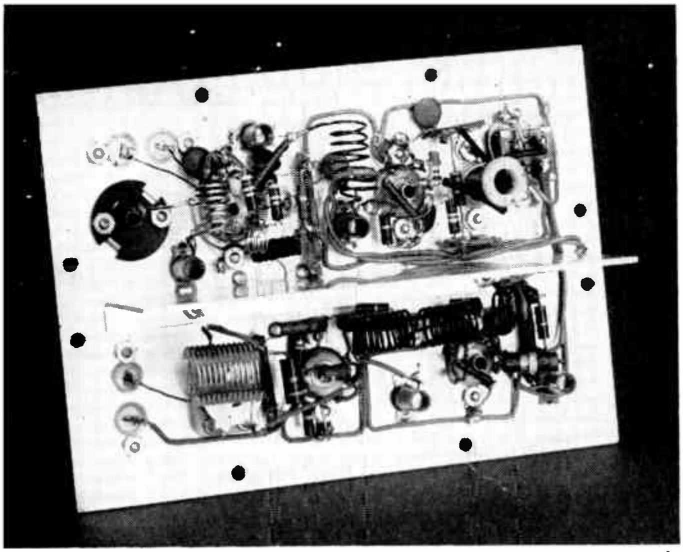


Fig. 16-11 — The 144-Mc. converter is separated into two parts by a shield partition. At the top are the r.f. and mixer stages, with the oscillator and multiplier portion below the shield.

are inserted. The loop can be made small enough to slip through between the strips of polystyrene on the Miniductor, and then spread to give the desired coupling. Cement the loops in place when this is achieved.

THE 144-MC. CONVERTER

The 2-meter converter is shown in Figs. 16-11 and 16-12. From the photograph it may be seen that the r.f. and mixer components are separated from the oscillator-multiplier chain by a shield partition. The r.f. portion is in the upper half of the picture. Use of small plastic trimmers for the tuned circuits saves enough space so that the additional tube is handled without crowding.

The r.f. circuit is the simplified cascode, using any of the several dual triodes designed for this application. Double-tuned circuits in the r.f. plate and mixer grid provide bandpass response

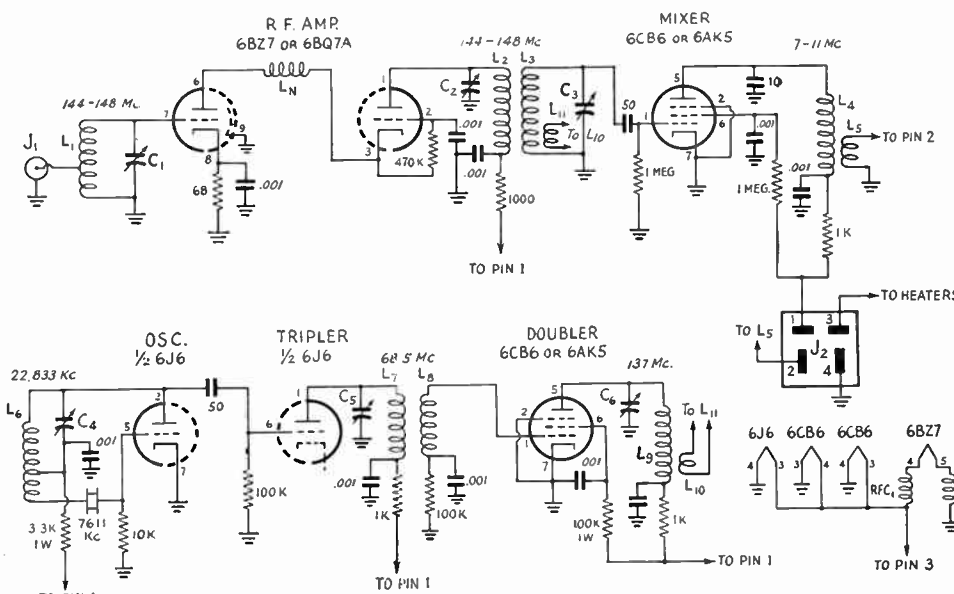


Fig. 16-12 — Schematic diagram and parts information for the 144-Mc. converter.

C_1, C_2, C_3, C_5, C_6 — 1- to 8-μμf. plastic trimmer (Erie 832-10).

C_4 — 50-μμf. min. trimmer (Hammarlund MAPC-50).

L_N — 5 turns No. 20 tinned, ¼-inch diam. Adjust spacing for neutralizing: see text.

L_1 — 6 turns No. 20 tinned, ¼-inch diam., turns spaced diam. of wire. Tap at 2½ turns.

L_2 — 4 turns No. 20 enam. ⅜-inch diam., ⅜ inch long.

L_3 — 3 turns. No. 20 enam., ⅜-inch diam., 5/16 inch long. L_2 and L_3 are in line, with their cold ends ⅛ inch apart.

L_4 — No. 28 enam. close wound 1 inch on ⅜-inch slug-tuned form (National XR-91). Lacquer and dry before winding L_5. Wind on upper portion of form.

L_5 — 10 turns, same, wound over cold end of L_4.

L_6 — 12 turns No. 20 tinned, spaced diam. of wire, ⅝-inch diam. Tap at 3½ turns.

L_7 — 11 turns No. 20 enam., ⅜-inch diam., ¾ inch long.

L_8 — 8 turns like L_7, ⅝ inch long. L_7 and L_8 are in line with their cold ends 3/16 inch apart.

L_9 — 4 turns like L_7, ⅜ inch long.

L_{10}, L_{11} — 1 turn insulated wire at each end, linking L_3 with L_9.

J_1 — Coaxial fitting.

J_2 — 4-pin male chassis fitting (Jones P-304-AB).

RFC_1, RFC_2 — Bifilar-wound r.f. chokes. Twist two pieces of No. 26 enameled wire together and wind 15 turns on ¼-inch diameter.

and help to attenuate unwanted signals on other frequencies. The oscillator-multiplier circuit is similar to the 50-Mc. converter, except that the second half of the 6J6 is a tripler. This is coupled through another pair of double-tuned circuits to an additional doubler stage.

The order of frequency multiplication can be altered to take care of local interference conditions. Should it turn out that unwanted signals are brought in as a result of frequencies appearing in the multiplier chain, the second stage can be made a doubler and the pentode a tripler. The use of link coupling, and the isolation afforded by the shield, should reduce spurious responses to negligible proportions in most locations, however.

The first steps in adjustment of the 144-Mc. converter are similar to those outlined for the 50-Mc. model. The only additional work required is the neutralization of the 6BQ7 stage. This is done by adjusting the spacing of the turns in L_N for lowest noise figure, as indicated with a noise generator, or by best signal-to-noise ratio on a test signal. The inductance is not extremely critical, and it may be set somewhat on the low-inductance side of the largest value that can be used without oscillation developing in the r.f. stage.

Other than the neutralization, only the tuning of the input circuit will affect the noise figure materially. This is also best done with a noise generator. It will be found that best results will be obtained with L_1C_1 resonated somewhat on the low-frequency side of the point that produces maximum gain. The tap on L_1 should be set higher on the coil than the point that gives maximum signal response. The objective, as in the other adjustments outlined above, is best signal-to-noise ratio, rather than maximum gain.

Uniform response across the band can be attained by stagger-tuning the r.f. plate, mixer grid and mixer plate circuits. Injection coupling should be set as low as will deliver optimum performance. This can be controlled by the position of the coupling loops, L_{10} and L_{11}, or by varying the output of the pentode stage by raising or lowering the value of the screen dropping resistor.

THE 220-MC. CONVERTER

Circuitry and layout for the 220-Mc. converter, Figs. 16-13 and 16-14, are very similar to the 144-Mc. model, except that an additional stage is used following the cascode, and an additional shield divides the socket of this stage. This helps to make up for the somewhat lower gain of the cascode at the higher frequency, and it improves the rejection of unwanted signals considerably. The latter condition has been found to be troublesome in 220-Mc. work, particularly in areas where TV and FM broadcasting stations are in operation.

No tuning condensers are used in the r.f. circuits, the coils being tuned to the desired frequency by adjusting the turn spacing until they resonate properly with the tube capacitances that appear across them. A variation on the double-tuned circuit is used in which a center-tapped coil serves as both grid and plate inductance. This type of circuit is well adapted to use at frequencies where tube capacitance becomes a limiting factor in the performance of r.f. amplifiers.

A different form of i.f. output coupling is shown in this converter, though it works identically to the method used in the other models. Note that the mixer plate coil is loaded by a 4700-ohm resistance in this case. The i.f. must cover from 7 to 12 Mc. for the 220-Mc. band, so a broader response is required. The value of this resistance can be altered to attain the desired degree of uniformity, though lower values than the one shown will result in lower over-all gain.

The tuning condenser in the input circuit tunes out the reactance of the line to the antenna. It may not be necessary in some installations, but it is likely to be helpful in reducing spurious responses. The same technique may also be applied

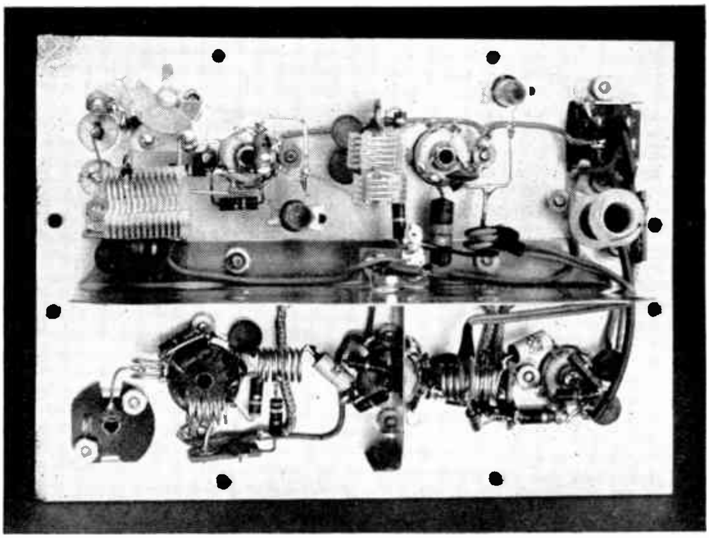


Fig. 16-13 — The 220-Mc. crystal-controlled converter. Note that two shields are used; one separating the injection and r.f. chains, the other dividing the socket for the 6AK5 r.f. stage. R.f. components occupy the lower half of the assembly.

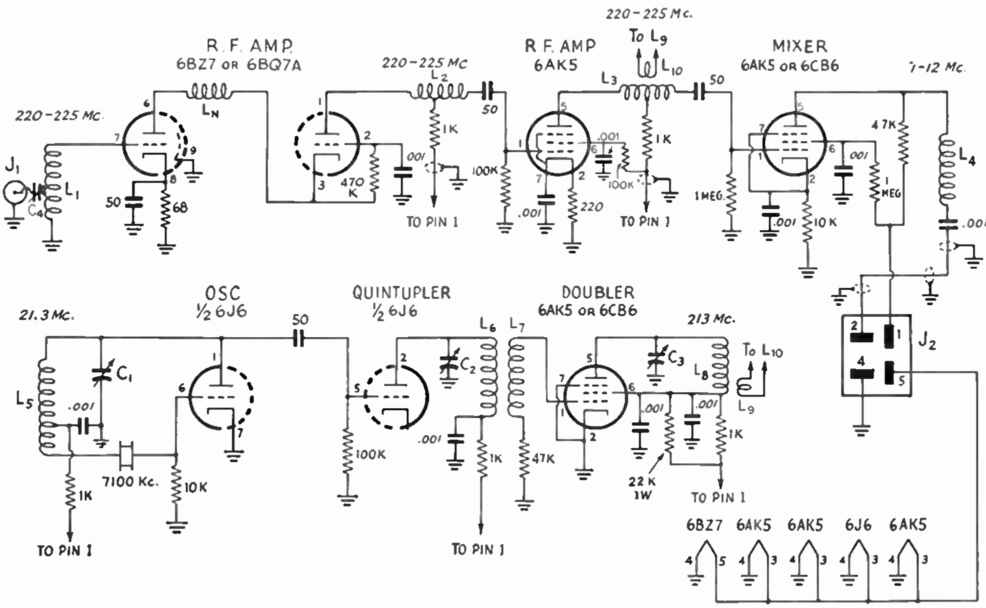


Fig. 16-14 — Schematic diagram and parts information for the 220-Mc. converter.

C1 — 50-μμf. miniature variable (Hammarlund MAPC-50).
C2 — 8-μμf. plastic trimmer (Erie 532-10).
C3 — 5-μμf. plastic trimmer (Erie 532-08-OR5).
L1 — 3 turns No. 20 tinned, ¼-inch diam., ¼ inch long, center tapped.
LN — 5 turns No. 20 tinned, ¼-inch diam. Adjust spacing for neutralization; see text.
L2, L3 — 7 turns No. 20 tinned, spaced 1 diam., ¼-inch diam., center-tapped.
L4 — No. 28 enam. wound one inch on ⅜-inch slug-tuned form (National XR-91).
L5 — 12 turns No. 20 tinned, spaced one diam., ⅝-inch diam., tapped at 3½ turns (B & W Miniductor No. 3007).
L6 — 4 turns No. 20 tinned, ½-inch diam., ¼ inch long (B & W Miniductor No. 3003).
L7 — 5 turns like *L6*. *L6* and *L7* are in line with their cold ends spaced ⅛ inch.
L8 — 2½ turns No. 20 enam., ¼ inch long.
L9, L10 — 2 turns insulated wire between turns of *L8* and *L3*, connected by link of same material.
J1 — Coaxial fitting.
J2 — Male 4-prong chassis fitting (Jones P-304-AB).

to advantage in the other converters, when spurious signals are bothersome.

Adjustment procedure is similar to that outlined for the 144-Mc. model, except that the spacing of the turns in the r.f. coils must be adjusted, rather than tuning them by capacitors. As in the 144-Mc. converter, the order of frequency multiplication can be altered to take care of any extreme local interference problems resulting from near-by TV, FM or other high-powered stations that may ride through as spurious responses. The oscillator can be operated on its fifth overtone instead of the third, making the second and third stages operate as doubler and tripler, or vice versa. Fifth-overtone operation of the oscillator will require more care in adjustment of feedback than is the case with the third.

The coupling between L_8 and L_3 will be a factor in holding down spurious responses. It should be set at the lowest value that will allow satisfactory performance, by altering the position of the coupling loops, L_9 and L_{10}, or by varying the value of the screen-dropping resistor in the last frequency-multiplier stage.

If a noise generator is available, and care is used in making the adjustments, it should be possible to achieve noise figures under 6 db. for the 220-Mc. converter and 5 db. for the 144- and 50-Mc. models.

V.H.F. RECEIVING BALUNS

As pointed out in the preceding converter descriptions, coaxial antenna input circuits are preferable in v.h.f. receivers where single-ended circuitry is employed. Where long transmission lines must be used, however, the losses in coaxial line discourage its use in feeding the antenna system. Particularly on 144 Mc. and higher, many amateurs prefer close-spaced open-wire lines for runs of 50 feet or more between the operating position and the antenna.

The advantages of coaxial input coupling and the low losses of open-wire balanced lines can both be retained if some means of coupling between the balanced line and the unbalanced receiver input circuit is provided. Such a device, usually called a "balun," is shown in Fig. 13-23D. V.h.f. receiver baluns are usually made of small coaxial line such as RG-59/U, and installed at the converter input terminal. The propagation factor of the line should be taken into account, making the actual length of the folded portion 65 per cent of a half-wave. The straight portion may be any convenient length, though it is usually a wavelength or less.

A 3-band balun for v.h.f. receiving use may also be made by using the coils from a so-called "elevator transformer" for this purpose that can

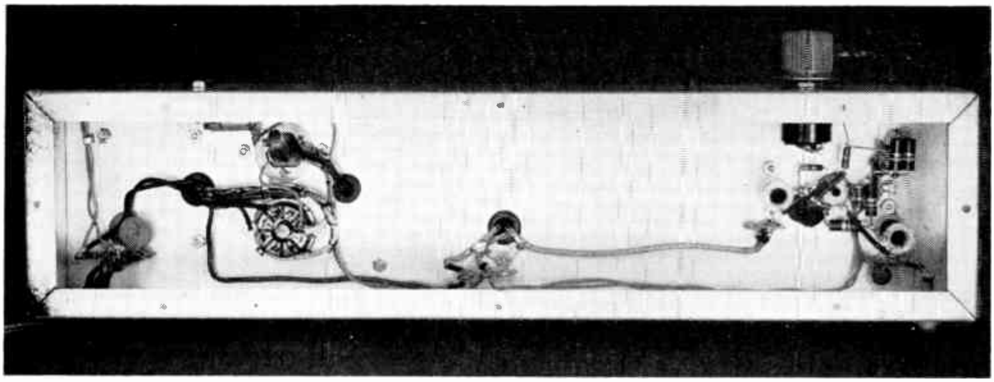


Fig. 16-15 — Bottom view of the i.f. and power supply unit with bottom cover removed. Power components are at the left. A smaller chassis may be used if less than the three converters are to be built.

be obtained from some TV receiver parts distributors. Such a balun would consist of two pairs of coils, connected in parallel at one end and in series at the other. The parallel end is wired to a coaxial connector and the series end to a crystal socket or a pair of binding posts. The assembly should be housed in a copper or aluminum box that may be as small as 1 × 1½ × 2½ inches.

Like the coaxial-line balun, this converts from balanced to unbalanced termination, and provides a 4-to-1 impedance transformation in the process. The coils are designed for use across the v.h.f. TV range, 54 to 216 Mc., so they will serve well for all three amateur v.h.f. bands, 50, 144 and 220 Mc. See Fig. 13-24 for connections.

THE I.F. AMPLIFIER AND POWER SUPPLY

The i.f. amplifier (Figs. 16-15 and 16-16) serves two useful purposes. It builds up the gain, for receivers that may be poor performers at 7 Mc., and it provides a means of controlling the over-all gain of the system without disturbing the gain or S-meter controls on the receiver itself. The receiver may thus be operated exactly as it would be on 7 Mc., and the gain of the converter adjusted so that v.h.f. signals will be received similarly to those on lower frequency bands.

It is obvious from the photographs that the i.f. and power supply unit could have been built in a smaller space. If the builder is considering only one or two of the converters he may wish to do this, but where all three are used the arrangement shown is a convenient one. The i.f. chassis is a standard size, 3 × 4 × 17-inch aluminum, to which a bottom plate is added for shielding. Rubber feet can be attached to the two ends of the base, and one on each of the converters at the rear, to prevent the combination from marring a receiver top.

The heater voltage, the plate voltage and the i.f. input lead are all carried on shielded wire to a 4-pin plug. This is connected to whichever converter is to be used at the moment, and no other changes other than plugging the antenna into the proper jack are required in changing from one v.h.f. band to another. The shielded wires in the cable are bonded together several times and then wrapped with plastic tape. The coaxial fitting for the connection to the receiver is at the extreme right on the rear wall of the i.f. chassis.

The only adjustment required in the i.f. unit is to set the coil slugs (on noise or signal) so that the response will be as nearly flat as possible across 7 to 11 Mc.

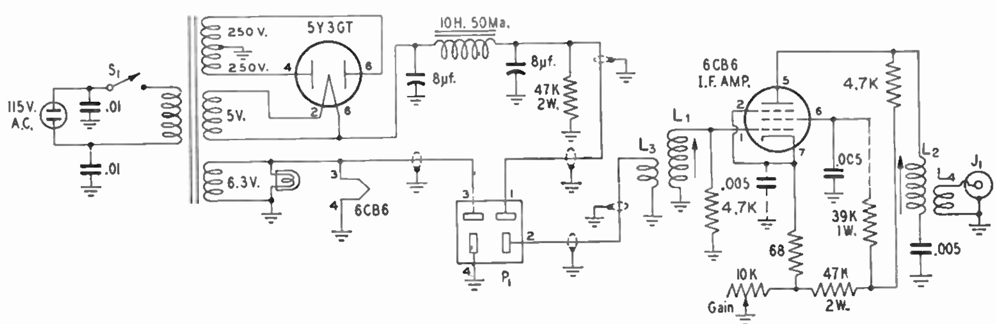


Fig. 16-16 — Schematic diagram and parts information for the i.f. and power supply unit used with the crystal-controlled converters.

L_1, L_2 — No. 28 enameled wire close wound on ⅜-inch slug-tuned form (National XR-91). Lacquer and dry before adding coupling winding. Wind on upper portion of form.

L_3, L_4 — 10 turns same wound over cold ends of L_1 and L_2.

J_1 — Coaxial fitting.

P_1 — Female 4-pin on end of cable (Jones S-304-CCT).

A Simple Converter for 50 and 144 Mc.

Though the more complex equipment already described is typical of the gear that must be used in order to attain top performance on the v.h.f. bands, it is possible to start with simpler devices and still do a good job. The converter shown in Figs. 16-17 through 16-20 provides the best performance that can be expected from simple equipment. It was not built to be the simplest possible receiving device; rather, it was designed to provide good results with a minimum of complication and cost.

It uses a dual triode, 6J6, as a combined mixer-oscillator, followed by a 6AK5 i.f. amplifier. The latter is necessary; do not try to do without it. The output of a triode mixer is too low to give adequate gain for most receivers. The i.f. amplifier stage makes the converter usable with even the simplest receivers, and provides a convenient means of controlling the over all gain of the system. Plug-in coils mounted inside tube-base type forms provide the means of changing bands.

Mechanical Details

Though it could be built in a much smaller space, the converter uses a 3 by 5 by 10-inch chassis, allowing plenty of room for the work that must be done underside. The main tuning condenser is a split-stator variable made from a double-bearing double-spaced 15-$\mu\mu$fd. type. Each section is reduced to three stator and two rotor plates. This unit is mounted under the chassis, as close to the top plate as possible, to make room for the vernier dial on the front panel. The mixer and i.f. plate coils, L_4 and L_5, are mounted under the chassis. Normally this will provide all the shielding necessary for the i.f. circuits. If trouble is experienced with signals on the intermediate frequency a bottom plate may be added to the chassis. The panel is set out from the chassis front with half-inch pillars.

A smooth-running dial on the oscillator tuning is a necessity in a v.h.f. converter when communications-receiver selectivity is used. The National type SCN has a good tuning rate, plus ample space for calibration scales for both bands.

The circuit is so simple that no trouble should be experienced if the general parts arrangement is followed. Look over the photographs closely before starting to lay out the chassis for drilling. In the rear view, Fig. 16-18, the oscillator coil, the 6J6 tube, and the mixer grid coil, L_1-L_2, appear in that order, from left to right, close to the panel. The 6AK5 tube is nearer the back, with the slug adjustment screws of the mixer plate coil, L_4, and the i.f. plate coils, L_5-L_6, at the right and left, respectively. Holes are drilled in spare space at the back of the chassis to provide for storage of the set of coils not in use.

Looking in the bottom view, Fig. 16-20, we see the oscillator tuning condenser, C_5, at the center, the 6J6 socket at the left and the coil socket at the right. Note that the latter is as close to C_5 as possible.

The only critical job in the adjustment procedure is involved in getting the inductance of the oscillator plug-in coils, L_3, to the correct value. There being only one parallel trimmer for the oscillator (C_4) the coils must be made and adjusted carefully in order to have the desired bandspread on both ranges.

Considerable care must be used in the placement of the oscillator and mixer components, so that all leads will be very short; otherwise it will not be possible to resonate these circuits at 148 Mc. The 6J6 socket is at the left of C_5 in the bottom view, and the mixer grid circuit components appear just to the left of the middle. The i.f. amplifier gain control, R_7, is at the right. The 300-ohm line from the crystal-socket antenna terminal, J_1, may be seen at the far left.

The mixer plate coil, the i.f. amplifier socket, and the output coil assembly are across the bottom of this view, from left to right. The antenna terminal, power plug and i.f. output connector are on the rear wall in the same order.

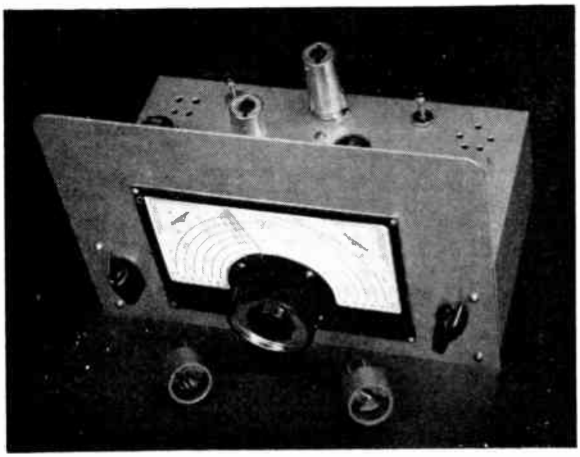


Fig. 16-17 — A 2-tube converter for 50 and 144 Mc. The vernier dial is for oscillator tuning. The two knobs are the i.f. gain control, right, and the mixer tuning condenser. In front are the 2-meter mixer and oscillator plug-in coils.

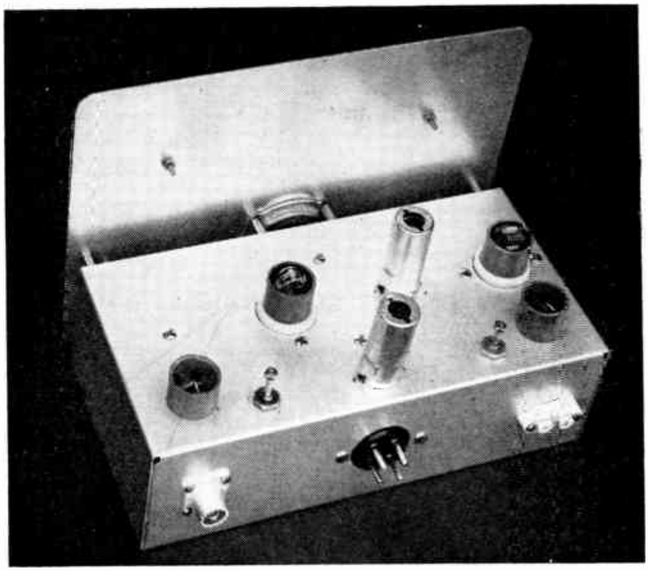


Fig. 16-18 — Rear view of the simple converter. Near the panel, left to right, the oscillator coils are shown in place. The i.f. amplifier tube is nearer the back of the chassis, with the slug-tuned mixer and i.f. plate coils at either side. Coils not in use are stored at the back of the chassis.

Test Procedure

When the assembly and wiring are completed, the oscillator operation should be checked. The power supply should deliver 6.3 volts a.c., at 1 ampere, and 150 volts d.c. at 30 ma., preferably regulated. Insert a milliammeter in series with R_3 and check for oscillation by touching any bare spot in the oscillator plate or grid circuit with a pencil. A change in current indicates oscillation.

Two types of bandspread are possible. With the coil values given in the parts list, the 50-Mc. band covers about 90 divisions of the dial. The 144-Mc. band covers about 50 divisions. The capacitance needed at C_4 is about 12.5 μμfd. in

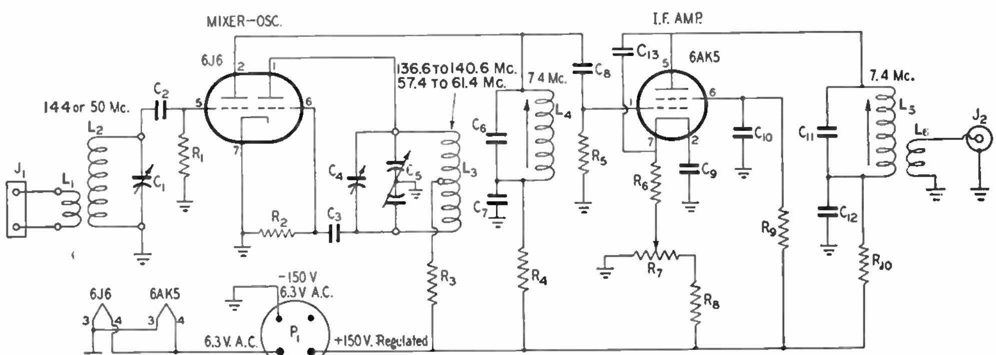


Fig. 16-19 — Schematic diagram of the two-tube converter for 50 and 144 Mc.

C_1 — 15-μμfd. midget variable (Hammarlund HF-15).
C_2 — 100-μμfd. mica or ceramic.
C_3, C_8 — 47-μμfd. mica or ceramic.
C_4 — 35-μμfd. ceramic trimmer (Centralab 820-C).
C_5 — Double-spaced split-stator variable, about 8 μμfd. per section (Hammarlund HFD-15-X, reduced to 3 stator and 2 rotor plates in each section).
C_6, C_{11} — 68-μμfd. mica or ceramic.
C_7, C_9, C_{10}, C_{12} — 0.01-μfd. disk ceramic.
C_{13} — 15-μμfd. ceramic. Connect directly from Pin 5 to Pin 7 on 6AK5 socket.
R_1, R_5 — 1 megohm, ½ watt.
R_2 — 10,000 ohms, ½ watt.
R_3, R_4, R_9, R_{10} — 1000 ohms, ½ watt
R_6 — 220 ohms, ½ watt.
R_7 — 2000-ohm 4-watt potentiometer
R_8 — 22,000 ohms, 1 watt.
L_1 — 50 Mc.: 2 turns No. 22 enam. interwound in cold end of L_2.
144 Mc.: 3 turns No. 22 enam. ¼-inch diam., close-wound at cold end of L_2.
L_2 — 50 Mc.: 7 turns No. 22 tinned, ½-inch diam., 7/16 inch long (B & W No. 3003).
144 Mc.: 2 turns No. 16 tinned, ¼-inch diam., ¼-inch long.
L_3 — 50 Mc.: 6 turns No. 22 tinned, ½-inch diam., 7/16 inch long center-tapped (B & W No. 3003, with end turns spread slightly). Alternate design for more bandspread, see text.
144 Mc.: U-shaped loop No. 12 wire, ¾ inch wide, 1 inch long, center-tapped.
Coils L_1 and L_2 are supported inside Millen 1-inch diameter 4-prong forms. L_3 in Millen 45005, 5-prong. Saw off to ¾-inch length.
L_4, L_5 — 23 turns No. 22 enam. close-wound on National XR-50 slug-tuned form.
L_6 — 3 turns No. 22 enam. close-wound at cold end of L_5.
J_1 — Crystal socket for antenna terminal.
J_2 — Coaxial fitting, female.
P_1 — 4-prong power fitting, male.

this case. If more bandspread is wanted on 144 Mc., the setting of C_4 can be increased to around 23 $\mu\mu$fd., and L_3 reduced to 4 turns. The 2-meter band will then cover around 72 divisions. It will not be possible to cover the whole of the 50-Mc. band with this arrangement, without resetting C_4, but this is no great handicap so long as activity is concentrated in the lower portion of the band, as at present.

The frequency of the oscillator may be checked with an absorption-type wavemeter or Lecher wires. For the 50-Mc. range, the oscillator should tune from 57.4 to 61.4 Mc. in order to beat with an incoming signal to produce a 7.4-Mc. i.f. (The oscillator is on the *high* side of the signal.) A kick in the oscillator plate current, or a flicker in the voltage-regulator tube in the power supply, can be used to show when the frequency is found with the measuring device.

Set the padder, C_4, so that 57.4 Mc. comes at about 5 divisions in from the maximum-capacity end of the tuning range, and check to see where 61.4 Mc. is found. It should come just inside the minimum-capacity end of the range. If the circuit will not tune to 61.4 Mc. the inductance of L_3 is too low. Move the turns closer together, and reset C_4 as before for 57.4 Mc. If the bandspread is too small, spread the turns and increase the capacitance of C_4 to compensate, for the desired amount of spread, about 90 divisions on the dial.

Next check the 2-meter range. Here the coil must be adjusted in inductance until the oscillator will hit 136.6 Mc. somewhere between the middle and the maximum-capacity end of the tuning range of C_5. The high end, 140.6 Mc., will then appear about 50 divisions higher on the dial. The oscillator is on the *low* side of the signal on this range. Do not change the setting of C_4 in this process, or it will be necessary to alter the 50-Mc. coil again.

Once the oscillator covers the proper frequency ranges the converter may be tested in actual reception. Connect the output through a coaxial cable to a receiver tuned to approximately 7.4 Mc. There should be an increase in noise as the gain control is turned up. The mixer and i.f. amplifier plate windings can be tuned to the proper frequency merely by adjusting the core screws for maximum noise.

The mixer grid circuit may also be peaked on noise, though care should be taken to see that it is not peaked on the image, 14.8 Mc. away from the signal frequency. If the grid circuit is tuned to the desired frequency there will be a considerable increase in the strength of a signal as the grid condenser, C_1, is tuned through resonance. If the circuit is tuned to the image frequency the noise will peak up, but an amateur-band signal will drop in strength as the noise peak occurs. Tuning the mixer grid circuit shifts the oscillator frequency slightly, so it may be peaked more accurately on noise than when listening to a signal.

A final check of the dial calibration may be made by tuning in signals of known frequency, or by means of an accurate signal generator. Few wavemeters are sufficiently accurate for final calibration by the method outlined earlier.

If trouble is encountered with signals in the 7-Mc. region leaking through, the i.f. can be shifted slightly to tune out the interference. In some instances it may be necessary to put a bottom plate on the chassis. Small changes in intermediate frequency can be made without resetting either the oscillator padder or the i.f. coils. With the i.f. amplifier built into the converter, the setup will have adequate gain for use with almost any receiver. Reception will be nearly as good as with more complex designs, the principal difference being a somewhat higher noise figure (slightly degraded signal-to-noise ratio) in the simpler job. The use of a low-noise r.f. amplifier ahead of the converter (an example is the preamplifier of Fig. 16-22) will make possible reception equal to the best obtainable in a converter having a tunable oscillator.

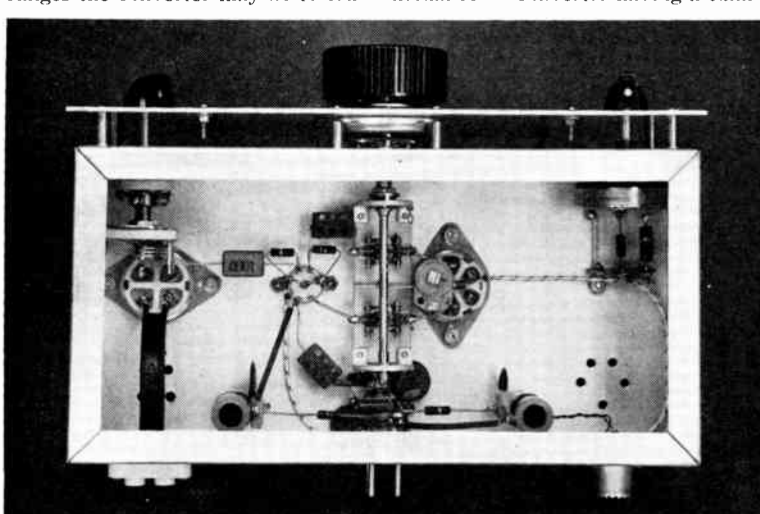


Fig. 16-20 — **Bottom view of the two-band converter. The split-stator condenser at the center is for oscillator tuning. The oscillator coil socket is at the right and the 6J6 socket at the left. The mixer tuning condenser and grid coil socket are in the upper left corner, with the i.f. coils and tube socket at the rear.**

Low-Noise Preamplifier for 144 Mc.

The triode preamplifier shown in Figs. 16-21 to 16-23 will improve the sensitivity and signal-to-noise ratio of receivers or converters for 144 Mc. that are deficient in these respects. Two separate triode tubes are shown, but any of the dual triodes designed for v.h.f. amplifier service may be used similarly. The circuit may be adapted to use on 50 or 220 Mc., by suitable alteration of coil and condenser values.

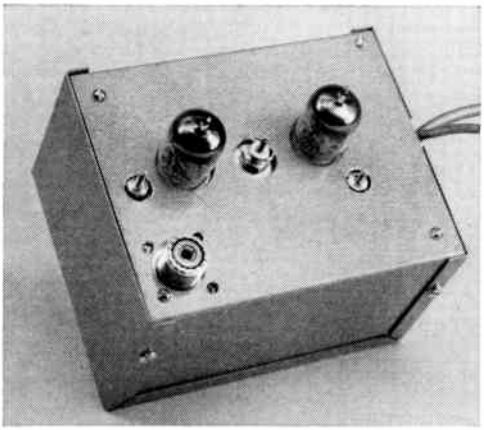


Fig. 16-21 — **Two-meter preamplifier using two 6AJ4 tubes. Adjustments are *(left to right)* input tuning capacitor, slug of neutralizing winding, and the plate tuning capacitor of the second stage.**

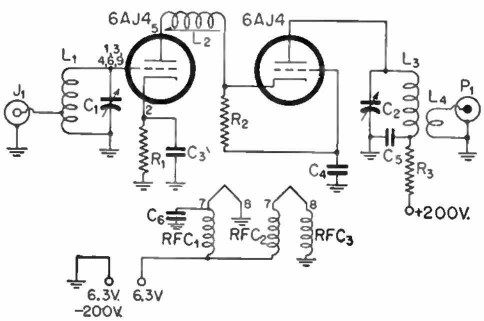


Fig. 16-22 — **Schematic diagram and parts list for the low-noise preamplifier.**

C_1, C_2 — Plastic trimmer, 1 to 8 μμfd. (Erie style 532-10).
C_3, C_4, C_5, C_6 — 0.001-μfd. disk ceramic.
R_1 — 68 ohms, ½ watt, carbon.
R_2 — 0.47 megohm, ½ watt.
R_3 — 470 ohms, ½ watt, carbon.
L_1 — 4 turns No. 16 tinned, ¼-inch diam., spaced 1 diameter, tapped at 1¾ turns from ground end.
L_2 — 4 turns No. 24 on ¼-inch slug-tuned form.
L_3 — 5 turns No. 18 enam., ¼-inch diam., spaced half diameter.
L_4 — 2 turns insulated wire wound over cold end of L_3.
J_1 — Coaxial antenna fitting.
P_1 — Coaxial plug on cable of suitable length to reach converter input.
RFC_1 — 22 turns No. 22 enam., ⅜-inch diam., close-wound.
RFC_2, RFC_3 — 18 turns each, No. 24 enam., ¼-inch diam. Twist wires together before winding. Coat turns with household cement.

Pin connections given on the schematic diagram, Fig. 16-22, are for the 6AJ4 or 6AM4. Other tubes such as the 6AN4 and 417A will work equally well, if pin connections shown in the tube data section of this *Handbook* are followed. Slightly different values of cathode bias resistor may be needed if tubes other than the 6AJ4 are used.

The preamplifier is housed in a standard 3 × 4 × 5-inch aluminum utility box. The components were mounted on a sheet of flashing copper and the preliminary work of wiring was done with this plate as a chassis. The plate was later fastened to the inside of the top of the box. The parts could be mounted on the box directly, but they are more accessible if the work is done as described above.

Looking at the interior view, Fig. 16-23, we see the coax fitting, the first tube socket and the input circuit at the left. Between the tube sockets, at the center of the copper base plate, is the slug-tuned neutralizing winding, L_2. A small copper shield divides the second socket, isolating the input and output circuits. This shield is not always needed, but it may be an aid to neutralization. At the far right are the output circuit and the bifilar-wound r.f. chokes for the heater circuit of the second stage. The tuning condensers, C_1 and C_2, are plastic trimmers of a design that allows a saving in space and offers lower minimum capacitance and lead inductance than conventional flat-plate trimmers.

The five grid pins of the 6AJ4 may be strapped together or used individually, as layout requirements dictate. In this instance, Pin 4 is used for

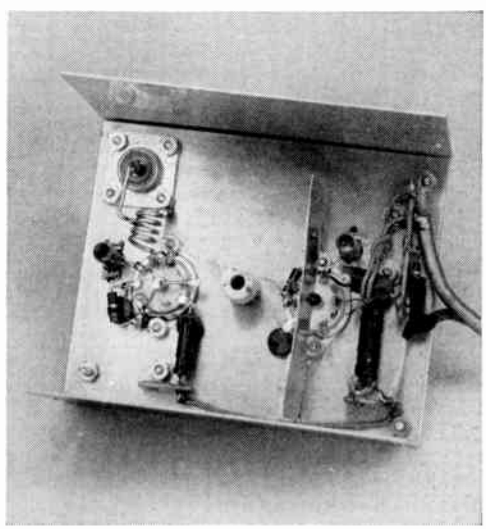


Fig. 16-23 — **Interior view of the 144-Mc. r.f. amplifier. A small shield across the second tube socket isolates the input and output circuits. The amplifier is built on a copper plate, which is then fitted to the top of a standard aluminum utility box.**

the hot end of L_1, with the trimmer, C_1, connected to Pin 3. In the second stage, Pins 3 and 4 are tied to the grid side of R_2, and Pin 1 is by-passed by C_4.

Adjustment

A noise generator will make the adjustment of the amplifier easy, as it is then only necessary to peak the plate circuit (by C_2) for maximum gain, and then adjust the inductance of L_3 and the setting of C_1 for lowest noise figure. It is possible to follow this routine using signals or a signal generator, but it is a more difficult process.

If a signal is to be used, peak the second plate circuit for maximum response first. Then tune the input circuit for maximum also, if the amplifier does not oscillate. If it should oscillate, vary the setting of the slug in L_2 to stop it, before attempting to peak any other adjustments. In adjusting the input circuit, watch for best signal-to-noise ratio, now, rather than for maximum gain. This will show up somewhat on the high-capacity side of the maximum-gain point, as the rotor of C_1 is turned into the stator.

The position of the tap on L_1 can be adjusted in the same way. The optimum point will be higher on the coil than the point at which maximum gain is observed. If the amplifier is adjusted at 146 Mc. it should not be necessary to repeak it across the entire band.

An amplifier of this sort should not be expected to produce a large improvement in reception when it is used ahead of a converter that already has a good triode front end, but installed ahead of a pentode amplifier, and particularly a converter having a bandswitching r.f. circuit, it will help considerably in the reception of weak signals, by increasing the margin of the signal over noise.

Receivers for 420 Mc.

For best signal-to-noise ratio, receivers for any frequency should have the highest degree of selectivity that can be used successfully at the frequency in question. With crystal control or its equivalent in stability accepted as standard practice on all bands up through 148 Mc., there is little point in using more bandwidth in receivers for these frequencies than is necessary for satisfactory voice reception, a maximum of about 10 kc. Such communication selectivity is now being used successfully by most workers on 220 and 420 Mc., too, but it imposes several problems not encountered on lower bands.

First is the matter of oscillator instability in the converter. Even the best tunable oscillator at 420 Mc. suffers from vibration and hand-capacity effects sufficiently to make it difficult to hold the signal in a 10-kc. i.f. bandwidth.

Then, there are still some unstable transmitters being used in work on 220 and 420 Mc. It is out of the question to copy these on a selective receiver.

Last, searching a band 30 megacycles wide is excessively time-consuming when communications-receiver selectivity is used in the i.f. system.

There is no single solution to these problems, but the best approach appears to be that of breaking up of the band into segments for different types of operation. This is being done by mutual agreement among 420-Mc. operators at present, as follows: 420 to 432 Mc. — modulated oscillators and wideband FM; 432 to 436 Mc. — crystal-controlled c.w., AM and narrow-band FM; 436 to 450 — television.

The first segment can be covered with a superregenerative receiver, a superheterodyne having a wideband i.f. system, or a converter used ahead of an FM broadcast receiver. The high selectivity required for best use of the middle portion makes a crystal-controlled or otherwise highly stable converter and communications receiver combination almost mandatory. Amateur TV is usually received with a converter ahead of a standard TV receiver, tuned to some channel that is not in use locally.

Many of the tubes used on the v.h.f. bands are useless at 420 Mc., and the performance of even the best u.h.f. tubes is down compared to lower bands. Only the lighthouse or pencil-triode tubes and a few of the miniatures are usable, and these require modifications of conventional circuit technique to produce satisfactory results.

Crystal diodes are often used as mixers in 420-Mc. receivers, as in this frequency range they work nearly as well as vacuum tubes. The over-all gain of a converter having a crystal mixer is about 10 db. lower than one using a tube, so this difference must be made up in the i.f. amplifier. The noise figure of a receiver having a crystal mixer and no r.f. stage includes the noise figure of the i.f. amplifier following the mixer, so best results require that the i.f. amplifier employ low-noise techniques discussed earlier in this chapter. If the i.f. is 50 Mc. or higher it is particularly important that a low-noise triode be used for the first i.f. stage.

Crystal diodes of the type used in radar mixers, such as the 1N21 series, are well suited to 420-Mc. mixer service, though care must be taken to avoid damage from transmitter r.f. energy. Other types of crystal diodes such as the 1N72 and CK710 will stand higher values of crystal current, and their use is recommended.

Few conventional vacuum tubes work well as mixers at 420 Mc. and higher. The 6J6 is useful where a balanced input circuit is desired, as in Fig. 16-5B. For single-ended circuitry the 6AM4 and 6AN4 are recommended. They may be used in grounded-grid or grounded-cathode circuits.

For high-selectivity coverage of the 432- to 436-Mc. segment of the band, a common practice is to use a crystal-controlled converter working into another converter for either the 50- or 144-Mc. band, tuning the latter for the four-megacycle tuning range.

A 420-MC. R.F. AMPLIFIER

The r.f. amplifier shown in Figs. 16-24 through 16-26 is capable of a gain or more than 15 db. and its noise figure can be as low as 6 db. with careful adjustment. It will make a large improvement in the sensitivity of any converter or receiver that has no r.f. stage, or one that is working poorly.

The design shown is for either the 6AJ4 or 6AM4, but with suitable socket and pin-connection changes the 417A and 6AN4 will work equally well. It is a grounded-grid amplifier with a half-wave line in the plate circuit. The antenna is connected to the cathode of the tube through a coupling condenser. As the input impedance of the grounded-grid stage is low, nothing is gained by the use of a tuned circuit in the cathode lead. Output is taken off through a coupling loop at the point of lowest r.f. voltage along the line.

The amplifier is built in a frame of flashing copper that serves as the outer conductor of the tank circuit. The whole assembly is 10 inches long and 1¼ inches square, except for the bottom, which is about 1¾ inches wide. Edges are folded over with lips ¼ inch wide which slide into a bottom cover made from copper sheet 2¼ by 10 inches in size, with its edges bent up ¼ inch wide on each side.

The plate circuit is made of ¼-inch copper tubing tuned by a copper-tab capacitor at the far end from the tube. Plate voltage is fed in at the point of minimum r.f. voltage, which in this instance is about 5 inches from the open end. The antenna is connected to the cathode through a coupling condenser. The input impedance of the grounded-grid amplifier is so low that nothing is gained by using a tuned circuit at this point. The cathode and heater are maintained above ground potential by small air-wound r.f. chokes.

The tube socket is two inches in from the end of the trough, and is so oriented that its plate connection, Pin 5, is in the proper position to connect to the line with the shortest possible lead. A copper shielding fin is mounted across the interior of the trough 2⅛ inches from the end, dividing the socket so that Pins 3, 4, 5 and 6 are on the plate side of the partition.

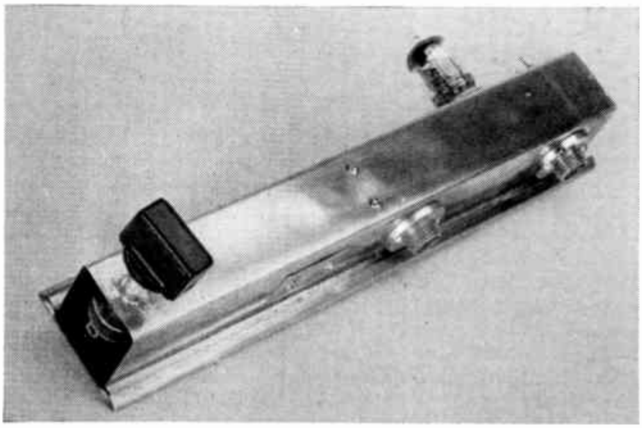


Fig. 16-24 — A highly effective r.f. amplifier for 420 Mc. The tank circuit is a half-wave line made of flashing copper. Coaxial fittings are for input and output connections. Heater and plate voltages are brought in on feed-through by-pass capacitors just visible on either side of the 6AJ4 tube.

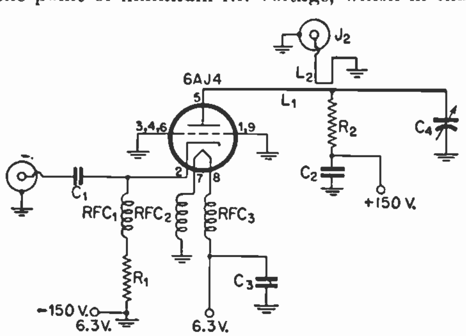


Fig. 16-25 — Schematic diagram of the 420-Mc. r.f. amplifier.

C_1 — 500-μμfd. ceramic.
C_2, C_3 — 1000-μμfd. ceramic feed-through (Erie style 2404).
C_4 — Copper tabs, ⅞-inch diam.; see text and photographs.
R_1 — 150 ohms, ½ watt.
R_2 — 470 ohms, ½ watt.
L_1 — ¼-inch copper tubing, 7⅜ inches long, tapped 2⅜ inches from plate end.
L_2 — Loop of insulated wire adjacent to L_1 for ¾ inch.
J_1, J_2 — Coaxial fitting.
RFC_1, RFC_2, RFC_3 — 9 turns No. 22, ⅜-inch diam., spaced one diam.

Minimum grid-lead inductance is important. This was insured by bending all the grid prongs down against the ceramic body of the socket, and then making the mounting hole just big enough to pass this part of the socket and the prongs. They were soldered to the wall of the trough.

Input and output connections are coaxial fittings mounted on the side wall of the trough. B-plus and heater voltage are brought into the assembly on feed-through capacitors mounted on the same side of the trough as the tube. Connection to the inner conductor of the line is made with a grid clip, so that the point of connection can be adjusted for optimum results.

The copper tubing is slotted at the plate end with a hack saw to a depth of about ¼ inch, and a strip of flashing copper soldered into this slot to make the plate connection. A copper tab about the size of a one-cent piece is soldered to the other end of the tubing to provide the stationary plate of C_4. The line is supported near the low-voltage point by a ¼-inch-thick block of polystyrene. This is centered at a point 5¼ inches in from the tube end of the trough assembly. The hole for the B-plus feed-through is 4¼ inches from the same end.

The movable plate of C_4 is soldered to a screw running through a nut soldered to the upper

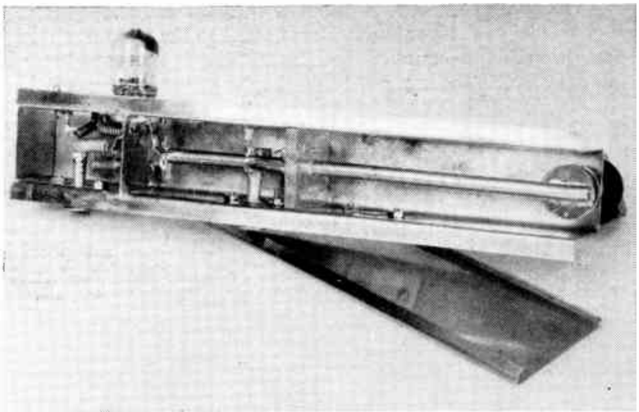


Fig. 16-26 — Bottom view of the 420-Mc. r.f. amplifier, with the slip-on cover removed. The inner conductor of the tank circuit is held in place by a block of polystyrene, mounted near the low-voltage point on the line. The plate-voltage feed-through and output coupling loop may be seen at the left of this support. Heater, cathode and antenna-circuit components are in a separate compartment at the tube end of the assembly. The line is tuned at the opposite end by a handmade copper-tab capacitor.

surface of the trough at a point 3/8 inch in from the open end. If a fine-thread screw is available for this purpose it will make for easier tuning, though a 6/32 thread was used in this model. This made a wobbly contact, so a coil spring was installed between the top of the trough and the knob to keep some tension on the adjusting screw.

Adjustment of the 420-Mc. amplifier is made easier if a noise generator is used, though it is not as important as in the case amplifiers with tuned input circuits. If the amplifier is working properly there will be an appreciable rise in noise as the plate circuit is tuned through resonance, and it may break into oscillation if operated without load. When connected to a following stage, with a reasonably-matched antenna plugged into J_1, the amplifier should not oscillate unless the coupling loop, L_2, is much too far from the inner conductor.

When the amplifier is operating stably and tuned to a test signal (or to a peak of response to a noise generator), the next step is to locate the optimum position for feeding the plate voltage into the line. This may be done by running a pencil lead slowly up and down the inner conductor, until a spot is found where touching the lead to the line has little or no effect on the operation of the amplifier. The plate voltage clip should be placed at this point and the process repeated, moving the clip slightly until it is at the minimum-voltage point precisely. This adjustment should be made at the midpoint of the tuning range over which the amplifier is to be used.

The position of the coupling loop should then be adjusted for best signal-to-noise ratio. This will probably turn out to be with the insulated wire lying against the inner conductor for a distance of about 3/4 to 1 inch, starting at the minimum-voltage point just located.

A CRYSTAL-CONTROLLED CONVERTER FOR 432 MC.

The converter shown in Figs. 16-27 through 16-30 is designed to provide high sensitivity and signal-to-noise ratio in reception of signals in the 432- to 436-Mc. range. It uses a grounded-grid r.f. amplifier stage similar to the one shown in Fig. 16-24, working into a crystal-diode mixer. The intermediate frequency, with the design constants given, is 50 to 54 Mc., though lower frequencies could be used by suitable modification of the injection chain.

Crystal-controlled injection on 382 Mc. is provided by two 6J6s operating as overtone oscillator-tripler and tripler-doubler, respectively. As only a small amount of r.f. is required at 382 Mc.,

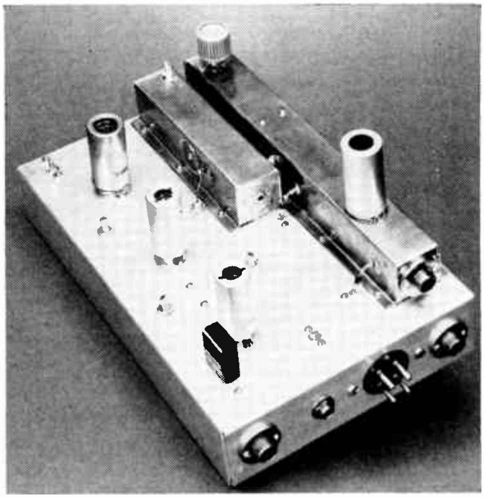


Fig. 16-27 — A crystal-controlled converter for 432 to 436 Mc. R.f. and mixer stages are in copper subassemblies at the right. Oscillator, multiplier and i.f. amplifier are on the left side.

this line-up is not difficult to build or adjust. An inexpensive 7-Mc. crystal is used. An i.f. preamplifier stage follows the crystal mixer. This may or may not be needed, depending on the performance of the receiver or converter that will serve as the tunable i.f. Low-noise amplification in the i.f. stage is a factor in the over-all performance of the system, so use of the built-in i.f. stage is recommended.

Construction

The converter is built on a 7 × 11 × 2-inch aluminum chassis, with the r.f. and mixer portions in a copper subassembly that mounts on the top of the chassis, at the right side as seen in

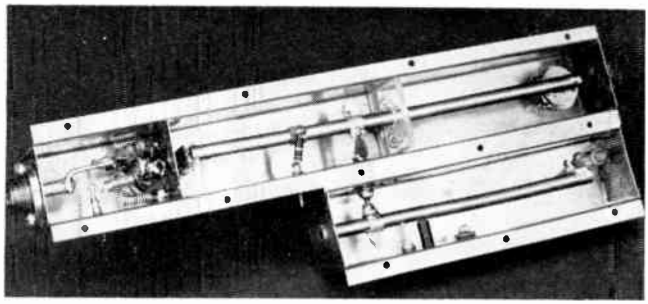


Fig. 16-28 — Interior view of the r.f. amplifier and mixer assemblies. The r.f. circuit is a half-wave line. The shorter assembly is the quarter-wave line using a crystal diode mixer.

Fig. 16-27. The oscillator-tripler and tripler-doubler 6J6s are at the left front, with the 6BQ7A i.f. amplifier at the rear. The mixer line is the short portion of the copper assembly, with the r.f. amplifier line at the right. In the bottom view, Fig. 16-28, the injection-chain and i.f. amplifier components are visible.

Fig. 16-28 is an interior view of the r.f. and mixer lines. These are made as two separate assemblies, joined by short length of copper tubing that is visible in the top view. Both tank circuits are 1¼ inches square, with ¼-inch copper tubing inner conductors. They are made from sheets of flashing copper 4¼ inches wide. The mixer compartment is 5½ inches long and the r.f. portion is 10 inches long.

The r.f. amplifier is similar structurally to the one described previously, except for the method of coupling between it and the crystal mixer. This is done with a grid clip on each line and a ceramic coupling condenser. The lead from the capacitor, inside the amplifier line, is brought through a half-inch length of copper tubing that is soldered into the walls of both lines. The lead is insulated with spaghetti sleeving.

The B-plus feed to the r.f. stage should be at the point of minimum r.f. voltage, 1⅞ inches from the plate end of the copper tubing. The coupling tap is one inch out from the B-plus feed-point. The coupling point on the mixer line is 1 inch from the ground end. The crystal diode is inserted in a small hole in the mixer inner conductor, 1¾ inches from the ground end. The inner conductors of the r.f. and mixer lines are 7 3/16 and 5 inches long, respectively. Mixer tuning is done with a small plastic trimmer, C_{10}, while the r.f. plate circuit is tuned with a hand-made tab capacitor, C_9, similar to C_4 in Fig. 16-25.

Note the r.f. by-pass, C_8, on the outside of the mixer line. This is made from a piece of copper ⅞ inch in diameter, insulated from the line housing by a piece of vinyl plastic. Two thicknesses of the material commonly used for small parts envelopes are satisfactory. The crystal, which may be any of the u.h.f. diodes, is slipped through a close-fit hole and is held in place by the wire soldered to its outside terminal.

Plate and filament voltages are fed into the assembly on feed-through by-pass capacitors, visible in the top-view photograph. Antenna connection is made through a coaxial fitting on the end of the r.f. assembly. A crystal-current jack, a 4-pin power fitting and two i.f. connectors are on the end wall of the chassis. The second coaxial connector was installed so that tests could be made with and without the i.f. amplifier stage.

Wiring in the power circuits is done with shielded wire, in case that TVI might result from the oscillator or multiplier stages. The addition of a bottom plate and power-lead filtering would then be effective. Injection and i.f. coupling leads are also made of shielded wire, this serving in place of coax line that is harder to handle.

The output of the injection chain is coupled into the mixer line by means of a loop, L_8, that is not visible in the photographs. This loop is mounted on the copper base plate that is under

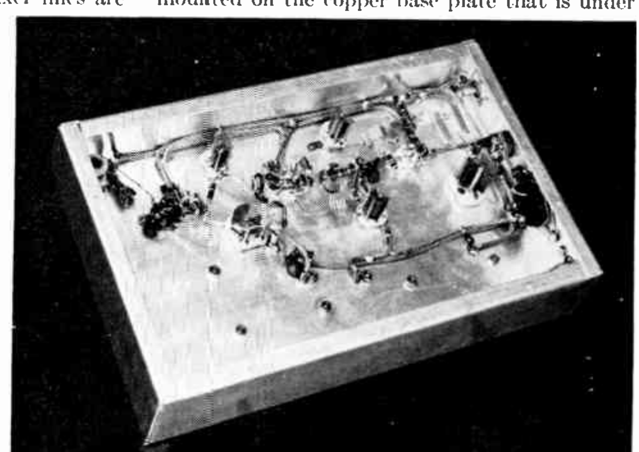


Fig. 16-29 — Bottom view of the 432-Mc. converter, showing the oscillator, multiplier and i.f. amplifier circuits.

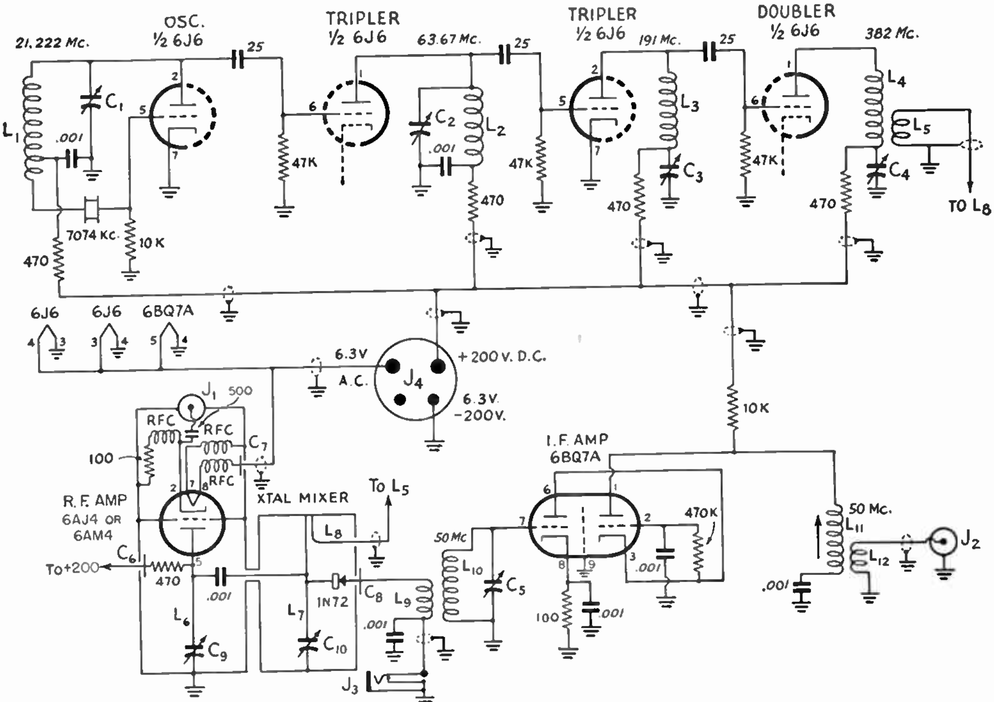


Fig. 16-30 — Wiring diagram and parts list for the 432-Mc. crystal-controlled converter. Values given are for an i.f. of 50 to 54 Mc.

C_1 — 75-μμf. miniature trimmer (Hammarlund MAPC-75).
C_2, C_3, C_4 — 2-μμf. miniature trimmer (Johnson 20M11).
C_5 — 25-μμf. miniature trimmer (Hammarlund MAPC-50).
C_6, C_7 — 500-μμf. feed-through ceramic (Centralab MFT-500).
C_8 — Handmade copper-tab by-pass; see text.
C_9 — Handmade copper-tab variable; see text.
C_{10} — 0.5- to 5-μμf. plastic trimmer (Erie style 532-08-0R5).
L_1 — 13½ turns No. 20 tinned, ⅝-inch diam., ⅞ inch long, tapped at 4½ turns (B & W Miniductor No. 3007).
L_2 — 5 turns No. 20 tinned, ½-inch diam., ⅜ inch long (B & W Miniductor No. 3003).
L_3 — 2¾ turns similar to L_2.
L_4 — 2 turns No. 12 tinned, ¼-inch diam., ¼ inch long.
L_5 — 1 turn ins. wire between turns of L_4. May be inner conductor of shielded wire, with braid removed.
L_6 — Half-wave line, ¼-inch copper tubing, 7 3/16 inches long.
L_7 — Quarter-wave line, ¼-inch copper tubing, 5 inches long.
L_8 — Loop of insulated wire 1 inch long and ½ inch high projecting through base plate on which line assemblies are mounted. May be made from inner conductor of shielded wire, with braid removed from last two inches.
L_9 — 2 turns No. 22 enam. around cold end of L_{10}.
L_{10} — 6 turns similar to L_2.
L_{11} — 11 turns No. 22 enam. close-wound on ⅜-inch slug-tuned form (National XR-91).
L_{12} — 4 turns No. 28 silk or enamel wound over cold end of L_{11}.
J_1, J_2 — Coaxial fitting.
J_3 — Closed-circuit jack.
J_4 — 4-pin male chassis fitting.
RFC — 10 turns No. 22 tinned, ⅛-inch diam. Space turns diam. of wire.

the mixer and r.f. assembly. Its size and proximity to the mixer inner conductor are not particularly critical, as there is a surplus of injection under ordinary conditions of operation.

Adjustment

The first step in putting the converter into operation is to tune up the oscillator and multiplier stages. This process is similar to the adjustment of a transmitter and will not be detailed here. Check to see that the proper frequencies appear as indicated on the schematic diagram. Only enough power at 382 Mc. is needed to develop about 0.5 ma. of crystal current. Anything from 0.2 to 1.0 ma. is satisfactory. Adjustments should be made with no plate voltage on the r.f. stage.

Now connect the converter to a 50-Mc. receiver or converter and peak the i.f. amplifier circuits at about 52 Mc. on noise. Next apply plate voltage and feed a signal into the r.f. stage. Peak the r.f. and mixer capacitors for maximum response at about 434 Mc. These adjustments can be made on noise also, if the circuits were close to resonance originally. A noise generator will give the best check on converter performance, but the margin of signal over noise that is obtained on a received signal is also usable, if adjustments are made with care.

The points of connection for the B-plus and the coupling taps on the r.f. and mixer lines are critical adjustments, but if the dimensions given above are followed carefully the points should be close to optimum. Adjustments can be made and checked readily if the r.f.-mixer assembly is mounted in place temporarily with a few self-tapping screws.

V.H.F. Transmitters

Beginning with the v.h.f. region, amateur frequency assignments are not in direct harmonic relationship with our lower-frequency bands. This fact, coupled with the necessity for extreme care in selection and placement of components for low circuit capacitance and minimum lead inductance, makes it highly desirable to construct separate gear for v.h.f. work, rather than attempt to adapt for v.h.f. use a transmitter designed for the lower amateur frequencies.

Transmitter stability regulations for the 50-Mc. band are the same as for lower bands, and proper design may make it possible to use the same rig for 50, 28, 21, and even 14 Mc., but incorporation of 50 Mc. and higher in the usual multiband transmitter is generally not feasible. Rather, it is usually more satisfactory to combine 50 and 144 Mc., since the two bands are close to a third-harmonic relationship. At least the exciter portion of the transmitter may be made to cover the requirements for both these bands very readily.

Though no stability restrictions are imposed by law on operation at 144 Mc. and higher amateur bands (other than that the entire emission must be kept within the limits of the band in question), experience has demonstrated the value of using crystal control or its equivalent in v.h.f. work. Crystal-controlled transmitters and receivers having the minimum bandwidth necessary for voice communication make it possible for hundreds of stations to operate without undue interference in a band that would appear crowded if occupied by a dozen or less stations using broadband receivers and unstable transmitters.

The use of narrow-band communications systems also pays off in the form of improved efficiency in both transmitter and receiver. It is this factor, perhaps more than the interference potentialities of the wide-band systems, which makes it desirable to employ advanced techniques at 220 and even 420 Mc. Stabilized transmitters for 220 Mc. are not too difficult to build, and their use at this frequency is highly recommended.

Construction of multistage rigs for 420 Mc. is not easy, and the choice of tubes suitable for this type of work is quite limited, but the advanced amateur who is interested in making the most of the interesting possibilities afforded by this developing field will be satisfied with nothing less. The 420-Mc. band is much wider than our lower v.h.f. assignments, however, and interference is not likely to become a limiting factor in this band for a long time to come. Thus it may be more important, in many localities, to get activity rolling with any sort of gear, leaving perfection in design to come along as the need develops.

At 420 Mc. and in the higher amateur assignments most standard tubes cannot be used with any degree of success, and special tubes designed for these frequencies must be employed. These types have extremely close electrode spacing, to reduce transit-time effects, and are constructed with leads having virtually no inductance. Several more-or-less conventional tubes are now available which will operate with fair efficiency up to about 500 Mc., but best performance is obtained with the "lighthouse," "pencil tube," or coaxial-electrode types built especially for u.h.f. applications, and requiring specially-designed tank circuits.

Frequency modulation may be used throughout the v.h.f. and higher bands, wide-band emission being permitted above 52.5 Mc. and narrow-band FM anywhere. Where suitable receivers are available to make best use of such emissions, either wide-band or narrow-band FM can provide effective v.h.f. communication. Their use is particularly advantageous in congested areas where the freedom from interference to broadcast and television reception they enjoy may permit operation when an amplitude-modulated transmitter of any power would be a constant source of trouble.

Transmitter Technique

The low-power stages of a transmitter for the v.h.f. bands need not be greatly different in design from those used for lower bands, and many of the ideas in Chapter Six may be used to good advantage in the initial stages of the v.h.f. rig. The constructor has the choice of starting at some lower frequency, usually around 6, 8 or 12 Mc., multiplying to the operating frequency in one or more additional stages, or he can use a high initial frequency and thus reduce the number of multiplier stages required or eliminate them entirely. The first approach has the virtue of employing low-cost crystals, and it usually results in better stability when methods other than crystal control are used, but high-frequency crystals may effect a considerable economy in power consumption, an important factor in portable or emergency-powered gear.

A high starting frequency may be helpful in preventing TVI that can result from amplification of unwanted harmonics from a crystal oscillator on 6, 8 or 12 Mc. Several troublesome harmonics are eliminated if a crystal frequency of 24 Mc. or higher is used.

CRYSTAL OSCILLATORS

Crystal oscillator stages for v.h.f. transmitters may make use of any of the circuits shown in Chapter 6, when crystals up to 12 Mc. are employed, but certain variations are helpful for higher frequencies. Crystals for 12 Mc. or higher are usually of the overtone variety. Their frequency of oscillation is an approximate multiple of some lower frequency, for which the crystal is actually ground. Thus 24-Mc. crystals commonly used in 144-Mc. work are 8-Mc. cuts, specially treated for overtone characteristics. Until recent years such crystals were tricky in operation and subject to excessive drift if operated at high crystal current. The overtone crystals now being supplied are approximately as stable as those designed for fundamental operation, and they are easy to handle in properly designed circuits.

Best results are usually obtained with overtone crystals if some regeneration is added. This makes for easy starting under load and greater output than would be obtainable in a simple triode or tetrode circuit. Two regenerative circuits, with constants for 24- or 25-Mc. crystals, are shown in Fig. 17-1. Triodes are shown, but the same arrangement may be used with tetrode or pentode tubes. The important point in either case is the amount of regeneration, controlled by the position and number of turns in the feed-back winding, L_2, in Fig. 17-1-A or the position of the tap on L_1 in B. There should be only enough feed-back to assure easy crystal starting and satisfactory operation under load; too much will result in random oscillation not under the control of the crystal.

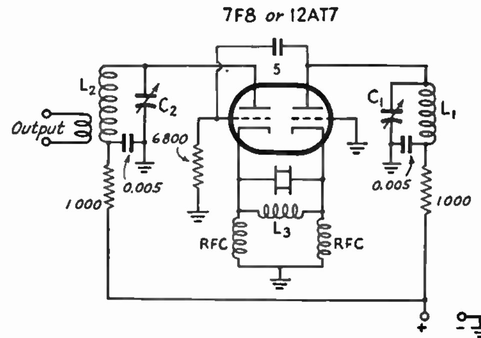


Fig. 17-2 — The functions of crystal oscillator, cathode follower and frequency multiplier are combined in this dual-triode circuit. The circuit L_1C_1 tunes to the desired overtone frequency, and L_2C_2 its second or third harmonic. L_3 should resonate with tube and crystal capacitance just below the frequency of oscillation. The value of the r.f. chokes in the cathode circuit is not critical. Values for obtaining 144-Mc. output with a 24-Mc. crystal are given below.

C_1 — 20-μμfd. variable.
C_2 — 10-μμfd. variable.
L_1 — 5 turns No. 18, ½-inch diam., ½ inch long.
L_2 — 2 turns No. 18, ½-inch diam., ½ inch long.
L_3 — 4 turns No. 18, ⅜-inch diam., ¼ inch long.

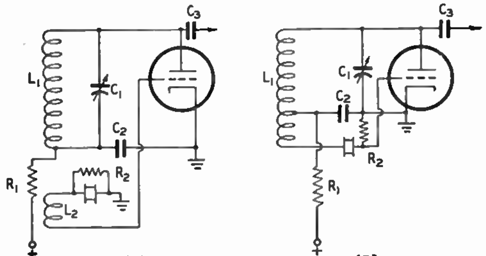


Fig. 17-1 — Regenerative crystal oscillator circuits for v.h.f. use. Feed-back is controlled by the position of L_2 with respect to L_1 in A, or by the position of the tap on L_1 in B. Constants below are for 24 to 27 Mc.

C_1 — 50-μμfd. variable.
C_2 — 0.005-μfd. ceramic or mica.
C_3 — 25-μμfd. ceramic or mica.
R_1 — Decoupling resistor, 1000 to 5000 ohms, carbon.
R_2 — Grid leak, to suit tube used.
L_1 (A) — 18 turns No. 18, ½-inch dia., 1¼ inches long.
L_2 (A) — 3 turns similar to A, mounted on same axis, about ⅛ inch apart.
L_1 (B) — 14 turns No. 18, ½-inch dia., 1 inch long. Tap at about 4½ turns (see text).

Overtone operation is possible with standard fundamental-type crystals, using the circuits of Fig. 17-1. Practically all will oscillate on their third overtones, and fifth and higher odd overtones may be possible. Adjustment of regeneration is more critical, however, if the crystals are not ground for overtone characteristics. It should also be noted that the frequency may not be an exact multiple of that marked on the crystal holder, so care should be used in working with crystals that are near a band edge.

Crystals ground for overtone service can be made to oscillate on other overtones than the one marked on the holder. A 24-Mc. crystal, actually an 8-Mc. cut, may be made to oscillate on 40, 56, 72 Mc. or even higher odd multiples of its 8-Mc. fundamental frequency. The circuits of Fig. 17-1 may be used, but for high-order overtones the dual triode circuit of Fig. 17-2 is more reliable. Values for achieving 144-Mc. output with a 24-Mc. crystal (9th overtone instead of 3rd) are given.

The crystal is resonated, by means of L_3 connected across it, at a frequency just *below* the desired overtone, or about 70 Mc. in this example. Circuit L_1C_1 tunes to the desired overtone, 72 Mc.; L_2C_2 to a harmonic, in this case 144 Mc. Regeneration is controlled by varying the coupling between L_1 and L_3, so that only crystal oscillation is developed. Polarity of these windings is important; bringing them closer should *reduce* the tendency to self oscillation.

Crystals are now available for frequencies up to around 100 Mc. They are somewhat more expensive than those for 30 Mc. and lower, however, so they have not been used widely in amateur work, except where a saving in power is important. Use of 50-Mc. crystals is made occasionally as a means of preventing radiation of the harmonics of lower frequency crystals that might cause interference to television reception.

FREQUENCY MULTIPLIERS

Frequency multiplying stages in a v.h.f. transmitter follow standard practice, the principal precaution being arrangement of components for short lead length and minimum stray capacitance. This is particularly important at 144 Mc. and higher. To reduce the possibility of radiation of oscillator harmonics on frequencies that might interfere with television or other services, the lowest satisfactory power level should be used. Low powered stages are easier to shield or filter, in case such steps become necessary.

Common practice in v.h.f. exciter design is to make the tuned circuits capable of operation over the whole range from 48 to 54 Mc., so that the output stage can drive either a 50-Mc. amplifier or a tripler from 48 to 144 Mc. Tripling is often done with push-pull stages, particularly when the output frequency is to be 144 Mc. or higher. The output capacitances of the tubes in such a circuit are in series, permitting a better L/C ratio than is possible with single-ended circuits.

AMPLIFIERS

Most transmitting tubes now used by amateurs will work on 50 Mc., but for 144 Mc. and higher the tube types are limited to those having low input and output capacitances and compact physical structure. Leads must be as short as possible, and soldered connections should be avoided in high-powered circuits, where heating may be great enough to reach the melting point of the solder used.

Plug-in coils and their associated sockets or jack bars are generally unsatisfactory for use at 144 Mc. and higher because of the stray inductance and capacitance they introduce. One way around this trouble is the dual tank circuit shown in Fig. 17-3. Here the tank circuit for 144 Mc. is a conventional tuned line, with its shorting bar made removable by plugs or clips. When the stage is to be used on another band the shorting bar is removed and a coil is plugged into the jack bar, the line then serving as a pair of plate leads.

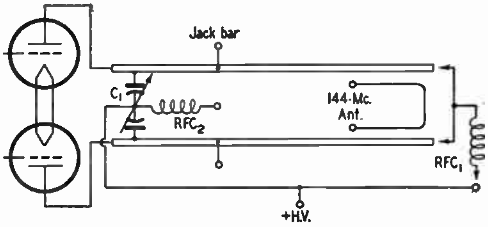


Fig. 17-3 — **An efficient two-band tank circuit for 50 and 144 Mc. For operation on 144 Mc. the shorting bar is plugged into the end of the line. For 50 Mc. a suitable tank coil is plugged into the jack bar. The line then serves merely as a pair of plate leads. RFC_1 is a 144-Mc. choke; RFC_2 a 50-Mc. choke. The split-stator variable, C_1, tunes either circuit.**

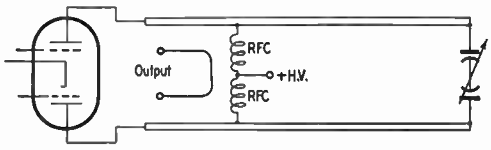


Fig. 17-4 — **Half-wave line tank circuit, for use at 220 or 420 Mc., where tube and circuit capacitances prohibit the use of an ordinary tuned circuit. Plate voltage is fed into the line at the point of lowest r.f. voltage (see text).**

Such an arrangement will operate as efficiently on 144 Mc. as if it were designed for that band alone, yet it can be made to work properly on any lower band.

At 220 Mc. and higher it may be necessary to employ half-wave lines as tuned circuits, as shown in Fig. 17-4. Here the tuning capacitance, instead of being connected directly in parallel with the

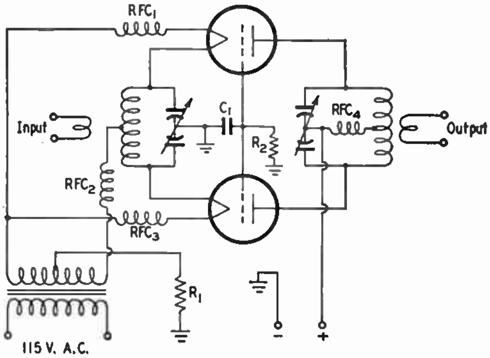


Fig. 17-5 — **Grounded-grid r.f. amplifier. Driving voltage is fed into the cathode circuit, with the control grids maintained at ground potential.**

output capacitance of the tube, is at the far end of a half-wave line. Plate voltage is fed into the line near the middle, at the point where the r.f. voltage is lowest. The proper point can be located by first operating the stage with the voltage fed in near the middle of the line, and then touching a pencil point along the line to locate the spot where the least effect on the grid or plate current is noted. This check should be made with the pencil in an insulating mount, if dangerous values of plate voltage are used.

Neutralization of triode amplifiers for 50 and 144 Mc. can follow standard practice, but the stray inductance and capacitance introduced by the neutralizing circuits may be excessive for 220 Mc. and higher. In such instances grounded-grid amplifiers may be used as shown in Fig. 17-5. Driving power is applied to the cathode circuit, with the grid acting as a shield. Grounded-grid amplifiers are stable, but they require high driving power. Some of the drive appears in the output, so both the driver and amplifier must be modulated when amplitude modulation is used. For this reason the grounded-grid amplifier is used mainly for FM applications.

Tetrode and pentode amplifiers may operate without neutralization, but it is advisable to

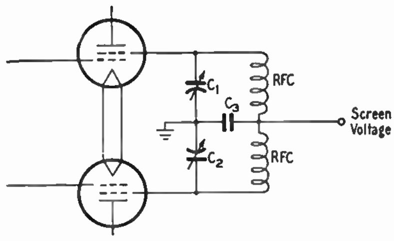


Fig. 17-6 — Tuned screen circuit for stabilizing a v.h.f. tetrode push-pull amplifier. C_1 and C_2 may be the two halves of a split-stator variable condenser, if the circuit is symmetrical electrically. The r.f. choke and condenser values vary with frequency, making this form of neutralization essentially a one-band device. C_3 should be about 0.001 μfd. for v.h.f. applications.

plan for it in the original layout. With such tubes as the 829 or 832 enough neutralizing capacitance can be obtained by running short lengths of stiff wire up through the chassis alongside the tube plates, crossing them over to the opposite grid terminals below the chassis. Neutralization is adjusted by trimming or bending the wires.

Instability may show up in tetrode amplifiers as the result of ineffective screen by-passing, in which case conventional cross-over neutralization will accomplish little or nothing. The solution lies in series-resonating the screen circuits to ground, as shown in Fig. 17-6. A small split-stator variable can be used for C_1 and C_2 if the layout is completely symmetrical. The r.f. choke and condenser values vary with frequency, so screen neutralization is essentially a one-band device.

FREQUENCY MODULATION

Though FM has not enjoyed great popularity in v.h.f. operation, probably because of lack of suitable receivers in most v.h.f. stations, its possibilities should not be overlooked, particularly for the higher bands. At 420 Mc., for instance, the efficiency of most amplifiers is so low that it is often difficult to develop sufficient grid drive for proper AM service. With FM any amount of grid drive may be used without affecting the audio quality of the signal, and the modulation process adds nothing to the plate dissipation. Thus considerably higher power can be run with FM than with AM before damage to the tubes develops or the signal is of poor quality.

Frequency modulation also simplifies transmitter design. The principal obstacle to greater use of FM in v.h.f. work is the wide variation in selectivity of v.h.f. receivers, making it difficult for the operator to set up his deviation so that it will be satisfactory for all listeners.

TVI PREVENTION AND CURE

Interference to television reception is not ordinarily so serious a problem with v.h.f. gear as with equipment for lower amateur bands, where more harmonics of the operating frequency fall within the television channels. The principal causes of TVI from v.h.f. transmitters are as follows:

1) Adjacent-channel interference in Channel 2 from 50 Mc.

2) Fourth harmonic of 50 Mc. in Channels 11, 12 or 13, depending on the operating frequency.

3) Radiation of unused harmonics of the oscillator or multiplier stages. Examples are 9th harmonic of 6 Mc., and 7th harmonic of 8 Mc. in Channel 2; 10th harmonic of 8 Mc. in Channel 6; 7th harmonic of 25-Mc. stages in Channel 7; 4th harmonic of 48-Mc. stages in Channel 9 or 10; and many other combinations. This may include i.f. pick-up, as in the cases of 24-Mc. interference in receivers having 21-Mc. i.f. systems, and 48-Mc. trouble in 45-Mc. i.f.'s.

4) Fundamental blocking effects, including modulation bars, usually found only in the lower channels, from 50-Mc. equipment.

5) Image interference in Channel 2 from 144 Mc., in receivers having a 45-Mc. i.f.

6) Sound interference (picture clear in some cases) resulting from r.f. pick-up by the audio circuits of the TV receiver.

There are many other possibilities, and u.h.f. TV in general use will add to the list, but nearly all can be corrected completely, and the rest can be substantially reduced.

Items 1, 4 and 5 are receiver faults, and nothing can be done at the transmitter to reduce them, except to lower the power or increase separation between the transmitting and TV antenna systems. Item 6 is also a receiver fault, but it can be alleviated at the transmitter by using FM or c.w. instead of AM 'phone.

Treatment of the various harmonic troubles, Items 2 and 3, follows the standard methods detailed elsewhere in this *Handbook*. It is suggested that the prospective builder of new v.h.f. equipment familiarize himself with TVI prevention techniques, and incorporate them in new construction projects.

Use as high a starting frequency as possible, to reduce the number of harmonics that might cause trouble. Select crystal frequencies that do not have harmonics in TV channels in use locally. Example: The 10th harmonic of 8-Mc. crystals used for operation in the low part of the 50-Mc. band falls in Channel 6, but 6-Mc. crystals for the same frequency range have no harmonic in that channel.

If TVI is a serious problem, use the lowest transmitter power that will do the job at hand. Much interesting work can be done on the v.h.f. bands with but a few watts output, particularly if a good antenna system is used.

Keep the power in the multiplier and driver stages at the lowest practical level, and use link coupling in preference to capacitive coupling, particularly in the later stages.

Plan for complete shielding and filtering of the r.f. sections of the transmitter, should these steps become necessary.

Use coaxial line to feed the antenna system, and locate the radiating portion as far as possible from TV receivers and antenna systems.

A Complete Transmitter for 144 Through 21 Mc.

The rack-mounted equipment shown in Fig. 17-7 is an example of the way in which the low-power stages of a rig can be designed to provide for several bands. Each piece of equipment can be used alone, or they combine readily to cover 21, 28, 50 and 144 Mc., at a power level approaching the legal maximum.

At the bottom is a VFO unit tailored to the needs of the v.h.f. man, but useful on lower frequencies as well. Next is an exciter capable of up to 40 watts output on 21, 28 or 48 to 54 Mc. It is a fine low-powered rig for use on 15, 10 or 6 meters as well. Above the exciter are two units designed for high-power operation on 144 and 50 Mc.

THE EXCITER

The transmitter-exciter shown in Figs. 17-8 through 17-10 was designed for the v.h.f. man who likes to work some of the lower bands as well. It delivers up to 40 watts output on 21, 28 or 50 Mc., and covers the range down to 48 Mc. so that it may be used as a source of excitation for additional stages that multiply to 144 Mc. Though it was intended for use with the high-powered amplifiers described later, it may be used effectively as a complete transmitter in itself.

Shielding for TVI reduction was achieved by building the unit inside a standard aluminum chassis. Each power lead is by-passed at the power plug, and all wiring was done with shielded wire. Output is taken off through a coaxial fitting, so that a low-pass filter can be inserted in the line for harmonic attenuation if needed.

Circuit Details

The exciter circuit follows standard practice. The oscillator is a 5763 grid-plate type with provision for 10 crystals and VFO input. Crystals may be in the 3.5-, 6-, 7-, 8-, 14- or 24-Mc. ranges. On 21 Mc. the oscillator output is on the signal frequency, and best results are obtained with 7-Mc. crystals, tripling in the plate circuit. For 28 Mc. the oscillator doubles to 14 Mc. with 7-Mc. crystals, quadruples from 3.5 Mc., or works straight through with 14-Mc. overtone crystals. For operation on 50 or 144 Mc., the oscillator output is on 24 to 27 Mc., quadrupling, tripling or working straight through, for 6-, 8- or 24-Mc. crystals, respectively. The 100-$\mu\mu$fd. tuning capacitor at C_6 tunes the oscillator plate circuit from 14 to 27 Mc., so no bandswitching is needed in this stage.

Another 5763 follows the oscillator, working straight through on 21 Mc., or doubling to 28 or 48 to 54 Mc. Two coils, L_2 and L_3, and a 50-$\mu\mu$fd. condenser, C_{10}, cover 21 to 30 Mc., and 48 to 54 Mc., respectively. In case trouble is encountered in making the 5763 run stably as a 21-Mc. amplifier, a third switch position is available for connecting a damping resistor, R_8, in series with L_2.

The output stage uses a 6146, with a tapped coil for 21 and 28 Mc., and a second coil for 48 to 54 Mc. Output coupling links in these two

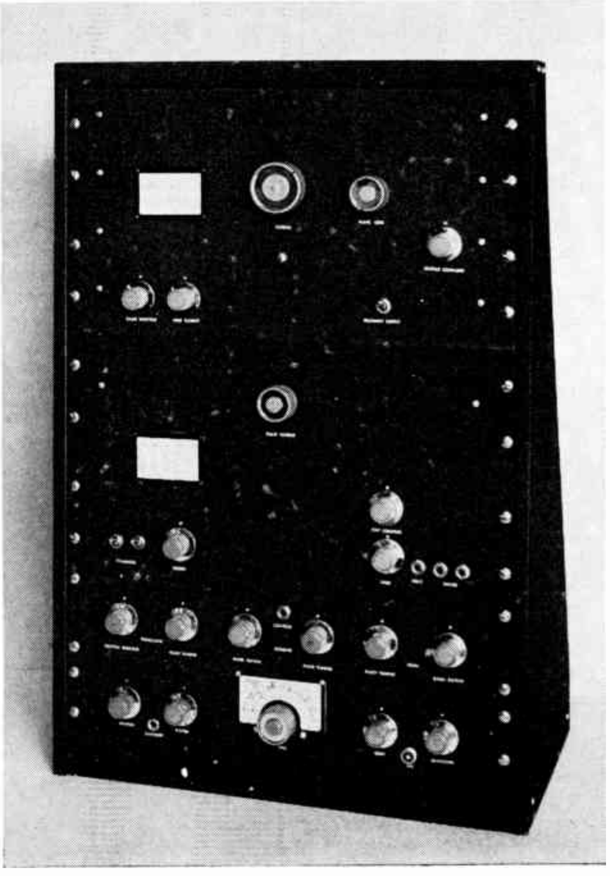


Fig. 17-7 — A complete transmitter for 144 through 21 Mc. The four units are, from the bottom up, a VFO with reactance modulator; an exciter-transmitter with up to 40 watts output; a tripler-driver-amplifier for 144 Mc.; and a shielded amplifier for 50, 28 and 21 Mc.

Fig. 17-8 — Looking into the bandswitching exciter-transmitter from the top front. Oscillator components are in the left compartment, the doubler and power connector in the center, and the output stage at the right. Note that the 6146 socket is mounted inside the output stage compartment.

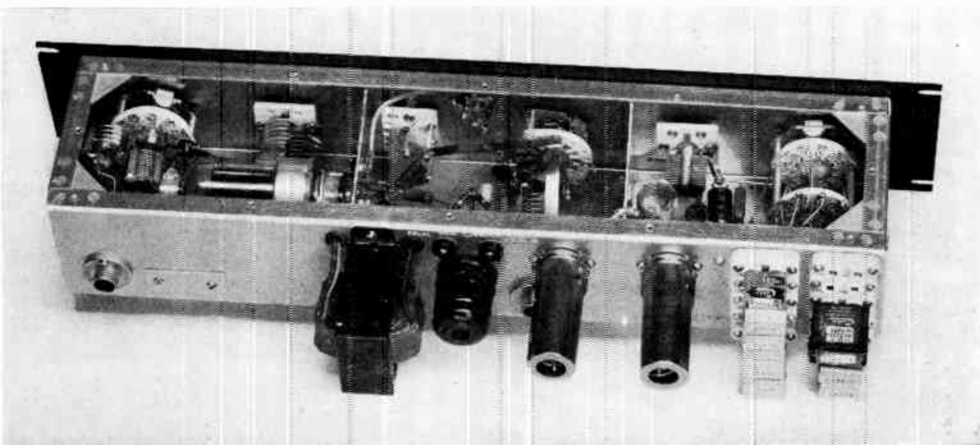
coils are also switched. The 6146 works nicely over a wide range of plate voltages, so this rig may be used in exciter service with as little as 300 volts on the final, or it may be used as a complete transmitter at up to 500 volts. A 2E26 may be used in the final stage where its power output is adequate for the job at hand.

The exciter is built largely inside a 3 × 5 × 17-inch aluminum chassis and is fitted with a standard 3½-inch rack panel. Only the crystals, the first two tubes and the filament transformer are outside, and these are mounted on the rear wall of the chassis to keep down the vertical dimension.

Arrangement of parts is not particularly critical, the principal consideration in the first two stages being to mount the tubes in such position that the coupling lead (C_{25} to the grid of the second 5763) is short. The grid circuit of the second stage should be isolated from the rest of the components to reduce the tendency toward self-oscillation when the stage is operated straight through on 21 Mc. The lead to the grid is made with a short piece of RG-59/U coax, run through a slot in the top of the partition, and a small piece of flashing copper is soldered across the 5763 socket between Pins 1 and 9 to isolate the input and out circuits further. Leads from the tube plate to the bandswitch, S_2, and thence to the tuning condenser, C_{10}, are made with ¼-inch-wide copper strap, to hold down lead inductance.

Note the method of mounting the socket for the 6146. Contrary to common practice, this socket is mounted on the *tube side* of the partition. Cathode, heater and screen pins (Nos. 1, 3, 4, 6 and 7) are by-passed individually to separate points on the partition with the shortest possible leads. Heater and cathode leads are brought through the partition with shielded wire, and the control grid and screen leads are run through on short lengths of stiff wire insulated with spaghetti sleeving. Mounting the 6146 socket inside the final stage compartment provides a short plate-

Fig. 17-9 — Rear view of the exciter. On the rear wall at the right are 10 crystal sockets of various types. Then come the two 5763s, the power plug, the filament transformer, and the output coaxial fitting. On the inside front wall are, in the same order, the crystal switch, oscillator tuning, doubler bandswitch, doubler tuning, and final bandswitch.



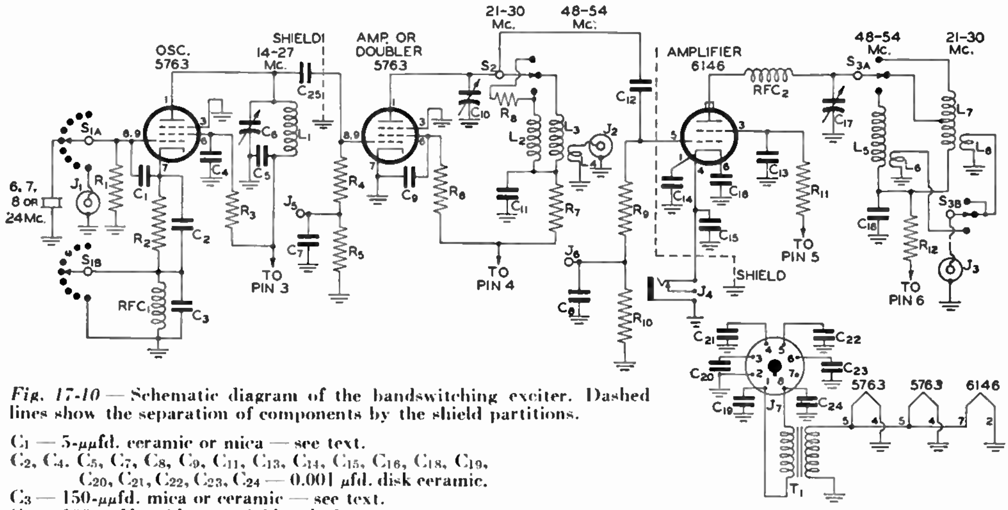


Fig. 17-10 — Schematic diagram of the bandswitching exciter. Dashed lines show the separation of components by the shield partitions.

C_1 — 5-μμfd. ceramic or mica — see text.
C_2, C_4, C_5, C_7, C_8, C_9, C_{11}, C_{13}, C_{14}, C_{15}, C_{16}, C_{18}, C_{19}, C_{20}, C_{21}, C_{22}, C_{23}, C_{24} — 0.001 μfd. disk ceramic.
C_3 — 150-μμfd. mica or ceramic — see text.
C_6 — 100-μμfd. midget variable, shaft-mounting type.
C_{10} — 50-μμfd. midget variable, shaft-mounting type.
C_{12} — 15-μμfd. mica or ceramic.
C_{17} — 20-μμfd. double-spaced midget variable, shaft-mounting type.
C_{25} — 50-μμfd. ceramic or mica.
R_1, R_4 — 0.1 megohm, ½ watt.
R_2 — 220 ohms, ½ watt.
R_3, R_6 — 22,000 ohms, 1 watt.
R_5, R_{10} — 1000 ohms, ½ watt.
R_7 — 100 ohms, ½ watt.
R_8 — 7.5 ohms 1 watt (two 15-ohm ½-watt resistors in parallel).
R_9 — 33,000 ohms, 1 watt.
R_{11} — 20,000 ohms, 10 watts.
R_{12} — 68 ohms, ½ watt.
L_1 — 8½ turns No. 20 tinned, ¾-inch diam., ½ inch long (B & W Miniductor No. 3011).
L_2 — 7 turns like L_1, ⅞ inch long.
L_3 — 4 turns No. 20 tinned, ⅝-inch diam., ½ inch long (B & W No. 3006).
L_4 — 2 turns No. 18 push-back, ⅝-inch diam., coupled to cold end of L_3.
L_5 — 4 turns No. 20 tinned, ¾-inch diam., ½ inch long (B & W No. 3010).
L_6 — 4½ turns No. 20 tinned, ½-inch diam., ½ inch long, mounted inside cold end of L_5. (B & W Miniductor No. 3003.)
L_7 — 11 turns like L_1, tapped at 7 turns, ¾ inch long.
L_8 — 9 turns B & W No. 3004, ½-inch diam., ⅝ inch long, mounted inside cold end of L_7.
J_1, J_2, J_3 — Coaxial fitting. J_1 is for VFO input.
J_4 — Closed-circuit jack.
J_5, J_6 — Tip jack.
J_7 — 8-pin male chassis fitting.
RFC_1 — 2.5-mh. r.f. choke (National R-100-S).
RFC_2 — Parasitic choke, 6 turns No. 20 enamel, ¼-inch diam., ⅜ inch long.
S_{1A}, S_{1B} — 11-position 2-section ceramic wafer switch. (Made from centralab P-122 index assembly and 2 centralab type Y switch sections. Complete assembly CRL 2513.)
S_2 — Similar to above, but single section (CRL 2501 on 2503, wafer type X or Y).
S_{3A}, S_{3B} — Same but 2-pole 3-position single section (CRL 2505, wafer type RR).
T_1 — 6.3-v. 3-amp. filament transformer.

to-cathode return. The stage may possibly be unstable if the socket is mounted on the opposite side of the partition from the tube, as is usually done.

The three tuning condensers should be the shaft-mounting type, not the sort that mount on small pillars. Unless the rotor shaft is grounded solidly to the panel it will act as an "antenna" to radiate harmonic energy that is almost certain to cause TVI. The meter tip jacks, J_5 and J_6, may also turn out to be harmonic radiators, unless by-passed right at the point where they come through the rear wall.

The output coupling links, L_6 and L_8, are the smallest diameter B & W Miniductor, which makes a close fit inside the larger size used for L_5 and L_7. They are held in place with household cement. A coupling link is also provided for L_3, so that a small amount of power can be taken off at 48 Mc. if desired. This is made of self-supporting stiff insulated wire, coupled closely to the cold end of L_3.

Note that the front-panel appearance is completely symmetrical, the controls being spaced at regular intervals horizontally, and in the center of the panel vertically. The chassis is bottom up, with the cover at the top. This allows ready access to the inside when the unit is in its normal operating position, but it may be used the other side up, if the builder so desires. Ventilation of the 6146 is afforded by twenty ¼-inch holes drilled in the top and bottom surfaces over and under the tube.

Testing and Use

For initial tests a power supply delivering 200 to 250 volts is adequate. Each stage has its plate-screen power lead brought out to the plug separately, so that individual metering is possible. Applying voltage through Pin 3, we note that the stage draws low current until oscillation is obtained, because of the cathode bias. Plug a low-range meter into J_5 to read the grid current of the following stage, and tune C_6 for maximum indication, which will be about 0.5 to 1 ma. at normal operating voltage. The oscillator plate-screen current will be around 20 ma.

Should the oscillator refuse to start, try other crystals, and then experiment with the values of C_1 and C_3. The grid-to-cathode capacitor, C_1, may not be necessary, particularly if crystals no lower than 6 Mc. are used. Use the lowest value that will permit oscillation with all crystals. The value of C_3 may be critical when overtone-type crystals are used. Improper values at either of these positions may result in intermittent oscillation, or none at all.

Check the output frequency with a calibrated wavemeter, or by listening with a receiver whose calibration can be relied upon, and proceed to the following stage. Plug the grid meter into J_6, apply power through Pin 4, and check the output frequency when C_{10} is tuned for maximum grid current. At least 2 ma. should be available. Check for self-oscillation by removing excitation. Should self-oscillation occur on the 21-Mc. range, switch in the damping resistor, R_8. This should be the lowest value permissible, as the output from the stage drops rapidly as the series resistance is increased above a few ohms.

When around 2 ma. of grid current is obtained the output stage may be checked. This may be done initially with 250 to 300 volts applied through Pins 5 and 6, using a 25-watt lamp plugged into J_3 for a dummy load. Cutting the excitation (do it only briefly — 6146s draw a tremendous amount of plate current!) should result in zero grid current. If the stage is operating correctly the output should be around 15 watts with 300 volts on the plate.

Increasing to 400 to 450 volts it should be possible to get at least 35 watts output on all frequencies. In an enclosed layout of such small dimensions it is not advisable to go much beyond this level, as the heat dissipation may be high enough to damage the small coils used. Where the exciter is used to drive a high-powered tetrode final stage, 300 volts on the 6146 and 200 to 250 volts on the 5763s is plenty. The rig may be used as a complete transmitter, modulating the output stage on 28 or 50 Mc., at 30 to 50 watts input. The operating conditions in all stages can be adjusted to suit the builder's own requirements by varying the screen resistor values. The exciter is keyed in the 6146 cathode lead for c.w. operation.

A 144-MC. DRIVER-AMPLIFIER

Shown just above the exciter in the composite photograph, Fig. 17-7, and separately in Figs. 17-11 through 17-13 is a three-stage tripler-driver-amplifier for high-power operation on 144 Mc. It may be used with any exciter that is capable of delivering 5 watts or more on 48 Mc. If a 2-meter exciter is available the tripler may be omitted. The driving power required in that case would be about 10 watts on 144 Mc.

As may be seen from the schematic diagram, Fig. 17-12, a push-pull tripler stage with a pair of 5763s drives a tetrode amplifier using an AX-9903/5894A, which, in turn, drives a pair of 4-125As in the final stage. Input to the final can be up to slightly over 600 watts on AM 'phone, or 750 watts on c.w. By suitable adjustment of the grid drive and the final-amplifier screen and plate voltages, the input can be run as low as 150 watts with good efficiency. Some method of varying the input is recommended, as much of the operation on 144 Mc. can be carried on satisfactorily with moderate power.

Electrical and Mechanical Details

The tripler uses two tubes in push-pull in preference to a single tube, as this allows the tubes to be operated at low input and still deliver adequate drive to the succeeding stage without critical adjustments. The tripler grid circuit is self-resonant. The tripler and driver plate tuning adjustments are ganged. Straps of flashing copper 5/16 inch wide are used for the leads from the 5763 plates to the tuning condenser, C_1, to hold down lead inductance.

From the bottom view, Fig. 17-13, it will be seen that sheets of flashing copper are fastened to the bottom of the chassis, covering the area of the driver and final stages, to improve grounding circuit conductivity. Note that the rotor of the driver tuning condenser, C_2, is grounded through a 100-ohm resistor, R_5. This was done to cure a 250-Mc. parasitic oscillation. Ventila-

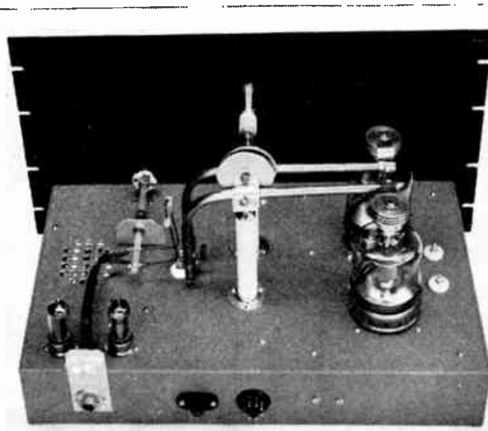


Fig. 17-11 — Rear view of the 4-125A amplifier for 144 Mc., showing details of the parallel-line plate circuit. The 5763 tripler tubes are at the left. Note ventilation holes, below which is mounted the driver tube, out of sight under the chassis.

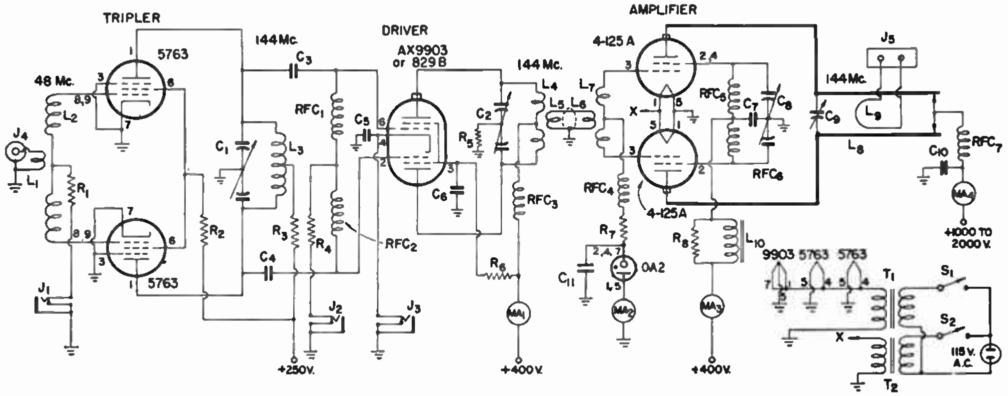


Fig. 17-12 — Wiring diagram and parts list for the high-powered 144-Mc. transmitter.

C_1, C_2 — 10-$\mu\mu$fd.-per-section butterfly variable (Cardwell ER-6-BF/S. Johnson 10LB15 alternate: see text).
C_3, C_4 — 10-$\mu\mu$fd. mica.
C_5, C_6 — 0.001-μfd. disk ceramic.
C_7 — 0.005-μfd. disk ceramic.
C_8 — 50-$\mu\mu$fd.-per-section split-stator variable (made from Millen 19140; see text).
C_9 — Plate-line tuning adjustment (made from neutralizing condenser; see text).
C_{10} — 0.001-μfd. 5000-volt mica.
C_{11} — 0.25-μfd. tubular.
R_1 — 150,000 ohms, 1 watt.
R_2 — 18,000 ohms, 1 watt.
R_3 — 100 ohms, ½ watt.
R_4 — 10,000 ohms, 1 watt.
R_5 — 100 ohms, 1 watt.
R_6 — 10,000 ohms, 10 watts.
R_7 — 5000 ohms, 10 watts.
R_8 — 27,000 ohms. Use only if needed; see text.
L_1 — 1 turn No. 14 enam., ¾-inch diam.
L_2 — 6 turns each side of center, No. 20, ⅝-inch diam., spaced wire diam., ¼-inch space at center for L_1 (B & W Miniductor No. 3007).
L_3 — 2 turns No. 14 enam., spaced ⅛ inch, ½-inch diam.
L_4 — 2 turns No. 14 enam., spaced ⅜ inch, 1⅜-inch diam.
L_5 — 2 turns No. 18 push-back, close-spaced, inserted between turns of L_4.
L_6 — Loop of No. 14 enam., 4 inches long, inside L_7.
L_7 — Copper strap 5/16 inch wide and 8 inches overall from grid to grid; see text and bottom-view photograph.
L_8 — Plate line, ⅜-inch o.d. copper tubing 12 inches long, spaced 1⅜ inches center-to-center. Bend on 1-inch radius to make inverted "L" 4½ inches high.
L_9 — Output coupling loop, made from 13½-inch piece of No. 14 enam. Sides ⅞ inch spaced. Vertical portion 2½ inches high.
L_{10} — 5-hy. (min.) choke, 100 ma. or more rating.
J_1, J_2, J_3 — Closed-circuit jack.
J_4 — Coaxial fitting.
J_5 — Crystal socket for output terminal.
MA_1, MA_2, MA_3, MA_4 — External meters, not shown in photographs, 200, 50, 100 and 500 ma., respectively.
RFC_1, RFC_2, RFC_3, RFC_4, RFC_7 — 1.8-μhy. solenoid v.h.f. choke (Ohmite Z-144).
RFC_5, RFC_6 — 7-μhy. solenoid v.h.f. choke (Ohmite Z-50).
S_1, S_2 — S.p.s.t. toggle switch.
T_1 — 6.3-volt 4-amp. filament transformer.
T_2 — 5-volt 13-amp. filament transformer (Chicago FO-513).

tion for the driver tube is provided by drilling holes through the copper plate and chassis over the tube. An 829B may be used in place of the 9903/5894A, with some sacrifice in driver stage efficiency.

If the 9903 is used, the tube plate leads should be very pliable material, as the tube structure is fragile. The 5894A, an improved version of the 9903, is considerably more rugged mechanically. If standard heat-dissipating connectors are used they should be filed down by about one-third of their diameter because of the close pin spacing. Cardwell butterfly capacitors were used for C_1 and C_2 because of their inherent provision for ganging. Other types such as the Johnson 10LB15 can be substituted by soldering a ganging extension to the rear end of the rotor shaft of C_2.

The driver plate and final grid circuits are widely separated so that coupling between them will be confined to the link circuit. This helps to keep unwanted harmonics from being transferred to the final grids. This potential source of TVI can be further reduced by installing link-coupled tuned circuits in the tripler plate and driver grid positions, if the station location is one where 192-Mc. energy might cause TVI in Channels 9 or 10.

The relatively high input and output capacitances of the 4-125As rule out conventional coil-and-condenser circuits at 144 Mc., so no grid tuning capacitor is used in the final stage, and only a very small variable capacitance is used in the plate circuit. The entire grid circuit is made of 5/16-inch-wide copper strap. Two pieces each 1½ inches long connect the grid terminals to feed-through bushings that are provided for mounting neutralizing tabs, if needed. The center portion of the grid circuit is an egg-shaped loop mounted on the feed-throughs, as seen in the bottom view. The bushings are mounted near the inner corners of the 4-125A sockets. The holes for them are drilled larger than needed to pass the ceramic portions, to keep the grid-to-ground capacitance at a minimum.

The principal neutralizing adjustment is the split-stator variable condenser, C_8, connected from the screens to ground. A single-section variable (Millen 19140 or Hammarlund MC-140) having supports at each end of the rotor shaft was modified for this purpose as these types provide a symmetrical path from rotor to ground

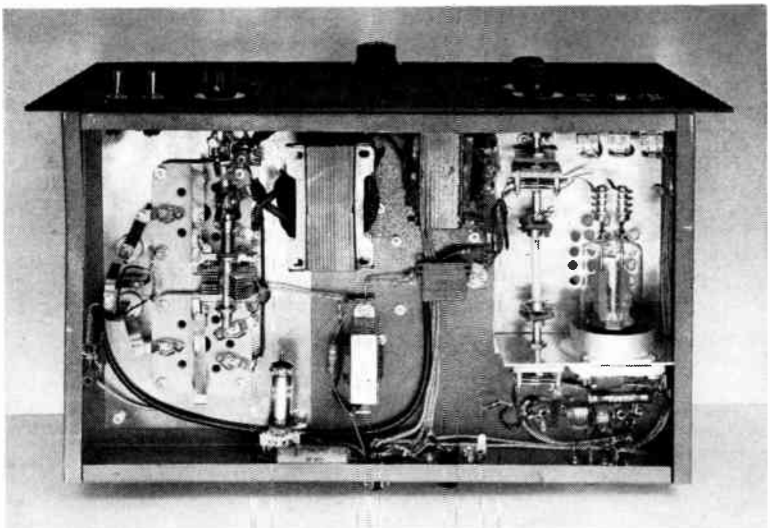


Fig. 17-13 — Looking under the chassis of the high-power 2-meter rig. At the lower right are the components of the tripler stage, with the AX-9903 driver tube just above the aluminum partition. The 4-125A sockets, grid circuit, and screen-neutralization capacitor are at the left. The VR-tube bias system is mounted on the rear chassis wall.

for each side of the circuit. A strip of brass or aluminum is first screwed to the metal mounting brackets at each end, tying them together electrically and mechanically. Then the stator bars are sawed in half, leaving an equal number of plates on each side. These condensers have 9 plates each on stator and rotor originally. The middle stator plate is cut out and the front rotor plate removed, leaving a split-stator condenser with 4 plates on each stator and 8 on the rotor. The two screen terminals on each socket are strapped together, and the connection to the stators of C_8 is made with copper strap. Symmetry and low inductance are extremely important in this circuit.

The screen circuit also includes two solenoid-type r.f. chokes connected directly to the screen terminals. These are under C_8 and do not show in the bottom view. Their common connection is by-passed, and a small filter choke is connected in the screen voltage lead for modulation purposes. The screen variable capacitor is driven through two universal joint couplings to bring the drive shaft out to a point that provides a pleasing front panel appearance.

Fixed bias for the final stage is provided without use of batteries or an external supply by inserting a voltage regulator tube in series with the grid leak and by-passing the tube with a low-leakage capacitor. When the gas tube fires with application of excitation, C_{11} charges. Removing excitation stops the current flow through the VR tube and leaves the charge in C_{11} applied to the 4-125A grids. This cuts off the plate and screen current until the charge in C_{11} leaks off. The cut-off time varies with the leakage characteristics of C_{11} and associated components, and some experimentation may be necessary. An external bias source of 90 volts or more may, of course, be substituted.

The construction of the final plate circuit is obvious from the top-view photograph. The tuning device, C_9, is made from parts of a standard neutralizing capacitor (Millen 15011) mounted on 4-inch ceramic stand-offs (National GS-4) in the center of the chassis. The lead screw on the adjustable plate is extended by means of a short length of ¼-inch diameter brass rod soldered to its end, and this is connected through an insulating coupling and a polystyrene rod to a knob on the front panel. This tuning arrangement provides no logging scale or reset indicator of any sort, but it results in a very worth-while improvement in tank-circuit efficiency over conventional tuning methods.

The copper tubing tank circuit is mounted in place by means of straps of aluminum wrapped around the lines and fastened to the top of the stand-offs. Connection to the tube plates is made with ¾-inch-wide copper straps that are bolted to the plate lines. No solder is used anywhere in this plate line assembly: the heat dissipated at the tube end of the line would be sufficient to melt soldered connections. The heat-dissipating connectors for the 4-125A plates were cut down to four fins high to reduce plate lead length. Just beyond the stand-off insulators and C_9 the plate lines are bent to a vertical position around a radius of about one inch, the bottom of the line ending about a half inch above the chassis. Here an adjustable strap of flashing copper is wrapped around the lines, and an r.f. choke is connected through a lug to a feed-through bushing carrying the high-voltage d.c. The by-pass, C_{10}, is under the chassis.

Details of the antenna coupling loop are visible in the top view. The pick-up loop is made adjustable by mounting it through a polystyrene rod that can be rotated from the front panel. This rod passes through a shaft bearing and a tension adjusting device (National SB and Millen 10061) mounted on a small aluminum bracket. Note that a short length of rod is fastened at the top of the loop, so that no adjustment of the coupling will allow it to come in contact with the line electrically.

Adjustment and Operation

This rig contains its own filament transformer so only plate and screen supplies are external. These should be capable of furnishing 250 volts at 75 ma. for the tripler, 400 volts at 200 ma. for the driver, 300 to 400 volts at 75 ma. for the final screens, and 1000 to 2000 volts at 400 ma. for the amplifier plates. The screens of the final and the driver plates may be run from the same supply, though a more flexible set-up is possible if the voltage applied to the final screens is adjustable separately.

The tripler should be tuned up first. Plug a low-range milliammeter in the tripler grid current jack, J_1, and apply grid-drive through a coaxial cable and J_4. Adjust the spacing between the two halves of the grid coil, L_2, and the position of L_1, for maximum grid current. This should be 1 to 2 ma. Transfer the meter to the driver grid jack, J_2, and apply plate voltage through R_3, tuning C_1 for maximum grid current, which should be between 3 and 5 ma. The inductance of L_3 should be adjusted so that the low end of the band is reached with C_1 set somewhere between the mid-point and the maximum end of its range. Total plate-screen current to the 5763s need not be more than about 50 ma.

Next, tune C_2 through resonance and note whether the grid current changes. Should it dip down at resonance the stage will require neutralization. This is unlikely with the 9903 or 5894A, however, as these tubes are designed to be inherently neutralized at frequencies around 150 Mc. Next, plug a 200-ma. meter into J_3, or connect one externally in series with the plate-screen supply, as shown in Fig. 17-12, and apply plate voltage, preferably with a lamp load coupled to L_4. If the stage is working correctly, it should be possible to light a 40-watt lamp to full brilliance. Check for self-oscillation by removing excitation briefly. To protect the driver tube, it might be well to make these initial tests at 250 volts or so, increasing to 400 to 500 volts only when the stage is found to be working correctly.

Next, couple the output from the driver stage to the grid circuit of the final, by means of a coaxial cable and L_5 and L_6. The latter should be the same general shape as L_7, and mounted inside or just above it, with about ⅛-inch separation. The resonant frequency of the grid circuit can be changed slightly by altering the shape of the grid inductance. Squeezing the sides together raises the frequency; making the tank more nearly round lowers it. When the circuit is properly resonated, it should be possible to develop 25 to 30 ma. grid current, measured in series with the VR tube and ground (MA_2 in Fig. 1). The setting of the screen-to-ground capacitor, C_8, will affect the grid current, but it may be set approximately to the proper point by adjusting it for maximum grid current with the plate voltage off. The total plate and screen current should be 175 to 200 ma. When the coupling loops at both ends of the coax have been adjusted so as to give maximum grid current, adjust the turn spacing of L_4 so that its tuning capacitance will be the same as that of C_1. The two condensers may then be ganged by means of flexible couplings and an insulating shaft.

Now connect a 100-watt lamp at the output terminals and apply about 500 volts to the final plates and 200 or less to the screens, metering both circuits as shown in the schematic diagram. Adjust C_9 for maximum output, watching the grid and plate meters. Move the setting of the screen adjustment in small steps until maximum output, minimum plate current, and maximum grid current all occur at the same setting of the plate tuning. This is the screen adjustment at which the amplifier will operate most stably. Neutralization can also be done by running the amplifier without excitation, adjusting C_8 until there is no evidence of oscillation, but this gives a broader indication than the first method.

Should it be impossible to achieve complete stability by the screen adjustment alone, it may be necessary to add grid-plate capacitance by mounting stiff wires or tabs on the feed-through bushings. In this amplifier, the capacitance added by the feed-through rods alone was just about the right amount, however. This is not the conventional cross-over neutralization, but rather additional *grid-plate* capacitance. The amount of capacitance added is adjusted in the same way as for triode neutralizing circuits of the crossover type.

Once the amplifier is stabilized at low voltages, proceed to final checks at normal plate and screen operating conditions. A suitable load for high-power tests is something of a problem, as no lamp combination represents a load that simulates an antenna system at this frequency. A fair load can be made, however, by connecting three or four 100-watt lamps in parallel. Lamps larger than the 100-watt variety are useless for load purposes, as they tend to develop filament hot spots and burn out before reaching anything like normal brilliance.

A method of varying the screen voltage continuously is extremely useful at this juncture, as the final tubes can be made to draw any desired plate current by suitable variation of the screen voltage. Screen dissipation should be watched closely to see that it does not run much over 20 watts in plate-modulated service or 30 watts on c.w., and it is strongly recommended that a screen-current meter be made a permanent part of the metering system. Efficient operation is possible over a range of 800 to 2500 volts on the plates.

The tetrode amplifier with separate screen voltage supply should *never* be operated without load, or with no plate voltage applied. Screen dissipation is certain to be excessive in either case and tube damage or failure is invited.

Tests with the lamp load should be monitored for freedom from modulation. With some types of chokes for L_{10}, there may be a tendency to oscillation at some audible frequency. Should this develop, it can be damped by loading the choke slightly with a resistor, as shown by R_8 in Fig.

17-12. The highest value of resistance that will stop the oscillation should be used, if any is necessary. Substituting another choke is a better method. It should have a minimum of 5 henrys inductance, but a wide variety of small filter chokes may be satisfactory.

In general the manufacturer's typical operating conditions for the 4-125As can be followed with good results, but many variations are possible. In v.h.f. work there is no need to run high power at all times, so provision should be made to drop the plate and screen voltages. Efficient operation at plate voltages as low as 800 is possible, if the screen voltage is altered in proportion. Considerable latitude in grid drive is also possible. The principal precaution is to see that none of the tube elements is operated above the maximum safe dissipation given in the manufacturer's literature.

A FINAL AMPLIFIER FOR 50, 28 AND 21 MC.

The top unit in the rack of v.h.f. equipment, Fig. 17-7, shown in detail in Figs. 17-14 through 17-16, is a high-powered companion to the exciter described earlier. It covers the same three bands, with a maximum power rating of 600 watts input on AM 'phone, or 800 on c.w., and may be used with any exciter capable of delivering 15 to 25 watts output in the proper frequency range. It is completely shielded, for TVI reduction, and may be changed from band to band without opening the enclosure.

The plate circuit is a pi network, with a variable inductor as the main element. Conventional bandswitching is employed in the grid circuit. Parasitic suppression and neutralizing methods are the principal departures from familiar practice. The aluminum enclosure calls for forced-air cooling.

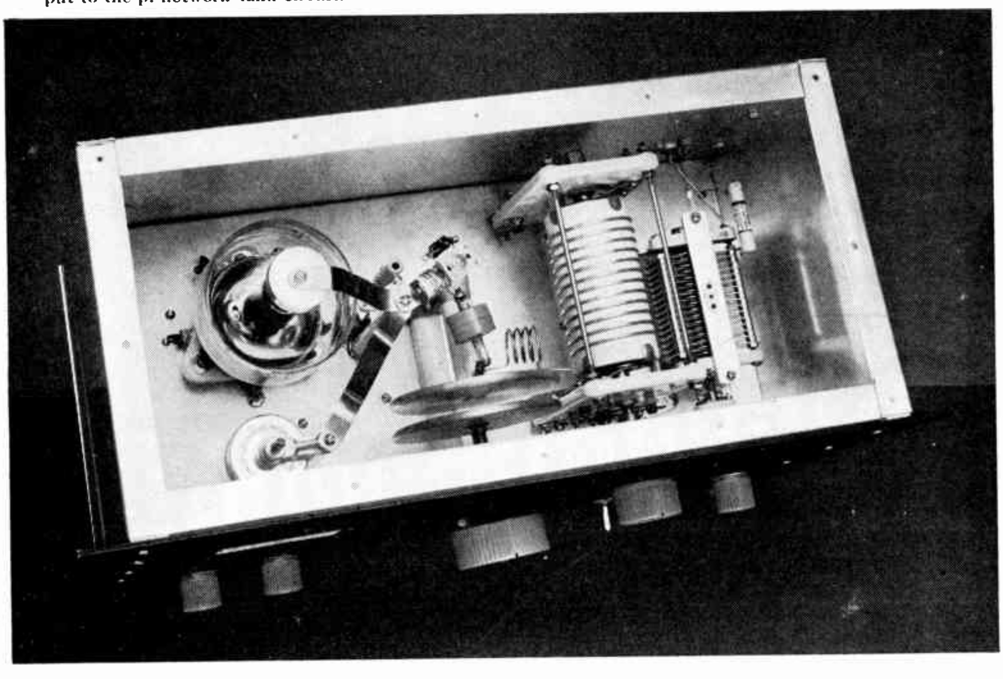
Electrical and Mechanical Features

Looking into the top of the amplifier, as in Fig. 17-14, we see the 4-250A tetrode tube at the left. Just below it is the neutralizing capacitor. At the center of the chassis is the input tuning condenser, C_9, of the pi-network tank circuit, with the variable inductor at its right. The variable condenser at the far right is the output condenser, C_{10}. The small components to the right of the tube comprise the parasitic suppression circuit. The coupling capacitor, C_8, and the 50-Mc. auxiliary coil, L_8, are near the center of the photograph. Grid-circuit components are visible in the bottom view, along with the filament transformer, cooling fan, and modulation choke.

In order to obtain a satisfactory tuning range and minimum stray inductance, a large neutralizing-type condenser is used for tuning the input to the pi-network plate circuit. The capacity range is about 5 to 20 $\mu\mu$fd. The output tuning range needed for C_{10} is roughly 50 to 150 $\mu\mu$fd., so a conventional transmitting variable may be used. With a properly matched load the r.f. voltage across J_2 is low, and a plate spacing of 0.047 inch is adequate, even with high power.

The variable inductor assembly has considerable stray capacitance, which would make it

Fig. 17-14 — Looking inside the 3-band amplifier. Note the neutralizing condenser used for tuning the input to the pi-network tank circuit. The small air-wound coil, center, is the 50-Mc. portion of the tank, L_8.



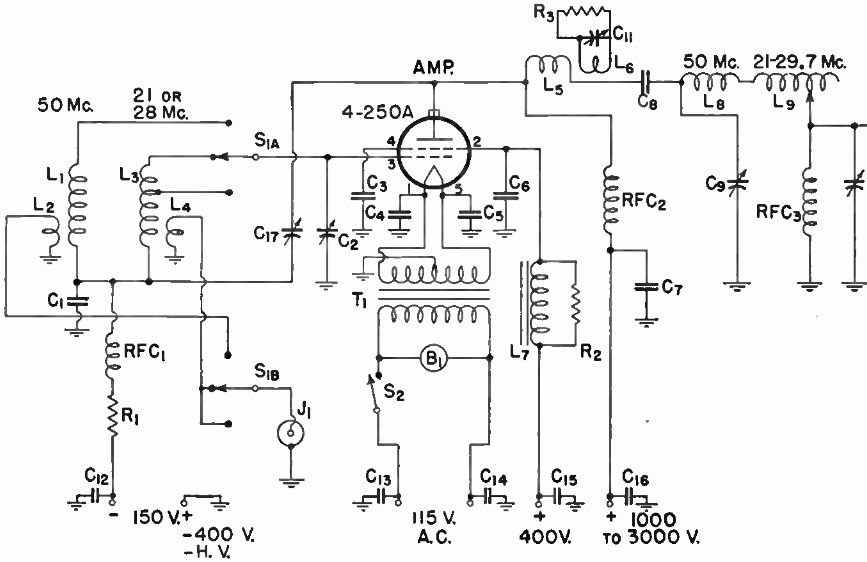


Fig. 17-15 — Schematic diagram and parts list for the 4-250A amplifier.

C_1 — 220-μμfd. silver mica.
C_2 — 30-μμfd. miniature variable, double-spaced (Hammarlund HF-30-X, shaft-mounted).
C_3, C_4, C_5, C_6, C_{12}, C_{13}, C_{14}, C_{15} — 0.001-μfd. disk ceramic.
C_7, C_8, C_{16} — 500-μμfd. 10,000-volt ceramic (Centralab TV3-501).
C_9 — 5–20-μμfd. disk-type variable (National NC-500 neutralizing condenser, with mounting bracket reversed).
C_{10} — 200-μμfd. variable, 0.047-inch spacing (National TMK-200).
C_{11} — 3–30-μμfd. mica trimmer.
C_{17} — 2–8-μμfd. neutralizing condenser (National NC-800A).
R_1 — 10,000 ohms, 5 watts.
R_2 — See text — use only if needed.
R_3 — Approximately 100 ohms, 6 watts (three 330-ohm 2-watt resistors in parallel).
L_1 — 2½ turns No. 20 tinned, ¾-inch diam.; turns spaced ⅛ inch (B & W Miniductor No. 3010).
L_2 — 4 turns B & W No. 3004 cemented inside cold end of L_1.
L_3 — 8 turns No. 20 tinned, ¾-inch diam., 9/16 inch long, tapped at 6 turns (No. 3011).
L_4 — 7 turns B & W No. 3001 cemented inside cold end of L_3.
L_5 — 3 turns No. 16 tinned, spaced 1/16 inch, on ½-inch diam. ceramic stand-off, 1 inch long.
L_6 — 2 turns similar to L_5, and about ¼ inch away from it on same form.
L_7 — 10-hy. 100-ma. filter choke.
L_8 — 4 turns No. 14 tinned, ⅝-inch diam., spaced ⅛ inch.
L_9 — 6.2-μh. variable inductor (B & W No. 3851).
B_1 — Blower motor and fan (Allied Catalog Nos. 72–702 and 72–703).
J_1, J_2 — Coaxial fitting, female.
RFC_1, RFC_2, RFC_3 — 20-μh r.f. choke (Ohmite Z-28).
S_{1A}, S_{1B} — 2-pole 3-position ceramic wafer swltch (Centralab 2505, wafer type RR).
S_2 — Single-pole single-throw toggle switch.

impossible to develop proper circuit Q at 50 Mc. if the variable coil alone were used, so a small air-wound coil, L_8, is connected ahead of the variable unit. Its inductance is such that only a small portion (one turn or less) of L_9 is used at 50 Mc.

Parallel feed of the high voltage, through RFC_2, permits the tank circuit to be operated with no d.c. applied to its components. The purpose of RFC_3 is to provide a path to ground for the high voltage in case C_8 should break down. The coils L_5 and L_6, the capacitor C_{11}, and the resistor R_3 comprise a parasitic-suppression circuit that will be discussed later.

The grid circuit is largely self-explanatory, with the possible exception of the neutralizing method used. C_1 and C_{17} make up a capacity bridge, by means of which energy is fed back into the grid circuit from the plate. In this method, C_1 has a critical value. It should be such that the amplifier can be neutralized with C_{17} at approximately the midpoint of its range. It is possible that some variation in layout might eliminate the need for neutralization, though provision should be made for it when the amplifier is built.

Note that the 4-250A socket is mounted above the chassis, with the control grid toward the front. It is raised so that the prongs just clear the chassis. Each contact, with the exception of the control grid, is then by-passed individually to the chassis with the shortest possible leads.

The screen voltage is obtained from a separate source, in preference to the use of a dropping resistor connected to the plate supply. The modulation choke, L_7, should have a minimum of 10 henrys inductance, and a current-carrying capacity of about twice the expected screen current. The resistor connected across the choke should be added only if needed to suppress "singing" resulting from choke resonance in the audio range. It should be the highest value that will stop such tone modulation of the transmitted signal.

Arrangement of parts should be such that r.f. leads are short, and copper or silver strap should be used in preference to wire in r.f. circuits wherever it is mechanically feasible. The by-pass, C_7, and the blocking capacitor, C_8, are high-voltage ceramic units of the type used in TV receiver power supplies. The parasitic-suppression circuit and the parallel-feed r.f. choke are mounted on a ceramic pillar made from two 3-inch stand-off insulators. The r.f. choke should be as far from the tube envelope as possible, to prevent blistering of the paint by heat radiated from the tube.

The filament transformer, modulation choke, grid-circuit components and cooling fan are mounted below the chassis, which is a standard 3 × 10 × 17-inch job. The fan may be placed at any point where the blades can rotate close to an intake hole. If this is not possible, a duct just larger than the area of the fan blades can be used to channel the air to the fan. The blades must be bent so that air will be drawn inward. Holes in the chassis just below the tube socket and in the top cover over the tube provide the only air path out of the enclosure. Any other holes should be plugged, and the shielding of the upper portion of the amplifier should make a good fit to the chassis. Circulation may be checked by placing a smoke source near the intake hole. The smoke should be drawn in rapidly, flowing out through the top holes only. A light piece of paper placed over the holes in the top cover should rise perceptibly when the fan is started.

The shielding of the main assembly is made in four pieces, fitted to the front, back and sides of the chassis. The edges are folded over three quarters of an inch and drilled and tapped, or the assembly may be made with self-tapping screws. The entire job should make good contact electrically and mechanically, if cooling and TVI prevention measures are to be effective.

Adjustment and Operation

Initial tests may be made on the amplifier with the parasitic suppression and neutralizing circuits omitted, though both will probably be needed. Start with resistor bias only, as instability will be more evident if the plate current is not cut off in the absence of excitation. The plate and screen voltages should be such that the dissipation by these elements is below the permissible maximum for the tube. A suitable load for the first tests can be made by connecting three 100-watt lamps in parallel at J_2.

With a 25- or 50-ma. meter connected between R_1 and ground, apply plate and screen voltages (but not grid drive) and watch for signs of grid current. If any appears it will indicate oscillation, either a v.h.f. parasitic, or tuned-plate tuned-grid feed-back near the operating frequency. If a v.h.f. parasitic is encountered, it can be suppressed with the *LCR* combination shown in the schematic diagram. L_6 and C_{11} tune to the parasitic frequency. L_5 should be as low inductance as possible, in order to keep the frequency of the parasitic high. The lower the parasitic frequency the greater will be the 50-Mc. energy dissipated in the suppression circuit. With the values given in the parts list there is no overheating of the resistors by dissipation of 50-Mc. energy, yet the loading at the parasitic frequency is sufficient to prevent oscillations from starting up, if the tuning of C_{11} and the coupling between L_5 and L_6 are adjusted carefully.

A check on the need for neutralization may be made by operating the amplifier normally and observing the grid and plate currents simul-

Fig. 17-16 — **Bottom view of the amplifier for 50, 28 and 21 Mc., with bottom cover removed. Note method of mounting the ventilating fan. The chassis should be made as nearly airtight as possible, except for the fan hole and holes drilled under the tube socket. Air is thus drawn in through the base and forced up around the base seal of the tube, leaving through holes in the top cover. Screening of the fan hole may be required for TVI prevention.**

taneously. Maximum grid current and minimum plate current should occur at the same setting of C_9. If the grid current rises as the plate circuit is tuned to the high-frequency side of resonance, more neutralizing capacitance is needed. If neutralization cannot be achieved at any setting of C_{17} it may be necessary to use a different value of capacitance at C_1. Perfect neutralization may not be possible on all three bands with one setting of C_{17}, but it should be possible to find a satisfactory compromise.

With the amplifier operating stably, actual on-the-air conditions can be set up. The typical operating conditions given by the tube manufacturer can be used as a guide, but any of the values can be varied considerably, provided the maximum safe figure for each of the tube elements is not exceeded. Thus it may be desirable to lower the grid bias when operating at low plate voltage, in order to get the amplifier to draw more plate current. As little as 1000 volts on the plate works well, provided that the grid drive and screen voltage are properly altered.

If the antenna system has an open-wire or other balanced line, the output of the amplifier should be fed through an antenna coupler that provides for coaxial input and balanced output. A low-pass filter can then be used, if needed, between the amplifier and the antenna coupler, to reduce harmonic radiation that might cause TVI.

Though the adjustments are not critical, there are certain optimum values of C_9 and L_9. Their selection is explained in the discussion of tank circuit Q elsewhere in this *Handbook*. Capacitance required at C_9 will be of the order of 7 to 12 μμfd. for 50 Mc., 10 to 15 for 28 Mc., and around 20 μμfd. for 21 Mc. This will be nearly "all out" for 50 Mc., near the midpoint for 28, and down to about ¼ inch for 21. The variable coil can be adjusted for resonance for each band, and the approximate number of turns required can be logged for future reference. Logging of settings for C_9 can be done similarly. Adjustment of the variable coil should be made at low power level, to avoid arcing at the contact surface and possible damage to the roller and coil.

The capacitance needed at C_{10} will be about 50 μμfd for 50 Mc., 100 for 28 and 150 for 21 Mc. Adjustment of this control is similar to the use of the familiar swinging link. It is an output coupling adjustment only, and either L_9 or C_9 should be reset for resonance whenever C_{10} is varied. Adjustment should be made with a standing-wave bridge connected in the coaxial line between J_2 and the antenna coupler, taking care to see that the load is properly matched.

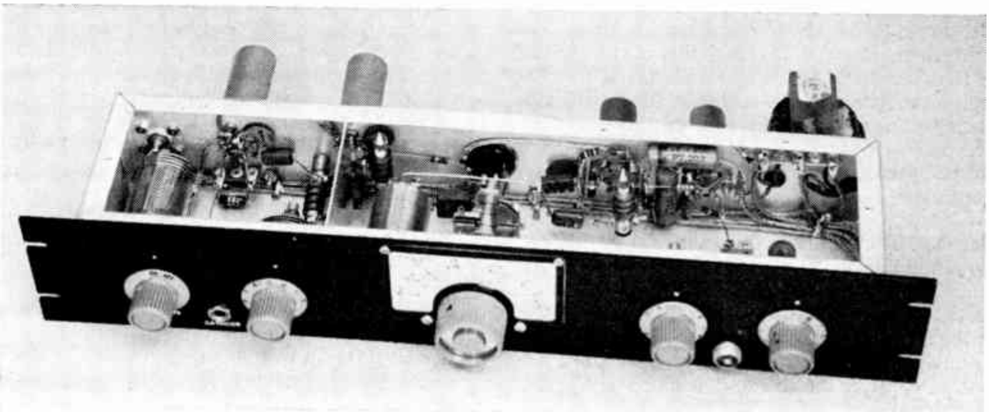
● A V.H.F. MAN'S VFO

The frequency-control unit shown in Figs. 17-7 and 17-17–17-19 is designed for the v.h.f. operator, though it may be used on all bands from 3.5 Mc. up as well. When used with the other equipment described in these pages it converts the crystal oscillator stage of the exciter to a frequency multiplier. The VFO unit has a speech amplifier and a reactance modulator for narrow-band FM built in.

The oscillator is a 5763, with a series-tuned Colpitts circuit having a tuning range of 3000 to 4000 kc. Its plate circuit is untuned, and the output is fed to another 5763 that serves as either amplifier or doubler. The plate circuit of the second stage may be tuned to the oscillator frequency or to its second harmonic.

With the values given in the parts list, one sweep of the vernier dial tunes the oscillator from 3000 to 3713 kc., with a little leeway at each end. The second stage is normally tuned from 6000 to 7425 kc., taking care of the 21-, 27-, 28-, 50- and 144-Mc. requirements of the complete station as desired. By resetting the band-set condenser, C_2, slightly the oscillator range can be extended to 4000 kc., permitting use of the VFO over the entire 3.5-Mc. band, as well as the 7- and 14-Mc. bands if the user so desires.

Fig. 17-17 — Top view of the VFO unit, with cover removed. Speech-amplifier and reactance-modulator components are at the right, with the oscillator tuning condenser and coil near the center. An aluminum partition divides the oscillator socket. The amplifier stage is at the left end.



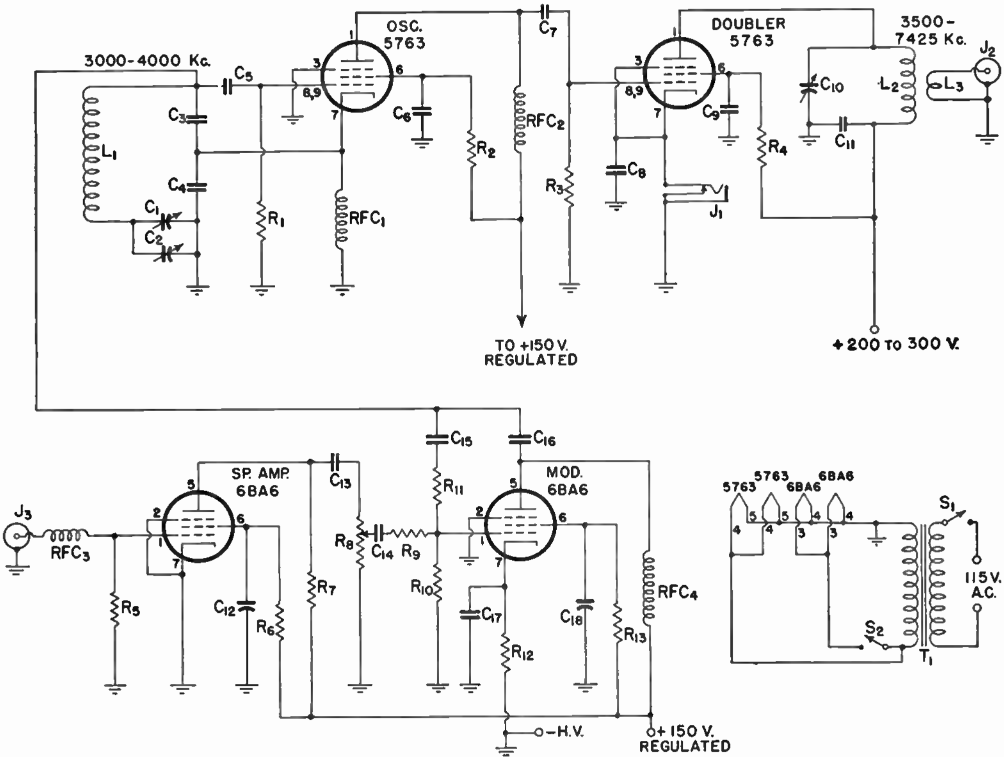


Fig. 17-18 — Schematic diagram and parts list for the VFO and reactance modulator.

C_1, C_2 — 50-μμfd. variable with rotor bearing at each end of shaft (Hammarlund MC-50). Remove plates in C_1 for desired bandspread — see text.
C_3, C_4 — 680-μμfd. silver mica.
C_5, C_{15}, C_{16} — 47-μμfd. silver mica.
C_6, C_8, C_9, C_{11}, C_{13}, C_{17} — 0.01-μfd. disk ceramic.
C_7 — 25-μμfd. ceramic or mica.
C_{10} — 140-μμfd. variable (Hammarlund MC-140).
C_{12}, C_{14}, C_{18} — 0.1-μfd. tubular.
R_1 — 68,000 ohms, ½ watt.
R_2 — 1000 ohms, ½ watt.
R_3 — 33,000 ohms, ½ watt.
R_4 — 22,000 ohms, 1 watt.
R_5 — 1 megohm, ½ watt.
R_6, R_{10}, R_{11} — 0.47 megohm, ½ watt.
R_7 — 0.22 megohm.
R_8 — 0.5-megohm potentiometer, with switch.
R_9 — 0.1 megohm, ½ watt.
R_{12} — 820 ohms, ½ watt.
R_{13} — 10,000 ohms, ½ watt.
L_1 — 40-μh. 25-watt transmitting coil (B & W Baby Inductor, type 80M, with plug-in base removed)
L_2 — 14-μh. 25-watt transmitting coil, end-linked (B & W type 40-MEL, with plug-in base removed).
L_3 — 4-turn link, part of L_2 assembly.
J_1 — Closed-circuit jack.
J_2, J_3 — Coaxial fitting, female.
RFC_1, RFC_2, RFC_4 — 2.5-mh. r.f. choke, stand-off type (National R-100S or R-100U).
RFC_3 — 2.5-mh. r.f. choke (National R-100).
S_1 — S.p.s.t. switch, shaft type.
S_2 — Switch or gain control, R_9.
T_1 — 6.3-volt 3-amp. filament transformer (Chicago FO-63).

Construction

Mechanically, the VFO is similar to the exciter, in that it is built inside a standard 3 × 4 × 17-inch aluminum chassis, with the tubes and filament transformer projecting from the rear wall. This makes a compact shielded unit that mounts on a 3½-inch rack panel. Looking into the top front view, Fig. 17-17, we see the oscillator tuning condenser, C_1, at the center, driven by the vernier dial. The oscillator inductance is to the left. An aluminum partition splits the oscillator tube socket, with pins 4 to 7 on the right side of the partition. Components of the output stage are at the far left. On the right side are the reactance modulator and speech-amplifier sockets, the deviation control, the band-set condenser, C_2, and the microphone jack.

The inductances in both stages are made from commercial plug-in coil assemblies. The plug-in bases are removed, and the coils mounted on pillars. The oscillator coil should have at least one half its diameter in all directions clear of metal objects of appreciable size. Wiring should be done with stiff wire, and all components connected with the oscillator circuit should be mounted rigidly.

Where the cable between the VFO and the following equipment is very short, the output from J_2 may be fed directly into the crystal socket. For more remote operation it may be necessary to install a tuned circuit and link coupling at the exciter end in order to insure efficient transfer of energy between the two units.

The reactance modulator follows standard practice. The gain of the first 6BA6 stage is suffi-

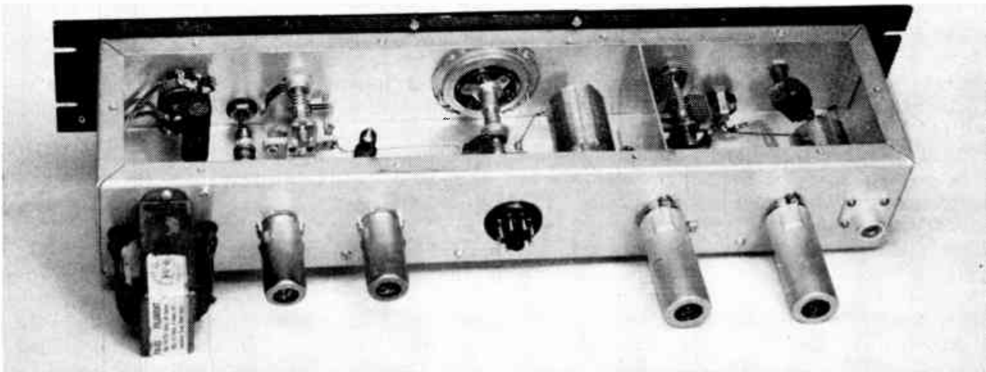


Fig. 17-19 — Looking into the VFO from the rear. The variable condenser at the left is C_2, for setting the band on the vernier dial. The large variable at the right allows the output circuit to be tuned to the oscillator frequency or its second harmonic.

cient to permit NFM operation on 10, 6 or 2 meters, with a crystal microphone. With the method of connection between the modulator and the oscillator shown in the schematic, the deviation is too low for use on frequencies lower than the 27-Mc. band. More deviation can be obtained by connecting the lead from the coupling capacitors, C_{15} and C_{16}, to the stators of C_1 and C_2, instead of across the tuned circuit. If the FM is to be used only above 27 Mc., however, the method shown is recommended.

Provision is made for turning off the heaters of the 6BA6s when the FM portion of the VFO is not in use. There is some frequency shift when the heaters are turned on and off in this way, however, and if the user expects to change frequently from FM to other modes it would be well to have S_2 break the B-plus lead, rather than the heaters. Where the deviation control is connected in the reactance-modulator grid circuit, as is done here, a blocking capacitor, C_{14}, must be added in series with the arm of the potentiometer. Otherwise, variation of the control will affect the frequency of the oscillator.

Operation

Deviation should be adjusted by listening to the signal on the band where the transmitter is to be used, as it increases with each frequency multiplication. Monitoring the signal is easy, as the proper harmonic of the VFO can be used, and all the rest of the rig left inoperative, thus preventing blocking of the receiver. Deviation requirements of various receivers will vary widely, but a safe starting point is to set the control so that speech sounds clean in a communications receiver with its crystal filter in the broadest "on" position.

The VFO dial (National MCN) can be calibrated with the aid of a receiver capable of tuning the oscillator or doubler range. Set the vernier dial so that the variable condenser is at maximum. Then adjust the bandset condenser until the oscillator frequency is 3000 kc. Check the tuning range before removing plates from C_1. The tuning range can be made to cover 3000 to 4000 kc. without resetting the bandset condenser, or if the user is interested in the v.h.f. bands only, it can be reduced to 3000 to 3375 kc., multiples of which cover the 50- and 144-Mc. bands. Plates can be removed from C_1, one at a time, resetting C_2 each time so that the frequency of the oscillator is 3000 kc. with C_1 at maximum, and checking the tuning range on the calibrated receiver. To cover 3000 to 3713 kc., C_1 was reduced to 3 stator and 2 rotor plates.

To use the VFO with the exciter described earlier, no more than 150 to 200 volts is needed on the second stage. Cathode current, metered at J_2, will be around 10 ma. when the doubler plate circuit is tuned to resonance. At this low input the tuning is unimportant, so long as the stages following receive sufficient excitation. It is not necessary to retune the doubler plate circuit for frequency shifts normally made within any one band.

The construction of the VFO is such that there should be little frequency drift due to heating as the tubes are operated far below ratings, and being mounted outside the main assembly they cause little temperature change in the frequency-controlling elements of the oscillator circuit. No special TVI precautions were taken, other than the shielding inherent in the design, and the use of shielded wire for all power wiring.

It is important that the power supply used on the VFO and modulator be well filtered and free from hum. Particularly where FM is used, the slightest a.c. ripple will show up in objectionable proportions. With sufficient filtering in the power supply, the note should be nearly comparable to crystal control, even on the v.h.f. range.

Note that no mention is made of keying the VFO unit. Experience has shown that oscillator keying results in too much frequency shift to be usable in v.h.f. work without precautions that are out of line for a simple unit such as this. In v.h.f. work, at least, keying should be done two stages or more away from the oscillator unless extensive stability measures are taken.

Transmitter-Exciters for 50 and 144 Mc.

The units shown in Figs. 17-20 through 17-25 are designed to serve several purposes. They may be used individually or together, depending upon whether the builder wishes to operate on both 50 and 144 Mc. or on either band alone. They may serve as complete transmitters for either mobile or home-station service, or they may be used as exciters for driving higher powered stages. The dual tetrode amplifier of Fig. 17-26 would be a suitable following stage for up to 100 watts input.

Overtone oscillator circuits are employed in the interest of low power consumption, circuit simplicity and ease of TVI prevention. Power wiring is done with shielded wire, and the physical arrangement of the parts is such that nearly complete shielding is obtained. If further enclosure is needed to prevent TVI it is merely necessary to cover the top of the unit. Power output is taken off by means of coaxial fittings, for convenience in mobile operation, and for complete shielding.

The two units are as similar, both mechanically and electrically, as possible. Both are built entirely on their 5 × 10-inch sheet aluminum top plates. These are screwed onto inverted 3 × 5 × 10-inch steel or aluminum chassis. Both use a 12AU7 dual triode as oscillator and frequency multiplier, with a 2E26 final amplifier. The 144-Mc. unit has a 5763 doubler stage between the 12AU7 and the 2E26, and the operating conditions of the stages vary somewhat.

The necessary driving power for the final is more readily obtained on 50 Mc., so the oscillator-multiplier is set up to run at lower input. Inductive neutralization (L_4 and L_5 in Fig. 17-22) was used to stabilize the 50-Mc. unit, whereas a small capacitance accomplishes the same end in the 144-Mc. amplifier. An end-linked tank circuit works well on 50 Mc., but a balanced tank with center link is more satisfactory for 144 Mc.

Both transmitters are set up to permit complete metering of all stages. Looking at the male chassis fittings in the schematic diagrams, it may be seen that each grid return, screen and plate lead is brought out to a separate pin. It is helpful during the adjustment of the rigs to be able to meter each stage without breaking into the main wiring. This is done by connecting a meter temporarily between the proper power plug pins. After adjustment is completed the meter can be replaced with a jumper in the plug. The exciter stages require 250 to 300 volts. The amplifier may be operated at the same level, or if more power is wanted the final plate voltage may be raised to 400 volts.

Adjustment and Operation

With either rig the oscillator stage should be checked first. This should be done with 150 to 200 volts until correct operation is established, and with no voltage on the following stages. Proper operation of the oscillator depends on the amount of feed-back, which can be adjusted by varying the position of L_2 with respect to L_1, or by changing the number of turns in either winding. For best mechanical stability, the two coils are made from a single piece of B & W Miniductor, breaking the wire to give the specified number of turns in each winding. Because the characteristics of tubes and crystals vary somewhat, it is well to start with at least one extra turn on each winding.

The feed-back should be only enough to insure easy starting of the oscillator under load. Adjustments should be made with the grid circuit of the following stage completed, with a low-range milliammeter connected to the proper terminals on the plug to read grid current. Oscillation will be evidenced by the sudden appearance of grid current as C_1 is rotated. If the feed-back is correct, this will occur at only a small portion of the tuning range of C_1. Listen to the oscillation at 24 or 25 Mc. It should vary only slightly in frequency, if at all, as C_1 is tuned. If the frequency changes gradually across the tuning range the oscillator is not crystal controlled, and too much feed-back is indicated. Remove a turn at a time from L_2 until only crystal-controlled oscillation remains. If there is insufficient feed-back there will be no oscillation. Feed-back can be increased by removing turns from L_1, or adding turns to L_2. If several crystals are available, try to find a median setting that will work with all of them.

Crystals may be the overtone variety, marked

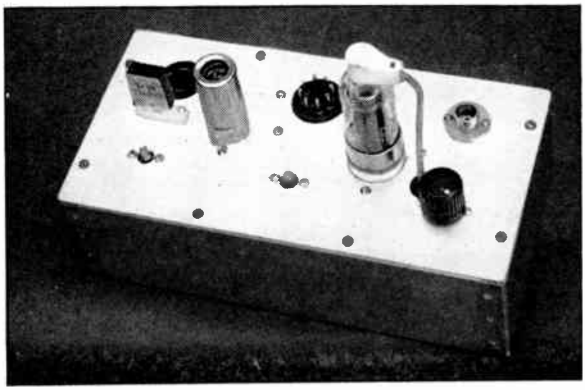


Fig. 17-20 — A 25-watt transmitter or exciter for 50 Mc. Oscillator and doubler are tuned by screwdriver adjustments at lower left and center of top plate. The amplifier control is the knob at the right. The 11-pin power fitting is at the center, rear, and the antenna output fitting is in the upper right.

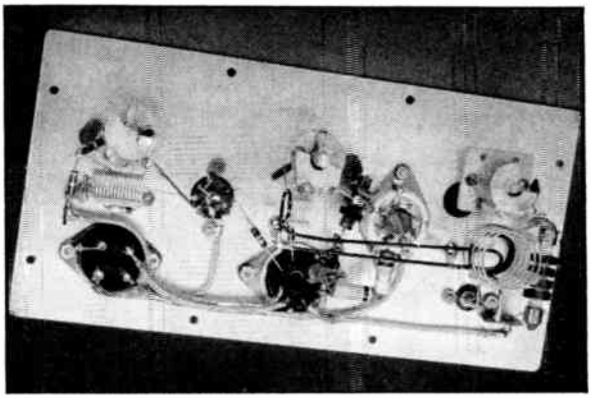


Fig. 17-21 — Bottom view of the 50-Mc. transmitter-exciter. Oscillator, doubler and final circuits are from left to right. Note the inductive neutralization link between L_3 and L_4. Disregard the power fitting at the lower left and follow Fig. 17-22 for power connections.

for frequencies between 24 and 27 Mc., or they may be fundamental-type cuts for 8 to 9 Mc., working on their third overtone. Much less feedback is needed for overtone crystals ordinarily, and if they are to be used exclusively L_2 may be reduced to as little as three turns. If difficulty with starting under load is encountered, the size of the coupling capacitor, C_3, can be reduced, and it may be advantageous to connect an r.f. choke between Pin 2 of the frequency multiplier and the grid leak, R_3.

The second half of the 12AU7 is operated as a doubler to 50 Mc. in the unit for that band, and as a tripler to 72 Mc. in the 144-Mc. model. It has no unusual features in either case. The amplifier is so easy to drive on 50 Mc. that input to both the oscillator and doubler stages can be kept at quite low level — not more than about 10 ma. plate current for each section. In the 144-Mc. unit the current drains will run about 12 to 15 ma. for each stage. Grid current should be 1 ma. or more in either case.

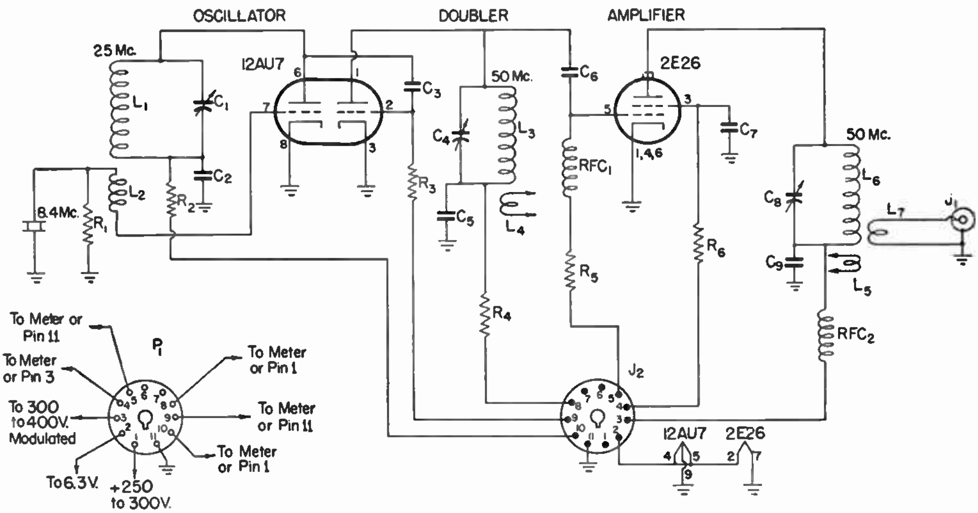


Fig. 17-22 — Schematic diagram and parts list for the 50-Mc. transmitter-exciter.

C_1 — 50-μμfd. trimmer (Millen 26050-LN).
C_2, C_5, C_7, C_9 — 0.001-μfd. disc ceramic.
C_3, C_6 — 50-μμfd. ceramic.
C_4 — 25-μμfd. trimmer (National MSR-25).
C_8 — 20-μμfd. double-spaced shaft-type trimmer (Millen 20920).
R_1 — 39,000 ohms, ½ watt.
R_2, R_4 — 470 ohms, ½ watt.
R_3 — 100,000 ohms, ½ watt.
R_5 — 68,000 ohms, ½ watt.
R_6 — 30,000 ohms, 3 watts. (3 10,000-ohm 1-watt resistors in series. May be reduced in resistance and wattage for 300-volt operation.)
L_1 — 9 turns No. 20, ½-inch diam., 9⁄16 inch long (B & W Miniductor No. 3003).
L_2 — 4 turns No. 20, ½-inch diam., ¼ inch long. L_1 and L_2 are made from a single piece of B & W Miniductor No. 3003, 13 turns total. See text and Fig. 17-21.
L_3 — 5 turns No. 20, ½-inch diam., 5⁄16 inch long (B & W No. 3003).
L_4, L_5 — 1-turn neutralizing loops connected by link, No. 14 enam. See Fig. 17-21.
L_6 — 5 turns No. 16, 1-inch diam., 1¼ inch long (B & W No. 3021).
L_7 — 3 turns No. 14 enam., ¾-inch diam., inside cold end of L_6.
J_1 — Coaxial output fitting.
J_2 — 11-pin male chassis fitting (Amphenol 86RCP11).
RFC_1 — 1-mh. r.f. choke (National R-50).
RFC_2 — 2.5-mh. r.f. choke (National R-100).

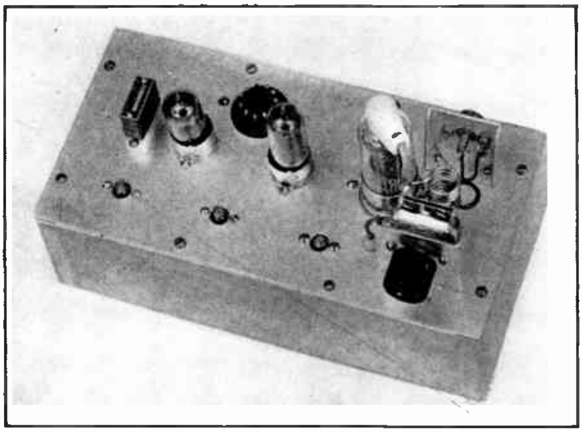


Fig. 17-23 — Top view of the 25-watt 144-Mc. transmitter. Layout is similar to the 50-Mc. model, except for the additional doubler stage and the mounting of the final tank circuit above the chassis.

The 5763 doubler stage in the 2-meter unit is of conventional design. Care must be taken in layout to keep down lead inductance. Note that the lead from the plate to the tuning condenser is made of quarter-inch wide copper strip.

Because of the difference in layouts required for the two frequencies, the two amplifiers operate somewhat differently. The 50-Mc. unit has the final tank coil and antenna coupling underneath the chassis. There is thus more feed-back, and neutralization was needed. This is furnished by the link that may be seen in the bottom view, Fig. 17-21. A loop of No. 14 enameled wire is mounted on stand-offs, with one turn coupled to L_3 and the other end to L_6. The position of the coupling loop at either end is adjusted for neutralization in the same way as for capacitively neutralized amplifiers. The loop (L_5) is between the second and third turns of L_6, with the antenna coupling coil below. Slight variations in layout may eliminate the need for neutralization, so the amplifier operation should be checked without it at first.

In order to shorten the plate lead, the plate circuit of the 2-meter unit was mounted above the chassis. This permits use of a balanced tank cir-

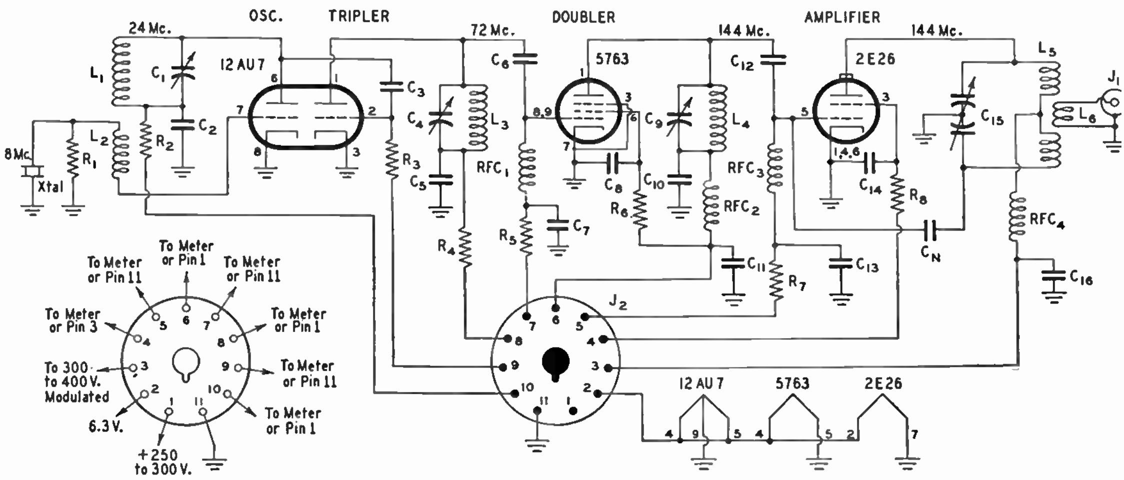


Fig. 17-24 — Schematic diagram of the 144-Mc. transmitter. Bottom views of both power plug and socket are shown.

C_1 — 50-μμfd. trimmer (National PSR-50).
C_2, C_5, C_7, C_8, C_{10}, C_{11}, C_{13}, C_{14}, C_{16} — 0.001-μfd. disc ceramic.
C_3, C_6 — 25-μμfd. ceramic.
C_4 — 25-μμfd. trimmer (National PSR-25).
C_9 — 10-μμfd. double-spaced trimmer (Millen 26920 cut down to 2 rotor and 3 stator plates).
C_{12} — 10-μμfd. ceramic.
C_{15} — 10-μμfd. per section butterfly (Johnson 10LB15).
R_1 — 10,000 ohms, 1 watt.
R_2, R_4 — 470 ohms, ½ watt.
R_3 — 100,000 ohms, ½ watt.
R_5 — 68,000 ohms, ½ watt.
R_6 — 12,000 ohms, ½ watt.
R_7 — 22,000 ohms, ½ watt.
R_8 — 22,000 ohms, 1 watt. Make like R_6 in Fig. 17-22 if using more than 300 volt plate supply.
L_1, L_2 — Similar to Fig. 17-22.
L_3 — 4 turns No. 18, ½-inch diam., ½ inch long (B & W No. 3002).
L_4 — 4 turns No. 14, ¼-inch diam., ⅝ inch long.
L_5 — 6 turns No. 14, 3 turns each side of center spaced diameter of wire, ½-inch diam., ¼-inch space at center of L_6.
L_6 — 2 turns No. 14 enam., ½-inch diam.
J_1 — Coaxial output fitting.
J_2 — 11-prong male chassis fitting (Amphenol 86 RCP11).
RFC_1 — 7-μh. r.f. choke (Ohmite Z-50).
RFC_2, RFC_3, RFC_4 — 1.8-μf. r.f. choke (Ohmite Z-144).

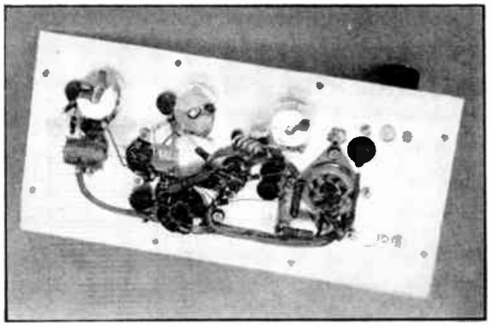


Fig. 17-25 — Under-chassis view of the 144-Mc. transmitter. Oscillator, tripler and doubler tuned circuits are from left to right.

cuit and practically eliminates the need for neutralization. To make up the difference in capacitance on the two sides of the circuit, a lead from the low side is run through a chassis bushing to just below the chassis level. If there is instability, the length of the lead below the chassis can be varied to effect neutralization. Contact is made to the 2E26 metal ring externally by means of a spring clip mounted under one of the socket-mounting screws. This contributes to more stable operation of the amplifier, though connection is made to the ring internally through Pin 8. Shielding may or may not be necessary on the 5763. Operation of the tube without a complete shield results in more effective cooling, and is recommended if possible.

Operating conditions for the various stages follow the tube manufacturer's recommendations closely. If more or less input to the final stage is desired it can be controlled by variation of the screen voltage, with a smaller or larger dropping resistor value.

If both transmitters are to be used, their operation may be controlled by an external switch that furnishes heater voltage to the unit desired at the moment. Plate voltages may be left connected to both units in this case, as only the one whose heaters are energized will draw current. Loading on the amplifier is varied by adjusting the position of the output coupling winding. In some cases the insertion of a series tuning condenser between the coupling loop and ground may be desirable. Power output will be about 15 watts maximum on 50 Mc. and about 10 watts for the 144-Mc. unit. If the plug connections given in the schematic diagrams are followed it will be possible to interchange the two power plugs without affecting the operation of the rigs.

A 100-Watt R.F. Amplifier for 50 and 144 Mc.

The r.f. amplifier shown in Figs. 17-26, 17-27 and 17-28 is designed for use with a dual beam tetrode such as the 829B or AX-9903. It is capable of handling an input of up to 120 watts on c.w. or FM and about 100 watts on AM 'phone. The driver stage should have an output of 5 watts or more, to assure adequate driving power. The same general layout may be used with an 832A or 815, if a suitable value of grid resistor is used. The 815 also requires a different socket.

The amplifier is built on an aluminum chassis 3 by 4 by 17 inches in size, with practically all components mounted topside. The two-band tank circuit described in Fig. 17-3 is used, to facilitate easy band changing and assure efficient operation on 144 Mc. Only the plate circuit is tuned. The grid coils are made to resonate with the input capacitance of the tube. The plate tuning condenser is cut down to a capacitance suitable for 144-Mc. used by removing plates, leaving two stator and three rotor plates in each section. The two stator plates left are those on either side of the stator connection lug. One rotor plate is removed from each end of the shaft and four from the middle.

The tube socket is mounted on a bracket 3⅝

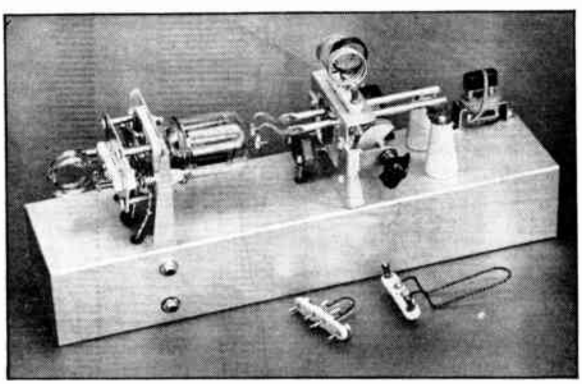


Fig. 17-26 — A dual-tetrode amplifier for 50 and 144 Mc., with 50-Mc. coils in place. In the foreground are the 144-Mc. grid coil and the antenna coupling loop used for 144-Mc. operation.

inches high, with the tube centered $2\frac{1}{2}$ inches above the chassis. The tuning condenser and coil socket are also mounted on brackets, the former $2\frac{3}{8}$ inches high. Both brackets have U-shaped cutouts to pass the plate lines with at least $\frac{5}{16}$ inch clearance all around.

The plate lines are $5\frac{1}{2}$ inches long, exclusive of the flexible portion at the plate end. This is of tinned braid, making $1\frac{1}{4}$ inches additional, from the end of the lines to the slip-on connectors. The flexible portion of the line is made fast by inserting the end of the braid in the tubing and crimping the tubing in a vise. The connection is soldered for added firmness, but the tubing should be squeezed tight enough to hold the braid in place, as long periods of operation may heat the line sufficiently to loosen soldered connections. Connections from the lines to the tuning condenser are made by wrapping the tubing with four turns of tinned wire and soldering this wrap to the line and the condenser tab. The far end of the line is mounted on 2-inch standoffs and small copper brackets, bringing the over-all height to $2\frac{1}{2}$ inches.

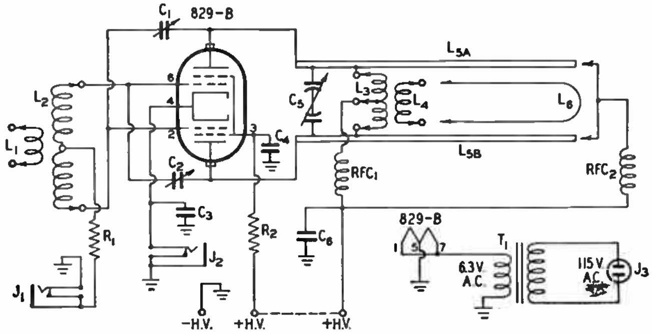


Fig. 17-27 — Schematic diagram of the two-band tetrode amplifier.

C_1, C_2 — Neutralizing capacitors, see text.
C_3, C_4 — 0.001-μfd. disc ceramic.
C_5 — Split-stator variable, approx. 15μμfd. per section (Millen 24935 with 2 stator and 3 rotor plates removed from each section).
C_6 — 0.001-μfd. mica, 1200-volt rating.
R_1 — 4700 ohms, 1 watt.
R_2 — 10,000 ohms, 10 watts.
L_1 — 50 Mc.: 3 turns No. 18, $1\frac{1}{4}$-inch dia., turns spaced wire dia.
144 Mc.: U-shaped loop $\frac{1}{2}$ inch wide and $1\frac{1}{8}$ inch long, No. 14 tinned.
L_2 — 50 Mc.: 2 turns each side of L_1, same dia. and spacing, center tapped. Can be made by removing one turn from each end of a National AR-16 10-S assembly.
144 Mc.: U-shaped loop similar to L_1, but center tapped. See Fig. 17-26.
L_3 — 3 turns each side of center, No. 12 tinned, 1 inch dia., spaced 1 dia., center tapped. Leave $\frac{1}{2}$-inch space for L_4.
L_4 — 3 turns No. 14 enamel, 1-inch dia., spaced 1 dia.
L_{5A}, L_{5B} — $\frac{1}{4}$-inch o.d. copper tubing, $5\frac{1}{2}$ inches long, spaced $\frac{3}{4}$ inch on centers. Make 7 inches long for 9903/5894A.
L_6 — Hairpin coupling loop $3\frac{1}{2}$ inches long, $\frac{3}{4}$ inch wide, No. 12 enamel.
J_1, J_2 — Closed-circuit jack.
J_3 — Male a.c. connector.
RFC_1 — 7.0-μh. r.f. choke (Ohmite Z-50).
RFC_2 — 1.8-μh. r.f. choke (Ohmite Z-144).
T_1 — Filament transformer, 6.3 volts, 3 amp.

The spacing of the lines, $\frac{3}{4}$ inch center to center, is determined by the spacing of the pins of the Millen 37212 plug used for a shorting bar. A short is placed across the terminals of the plug, and connection is made for the B-plus with a flexible lead. The Millen 37211 socket, mounted at the end of the chassis, serves as a convenient storage device for the plug and as a terminal strip for RFC_2. The plug may be used to adjust the line length; sliding it into or out of the line permits an adjustment of about $\frac{1}{4}$ inch in over-all length. This may be useful in counteracting for slight variations in tube characteristics.

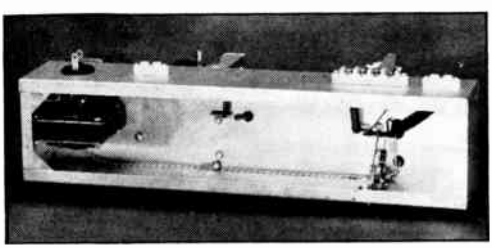


Fig. 17-28 — Bottom view of the tetrode amplifier.

The grid coil socket is mounted on a plate held in position by the screws on which the tube socket is mounted. It is positioned for minimum lead length — an important consideration. The input capacitance of the 829B is high enough so that it may be impossible to resonate the grid circuit at 148 Mc., if appreciable lead length or stray capacitance is introduced. If an 832A or AX-9903 is used the grid coil will be somewhat larger than that specified and neutralization may not be needed.

Neutralization is accomplished, when required, by means of leads brought through the bracket, adjacent to the tube plates. These are crossed over to the opposite grids at the socket. Feed-through bushings are used and soldering lugs are attached to the bushings to provide the neutralizing capacitance. If more is needed these can be replaced with small tabs of sheet copper.

There may be a slight change in neutralizing capacitance needed for the two bands. As neutralization is inclined to be more critical at the higher frequency, the adjustment should be made carefully on 144 Mc. This same setting may be satisfactory for 50-Mc. operation as well.

The plug-in coils are mounted on National PB-16 bases, fitting XB-16 sockets. When the stage is used on 144 Mc. the coupling is by means of a hairpin loop which plugs into the coil socket. The r.f. output is thus fed down to a crystal socket on the back of the chassis, for either band. A similar crystal socket is used for the r.f. input, at the tube end of the chassis.

Transmitter-Exciter for 220 Mc.

Construction of a stable transmitter for 220 Mc. is not difficult, and though simple oscillator-type rigs may suffice for short-range work, crystal control or its equivalent is highly worth-while. A low-powered transmitter need not be costly, as receiving tubes can be used throughout, and by selection of a frequency near the low edge of the band, a crystal can be obtained that will serve for the upper portion of the 144-Mc. band as well.

The transmitter shown in Figs. 17-29, 17-30 and 17-31 delivers about two watts. The final stage may be modulated for voice work, or the unit may be used as an exciter to drive higher-powered stages. Three 12AT7 dual triodes are used. The first serves as a third-overtone oscillator and frequency tripler. This drives a push-pull tripler to the operating frequency. The output stage is a neutralized push-pull amplifier.

Probably any of the several dual triodes having the same base connections as the 12AT7 could be used, and with minor modifications 6J6s will work well. The safe input for 6J6s is slightly lower, however, as their maximum plate voltage should not be higher than 250 volts. The 12AT7s will stand 300 volts, if the rig is adjusted properly.

Crystal frequencies should lie between 8.15 and 8.33 Mc., or 24.45 and 25 Mc. If the same crystal is to be used in 144-Mc. work, it should be between 8.15 and 8.222 or 24.45 and 24.66 Mc. Where crystals in the 8-Mc. range are used, it is suggested that values multiplying out to frequencies well inside the band edges be chosen, as the overtone frequency may not be exactly three times the frequency marked on the holder.

Construction and Adjustment

The transmitter is built on a sheet of aluminum 5 by 9½ inches in size, so that it may be mounted on a standard aluminum chassis of the same dimensions, 2 inches high. This makes for a minimum of mechanical work, and provides excellent shielding. Power leads are made with shielded cable, and each plate and grid lead is decoupled, to prevent radiation of harmonics through the power cabling. The shielded wire may not be necessary as a TVI-prevention measure, but it makes a neat assembly, and it is easy to install. Should TVI become a problem, most of the preventative measures will already have been taken.

In both top and bottom views the principal components may be identified readily. From left to right, we see the crystal and its associated circuit, the oscillator-tripler tube, the tripler plate circuit, the push-pull tripler tube, tripler plate and amplifier grid circuits, and final amplifier stage. Power is brought to the various circuits through an 8-pin power fitting, provision being made for metering all important circuits. The tube sockets, crystal socket, tuning condensers and output socket are centered on a line drawn down the middle of the base plate.

Initial adjustment of the transmitter can be done with any power supply that will deliver 150 to 250 volts. The rig can be operated at higher voltages than this, but for the first work it is well to stay below 250 volts. If a 300-volt supply is used, a 10-watt resistor of about 5000 ohms should be connected in series with the supply voltage temporarily. If the power fittings are wired as shown in Fig. 17-30, the various circuits can be metered during the testing operation.

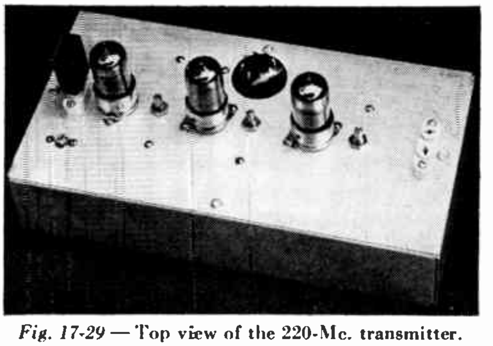


Fig. 17-29 — Top view of the 220-Mc. transmitter.

Start with the oscillator, by connecting a 50- or 100-ma. meter between Pins 3 and 4. Leave Pin 2 open for the present. Apply plate voltage and note the current as C_1 is rotated. There will be a sharp dip as the tube goes into oscillation. Check to see if this oscillation is controlled by the crystal. If there is self-oscillation, reduce the size of the feed-back winding, L_2, a half turn at a time. If there is no oscillation, more inductance may be needed in L_2. Feed-back may also be controlled by cutting the small winding loose from the larger one, and adjusting the spacing between them. They are made from one piece of B & W Miniductor, by cutting the wire at the fourth turn.

Once the oscillator is working correctly, solder a jumper wire between Pins 3 and 4, and connect the meter between Pins 3 and 5. Tune C_2 for a slight dip in plate current, and check the frequency to be sure that the stage is tripling. There should be enough output to light a 2-volt 60-ma. pilot lamp with a single-turn loop coupled at the center of L_3. The capacity of C_3 should be set at the point that gives the greatest output, readjusting C_2 each time C_3 is changed. The purpose of this capacitor is to balance the tank circuit. It will peak at a point that simulates the output capacitance of the tube appearing across the opposite end of L_3.

Now connect a jumper from Pin 3 to Pin 5, and connect the meter between Pins 3 and 6. A low-range meter (0–10 or 0–25 ma.) may be connected between Pins 8 and 7 also, to measure the final grid current. Adjust C_4 for maximum grid current, or for plate current dip, and move L_5 with respect to L_4, retuning C_4 as this is done, until the position that gives the highest grid cur-

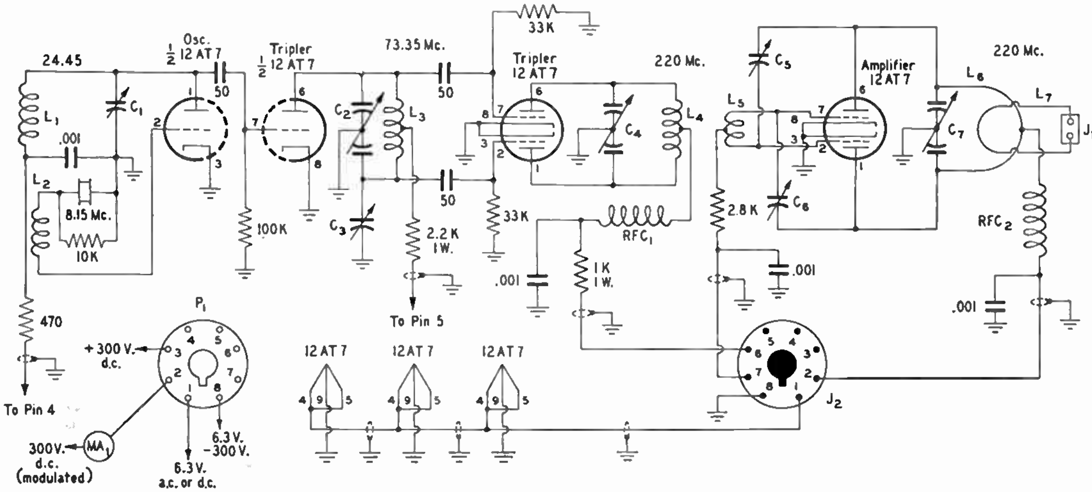


Fig. 17-30 — Wiring diagram of the low-powered 220-Mc. transmitter.

C_1 — 50-$\mu\mu$f. miniature trimmer (Hammarlund MAPC-50).
C_2 — 11-$\mu\mu$f. miniature butterfly variable (Johnson 11MB11).
C_3, C_5, C_6 — 3–30 $\mu\mu$f. mica trimmer.
C_4, C_7 — 8-$\mu\mu$f. miniature butterfly variable (Johnson 9MB11).
L_1 — 10 turns No. 20 tinned, ½-inch diam., spaced diam. of wire.
L_2 — 4 turns No. 20 tinned similar to L_1. L_1 and L_2 made from single piece of B & W Miniductor No. 3003; see text.
L_3 — 12 turns No. 18 tinned, ⅜-inch diam., spaced diam. of wire, center-tapped.
L_4 — 2 turns No. 18 enam., ½-inch diam., spaced ⅛ inch, center-tapped.
L_5 — 2 turns No. 18 enam., ⅜-inch diam., spaced 1/16 inch, center-tapped.
L_6 — U-shaped loop No. 16 tinned, made from 5 inches of wire. Sides of U are 1 inch apart, bent at right angles 1 inch from open end, center-tapped.
L_7 — Similar to L_6, but no center tap. Cover both loops with insulating spaghetti.
J_1 — Output terminal (crystal socket).
J_2 — Male power fitting, 8-pin (Amphenol 86-CP8).
P_1 — Matching cable fitting, 8-pin (Amphenol 78-PF8).
RFC_1, RFC_2 — 18 turns No. 22 enam., close-wound on 1-watt resistor of high value.

rent is found. There will be little change in plate current, so the final grid meter is the better indicator.

Add a jumper from Pin 3 to Pin 6 and proceed with neutralization of the final stage. This may be done in any of the conventional ways. Setting the neutralizing capacitors at the point where there is no change in grid current as the final plate tank is tuned through resonance is a satisfactory procedure. The two trimmers should be about the same setting, near minimum capacitance. Now apply plate voltage to the final stage, connecting the meter in series with Pin 2, and tune the final plate circuit for maximum output as indicated in a pilot lamp plugged into J_1. This may be a blue-bead 6.3-volt 250-ma. bulb, which will give a bright indication with about 2 watts output.

If L_6 will not resonate the desired frequency, its inductance may be varied somewhat by spreading or compressing the sides of the U-shaped tank. Making the U narrower lowers the inductance, broadening it lowers its resonant frequency. The position of L_7 with respect to L_6 should be adjusted for maximum antenna power. The degree of coupling will probably be somewhat different than that at which maximum lamp brilliance is found, as the lamp does not simulate the antenna load. Power input to the final stage should not exceed 10 watts.

Average operating currents, with 300-volt plate supply, will be about as follows: oscillator — 10 ma., tripler — 10 ma., push-pull tripler — 20 ma., final grid current, stage operating — 6–8 ma., final plate current, under load — 20–30 ma.

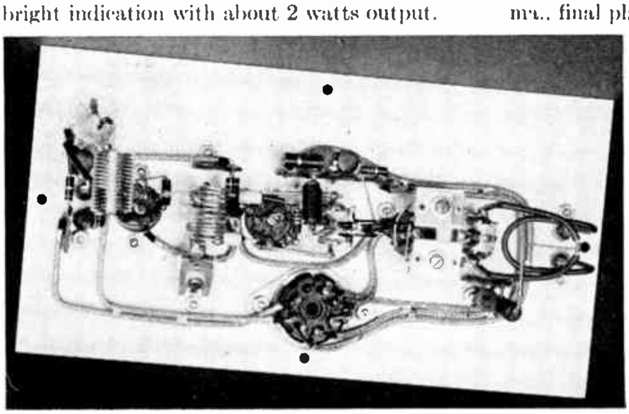


Fig. 17-31 — Interior view of the 220-Mc. transmitter. Components appear in the same order, left to right, as in external view.

Transmitting Equipment for 420 Mc.

As on lower frequencies, best results will be obtained in 420-Mc. work if the narrowest practical passband is used in the receiver. This dictates the use of stabilized transmitters, if the full possibilities of the 420-Mc. band are to be realized. The band is 30 megacycles wide, however, so there is plenty of room for the use of simple rigs and broadband receivers, both of which may be entirely adequate for short-distance experimental work.

Many descriptions of equipment in this category have appeared in *QST* in recent years. A bibliography at the end of this chapter lists these and various articles dealing with the conversion of war-surplus equipment for 420-Mc. use, as well as articles on more advanced equipment. Segregation of narrow and wideband techniques within the band appears desirable, however, and it is suggested that use of the 420-Mc. band be apportioned as follows:

420 to 432 Mc. — Modulated oscillators and wideband FM.

432 to 436 Mc. — Crystal-control AM, c.w. and narrowband FM.

436 to 450 Mc. — Amateur television.

A SIMPLE LOW-POWERED TRANSMITTER

The transmitter shown in Figs. 17-32 through 17-34 is typical of the sort of thing that can be used to good advantage in developing local activity on 420 Mc. It runs only a few watts input, and delivers only about one watt of output, but it is quite capable of working over a radius of several miles when used with a good antenna

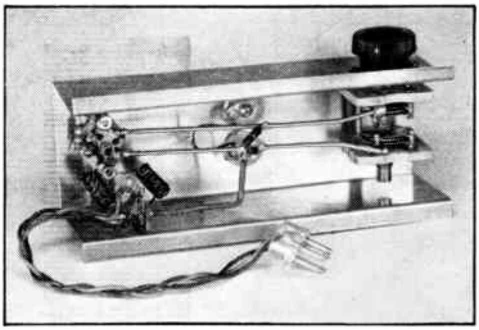
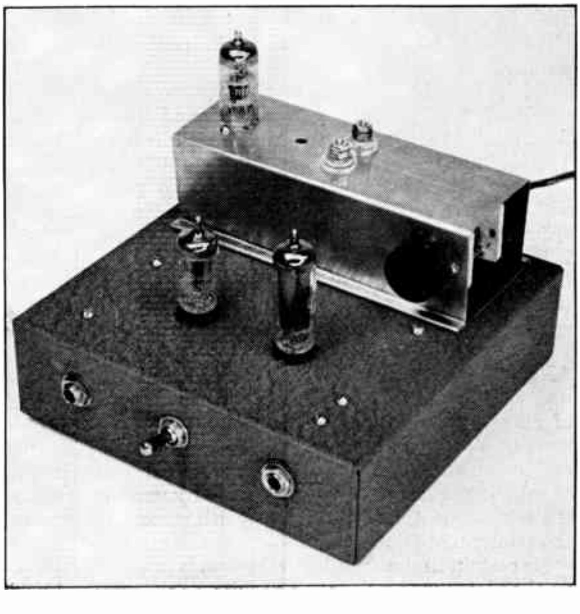


Fig. 17-33 — Bottom view of the oscillator assembly. The trough in which the components are mounted is made of flashing copper. It is 6 inches long, 1⅞ inches high, and 2¼ inches wide, with ¼-inch edges folded over for sliding into a clip attached to the main chassis.

system. A single 6J6 is used as a push-pull oscillator, with a half-wave line in its plate circuit. The complete oscillator assembly is built in a trough made of flashing copper. The 6AQ5 modulator and 6C4 speech amplifier are on the main chassis, at the back of which is a copper clip into which the oscillator unit is fitted. This arrangement permits experimenting with different types of r.f. sections without the necessity of making changes in the audio portion of the rig.

Only three adjustments are necessary in placing the unit into operation. The frequency should be checked with Lecher wires or a calibrated wavemeter, setting the frequency near the middle of the band. The method of determining the proper point for feeding the B-plus to the line is discussed earlier in this chapter. When this is

Fig. 17-32 — A 420-Mc. transmitter built in two units. The modulator portion, on a 7 × 7 × 2-inch chassis, uses a 6C4 driving a 6AQ5 modulator. The oscillator uses a 6J6 and is assembled on a removable trough-shaped chassis.



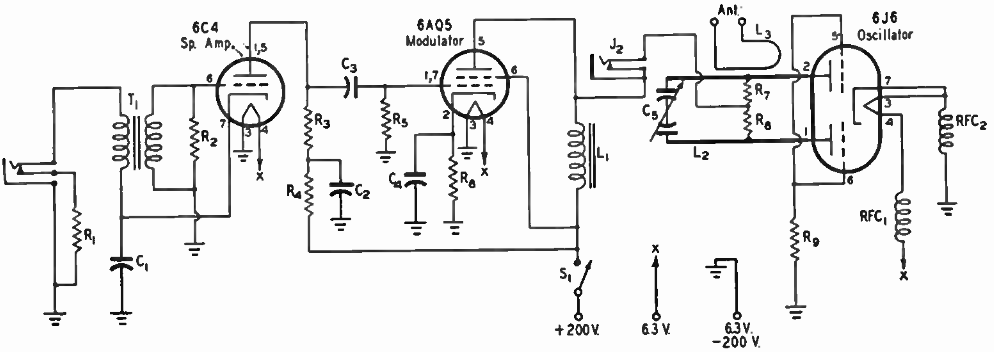


Fig. 17-34 — Schematic diagram of the 420-Mc. transmitter.

C_1, C_4 — 10-μfd. 25-volt electrolytic.
C_2 — 8-μfd. 450-volt electrolytic.
C_3 — 0.01-μfd. tubular.
C_5 — Miniature split-stator variable, 4 μμfd. per section. (Millen 21912D, with one rotor plate removed from each section.)
R_1 — 470 ohms, 1 watt.
R_2 — 0.33 megohm, ½ watt.
R_3, R_4 — 5000 ohms, 5 watts.
R_5 — 0.47 megohm, ½ watt.
R_6 — 680 ohms, 1 watt.
R_7, R_8 — 100 ohms, ½ watt, carbon.
R_9 — 2700 ohms, ½ watt.
L_1 — Midget filter choke.
L_2 — Plate line made of two pieces of No. 12 wire, 4¼ inches long, ⅜ inch apart, center to center.
L_3 — Hairpin of No. 18 wire. Portion which couples to L_2 is about ⅝ inch long. Position should be adjusted for maximum transfer of power to antenna.
J_1, J_2 — Closed-circuit jack.
RFC_1, RFC_2 — 12 turns No. 20 enameled wire, 3/16-inch diam., ¾ inch long.
T_1 — Single-button microphone transformer.

done the coupling loop should be adjusted for maximum power in the antenna and the transmitter is ready for use. Frequency checks should be made again, after the antenna is connected to be sure that the signal radiated is well inside the band limits.

AMPLIFIERS AND FREQUENCY MULTIPLIERS

Not many presently-available tubes work satisfactorily above 400 Mc. The 316A, 703A, 15E, 8012 and 8025, all triodes, work fairly well as oscillators, but are relatively ineffective as frequency multipliers. The 6J6 will deliver a small amount of power as a tripler, and more can be obtained with a pair connected in push-pull-parallel.

Of the tetrodes, the 832A and AX9903 are most used in 420-Mc frequency multipliers and amplifiers. One of these tubes as a push-pull tripler from 144 to 432 Mc. will drive another as a 432-Mc. amplifier. The 832A will give about 2 and 5 watts, while the AX9903 delivers 10 and 25 watts, respectively, in these applications. The 5675, 2C43, 2C39 and 4X150A are typical of the special u.h.f. tubes that are capable of high-efficiency operation, but their use involves the employment of special tank circuits and forced-air cooling.

The tripler-amplifier of Fig. 17-35 uses two AX9903/5894A dual tetrodes to deliver 25 to 30 watts output when driven by a 144-Mc. exciter of about 10 watts output. Half-wave lines are used in all 432-Mc. circuits, and a self-resonant coil in the grid circuit of the tripler. Adjustment of coupling between the stages is done by varying the position of the grid lines, L_4, with respect to the tripler plate lines.

Be certain that no mechanical stress is imposed on the plate pins by the tank circuits, as the 9903 is very easily broken. The 9903/5894A is a more rugged type recently introduced.

The point of connection for the plate voltage should be checked to be sure that it is at the minimum r.f. voltage point. A pencil lead may be touched along the line until the smallest effect on the output is observed. Initially, the plate voltage may be fed into the line at a point just toward the tube end from the center.

The position of the grid lines, L_4, is quite critical and must be adjusted carefully if maximum grid drive is to be obtained. Move the copper strips a small amount at a time, readjusting C_1 meanwhile, until at least 5 ma. of grid current is obtained. More may be used if obtainable. The grid circuit r.f. chokes are connected directly to the tube socket terminals, the input capacitance of the tube being high enough so that the nodal point is within the tube itself. Great care should be taken to see that the plate and grid lines do not come in contact with each other in the course of adjusting the coupling. This may be prevented by inserting thin sheets of mica or teflon between the plate and grid lines. Polystyrene is not usable for this purpose, as the heat radiated from the plate lines will melt it.

Adjustment of antenna coupling is also very critical, and can best be accomplished with a field-strength meter, which need be nothing more than a crystal diode inserted in a pick-up antenna. A line of any length may be run from the antenna to the meter, for remote indication.

Because of the relatively low efficiency obtainable at this frequency, the tubes should not be run at more than about 60 per cent of their normal ratings unless provision is made for forced-air cooling. The power capabilities can be stepped up by shielding the tubes and tank circuits and blowing air through the shields for cooling pur-

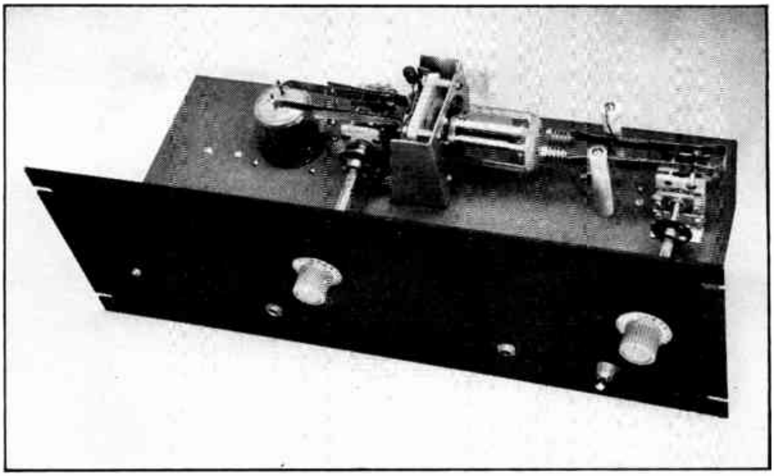


Fig. 17-35 — A tripler-amplifier for 420 Mc. Using two dual tetrodes, one as a tripler from 144 Mc. and the second as a straight-through amplifier, this unit delivers 25 watts output on 432 Mc. It can be driven by any 144-Mc. exciter having an output of 8 watts or more.

poses. Up to about 35 watts output can be developed safely in this way.

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"Four-Twenty Is Fun" (Tilton), Nov. 1947 *QST*, page 13.

"Operating the BC-645 on 420 Mc." (Ralph and Wood), Feb. 1947 *QST*, page 15.

"Fun on 420 with the BC-788" (Clapp), July 1948 *QST*, page 21.

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"Simpler Gear for the 420-Mc. Beginner" (Tilton), May 1949 *QST*, page 11.

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"Coaxial-Tank Amplifier for 220 and 420 Mc." (Brayley), May 1951 *QST*, page 39.

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"A 432-Mc. Converter from the Gold-Plated Test Oscillator," June 1952 *QST*, page 14.

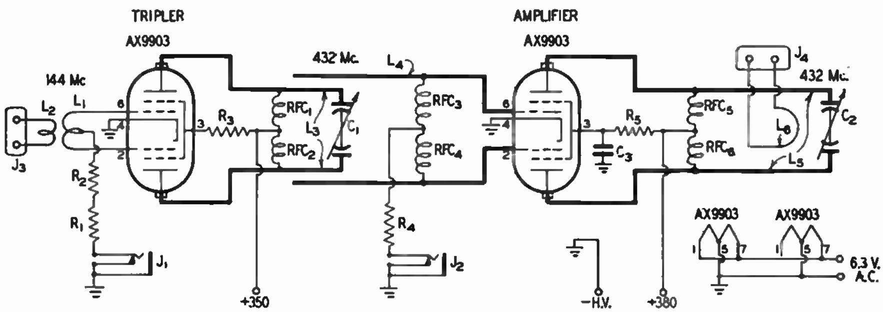


Fig. 17-36 — Schematic diagram of the tripler-amplifier for 432 Mc.

C_1, C_2 — Midget split-stator variable, about 4 μμfd. per section (Millen 21912D).
C_3 — 250-μμfd. ceramic.
R_1 — 50,000 ohms, 2 watts.
R_2 — 100 ohms, ½ watt, at center tap of L_1.
R_3 — 25,000 ohms, 10 watts.
R_4 — 10,000 ohms, 1 watt.
R_5 — 20,000 ohms, 10 watts.
L_1 — 2 turns No. 14 enamel, 9/16-inch diameter, spaced twice wire diameter.
L_2 — 2 turns No. 20 enamel, 9/16-inch diameter, between turns of L_1.
L_3 — Flexible copper or silver ribbon, ½ inch wide and 4 inches long. Average spacing about ⅝ in.
L_4 — Stiff copper strips 3 inches long. Adjust spacing between L_3 and L_4 for maximum grid current, as read in J_2.
L_5 — Flexible copper or silver ribbon, ½ inch wide and 4¾ inches long, including ¼ inch bent over for fastening to heat-dissipating connectors. Average spacing of line is about ⅝ inch. Bend last half inch inward to form padder capacitance. (See Fig. 17-35.) The connectors must be filed down to provide a spacing of at least ¼ inch between their inside edges.
L_6 — Coupling loop of No. 14 enameled wire. U-shaped portion is about 1 inch long.
J_1, J_2 — Closed-circuit jack.
J_3 — Crystal socket (Millen 33102).
J_4 — Antenna terminal (National FWG). Not used in revised version. (See Fig. 17-35.)
RFC_1, RFC_2, RFC_5, RFC_6 — U.h.f. choke (Ohmite Z-235). Attach to plate lines at point of lowest r.f. voltage.
RFC_3, RFC_4 — 11 turns No. 22 enamel, 3/16-inch diameter, 1 inch long. Attach directly to socket tabs.

V.H.F. Antennas

While the basic principles of antenna design are essentially the same for all frequencies where conventional elements are used, certain features of v.h.f. work call for changes in antenna techniques above 50 Mc. Here the physical size of arrays is reduced to the point where an antenna system having some gain over a simple dipole can be used in almost any location, and experimentation with various types of arrays is an important part of the program of progressive v.h.f. amateurs. The importance of high-gain antennas in v.h f. work cannot be overemphasized. By no other means can so large a return be obtained from a small investment as results from the erection of a good directional array.

DESIGN CONSIDERATIONS

At 50 Mc. and higher the frequency range over which antenna systems should operate effectively is usually wider than that encountered on lower bands; thus more attention must be focussed on broad frequency response, possibly to the extent of sacrificing other qualities such as high front-to-back ratio.

As we go higher in frequency transmission-line losses rise sharply, and it becomes more important to match the antenna system to the line properly. Most v.h.f. transmission lines are long in terms of wavelength, so it may be more effective to use a high-gain array at relatively low height, rather than a low-gain system at great height, particularly if the antenna location is not completely shielded by heavy foliage, buildings or other obstructions.

The effectiveness of a v.h.f. array is almost directly proportional to *size*, rather than number of elements. A 4-element array for 432 Mc. may have as much gain over a dipole as a similarly-designed array for 144 Mc., but it will intercept only one-third as much energy in receiving. To be equal in communication, the array for 432 Mc. must equal the 144-Mc. system in *area*, requiring three times the number of elements, if similar element configurations are used.

Polarization

Early v.h.f. work was done with simple antennas, and since the vertical dipole gave as good results in all directions as its horizontal counterpart offered in only two directions, vertical polarization became the accepted standard. Later when high-gain antennas came into use it was only natural that these, too, were put up vertical in areas where v.h.f. activity was already well established.

When the discovery of various forms of long-distance propagation stirred interest in v.h.f. operation in areas where there was no previous experience, many newcomers started in with horizontal arrays, these having been more or less standard practice on frequencies with which these operators were familiar. As use of the same polarization at both ends of the path is necessary for best results, this lack of standardization resulted in a conflict that, even now, has not yet been completely resolved.

Tests have shown no large difference in results over long paths though evidence points to a slight superiority for horizontal in certain kinds of terrain, but vertical has other factors in its favor. Horizontal arrays are generally easier to build and rotate. Where ignition noise and other forms of man-made interference are present, horizontal systems usually provide better signal-to-noise ratio. Simple 3- or 4-element arrays are more effective horizontal than vertical, as their radiation patterns are broad in the plane of the elements and sharp in a plane perpendicular to them.

Vertical systems can provide uniform coverage in all directions, a feature that is possible only with fairly complex horizontal arrays. Gain can be built up without introducing directivity, an important feature in net operation, or in locations where the installation of rotatable systems is not possible. Mobile operation is simpler with vertical antennas. Fear of increased TVI has kept v.h.f. men in densely-populated areas from adopting horizontal as a standard.

The factors favoring horizontal have been predominant on 50 Mc., and today we find it the standard for that band, except for emergency net operation involving mobile units. The slight advantage it offers in DX work has accelerated the trend to horizontal on 144 Mc. and higher bands,

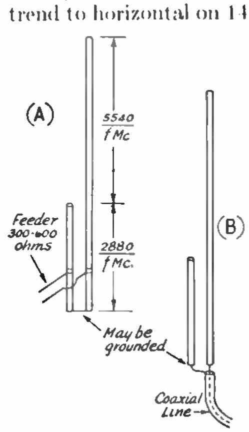


Fig. 18-1 — Two versions of the "J" antenna, often used in mobile installations, or in vertical arrays where parasitic elements may rotate around a fixed radiator.

though vertical polarization is still widely used.

The picture on 220 Mc. is still confused, the tendency being to follow the local 144-Mc. trend. Most 420-Mc. work is being done with horizontal. The newcomer to the v.h.f. bands should ascertain which is in general use in the areas he expects to work, and go along with the others in those areas. In setting up activity where there is no operation presently, it is recommended that horizontal polarization be used, principally as a step toward much-needed standardization.

IMPEDANCE MATCHING

Because line losses increase with frequency it is important that v.h.f. antenna systems be matched to their transmission lines carefully. Lines commonly used in v.h.f. work include open-wire, usually 300 to 500 ohms impedance, spaced one to two inches; polyethylene-insulated flexible lines, available in 300, 150 and 72 ohms impedance; and coaxial lines of 50 to 90 ohms impedance. Some of the methods by which these may be used to feed antennas of differing impedance are given below.

The "J"

Used mainly for feeding a vertical radiator around which parasitic elements are rotated, the "J" is a half-wave vertical radiator fed with a quarter-wave matching section, as shown in Fig. 18-1. For 50 or 144 Mc. the spacing of the matching section should be 2 inches or less. The point of attachment of the line will depend on its impedance. It should be slid along the matching section until the point is found that results in the lowest standing-wave ratio. The bottom of the matching section can be grounded, and it can be fed with balanced or coaxial line. The "J" is useful in 144-Mc. mobile applications, usually in the form shown in Fig. 18-1B.

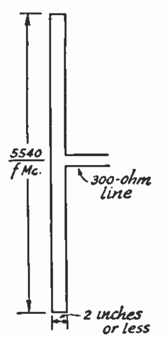


Fig. 18-2 — Details of the folded dipole.

The Delta or "Y" Match

A simple arrangement for feeding a dipole, either alone or as part of a parasitic array, is the delta or "Y" match, in which the line is fanned out and attached to the radiator at the points where the impedance along the element equals the line impedance. Dimensions for v.h.f. applications can be figured from data in the transmission-line chapter. Its chief weakness is the likelihood of radiation from the matching section, which may impair the effectiveness of a multi-element array.

The "T" Match

The principal disadvantages of the delta system can be overcome through the use of the "T" match, also detailed in the transmission lines chapter. It provides a means of adjustment, by sliding clips along the parallel conductors, yet the radiation from the matching section is negligible because of its close proximity to the main element. Its rigid construction is well suited to rotatable arrays. Because the matching is adjustable, the dimensions of the "T" section are not particularly critical. The system may be used with any balanced line, including a pair of coaxial lines, the outer conductors of which may be bonded together and grounded.

The Folded Dipole

A flexible means of matching a wide range of antenna impedances is the folded dipole, shown in its simplest form in Fig. 18-2. When made of uniform conductor size the impedance at the feed point is equal to the square of the number of elements in the folded dipole. Thus, the example of Fig. 18-2 has a feed-point impedance of 4 × 72, or approximately 288 ohms, making it a good match to 300-ohm line. A 3-wire dipole steps the impedance up 9 times.

Greater step-up can be obtained by making the fed portion of the dipole smaller in diameter than the solid portion. The spacing of the conductors affects the step-up in this case. Conductor ratios and spacings can be derived from the folded-dipole monogram in the transmission lines chapter. This principle is applied in the 4-element array of Fig. 18-6.

The Gamma Match

A simple device for feeding parasitic arrays with a single coaxial line is shown in Fig. 18-3. Known as the gamma match, it is a modification of the "T" system for unbalanced lines, well adapted to feeding arrays of all-metal construction. With the latter, the outer conductor of the coaxial line may be grounded to the metal boom, or to the center of the driven element. The inner conductor is then connected to a matching section, usually provided with a sliding clip for varying the point of connection to the driven element. The effectiveness of the system is improved if a condenser is connected in series with the gamma section, to tune out its reactance, as shown in Fig. 18-3. This should be mounted in a weatherproof box, which may be of metal and attached to the boom, or to the center of the driven element. A standing-wave bridge should be connected in the coaxial line, and the point of connection between the driven element and the matching section varied, readjusting the series condenser each time until minimum s.w.r. is ob-

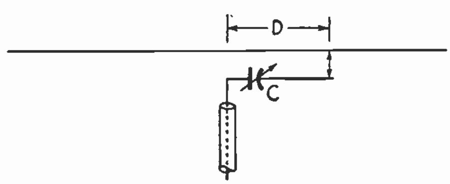


Fig. 18-3 — Schematic version of the gamma match. Values for *C* and *D* are given in the text.

tained. The distance out from the center of the driven element will be about 10 inches for 50 Mc. and 4 inches for 144. The maximum capacitance

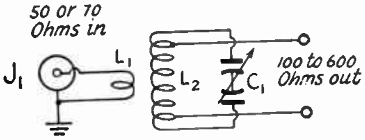


Fig. 18-4 — Antenna coupler for feeding a balanced load with coaxial line. The circuit L_2-C_1 must resonate at the operating frequency.

required at *C* will be about 75 and 25 μμfd. respectively. The r.f. voltage is low at this point so a receiving-type variable condenser may be used.

The Balun

Balanced loads such as are presented by a split dipole or folded dipole can be fed properly with coaxial line only if some form of balanced-to-coaxial line only if some form of balanced-to-unbalanced coupler (often called balun) is used at the feed point. Details of the various types of baluns may be found in the transmission lines chapter. One of these provides a 4-to-1 impedance step-up in addition to conversion from unbalanced line to balanced load.

The conversion may also be accomplished with a balanced circuit, link coupled to the coaxial line, as in Fig. 18-4. The balanced load is tapped onto the tuned circuit at the proper impedance points, in this case. Such a circuit can be in the array itself, or at any point between the transmitter and the antenna where such a conversion is convenient.

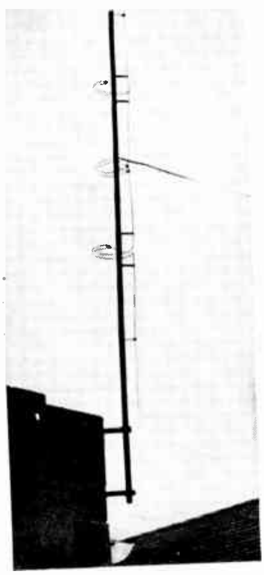


Fig. 18-5 — Collinear array for 144 Mc. made of TV ground wire mounted on a 1½-inch rug pole.

The "Q" Section

A quarter-wavelength of line known as a "Q" section may be used to match a low center impedance to a higher value of line impedance, as described in the transmission lines chapter. This may take the form of two pieces of tubing, ½ to ¼ inch in diameter, mounted so that their center-to-center spacing can be varied to achieve an impedance match between the antenna and the line, where the antenna impedance is not precisely known in advance. Lower values of "Q" section impedance than are available with tubing sizes can be made from lengths of insulated wire, or even coaxial line. The length of the "Q" section will take into account the propagation factor

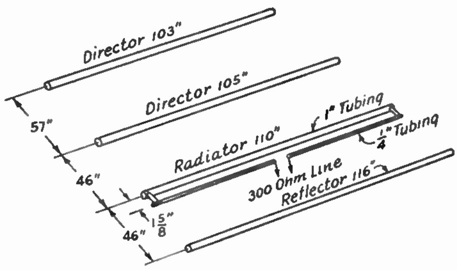


Fig. 18-6 — Dimensional drawing of a 4-element 50-Mc. array. Element length and spacing were derived experimentally for maximum forward gain at 50.5 Mc.

of the line, where such insulating materials are used.

In some installations it may be convenient to use "Q" sections longer than a single quarter wavelength, in which case any odd multiple of a quarter wavelength may be employed. The exact length for any such section may be determined by coupling the line to a source of r.f. energy of the proper frequency and trimming the line for maxi-

TABLE 18-I
Dimensions for V.H.F. Arrays, in Inches

Freq. (Mc.)	50	144	220	420
Driven Element	110	38	24⅞	12¾
Reflector	116	40	26⅛	13⅜
1st Director	105	36	23⅝	12⅛
2nd Director	103	35¾	23⅜	12
Phasing Section*	114	39½	25⅞	13¼
0.25 Wavelength	57	19¾	13	6⅝
0.2 Wavelength	46	15¾	10⅜	5⅜
0.15 Wavelength	34	11¾	7¾	4

* Open-wire line only.

mum loading. Such a "Q" section is often used as the flexible portion of a line feeding a rotatable array, to make connection from the array to a fixed transmission line anchor point at the top of the supporting tower.

Where it is desirable to repeat the antenna impedance at the anchor point, a section of flexible line any multiple of a *half* wavelength may be used.

ANTENNA SYSTEMS FOR 50 AND 144 MC.

The designing of v.h.f. array is both a mechanical and electrical problem. The electrical principles are basic, but a very wide range of mechanical ideas may be used, and the form that an array will take is usually dictated by the materials that are available. Most v.h.f. arrays can be built to formula dimensions given in Table 18-I. The driven element is usually cut from the formula:

$$\textit{Length} \text{ (in inches)} \quad \frac{5540}{\textit{Freq.} \text{ (Mc.)}}$$

Reflector elements are usually 5 per cent longer than the driven element. Directors are 5 per cent shorter, for the one nearest the driven element, and 6 per cent shorter for the next.

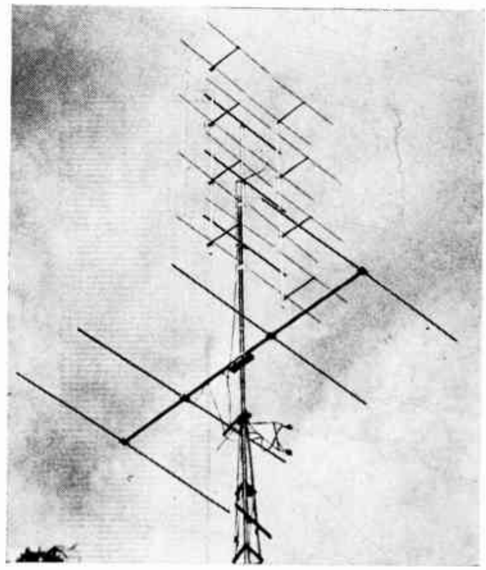


Fig. 18-8 — A 16-element array for 144 Mc. using the all-metal construction methods outlined in Figs. 18-11 to 18-13. The 4-element array for 50 Mc. below is also all-metal design.

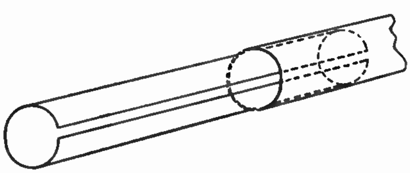


Fig. 18-7 — Detail drawing of inserts which may be used in the ends of the elements of a parasitic array to permit accurate adjustment of element length.

Parasitic element spacing from the driven element is usually 0.15 to 0.25 wavelength for a reflector, and 0.2 or more for directors. The closer the elements are spaced, the lower will be the feed impedance of the driven element. Close-spaced arrays are generally more difficult to tune up properly, and the frequency range over which they work is sharper, so they are seldom used in v.h.f. work.

Elements for 50 Mc. are usually ½ to 1 inch in diameter; 144-Mc. elements ¼ to ½ inch; 220- and 420-Mc. elements ⅛ to ⅜ inch.

A Collinear Array for 144 Mc.

Where some gain over a dipole is needed, yet directivity is undesirable, several half-wave elements may be mounted vertically and fed in phase, as shown in Fig. 18-5. The photograph shows three half-wave elements, but five may be used in a similar way. The center element is fed at its midpoint, either directly with 300-ohm Twin-Lead, or through a "Q" section. The two end elements are kept in phase with the center one by folded half-wave sections.

The array of Fig. 18-5 is built on a 1½-inch wooden rug pole, using aluminum TV ground wire for the elements and phasing sections. Inexpensive TV screw-eye insulators are used to support the elements, with the exception of the supports at the element ends. At these points better insulation is desirable, so ceramic pillars are used.

Two 117-inch pieces of wire or tubing are needed. The end elements are 38 inches long, the folded sections 40 inches over all, and the quarter-wave portions of the middle dipole are 19 inches. The "Q" section, if used, is 20 inches long. The phasing and "Q" sections are bent around into loops, as shown in the photograph. If the array is fed with 300-ohm line the "Q" section may be omitted without serious mismatch. With open-wire line, a "Q" section made of the element material, spaced about one inch, gives a good match. The spacing may be adjusted for minimum s.w.r.

A 4-Element Array for 50 Mc.

The array of Fig. 18-6 uses dimensions derived for maximum gain at 50.5 Mc. It will work well over the range from the low end of the band to nearly 52 Mc. If wider frequency response is desired, the driven element should be cut to the formula given above for the desired center frequency, and the reflector made slightly longer and the directors somewhat shorter than the dimensions given. The driven element is a folded dipole of nonuniform conductor size, stepping up the impedance so that the array can be fed with 300-ohm line. A 3-element array of similar dimensions could be matched with a 3-to-1 conductor ratio, instead of 4-to-1. The boom may be of metal or wood. The 50-Mc. array shown in

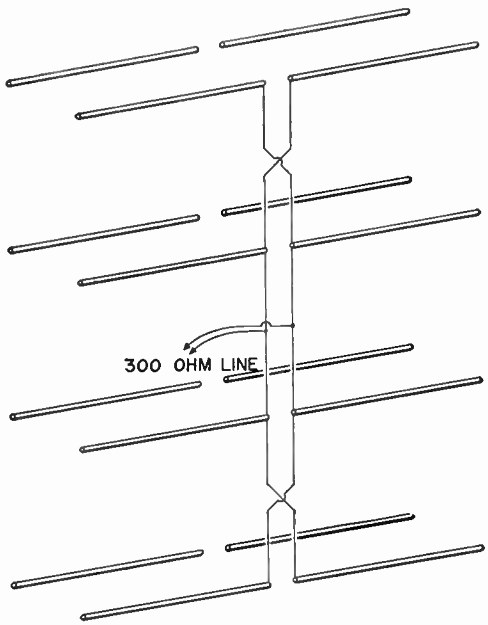


Fig. 18-9 — **Schematic drawing of a 16-element array. A variable "Q" section may be inserted at the feed point if accurate matching is desired. Reflector spacing is 0.2 wavelength.**

Fig. 18-8 uses 0.15-wavelength spacing for the reflector and 0.2 for the directors, resulting in slightly less gain than the wider spacing, but allowing considerably more compact construction.

Most v.h.f. arrays are erected to formula dimensions, but if the builder wishes to do so he may tune the array for optimum front-to-back ratio or forward gain. Adjustable inserts for tubing elements may be made by cutting short sections of the element stock lengthwise and inserting these extensions in the ends of the elements as shown in Fig. 18-7.

Stacking Parasitic Arrays

The radiation angle of a v.h.f. antenna system can be lowered and worthwhile gain obtained by stacking two parasitic arrays one above the other and feeding them in phase. The horizontal pattern of a vertically polarized array may be sharpened and gain added by mounting two arrays side by side and phasing them in the same way. The physical spacing between the two arrays is usually ½, ⅝ or 1 wavelength, depending on the phasing method used. Stacked arrays are usually fed at the center of the system to insure uniform current distribution between the driven elements.

In stacking 50-Mc. arrays the phasing line is usually 0.5 wavelength long. If the two arrays were set up originally for 300-ohm feed when used separately, the phasing line, which serves as a double "Q" section, should have an impedance of about 380 ohms, if the main transmission line is to be 300 ohms. No. 12 wires spaced one inch apart make a convenient phasing line. The gain of two arrays stacked 0.5 wavelength apart is approximately 4 db. over that of a single array.

Slightly more gain can be obtained by increasing the spacing to ⅝ wavelength. A phasing line for this spacing may be made of two pieces of coaxial line, with the outer conductors connected together and grounded, if desired. Because of the propagation factor of the coaxial line, such a phasing section is electrically a full wavelength long. The impedance at the midpoint between the two arrays is approximately half that of one array alone.

For 144 Mc. and higher, where the dimensions are within practical limits, the spacing between two stacked arrays may be increased to a full wavelength. This wide spacing is recommended only for arrays having three or more elements, and is most commonly used with 5-element arrays. The phasing line may be open wire, of any convenient wire size and spacing, and the impedance at the midpoint between the two arrays will be half that of one array alone. A "Q" section at the feed point is a convenient method of matching such a "5-over-5" array. Its dimensions will depend on the type of dipoles used in the individual arrays.

Phased Arrays

Superior performance is obtainable on 144 Mc. and higher by using curtains of 4, 6, 8 or more driven half-wave elements, arranged in pairs fed in phase, and backed up by reflectors. Figs. 18-8 and 18-9 show a 16-element array, while 18-10 is a 12-element array of similar design. The gains are about 14 db. for the 16-element and 12 db. for the 12-element. They may be used for either horizontal or vertical polarization. The pattern of the 12-element is similar in both planes.

The elements used in the 16-element array shown in the photograph are ¼-inch diameter dural, mounted in the manner shown in Figs. 18-11 and 18-12. The entire structure is of metal; the supports being at the low-voltage point of the elements, no insulation is required. The supporting structure for a 12-element array of similar

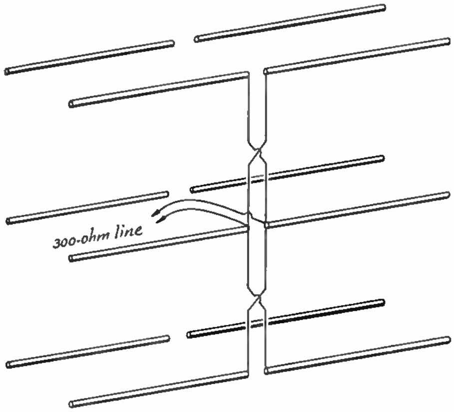


Fig. 18-10 — **Element arrangement and feed system of the 12-element array. Reflectors are spaced 0.15 wavelength behind the driven elements.**

design is shown in detail in Fig. 18-12, with the clamps for holding the array together made as shown in Fig. 18-13.

Element lengths and spacings are not particularly critical in arrays having many driven elements, and careful adjustment is not required for good results. The frequency response of these systems is broader than is the case in arrays where the gain is built up by the use of directors as well as reflectors. Either the 12- or 16-element array may be fed with 300-ohm line connected at the center of the system, as shown in the sketches. The reflectors in the 12-element array are spaced only 0.15 wavelength in back of the driven elements, in order to bring the feed impedance down to roughly 300 ohms. In the 16-element array 0.2-wavelength spacing is used for the reflectors, and even so, the feed impedance may be somewhat lower than 300 ohms. If a long feedline is necessary it may be desirable to insert a variable "Q" section at the feed point, in order to insure accurate matching for minimum s.w.r. In the 16-element array shown in the photograph, a "Q" section having an odd number of quarter-wavelengths of 300-ohm Twin-Lead is used to match the center impedance of around 200 ohms to the 450-ohm open wire line used for a 100-foot run to the operating position.

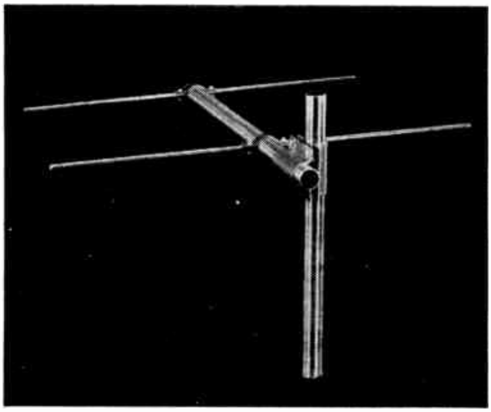


Fig. 18-11 — Model showing the method of assembling for all-metal construction of phased arrays. Dimensions of clamps are given in Fig. 18-13.

In all-metal construction it is important that the supporting structure be entirely in back of the reflector plane. This can be done readily by using the clamp method of assembly detailed in Figs. 18-11, 18-12 and 18-13. Dimensions given in Fig. 18-13 are for use with the tubing sizes given in Fig. 18-12. Suitable dimensions for other combinations can be worked out readily by making experimental clips from soft sheet copper, and using these for templates in making the clips to be used in the final assembly. When the array is completely assembled the screws holding it together should be drawn up as tightly as possible and then coated with durable lacquer or paint to prevent corrosion.

Long-Wire Antennas

Where long-wire systems designed for use on lower frequencies are available they may often be used on the v.h.f. bands with good results, particularly if the feed lines are not too long. "V" and rhombic antenna systems designed expressly for the v.h.f. bands are small enough in size to be used in many locations where similar arrays for lower frequencies would be out of the question. The polarization of long-wire systems is normally horizontal, but in locations where they have a downward slope they may also have a considerable vertical component. Their polarization discrimination is seldom as sharp as that of systems using half-wave elements.

Information on the various types of long-wire arrays will be found in an earlier chapter. At 144 Mc. and higher it is relatively easy to stack two or more "V" or rhombic arrays a half wave apart. This improves their performance considerably, but makes them essentially one-band devices.

Matching devices that permit feeding long-wire antenna systems with flat lines also introduce one-band limitation, so their use is not advisable except in the case of 50 and 144 Mc., two bands that are close to third-harmonic relationship. A "Q" section that is approximately three quarter-wavelengths long at 144 Mc. is one quarter-wavelength long at 50 Mc., so if the feed impedance of the antenna system is the same for both frequencies a "Q" section about

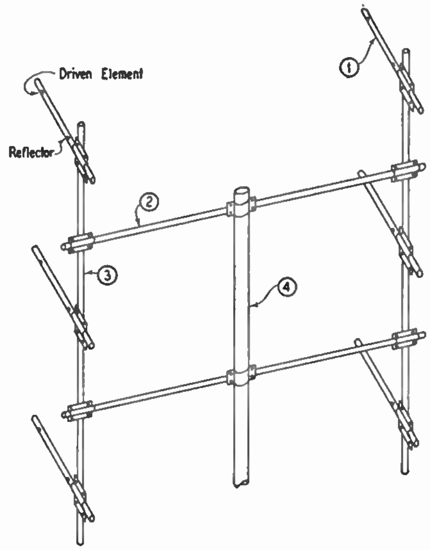


Fig. 18-12 — Supporting framework for a 12-element 144-Mc. array of all-metal design. Dimensions are as follows: element supports (*1*) ¾ by 16 inches; horizontal members (*2*) ¾ by 46 inches; vertical members (*3*) ¾ by 86 inches; vertical support (*4*) 1½-inch diameter, length as required; reflector-to-driven-element spacing 12 inches. Parts not shown in sketch: driven elements ¼ by 38 inches; reflectors ¼ by 40 inches; phasing lines No. 18 spaced 1 inch, 80 inches long, fanned out to 3½ inches at driven elements (transpose each half-wave section).

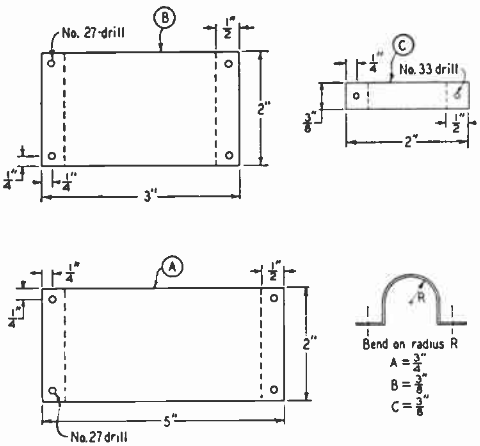


Fig. 18-13 — Detail drawings of the clamps used to assemble the all-metal 2-meter array. A, B and C are before bending into "U" shape. The right-angle bends should be made first, along the dotted lines as shown, then the plates may be bent around a piece of pipe of the proper diameter. Sheet stock should be 1/16-inch or heavier aluminum.

58 inches long may be used for both bands. In the case of a rhombic terminated in 800 ohms and fed with 300-ohm line, the matching section should have an impedance of about 500 ohms.

ARRAYS FOR 220 AND 420 MC.

The use of high-gain antenna systems is almost a necessity if work is to be done over any great distance on 220 and 420 Mc. Experimentation with antenna arrays for these frequencies is fascinating indeed, as their size is so small as to permit trying various element arrangements and feed systems with ease. Arrays for 420 Mc., particularly, are convenient for investigation and demonstration of antenna principles, as even high-gain systems may be of table-top proportions.

Any of the arrays described previously may be used on these bands, but those having large numbers of driven elements in phase are more readily adjusted for maximum effectiveness. The 12- and 16-element arrays of Figs. 18-9 and 18-10 are well adapted to use on 220 or 420. Suitable dimensions may be found in Table 18-I.

A 16-element array for 220 Mc. and a 24-element array for 420 Mc. are shown mounted back-to-back in Fig. 18-14. The 220-Mc. portion follows the 16-element design already described. It is fed at the center of the system with 300-ohm tubular Twin-Lead, matched to the center impedance of the array through a "Q" section of 7/16-inch tubing, spaced about 1½ inches center to center. This spacing was adjusted for minimum standing-wave ratio on the line.

Elements in the array shown are of 7/16-inch aluminum fuel-line tubing, which is very light in weight and easily worked. The supporting structure is dural tubing, using the clamp assembly methods of Fig. 18-12.

The 420-Mc. array uses two 12-element assemblies similar to Fig. 18-10, mounted one above the other, about one half wavelength separating the bottom of one from the top of the other. The two sets of phasing lines are joined by means of one-wavelength sections of Twin-Lead at the middle of the array. This junction, which has an impedance of around 150 ohms, is fed with 300-ohm tubular Twin-Lead through an adjustable "Q" section.

Elements in the 420-Mc. array are cut from thin-walled ¼-inch tubing. Their supports are the 7/16-inch stock used for the 220-Mc. elements. Slots were cut in the ends of these supports to take the elements, and a 4/40 screw was run through both pieces and drawn up tightly with a nut. The horizontal supports were fastened in holes drilled in the vertical members, and were also held in place with a 6/32 screw and nut. The small size and light weight of the 420-Mc. array did not require the use of clamps to make a strong assembly.

The two one-wavelength sections of 300-ohm line are 21¾ inches long, taking the propagation factor into account. The "Q" section may be of any convenient size of tubing, ¼ to ½ inch diameter. It should be made adjustable, as matching is important at this frequency. Dimensions for both arrays can be taken from Table 18-I.

Plane-Reflector Arrays

At 220 Mc. and higher, where their dimensions become practicable, plane-reflector arrays are widely used. Except as it affects the impedance of the system, as shown in Fig. 18-15, the spacing between the driven elements and the reflecting plane is not particularly critical. Maximum gain occurs around 0.1 to 0.15 wavelength, which is also the region of lowest impedance. Highest impedance appears at about 0.3 wavelength. A plane reflector spaced 0.22 wavelength in back of the driven elements has no effect on their feed impedance. As the gain of a plane-reflector array is nearly constant at spacings from 0.1 to 0.25 wavelength, it may be seen that the spacing may be varied to achieve an impedance match.

An advantage of the plane reflector is that it may be used with two driven element systems, one on each side of the plane, providing for two-band operation, or the incorporation of horizontal and vertical polarization in a single structure. The gain of a plane-reflector array is slightly higher than that of a similar number of driven elements backed up by parasitic reflectors. It also has a broader frequency response and higher front-to-back ratio. To achieve these ends, the reflecting plane must be larger than the area of the driven elements, extending at least a quarter wavelength on all sides. Chicken wire on a wood or metal frame makes a good plane reflector. Closely-spaced wires or rods may be substituted, with the spacing between them running up to 0.1

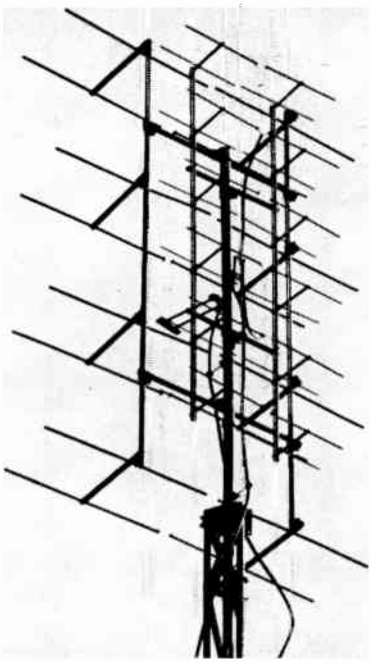


Fig. 18-14 — A 24-element array for 420 Mc. and a 16-element for 220 mounted back-to-back on a single support.

wavelength without appreciable reduction in effectiveness.

Corner Reflectors

In the corner reflector two plane surfaces are set at an angle, usually between 45 and 90 degrees, with the antenna on a line bisecting this angle. Maximum gain is obtained with the antenna 0.5 wavelength from the vertex, but compromise designs can be built with closer spacings. There is no focal point, as would be the case for a parabolic reflector. Corner angles greater than 90 degrees can be used at some sacrifice in gain. At less than 90 degrees the gain increases, but the size of the reflecting sheets must be increased to realize this gain.

At a spacing of 0.5 wavelength from the vertex, the impedance of the driven element is approximately twice that of the same dipole in free space. The impedance decreases with smaller spacings and corner angles, as shown in Fig. 18-15. The gain of a corner-reflector array with a 90-degree angle, 0.5 wavelength spacing and sides 1 wavelength long is approximately 10 db. Principal advantages of the corner reflector are broad frequency response and high front-to-back ratio.

MISCELLANEOUS ANTENNA SYSTEMS

Coaxial Antennas

With the "J" antenna, radiation from the matching section and the transmission line tends to combine with the radiation from the antenna in such a way as to raise the angle of radiation. At v.h.f. the lowest possible radiation angle is essential, and the coaxial antenna shown in Fig. 18-16 was developed to eliminate feeder radiation. The center conductor of a 70-ohm concentric transmission line is extended one-quarter wave beyond the end of the line, to act as the upper half of a half-wave antenna. The lower half is provided by the quarter-wave sleeve, the upper end of which is connected to the outer conductor of the concentric line. The sleeve acts as a shield about the transmission line and very little current is induced on the outside of the line by the antenna field. The line is non-resonant, since its characteristic impedance is the same as the center impedance of the half-wave antenna. The sleeve may be made of copper or brass tubing of suitable diameter to clear the transmission line. The coaxial antenna is somewhat difficult to construct, but is superior to simpler systems in its performance at low radiation angles.

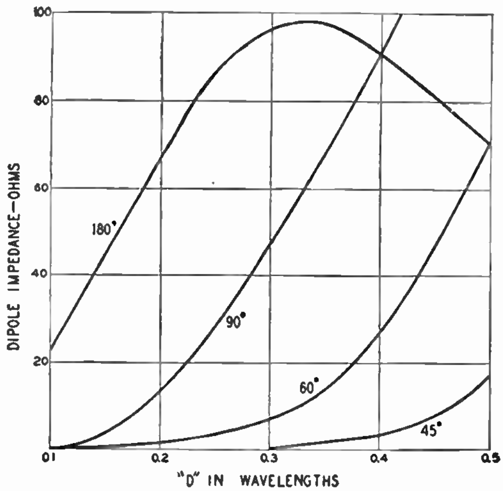


Fig. 18-15 — Feed impedance of the driven element in a corner-reflector array for corner angles of 180 (flat sheet), 90, 60 and 45 degrees. "D" is the dipole-to-vertex spacing.

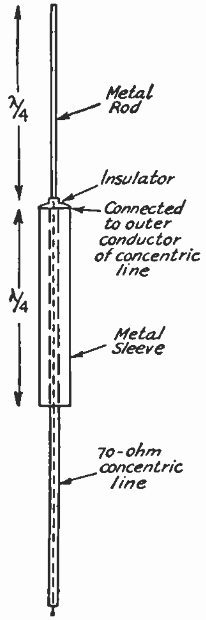


Fig. 18-16 — Coaxial antenna. The insulated inner conductor of the 70-ohm concentric line is connected to the quarter-wave metal rod which forms the upper half of the antenna.

Broadband Antennas

Certain types of antennas used in television are of interest because they work across a wide band of frequencies with relatively uniform response. At very-high frequencies an antenna made of small wire is purely resistive only over a very small frequency range. Its Q, and therefore its selectivity, is sufficient to limit is optimum performance to a narrow frequency range, and readjustment of the length or tuning is required for each narrow slice of the spectrum. With tuned transmission lines, the effective length of the antenna can be shifted by retuning the whole system. However, in the case of antennas fed by matched-impedance lines, any appreciable frequency change requires an actual mechanical adjustment of the system. Otherwise, the resulting mismatch with the line will be sufficient to cause significant reduction in power input to the antenna.

A properly designed and constructed wide-band antenna, on the other hand, will exhibit very nearly constant input impedance over several megacycles.

The simplest method of obtaining a broad-band characteristic is the use of what is termed a "cylindrical" antenna. This is no more than a conventional doublet in which large-diameter tubing is used for the elements. The use of a relatively large diameter-to-length ratio lowers the Q of the antenna, thus broadening the resonance characteristic.

As the diameter-to-length ratio is increased, end effects also increase, with the result that the antenna must be made shorter than a thin-wire antenna resonating at the same frequency. The reduction factor may be as much as 20 per cent with the tubing sizes commonly used for amateur antennas at v.h.f.

Cone Antennas

From the cylindrical antenna various specialized forms of broadly-resonant radiators have been evolved, including the ellipsoid, spheroid, cone, diamond and double diamond. Of these, the conical antenna is perhaps the most interesting. With large angles of revolution, the variation in the characteristic impedance with changes in frequency can be reduced to a very low value, making such an antenna suitable for extremely wide-band operation. The cone may be made up either of sheet metal or of multiple wire spines. A variation of this form of conical antenna is widely used in TV reception.

Parabolic Reflectors

A plane sheet may be formed into the shape of a parabolic curve and used with a driven radiator situated at its focus, to provide a highly-directive antenna system. If the parabolic reflector is sufficiently large so that the distance to the focal point is a number of wavelengths, optical conditions are approached and the wave across the mouth of the reflector is a plane wave. However, if the reflector is of the same order of dimensions as the operating wavelength, or less, the driven radiator is appreciably coupled to the reflecting sheet and minor lobes occur in the pattern. With an aperture of the order of 10 or 20 wavelengths, sizes that may be practical for microwave work, a beam-width of approximately 5 degrees may be achieved.

A reflecting paraboloid must be carefully designed and constructed to obtain ideal performance. The antenna must be located at the focal point. The most desirable focal length of the parabola is that which places the radiator along the plane of the mouth; this length is equal to one-half the mouth radius. At other focal distances interference fields may deform the pattern or cancel a sizable portion of the radiation.

CHAPTER 19

U.H.F. and Microwave Communication

In moving into the microwave region the amateur encounters marked differences in both the technical approach and the uses to which his frequency assignments may be put. Above 1000 Mc. we must discard most of our conventional circuitry and antenna ideas. Coils and condensers are replaced by coaxial tank circuits and resonant cavities. Parallel-wire transmission lines give way to coaxial lines or waveguide. Parasitic arrays are abandoned in favor of parabolic reflectors or horns. And in contrast to the random operating that has been so large a part of the amateur picture on our communication frequencies, microwave work is principally a matter of point-to-point communication between two cooperating stations.

These basic differences have tended to raise a natural boundary in the region around 500 Mc., beyond which relatively few communicating amateurs have ventured. The frequencies at the high end of the spectrum have a strong appeal to the experimenter, however, and the Technician Class license was developed to provide the means whereby this type of worker may legally engage in two-way communication on frequencies above 220 Mc.

At least some amateur work has been done in all the assignments now open to our use. The work of these pioneers in adapting the frequencies above 1000 Mc. to communication purposes has been in line with the best amateur tradition, and it is hoped that the bands beginning at 1215 Mc. will see much amateur exploration in the near future. The frequencies assigned to amateurs in the microwave region are as follows: 1215 to 1300 Mc., 2300 to 2450 Mc., 3300 to 3500 Mc., 5650 to 5925 Mc., 10,000 to 10,500 Mc., and 21,000 to 22,000 Mc. Any frequency above 30,000 Mc. may be used. Any type of emission may be used in any of these bands, except in the case of the lowest, where pulse transmission is prohibited.

U.H.F. Tank Circuits

In resonant circuits as employed at the lower frequencies it is possible to consider each of the reactance components as a separate entity. A coil is used to provide the required inductance and a condenser is connected across it to provide the needed capacitance. The fact that the coil itself has a certain amount of self-capacitance, as well as some resistance, while the condenser also possesses a small self-inductance, can usually be disregarded.

At the very-high and ultrahigh frequencies, however, it is no longer possible to separate these components. The connecting leads which, at lower frequencies, would serve merely to join the condenser to the coil now may have more inductance than the coil itself. The required inductance coil may be no more than a single turn of wire, yet even this single turn may have dimensions comparable to a wavelength at the operating frequency. Thus the energy in the field surrounding the "coil" may in part be radiated. At a sufficiently high frequency the loss by radiation may represent a major portion of the total energy in the circuit. Since energy which cannot be utilized as intended is wasted, regardless of whether it is consumed as heat by the resistance of the wire or simply radiated into space, the effect is as though the resistance of the tuned circuit were greatly increased and its Q greatly reduced.

For this reason, it is common practice to utilize resonant sections of transmission line as tuned circuits at frequencies above 300 Mc. A quarter-wavelength line, or any odd multiple thereof, shorted at one end and open at the other, exhibits large standing waves. When a voltage of the frequency at which such a line is resonant is applied to the open end, the response is very similar to that of a parallel resonant circuit; it will have very high input impedance at resonance and a large current flowing at the short-circuited end. The input impedance may be as high as 0.4 megohm for a well-constructed line.

The action of a resonant quarter-wavelength line can be compared with that of a coil-and-condenser combination whose constants have been adjusted to resonance at a corresponding frequency. Around the point of resonance, in fact, the line will display very nearly the same characteristics as those of the tuned circuit. The equivalent relationships are shown in Fig. 19-1. At frequencies off resonance the line displays qualities comparable to the inductive and capacitive reactances of the coil-and-condenser circuit, although the exact relationships involved are somewhat different. For all practical purposes, however, sections of resonant wire or transmission line can be used in much the same manner as coils or condensers.

In circuits operating above 300 Mc., the spacing between conductors becomes an appreciable fraction of a wavelength. To keep the radiation loss as small as possible the parallel conductors should not be spaced farther apart than 10 per cent of the wavelength, center to center. On the other hand, the spacing of large-diameter conductors should not be reduced to much less twice the diameter because of what is known as the *proximity effect*, whereby another form of loss is introduced through eddy currents set up by the adjacent fields. Because the cancellation is no longer complete, radiation from an open line becomes so great that the Q is greatly reduced. Consequently, at these frequencies coaxial lines must be used.

Construction

Practical information concerning the construction of transmission lines for such specific uses as feeding antennas and as resonant circuits in radio transmitters will be found in this

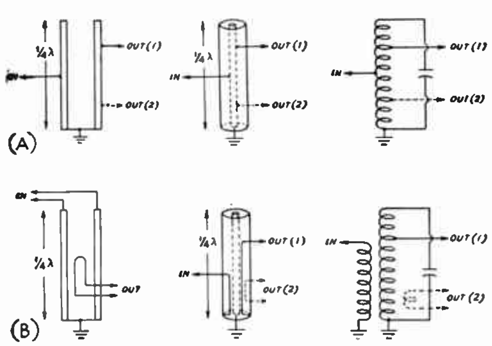


Fig. 19-1 — Equivalent coupling circuits for parallel-line, coaxial-line and conventional resonant circuits.

and other chapters of this *Handbook*. Certain basic considerations applicable in general to resonant lines used as circuit elements may be considered here, however.

While either parallel-line or coaxial sections may be used, the latter are preferred for higher-frequency operation. Representative methods for adjusting the length of such lines to resonance are shown in Fig. 19-2. At the left, a sliding shorting disk is used to reduce the effective length of the line by altering the position of the short-circuit. In the center, the same effect is accomplished by using a telescoping tube in the end of the inner conductor to vary its length and thereby the effective length of the line. At the right, two possible methods of mounting parallel-plate condensers, used to tune a "foreshortened" line to resonance, are illustrated. The arrangement with the loading capacitor at the open end of the line has the greatest tuning effect per unit of capacitance; the alternative method, which is equivalent to "tapping" the condenser down on the line, has less effect on the Q of the circuit. Lines with capacitive "loading" of the sort illustrated will be shorter, physically, than an unloaded line resonant at the same frequency.

The short-circuiting disk at the end of the line must be designed to make perfect electrical contact. The voltage is a minimum at this end of the line; therefore, it will not break down some of the thinnest insulating films. Usually a

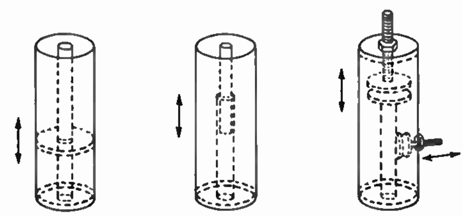


Fig. 19-2 — Methods of tuning coaxial resonant lines.

soldered connection or a tight clamp is used to secure good contact. When the length of line must be readily adjustable, the shorting plug is provided with spring collars which make contact on the inner and outer conductors at some distance away from the shorting plug at a point where the voltage is sufficient to break down the film between the collar and conductor.

Two methods of tuning parallel-conductor lines are shown in Fig. 19-3. The sliding short-circuiting strap can be tightened by means of screws and nuts to make good electrical contact. The parallel-plate condenser in the second drawing may be placed anywhere along the line, the tuning effect becoming less as the condenser is located nearer the shorted end of the line. Although a low-capacitance variable condenser of ordinary construction can be used, the circular-plate type shown is symmetrical and thus does not unbalance the line. It also has the further advantage that no insulating material is required.

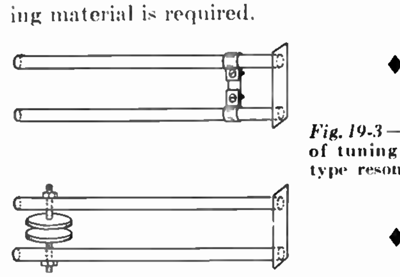


Fig. 19-3 — Methods of tuning parallel-type resonant lines.

Equivalent impedance points, for coupling or impedance-transformation purposes, are shown in Fig. 19-1 for parallel-line, coaxial-line, and conventional coil-and-condenser circuits.

Lumped-Constant Circuits

At the very-high frequencies the low values of L and C required make ordinary coils and condensers impracticable, while linear circuits offer mechanical difficulties in making tuning adjustments over a wide frequency range, and radiation from unshielded lines may reduce their effectiveness materially.

To overcome these difficulties, special high-Q

lumped-constant circuits have been developed in which connections from the "condenser" to the "coil" are an inherent part of the structure. Integral design minimizes both resistance and inductance and increases the C/L ratio.

The simplest of these circuits is based on the use of disks combining half-turn inductance loops with semicircular condenser plates. By connecting several of these half-turn coils in parallel, the effective inductance is reduced to a value appreciably below that for a single turn. Tuning is accomplished by interleaving grounded rotor plates between the turns. Both by shielding action and short-circuited-turn effect, these further reduce the inductance.

Another type of high-C circuit is a single-turn toroid, commonly termed the "hat" resonator. Two copper shells with wide, flat "brims" are mounted facing each other on an axially-aligned copper rod. The capacitance in the circuit is that between the wide shells, while the central rod comprises the inductance.

"Butterfly" Circuits

The tank circuits described in the preceding section are primarily fixed-frequency devices. The "butterfly" circuits shown in Fig. 19-4 are capable of being tuned over an exceptionally wide range, while still having high Q and reasonable physical dimensions. The circuit at A is derived from a conventional balanced-type variable condenser. The inductance is in the wide circular band connecting the stator plates. At its minimum setting the rotor plate fills the opening of the loop, reducing the inductance to a minimum. Connections are made to points *1* and *2*. This basic structure eliminates all connecting leads and avoids all sliding or wiping electrical contacts to a rotating member. A disadvantage is that the electrical midpoint shifts from point *3* to point *3′* as the rotor is turned. Constant magnetic coupling may be obtained by a coupling loop located at point *4*, however.

In the modification shown at D, two sectoral stators are spaced 180 degrees, thereby achiev-

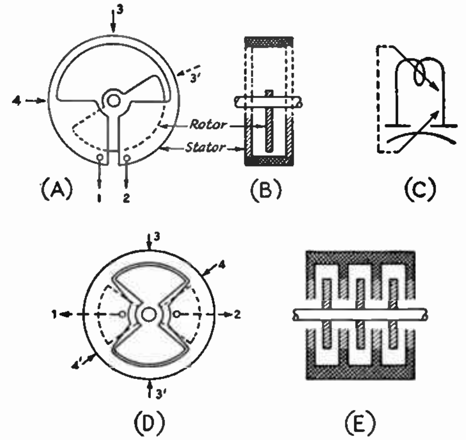


Fig. 19-4 — "Butterfly" tank circuits for v.h.f., showing front and cross-section views and the equivalent circuit.

ing the electrical symmetry required to permit tapping for balanced operation. Connections to the circuit should be made at points *1* and *2* and it may be tapped at points *3* and *3′*, which are the electrical midpoints. Where magnetic coupling is employed, points *4* and *4′* are suitable locations for coupling links.

The capacitance of any butterfly circuit may be computed by the standard formula for parallel-plate condensers given in the data chapter. The maximum inductance can be obtained approximately by finding the inductance of a full ring of the same diameter and multiplying the result by a factor of 0.17. The ratio of minimum to maximum inductance varies between 1.5 and 4 with conventional construction.

Any number of butterfly sections may be connected in parallel. In practice, units of four to eight plates prove most satisfactory. The ring and stator sections may either be made in a single piece or with separate sectoral stator plates and spacing rings assembled with machine screws.

Wave Guides and Cavity Resonators

A wave guide is a conducting tube through which energy is transmitted in the form of electromagnetic waves. The tube is not considered as carrying a current in the same sense that the wires of a two-conductor line do, but rather as a *boundary* which confines the waves to the enclosed space. Skin effect prevents any electromagnetic effects from being evident outside the guide. The energy is injected at one end, either through capacitive or inductive coupling or by radiation, and is received at the other end. The wave guide then merely confines the energy of the fields, which are propagated through it to the receiving end by means of reflections against its inner walls.

The difficulty of visualizing energy transfer without the usual closed circuit can be relieved somewhat by considering the guide as being evolved from an ordinary two-conductor line.

In Fig. 19-5A, several closed quarter-wave stubs are shown connected in parallel across a two-wire transmission line. Since the open end of each stub is equivalent to an open circuit, the line impedance is not affected by their presence. Enough stubs may be added to form a "U"-shaped rectangular tube with solid walls, as at B, and another identical "U"-shaped tube may be added edge-to-edge to form the rectangular pipe shown in Fig. 19-5C. As before, the line impedance still will not be affected. But now, instead of a two-wire transmission line, the energy is being conducted within a hollow rectangular tube.

This analogy to wave-guide operation is not exact, and therefore should not be taken too literally. In the evolution from the two-wire line to the closed tube the electric- and magnetic-field configurations undergo considerable

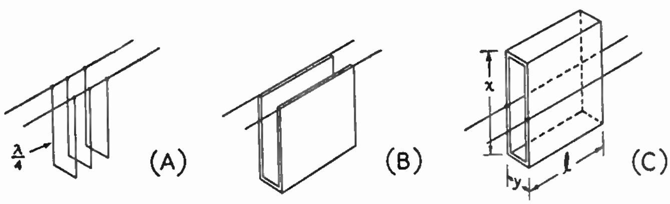


Fig. 19-5 — Evolution of a wave guide from a two-wire transmission line.

change, with the result that the guide does not actually operate like a two-conductor line shunted by an infinite number of quarter-wave stubs. If it did, only waves of the proper length to correspond to the stubs would be propagated through the tube, but the fact is that such waves do *not* pass through the guide. Only waves of shorter length — that is, higher frequency — can go through. The distance x represents half the *cut-off wavelength*, or the shortest wavelength that is unable to go through the guide. Or, to put it another way, waves of length equal to or greater than $2x$ cannot be propagated in the guide.

A second point of difference is that the apparent length of a wave along the direction of propagation through a guide always is greater than that of a wave of the same frequency in free space, whereas the wavelength along a two-conductor transmission line is the same as the free-space wavelength (when the insulation between the wires is air).

Operating Principles of Wave Guides

Analysis of wave-guide operation is based on the assumption that the guide material is a perfect conductor of electricity. Typical distributions of electric and magnetic fields in a rectangular guide are shown in Fig. 19-6. It will be observed that the intensity of the electric field is greatest at the center along the x dimension, diminishing to zero at the end walls. The latter is a necessary condition, since the existence of any electric field parallel to the walls at the surface would cause an infinite current to flow in a perfect conductor. This represents an impossible situation.

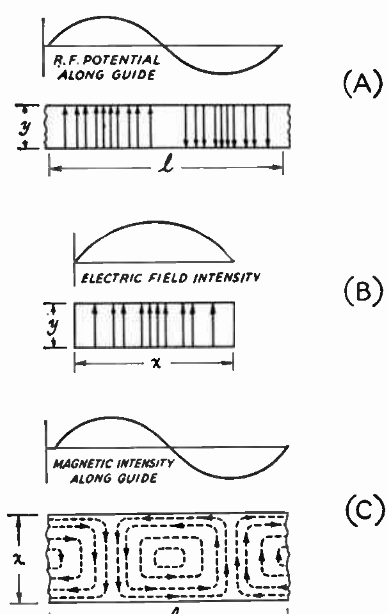


Fig. 19-6 — Field distribution in a rectangular wave guide. The $TE_{1,0}$ mode of propagation is depicted.

Zero electric field at the end walls will result if the wave is considered to consist of two separate waves moving in zigzag fashion down the guide, reflected back and forth from the end walls as shown in Fig. 19-7. Just at the walls, the positive crest of one wave meets the negative crest of the other, giving complete cancellation of the electric fields. The angle of

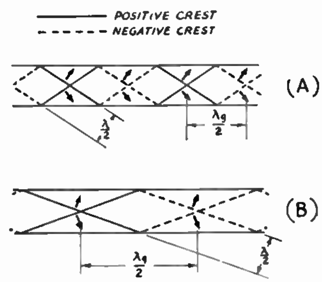


Fig. 19-7 — Reflection of two component waves in a rectangular guide. λ = wavelength in space, λg = wavelength in guide. Direction of wave motion is perpendicular to the wave front (crests) as shown by the arrows.

reflection at which this cancellation occurs depends upon the width x of the guide and the length of the waves; Fig. 19-7A illustrates the case of a wave considerably shorter than the cut-off wavelength, while B shows a longer wave. When the wavelength equals the cut-off value, the two waves simply bounce back and forth between the walls and no energy is transmitted through the guide.

The two waves travel with the speed of light, but since they do not travel in a straight line the energy does not travel through the guide as rapidly as it does in space. A further consequence of the repeated reflections is that the points of maximum intensity or wave crests are separated more along the line of propagation in the guide than they are in the two separate waves. In other words, the wavelength in the guide is greater than the free-space wavelength. This is also shown in Fig. 19-7.

Modes of Propagation

Fig. 19-6 represents a relatively simple distribution of the electric and magnetic fields. There is in general an infinite number of ways in which the fields can arrange themselves in a guide so long as there is no upper limit to the frequency to be transmitted. Each field configuration is called a *mode.* All modes may be separated into two general groups. One group, designated *TM* (*transverse magnetic*), has the magnetic field entirely transverse to the direction of propagation, but has a component of electric field in that direction. The other type, designated *TE* (*transverse electric*) has the electric field entirely transverse, but has a component of magnetic field in the direction of propagation. *TM* waves are sometimes called *E* waves, and *TE* waves are sometimes called *H* waves, but the *TM* and *TE* designations are preferred.

The particular mode of transmission is identified by the group letters followed by two subscript numerals; for example, $TE_{1,0}$, $TM_{1,1}$, etc. The number of possible modes increases with frequency for a given size of guide. There is only one possible mode (called the *dominant* mode) for the lowest frequency that can be transmitted. The dominant mode is the one generally used in practical work.

Wave-Guide Dimensions

In the rectangular guide the critical dimension is x in Fig. 19-5; this dimension must be more than one-half wavelength at the lowest frequency to be transmitted. In practice, the y dimension usually is made about equal to $\frac{1}{2}x$ to avoid the possibility of operation at other than the dominant mode.

Other cross-sectional shapes than the rectangle can be used, the most important being the circular pipe. Much the same considerations apply as in the rectangular case.

Wavelength formulas for rectangular and circular guides are given in the following table, where x is the width of a rectangular guide and r is the radius of a circular guide. All figures are in terms of the dominant mode.

	Rectangular	*Circular*
Cut-off wavelength	$2x$	$3.41r$
Longest wavelength transmitted with little attenuation	$1.6x$	$3.2r$
Shortest wavelength before next mode becomes possible	$1.1x$	$2.8r$

Cavity Resonators

At low and medium radio frequencies resonant circuits usually are composed of "lumped" constants of L and C; that is, the inductance is concentrated in a coil and the capacitance concentrated in a condenser. However, as the frequency is increased, coils and condensers must be reduced to impracticably small physical dimensions. Up to a certain point this difficulty may be overcome by using linear circuits but even these fail at extremely high frequencies. Another kind of circuit particularly applicable at wavelengths of the order of centimeters is the *cavity resonator*, which may be looked upon as a section of a wave guide with the dimensions chosen so that waves of a given length can be maintained inside.

The derivation of one type of cavity resonator from an ordinary *LC* circuit is shown in Fig. 19-8. As in the case of the wave-guide derivation, this picture must be accepted with some reservations, and for the same reasons.

Considering that even a straight piece of wire has appreciable inductance at very-high frequencies, it may be seen in Fig. 19-8A and B that a direct short across a two-plate condenser with air dielectric is the equivalent of a tuned circuit with a typical coiled inductance. With two wires between the plates, as shown in Fig. 19-8C, the circuit may be thought of

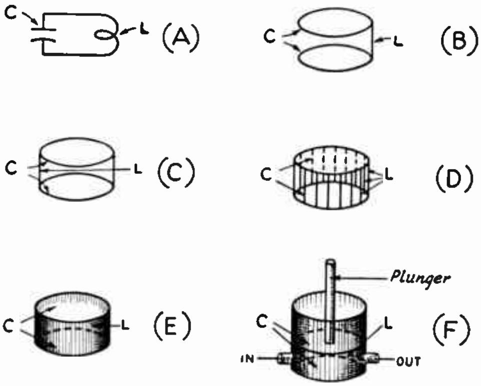


Fig. 19-8 — **Steps in the derivation of a cavity resonator from a conventional coil-and-condenser tuned circuit.**

as a resonant-line section. For d.c. or even low frequency r.f., this line would appear as a short across the two condenser plates. At the ultra-high frequencies, however, such a section of line a quarter wavelength long would appear as an open circuit when viewed from one of the plates with respect to the other end of the section.

Increasing the number of parallel wires between the plates of the condenser would have no effect on the equivalent circuit, as shown at D. Eventually, the closed figure at E will be developed. Since each wire which is added in D is like connecting inductances in parallel, the total inductance across the condenser becomes increasingly smaller as the solid form is approached, and the resonant frequency of the figure therefore becomes higher.

If energy now is introduced into the cavity in a manner such as that shown at F, the circuit will respond like any equivalent coil-condenser tank circuit at its resonant frequency. A cavity resonator may therefore be used as a u.h.f. tuning element, along with a vacuum tube of suitable design, to form the main components of an oscillator circuit which will be capable of functioning at frequencies considerably beyond the maximum limits possible when conventional tubes, coils and condensers are employed.

Other shapes than the cylinder may be used as resonators, among them the rectangular box, the sphere, and the sphere with re-entrant cones, as shown in Fig. 19-9. The resonant frequency depends upon the dimensions of the cavity and the mode of oscillation of the waves (comparable to the transmission modes in a wave guide). For the lowest modes the resonant wavelengths are as follows:

Cylinder	2.61*r*
Square box	1.41*l*
Sphere	2.28*r*
Sphere with re-entrant cones	4*r*

The resonant wavelengths of the cylinder and square box are independent of the height when the height is less than a half-wavelength. In other modes of oscillation the height must be a multiple of a half-wavelength as measured inside the cavity. Fig. 19-8F shows how a cylindrical cavity can be tuned when operating in such a mode. Other tuning methods include placing adjustable tuning paddles or "slugs" inside the cavity so that the standing-wave pattern of the electric and magnetic fields can be varied.

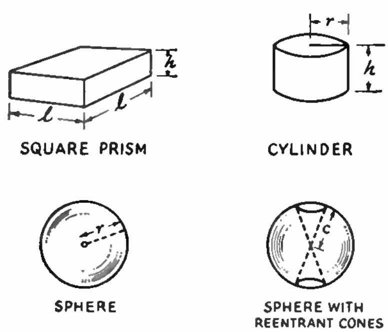


Fig. 19-9 — **Forms of cavity resonators.**

A form of cavity resonator in wide practical use is the re-entrant cylindrical type shown in Fig. 19-10. It is useful in connection with vacuum-tube oscillators of the types described for u.h.f. use elsewhere in this chapter. In construction it resembles a concentric line closed at both ends with capacitance loading at the top, but the actual mode of oscillation may differ considerably from that occurring in coaxial lines. The resonant frequency of such a cavity depends upon the diameters of the two cylinders and the distance *d* between the ends of the inner and outer cylinders.

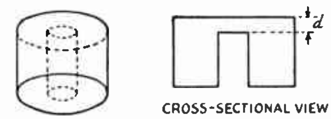


Fig. 19-10 — **Re-entrant cylindrical cavity resonator.**

Compared to ordinary resonant circuits, cavity resonators have extremely-high Q. A value of Q of the order of 1000 or more is readily obtainable, and Q values of several thousand can readily be secured with good design and construction.

Coupling to Wave Guides and Cavity Resonators

Energy may be introduced into or abstracted from a wave guide or resonator by means of either the electric or magnetic field. The energy transfer frequently is through a coaxial line, two methods for coupling to which are shown in Fig. 19-11. The probe shown at A is simply a short extension of the inner conductor of the coaxial line, so oriented that it is parallel to the electric lines of force. The loop shown at B is arranged so that it encloses some of the magnetic lines of force. The point at which maximum coupling will be secured depends upon the particular mode of propagation in the guide or cavity; the coupling will be maximum when the coupling device is in the most intense field.

Coupling can be varied by turning either the probe or loop through a 90-degree angle. When the probe is perpendicular to the electric lines the coupling will be minimum; similarly, when the plane of the loop is parallel to the magnetic lines the coupling will have its least possible value.

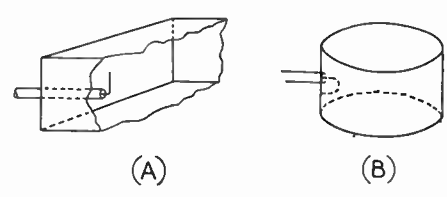


Fig. 19-11 — **Coupling to wave guides and resonators.**

U.H.F. and Microwave Tubes

At ultra-high frequencies, interelectrode capacitance and the inductance of internal leads determine the highest possible frequency to which a vacuum tube can be tuned. The tube usually will not oscillate up to this limit, however, because of dielectric losses, grid emission, and "transit-time" effects. In low-frequency operation, the actual time of flight of electrons between the cathode and the anode is negligible in relation to the duration of the cycle. At 1000 kc., for example, transit time of 0.001 microsecond, which is typical of conventional tubes, is only 1/1000 cycle. But at 100 Mc., this same transit time represents 1/10 of a cycle, and a full cycle at 1000 Mc. These limiting factors establish about 3000 Mc. as the upper frequency limit for negative-grid tubes.

With most tubes of conventional design, the upper limit of useful operation is around 150 Mc. For higher frequencies tubes of special construction are required. The "acorn" and "doorknob" types have been available for many years, these being useful up to 500 Mc. or more in special circuits. Newer miniature types, developed for use in u.h.f. television receivers, now provide good performance up to nearly 1000 Mc.

Very low interelectrode capacitance and lead inductance have been achieved in the newer tubes of modified construction. In some types the electrodes are provided with up to five separate leads which, when connected in parallel, have considerably-reduced effective inductance. In double-lead types the plate and grid elements are supported by heavy single wires which run entirely through the envelope, providing terminals at either end of the bulb. When a resonant circuit is connected to each pair of leads, the shunting capacitance divides between the two circuits. With linear circuits the leads become a part of the line and have distributed rather than lumped constants. Radiation loss is minimized and the effect of the transit time is reduced. In "lighthouse" tubes or *megatrons* the plate, grid and cathode are assembled in parallel planes, as shown in Fig. 19-12, instead of coaxially. The uniform coplanar electrode design and disk-seal terminals permit low interelectrode capacitance.

Velocity Modulation

In negative-grid operation the potential on the grid tends to reduce the electron velocity during the more negative half of the oscillation cycle, while on the other half-cycle the positive potential on the grid serves to accelerate them. Thus the electrons tend to separate into groups, those leaving the cathode during the negative half-cycle being collectively slowed down, while those leaving on the positive half are accelerated. After passing into the grid-plate space only a part of the electron stream follows the original form of the oscillation cycle, the remainder traveling to the plate at differing velocities. Since these contribute nothing to the power output at the operating frequency, the efficiency is reduced in direct proportion to the variation in velocity, the output reaching a value of zero when the transit time approaches a half-cycle.

This effect, such a disadvantage in conventional tubes, is an advantage in velocity-modulated tubes in that the input signal voltage on the grid is used to change the velocity of the electrons in a constant-current electron beam, rather than to vary the intensity of a constant-velocity current flow as is the method in ordinary tubes.

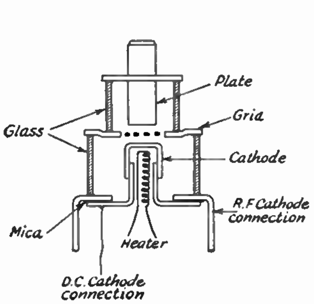


Fig. 19-12 — Sectional view of the "lighthouse" tube's construction. Close electrode spacing reduces transit time while the disk electrode connections reduce lead inductance.

A simple form of velocity-modulation oscillator tube is shown in Fig. 19-13. Electrons emitted from the cathode are accelerated through a negatively-biased cylindrical grid by a constant positive voltage applied to a sleeve electrode, shown in heavy lines. This electrode, which is the velocity-modulation control grid, consists of two hollow tubes, with a small space at each end between the inner tube, through which the electron beam passes, and the disks at the ends of the larger tube portion. With r.f. voltage applied across these gaps, which are small compared to the distance traveled by the electrons in one half-cycle, electrons entering the tube will be accelerated on positive half-cycles and decelerated on the negative half-cycles. The length of the tube is made equal to the distance covered by the electrons in one-half cycle, so that the electrons will be further accelerated or decelerated as they leave the tube.

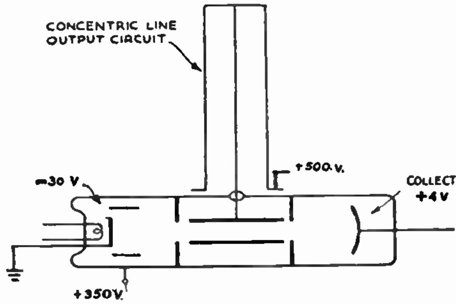


Fig. 19-13 — Simple form of cylindrical-grid velocity-modulated tube with retarding-field collector and coaxial-line output circuit, used as a superheterodyne high-frequency oscillator or as a superregenerative detector. Similar tubes can also be used as r.f. amplifiers and frequency converters in the 5–50-cm. region.

As the beam approaches the collector electrode, which is at nearly zero potential, the electrons are retarded, brought to rest, and ultimately turned back by the attraction of the positive sleeve electrode. The collector electrode is, therefore, also termed a *reflector*. The point at which electrons are returned depends on their velocity. Thus the velocity modulation is again translated into current modulation.

Velocity-modulated tubes operate satisfactorily up to 6000 Mc. (5 cm.) and higher, with outputs of 100 watts or more.

The Klystron

In the *klystron* velocity-modulated tube, the electrons emitted by the cathode are accelerated or retarded during their passage through an electric field established by two grids in a cavity resonator, or *rhumbatron*, called the "buncher." The high-frequency electric field between the grids is parallel to the electron stream. This field accelerates the electrons at one moment and retards them at another, in accordance with the variations of the r.f. voltage applied. The resulting velocity-modulated beam travels through a field-free "drift space," where the slowly-moving electrons are gradually overtaken by the faster ones. The electrons emerging from the pair of grids therefore are separated into groups or bunched along the direction of motion. The velocity-modulated electron stream is passed to a "catcher" rhumbatron. Again the beam passes through two parallel grids: the r.f. current created by the bunching of the electron beam induces an r.f. voltage between the grids. The catcher cavity is made resonant at the frequency of the velocity-modulated electron beam, so that an oscillating field is set up within it by the passage of the electron bunches through the grid aperture.

If a feed-back loop is provided between the two rhumbatrons, as shown in Fig. 19-14, oscillations will occur. The resonant frequency depends on the electrode voltages and on the shape of the cavities, and may be adjusted by varying the supply voltage and altering the dimensions of the rhumbatrons. The bunched beam current is rich in harmonics, but the output waveform is remarkably pure because the high Q of the catcher rhumbatron suppresses the unwanted harmonics.

Magnetrons

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

Magnetron oscillators are operated in two different ways. Electrically the circuits are similar, the difference being in the relation between electron transit time and the frequency of oscillation.

In the negative-resistance or dynatron type

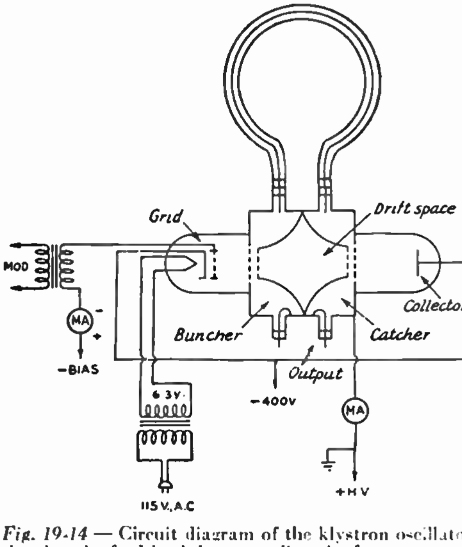


Fig. 19-14 — Circuit diagram of the klystron oscillator, showing the feed-back loop coupling the frequency-controlling rhumbatrons and the output loop in the catcher.

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

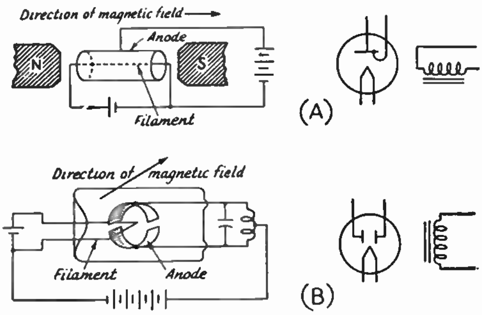


Fig. 19-15 — Conventional magnetrons, with equivalent schematic symbols at the right. A, simple cylindrical magnetron. B, split-anode negative-resistance magnetron.

travel to that half of the anode that is at the lower potential. In other words, a *decrease* in the potential of either half of the anode results in an *increase* in the electron current flowing to that half. The magnetron consequently exhibits negative-resistance characteristics. Negative-resistance magnetron oscillators are useful between 100 and 1000 Mc. Under the best operating conditions efficiencies of 20 to 25 per cent may be obtained. Since the power loss in the tube appears as heat in the anode, where it is readily dissipated, relatively large power-handling capacity can be obtained.

In the transit-time magnetron the frequency is determined primarily by its dimensions and by the electric and magnetic field intensities rather than by the tuning of the tank circuits. The efficiency is much better than that of a positive-grid oscillator and good power output can be obtained even on the superhighs.

In a nonoscillating magnetron with a weak magnetic field, electrons traveling from the cathode to the anode move almost radially, their trajectories being bent only slightly by the magnetic field. With increased magnetic field the electrons tend to spiral around the filament, their radial component of velocity being much smaller than the angular component. Under critical conditions of magnetic field strength, a cloud of electrons rotates about the filament. It extends up to the anode but does not actually reach it.

The nature of these electron trajectories is shown in Fig. 19-16. Cases A, B and C correspond to the nonoscillating condition. For a small magnetic field (A) the trajectory is bent slightly near the anode. This bending increases for a higher magnetic field (B) and the electron moves through quite a large angle near the anode before reaching it, signifying a large increase of space charge near the anode. For a strong magnetic field (C) electrons start radially from the cathode but are soon bent and curl about the filament in the form of a long spiral before reaching the anode. This means a very long transit time and a very large space charge in the whole region where the spiraling takes place. Under critical conditions (D), no current flows to the anode and no electron is able to move from cathode to anode, but a large space charge still exists between the cathode and anode. The spiraling becomes a set of concentric circles, and the entire space-charge distribution rotates about the filament.

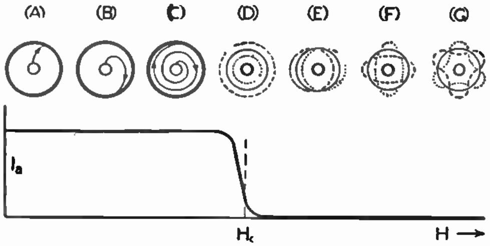


Fig. 19-16 — Electron trajectories for increasing values of magnetic field strength, *H*. Below is shown the corresponding curve of plate current, I_a. Oscillations commence when *H* reaches a critical value, H_c; progressively higher-order modes of oscillation occur beyond this point.

Fig. 19-16E, F and G depicts higher-order (harmonic-type) modes of operation in which the space charge oscillates not only symmetrically but in transverse directions contrasting to the vibrations of the fundamental.

In a transit-time magnetron oscillator the intensity of the magnetic field is adjusted so that, under static conditions, electrons leaving the cathode move in curved paths which just fail to reach the anode. All electrons are therefore deflected back to the cathode, and the anode current is zero. When an alternating voltage is applied between the two halves of the anode, causing the potentials of these halves to vary about their average positive values, the conditions in the tube become analogous to those in a positive-grid oscillator. If the period of the alternating voltage is made equal to the time required for an electron to make one complete rotation in the magnetic field, the a.c. component of the anode voltage reverses direction twice with each electron rotation. Some electrons will lose energy to the electric field, with the result that they are unable to reach the cathode and continue to rotate about it. Meanwhile other electrons gain energy from the field and are returned to the cathode.

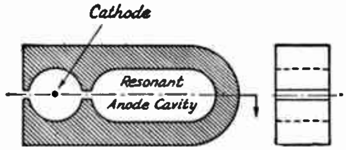


Fig. 19-17 — Split-anode magnetron with integral resonant anode cavity for use at u. h.f.

Since those electrons that lose energy remain in the interelectrode space longer than those that gain energy, the net effect is a transfer of energy from the electrons to the electric field. This energy can be applied to sustain oscillations in a resonant transmission line connected between the two halves of the anode.

Split-anode magnetrons for u.h.f. are constructed with a cavity resonator built into the tube structure, as illustrated in Fig. 19-17. The assembly is a solid block of copper which assists in heat dissipation. At extremely high frequencies operation is improved by subdividing the anode structure into from 4 to 16 or more segments, the resonant cavities for each anode coupled by slots of critical dimensions to the common cathode region, as in Fig. 19-18.

The efficiency of multisegment magnetrons reaches 65 or 70 per cent. Slotted-anode magnetrons with four segments function up to 30,000 Mc. (1 cm.), delivering up to 100 watts at efficiencies greater than 50 per cent. Using larger multiples of anodes and higher-order modes, performance can be attained at 0.2 cm.

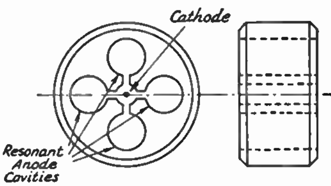


Fig. 19-18 — Multisegment magnetron with four resonant cavities. This construction is used for extremely high frequencies.

Traveling-Wave Tubes

Gain as high as 23 db. over a bandwidth of 800 Mc. at a center frequency of 3600 Mc. has been obtained through the use of a fairly-simple traveling-wave amplifier tube. Shown schematically in Fig. 19-19, the circuit consists of a helix, down which an electromagnetic wave travels. 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.

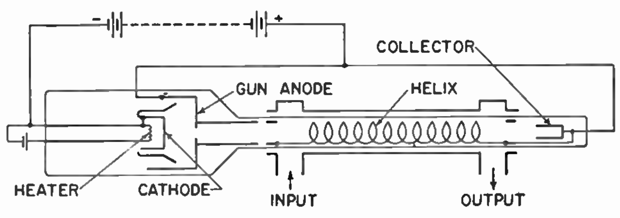


Fig. 19-19 — Schematic drawing of a traveling-wave amplifier tube.

The portions of Fig. 19-19 marked "input" and "output" are wave-guide 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 most valuable feature of the traveling-wave tube is its great bandwidth. The gain is high, though the efficiency is rather low. Typical power output is of the order of 200 milliwatts.

Amateur Microwave Technique

All the bands that have been assigned to amateurs in the microwave region have been used for experimental two-way communication. Complete descriptions of suitable equipment for all these bands is beyond the scope of this text, but examples of the techniques employed are shown below. Reference is made to various articles that have appeared in *QST*, describing microwave gear used by amateurs, for those who wish more details.

1215 Mc.

In this band it is possible to use a few more-or-less conventional triodes with linear circuits, though great care must be used in designing such layouts, and the efficiency will be very low. A transmitter for 1215 Mc., designed and built by W3MLN and W3HFW, is shown in Figs. 19-20 — 19-22. It uses a 703A doorknob triode, completely shielded, with the antenna as an integral part of the assembly. The tube is mounted at the end of a halfwave line. Output is capacitively coupled to the folded quarter-wave antenna by means of a probe mounted alongside the plate line.

It should be emphasized that complete shielding of the oscillating circuit (including the tube elements) is absolutely necessary. The circuit will not oscillate at all if the shield is removed from the grid and plate rods, and only very weakly if the tube shield is not in place. Output is only about one watt, with an input of 80 ma. at 350 volts, but two of these units have been used to communicate over distances up to 12 miles or so with S9 signals. The equipment is described in detail by the designers in *QST* for April, 1948, page 16.

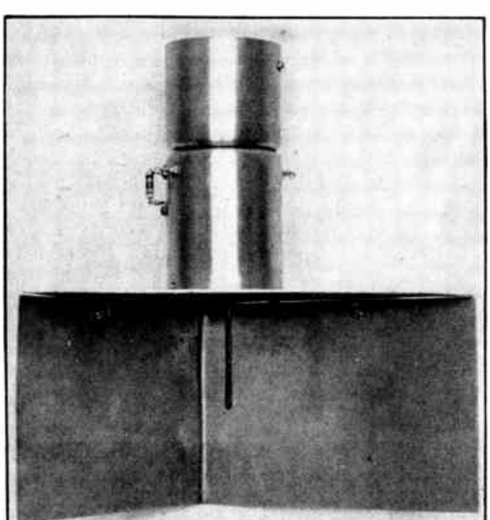


Fig. 19-20 — An oscillator and antenna system for 1215 Mc., built as one unit. (*W3HFW* – *W3MLN*)

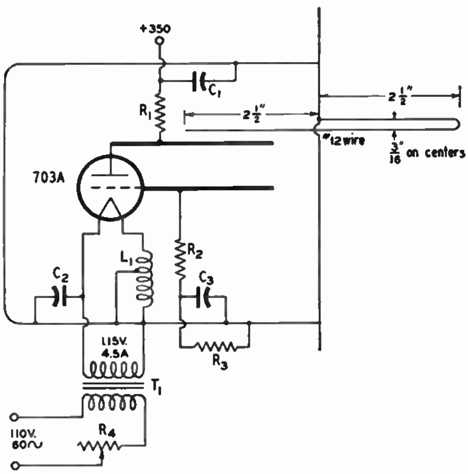


Fig. 19-21 — Schematic diagram of the 1215-Mc. oscillator.

Lighthouse tubes in suitably designed circuits are more efficient at this frequency. For best results cavities should be used, though trough-line and flat-plate circuits have been used.

Parabolic reflectors are usually employed for this and higher frequencies. It is desirable to make the transmitter or receiver an integral part of the antenna system if possible. If this cannot be done, coaxial line of the shortest usable length may be used. Air-insulated line is preferred to the flexible polyethylene-insulated variety, because of the higher losses in the latter.

2300 Mc.

Most of the work on 2300 Mc. has been done with lighthouse tubes in cavity oscillators, though some of the klystron types such as the 707B have been used. Cavities for this frequency may be a quarter wavelength, half wavelength or three-quarter wavelength long.

Details of a half-wave cavity oscillator using a 2C40 lighthouse tube are shown in Figs. 19-22 and 19-23. This oscillator was designed and built by W2RMA. It may be duplicated by any worker who has access to a few metal-working tools.

The main body of the cavity is 1-inch brass pipe, silver plated. The end that fits over the tube is cut out to an inside diameter of 1 1/32 inch, the

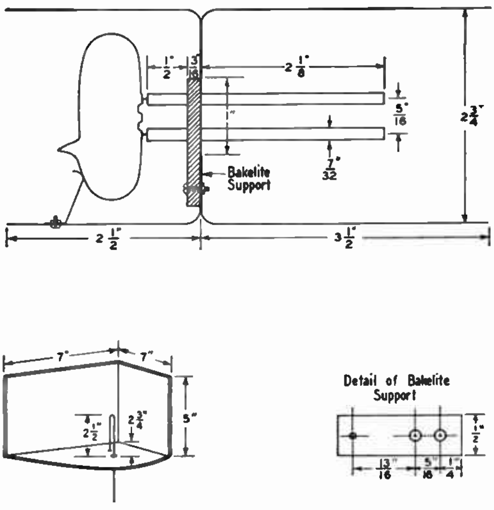


Fig. 19-22 — Detail drawing of the 703A oscillator for 1215 Mc.

only lathe work required. This end is also sawed crosswise at several points so that it may be clamped tightly to the tube with a brass strap, as seen in the photograph. Plate voltage is fed into the cavity through a feed-through capacitor mounted on the side of the tubing, and power is coupled out by means of a capacity probe and coaxial fitting at the hot end. The cavity is tuned with a screw mounted in the end, providing a variable capacitance to the anode post.

Output, with a 250-volt supply, will be 50 to 250 milliwatts. This seemingly small amount of power may be made to do very well with the antenna gain that is possible at this frequency with a parabolic reflector of reasonable dimensions. Gear for 2300 Mc. is described in *QST* for July, 1946, page 32, August, 1947, page 128, and February, 1948, page 11.

3300 Mc.

Lighthouse oscillators may be used on this frequency, but it is close to the top limit of their capabilities, so better results are obtainable with the klystron types. An advantage of the latter is that the frequency of oscillation may be varied over an appreciable range by changing the reflector voltage. This characteristic is also useful in providing a convenient means of obtaining frequency modulation. This sensitivity to voltage changes makes it desirable to use a regulated hum-free supply.

On this and higher frequencies a convenient system for two-way work is the use of a klystron as both transmitting oscillator and as a local oscillator for receiving. A crystal mixer is used in this case, its output being fed into a receiver serving as the i.f. system. If the receiver so used is capable of f.m. detection it is only necessary to modulate the klystron reflector voltage to provide f.m. communication of good quality. The oscillators of the two stations in communication are then operated on frequencies differing by the value of the intermediate frequency selected. A single antenna system is used for both transmitting and receiving, and no change-over arrangement is needed.

5650 Mc.

Amateur work in this range has been done largely with reflex klystrons, two types of which (2K43 and 2K44) are capable of operation within our band. The one-tube system described above may be used for each station, or of course separate tubes may be used for transmitter and local oscillator. In the latter case two antenna systems are required, but the transmitter efficiency is somewhat higher as some power is dissipated across the crystal in the one-tube arrangement.

Frequency modulation of klystrons is more practical than amplitude modulation. Modulation of the repellor voltage requires no audio power, as there is no current drawn by this tube element. A carbon microphone and a microphone transformer, with the repellor voltage fed through the secondary, will handle the audio requirements nicely.

The first two-way microwave communication in amateur history was carried out in this way by A. E. Harrison, W6BMS/2, and R. E. Merchant, W2LGF, who operated in the temporary 5300-Mc. band. Their equipment, described in *QST* for January, 1946, page 19, will also work in the present band.

10,000 Mc.

The 723A/B reflex klystron, available at low cost for some time on the surplus market, provided amateurs with a convenient and inexpensive means of operation on 10,000 Mc. As manufactured, the tube will not ordinarily operate in the amateur band without modification.

Like other tubes of the reflex klystron variety, the frequency of oscillation is varied by warping the built-in cavity. It is used with a modified octal socket, with pin No. 4 removed and the

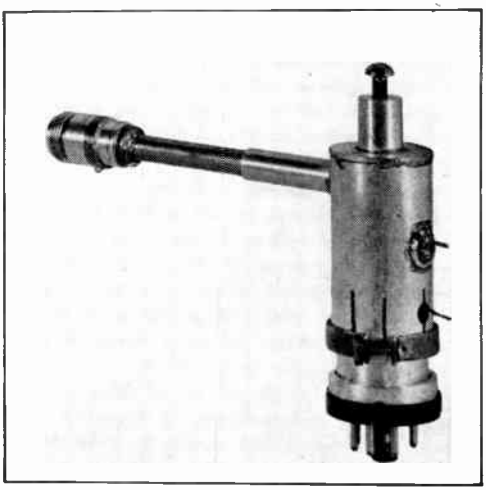


Fig. 19-23 — A half-wave cavity oscillator for 2300 Mc. (*W2RMA*)

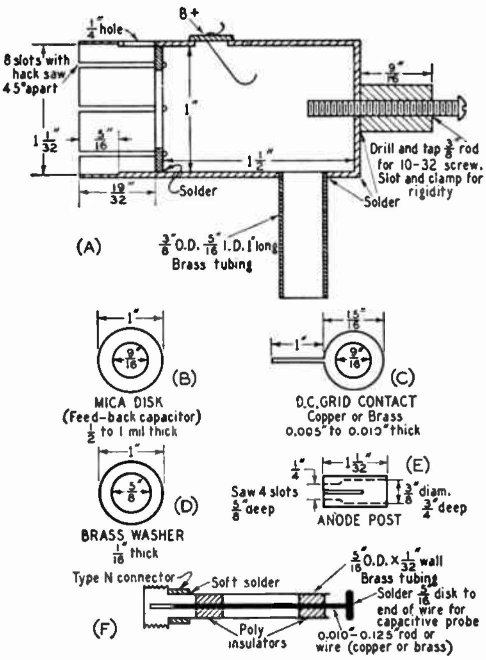


Fig. 19-24 — Mechanical details of the 2300-Mc. lighthouse oscillator.

hole enlarged to pass the coaxial line that is part of the tube. This line is terminated in an "antenna" which is ordinarily used to transfer power to a waveguide.

Two vertical struts are provided for tuning, one of which is already variable by means of a stud, which spreads or contracts the flexible strut on the right side, compressing or stretching the bellows, lowering or raising the frequency respectively.

The upper limit of frequency range, reached by rotating the tuning stud, will seldom be within the amateur band, hence it is necessary to perform the following operation. It may be seen that the top of the cavity is held in a fixed position on the strut on the side of the tube by two small nuts which, after having been tightened, have been spot-welded to each other. The spot weld should be filed away until each nut can be moved freely on the threaded stud. Next, the position of these nuts should be adjusted *very* carefully, to raise the top of the cavity as was done on the other side. Extreme care should be used in this operation, as excessive stretching of the bellows may break some of the seals and render the tube inoperative. It is advisable to move the lower nut only until a firm resistance is felt. The operating frequency should then be checked, and if it is still below the limit of the band another tube should be tried, as any further attempt to raise the frequency will almost certainly ruin the tube.

Equipment for use on 10,000 Mc. is described in detail in *QST* for February, 1947, page 58.

21,000 Mc.

Operation in this frequency, and in the unassigned region above 30,000 Mc. is still highly experimental in nature. Only once has the 21,000-Mc. band been used for amateur two-way communication. This was accomplished under laboratory conditions by two engineers whose specialty is development work in this field. Their work is detailed in *QST* for August, 1946, page 19. Type Z-668 reflex klystrons were used, with horn and parabolic antenna systems, to work two-way over a distance of 800 feet.

Mobile Equipment

The amateur who goes in for mobile operation will find plenty of room for exercising his individuality and developing original ideas in equipment. Each installation has its special problems to be solved.

Most mobile receiving systems are designed around the use of a h.f. converter working into a standard car broadcast receiver tuned to 1500 kc. which serves as the i.f. and audio amplifiers. The car receiver is modified to take a noise limiter and provide power for the converter.

While a few mobile transmitters may run an input to the final amplifier as high as 100 watts or more, an input of about 30 watts normally is considered the practical limit unless the car is equipped with a special battery-charging system. The majority of mobile operators use 'phone.

In contemplating a mobile installation, the car should be studied carefully to determine the most suitable spots for mounting the equipment. Then the various units should be built in a form that will make best use of that space. The location of the converter should have first consideration. It should be placed where the controls can be operated conveniently without distracting attention from the wheel. The following list suggests spots that may be found suitable, depending upon the individual car.

- On top of the instrument panel
- Attached to the steering post
- Under the instrument panel
- In a unit made to fit between the lower lip of the instrument panel and the floor at the center of the car
- On the left-hand door panel (detachable when not in use)
- Under the left-hand front seat
- In the motor compartment (controls extended through the instrument panel)

The transmitter power control can be placed close to the receiver position, or included in the converter unit. This control normally operates relays, rather than to switch the power circuit directly. This permits a minimum length of heavy-current battery circuit. Frequency within any of the 'phone bands sometimes is changed remotely by means of a stepping-switch system that switches crystals. In most cases, however, it is necessary to stop the car to make the several changes required in changing bands.

Depending upon the size of the transmitter unit, one of the following places may be found convenient for mounting the transmitter:

- In the glove compartment
- Under the instrument panel
- In a unit in combination with or without the converter, built to fit between the lower edge of the instrument panel and the floor at the center
- Under the right-hand or left-hand front seat
- On the ledge above the rear seat
- Fastened to the back of the front seat
- In the trunk
- In the motor compartment

Most mobile antennas consist of a vertical whip with some system of adjustable loading for the lower frequencies. Power supplies are of the vibrator-transformer-rectifier or motor-generator type operating from the car storage battery.

Units intended for use in mobile installations should be assembled with greater than ordinary care, since they will be subject to considerable vibration. Soldered joints should be well made and wire wrap-arounds should be used to avoid dependence upon the solder for mechanical strength. Self-tapping screws should be used wherever feasible, otherwise lock-washers should be provided. Any shafts that are normally operated at a permanent or semi-permanent setting should be provided with shaft locks so they cannot jar out of adjustment. Where wires pass through metal, the holes should be fitted with rubber grommets to prevent chafing. Any cabling or wiring between units should be securely clamped in place where it cannot work loose to interfere with the operation of the car.

Noise Elimination

Electrical-noise interference to reception in a car may arise from several different sources. As examples, trouble may be experienced with ignition noise, generator and voltage-regulator hash, or wheel and tire static.

A noise limiter added to the car b.c. receiver will go far in reducing some types, especially ignition noise from passing cars as well as your own. But for the satisfactory reception of weaker signals, some investigation and treatment of the car's electrical system will be necessary.

Ignition Interference

Fig. 20-1 indicates the measures that may be taken to suppress ignition interference. The condenser at the primary of the ignition coil should be of the coaxial type; ordinary types are not effective. It should be placed as close to the coil terminal as possible. In stubborn cases, two

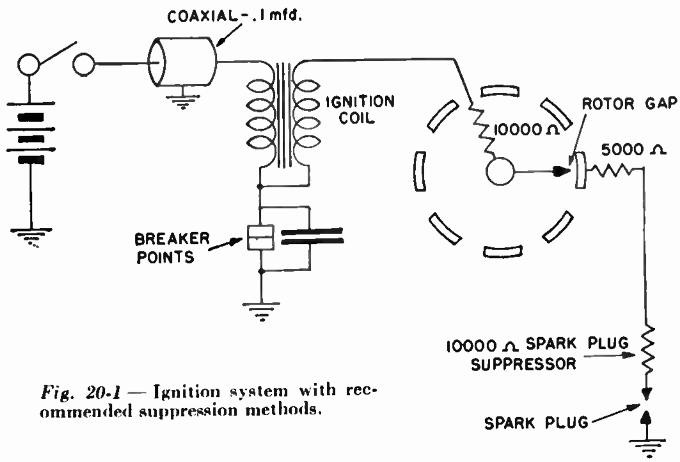


Fig. 20-1 — Ignition system with recommended suppression methods.

of these condensers with an r.f. choke between them may provide additional suppression. The size of the choke must be determined experimentally. The winding should be made with wire heavy enough to carry the coil primary current. A 10,000-ohm suppressor resistor should be inserted at the center tower of the distributor, a 5000-ohm suppressor at each spark-plug tower on the distributor, and a 10,000 ohm suppressor at each spark plug. The latter may be built-in or external. A good suppressor element should be molded of material having low capacitance. Erie type L7VR-10ME and L7VR-5ME are satisfactory. In extreme cases, it may be necessary to use shielded ignition wire. The 1951 Pontiac car was equipped with suppressor ignition wires, the resistance being distributed throughout the length of the wire. This is somewhat superior to lumped resistance and may be used if the lead lengths are right to fit your car. They should not be cut, but used as they are sold.

Generator Noise

Generator hash is caused by sparking at the commutator. The pitch of the noise varies with the speed of the motor. This type of noise may be eliminated by using a 0.1- to 0.25-μfd. coaxial condenser in the generator armature circuit. This condenser should be mounted as near the armature terminal as possible and directly on the frame of the generator.

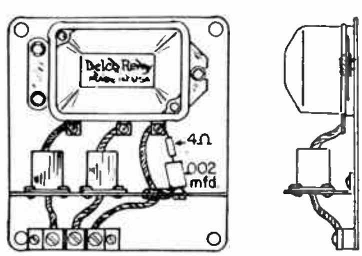


Fig. 20-2 — The right way to install by-passes to reduce interference from the regulator. A condenser should never be connected across the generator field lead without the small series resistor indicated.

To reduce the noise at 28 Mc., it may be necessary to insert a parallel trap, tuned to the middle of the band, in series with the generator output lead. The coil should have about 8 turns of No. 10 wire, space-wound on a 1-inch diameter and should be shunted with a 30-$\mu\mu$fd. mica trimmer. It can be pretuned by putting it in the antenna lead to the home-station receiver tuned to the middle of the band, and adjusting the trap to the point of minimum noise. The tuning may need to be peaked up after installing in the car, since it is fairly critical.

Voltage-Regulator Interference

In eliminating voltage-regulator noise, the use of two coaxial condensers, and a resistor-mica-condenser combination, as shown in Fig. 20-2, are effective. A 0.1- to 0.25-μfd. coaxial condenser should be placed between the battery terminal of the regulator and the battery, with its case well grounded. Another condenser of the same size and type should be placed between the generator terminal of the regulator and the generator. A 0.002-μfd. mica condenser with a 4-ohm carbon resistor in series should be connected between the field terminal of the regulator and ground. Never use a condenser across the field contacts or between field and ground without the resistor in series, since this greatly reduces the life of the regulator. In some cases, it may be necessary to pull double-braid shielding over the leads between the generator and regulator. It will be advisable to run new wires, grounding the shielding well at both ends. If regulator noise persists, it may be necessary to insulate the regulator from the car body. The wire shielding is then connected to the regulator case at one end and the generator frame at the other.

Wheel Static

Wheel static shows up as a steady popping in the receiver at speeds over about 15 m.p.h. on smooth dry streets. Front-wheel static collectors are available on the market to eliminate this variety of interference. They fit inside the dust cap and bear on the end of the axle, effectively grounding the wheel at all times. Those designated particularly for your car are preferable, since the universal type does not always fit well. They are designed to operate without lubrication and the end of the axle and dust cap should be cleaned of grease before the installation is made. These collectors require replacement about every 10,000 miles.

Rear-wheel collectors have a brush that bears against the inside of the brake drum. It may be necessary to order these from the factory through your dealer.

Tire Static

This sometimes sounds like a leaky power line and can be very troublesome even on the broadcast band. It can be remedied by injecting an antistatic powder into the inner tubes through the valve stem. The powder is marketed by Chevrolet and possibly others. Chevrolet dealers can also supply a convenient injector for inserting the powder.

Tracing Noise

To determine if the receiving antenna is picking up all of the noise, the shielded lead-in should be disconnected at the point where it connects to the antenna. The motor should be started with the receiver gain control wide open. If no noise is heard, all noise is being picked up via the antenna. If the noise is still heard with the antenna disconnected, even though it may be reduced in strength, it indicates that some signal from the ignition system is being picked up by the antenna transmission line. The lead-in may not be sufficiently-well shielded, or the shield not properly grounded. Noise may also be picked up through the 6-volt circuit, although this does not normally happen if the receiver is provided with the usual r.f.-choke-and-by-pass-condenser filter.

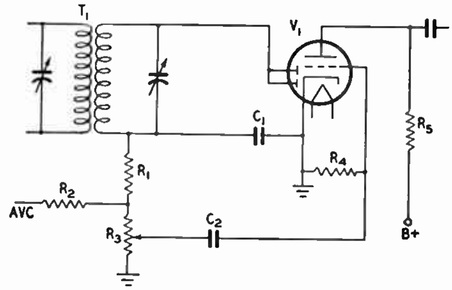
In case of noise from this source, a direct wire from the "hot" battery terminal to the receiver is recommended.

Ignition noise varies in repetition rate with engine speed and usually can be recognized by that characteristic in the early stages. Later, however, it may resolve itself into a popping noise that does not always correspond with engine speed. In such a case, it is a good idea to remove all leads from the generator so that the only source left is the ignition system.

Regulator and generator noise may be detected by racing the engine and cutting the ignition switch. This eliminates the ignition noise. Generator noise is characterized by its musical whine contrasted with the ragged raspy irregular noise from the regulator.

With the motor running at idling speed, or slightly faster, checks should be made to try to determine what is bringing the noise into the field of the antenna. It should be assumed that any control rod, metal tube, steering post, etc., passing from the motor compartment through an insulated bushing in the firewall will carry noise to a point where it can be radiated to the antenna. All of these should be bonded to the firewall with heavy wire or braid. Insulated wires can be stripped of r.f. by by-passing them to ground with 0.5-μfd. metal-case condensers. The following should not be overlooked: battery lead at the ammeter, gasoline gauge, ignition switch, headlight and taillight leads and the wiring of any accessories running from the motor compartment to the instrument panel or outside the car.

(A)



(B)

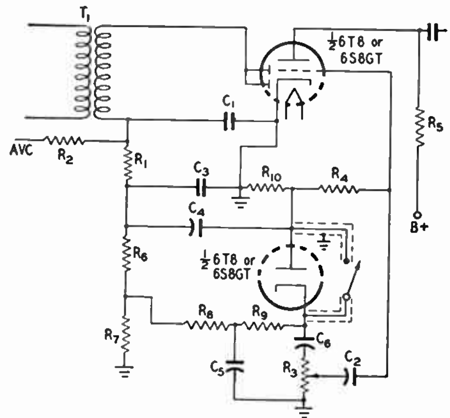


Fig. 20-3 — **Diagrams showing addition of noise limiter to car receiver. A — Usual circuit. B — Modification.**

C_1, C_3 — 100-μμfd. mica.
C_2, C_4, C_6 — 0.01-μfd. paper.
C_5 — 0.1-μfd. paper.
R_1 — 47,000 ohms.
R_2, R_{10} — 1 megohm.
R_3 — ½ megohm
R_7, R_8, R_9 — 0.47 megohm.
R_4 — 10 megohms.
R_5 — ¼ megohm.
R_6 — 0.1 megohm.
T_1 — I.f. transformer.
V_1 — Second detector.

The firewall should be bonded to the frame of the car and also to the motor block with heavy braid. If the exhaust pipe and muffler are insulated from the frame by rubber mountings, they should likewise be grounded to the frame with flexible copper braid.

Noise Limiter

Fig. 20-3 shows the alterations that may be made in the existing car-receiver circuit to provide for a noise limiter. The usual diode-triode second detector is replaced with a type having an extra independent diode. If the car receiver uses octal-base tubes, a 6S8GT may be substituted. The 7X7 is a suitable replacement in receivers using loktal-type tubes, while the 6T8 may be used with miniatures.

The switch that cuts the limiter in and out of the circuit may be located for convenience on or near the converter panel. Regardless of its placement, however, the leads to the switch should be shielded to prevent hum pick-up.

A Compact Multiband Mobile Converter

Figs. 20-4 through 20-9 show photographs and diagrams of a small mobile converter covering all bands from 3.5 to 29 Mc.

As the diagram of Fig. 20-6 indicates, the circuit includes an r.f. stage, mixer and h.f. oscillator, each using a 6AJ5, similar to the 6AK5 which can be used interchangeably in this circuit. The input circuit can be peaked up with the 50-$\mu\mu$fd. air trimmer, C_1. The grid circuit of the mixer is broadbanded, requiring no attention after preliminary adjustment. The main tuning control is C_{17} in the h.f. oscillator circuit. Fixed parallel padders are selected to spread each of the bands over a good share of the dial. All coils, including the i.f., are slug-tuned. Included in the bandswitch are the sections S_{1G} and S_{1H} which turn off the filament and plate power, as well as the dial lamps, when the gang is thrown to the b.c. position. A small relay, controlled from the transmitter panel, cuts the B supply to the converter while transmitting. The over-all dimensions are 3⅞ by 5⅛ by 6½ inches, not including protuberances, such as the r.f. tuning knob and the power plug. The panel is 5 by 3¾ inches and includes the dial, antenna-trimmer control and bandswitch. The chassis is 5 by 5¾ by 2. All parts of the enclosure are of aluminum sheet.

The dial mechanism is a planetary unit with a 5 to 1 ratio (National AVD). This is mounted on the panel 1¼ inches from the bottom edge. The dial face is a piece of ¼-inch Lucite or Plexiglas 3 by 5 inches. A semicircle is cut out of the bottom edge with a jig saw to clear the dial mechanism, and is also notched out on the right-hand side to pass the shaft of the antenna trimmer. Before making these cuts, however, the various dial scales should be laid out with a compass scriber, using the position of the dial shaft as the scribing center. The back side of the plastic is covered with ordinary black or other dark-colored paint to form a contrasting background for the calibration marks. A dial lamp is mounted in each upper corner of the panel and the plastic is drilled part way through at these points. The ends of the bulbs extend into these depressions and the transmitted light illuminates the panel. Twelve-volt lamps (operating at 6 volts, of course), or two 6-volt lamps in series, provide plenty of light at half normal voltage. A metal cover of light-gauge aluminum was fashioned to fit over the upper corners of the plastic to eliminate direct light from the lamps. The pointer is a piece of thin transparent plastic, cut to shape and fastened to the dial mechanism with the screws provided. A line is scribed down the center of the pointer.

Underneath, the main tuning-condenser shaft is matched up with the dial shaft and mounted in place. While the condenser shown in the photograph is a two-section job, only one of the sections is used. An L-shaped shield runs along the right-hand side and across the rear of the condenser to isolate it from the antenna trimmer mounted nearby on the right-hand edge of the chassis.

The bandswitch gang is made up from Cen-

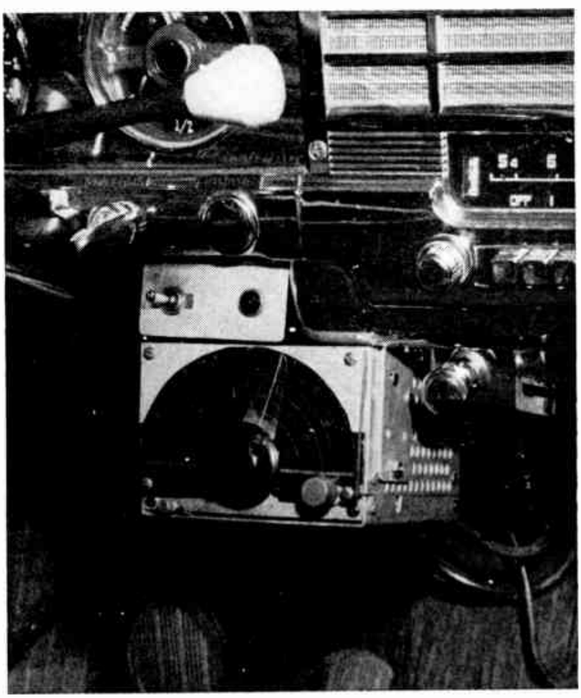


Fig. 20-4 — Bandswitching converter designed by W3MNR and W3DZZ installed under the dashboard near the b.c. receiver.

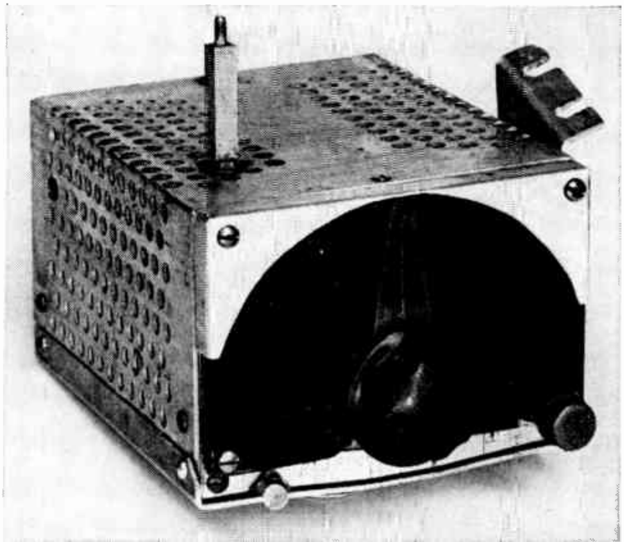


Fig. 20-5 — The dial of the band-switching mobile converter is a piece of clear plastic with calibration marks inscribed. The bandswitch control is at the lower left and the antenna trimmer to the right.

tralab switch-kit parts and consists of five ceramic wafers. Three wafers carry two circuits of five positions (Centralab type RR). The sixth position, shown in the diagram, is the arm slider contact which can be used in this case because the last switch position for all but S_{1D} is an open-circuit position. S_{1C} and S_{1D} are separate wafers each having one circuit and six positions (Centralab type X). The switch is mounted directly behind the main tuning condenser in a vertical position, its shaft 3⅜ inches from the front edge of the chassis. This unusual mounting is convenient for grouping tubes and coils around the switch sections. Only the switch index head and the first wafer are below the chassis. The two circuits of this wafer, comprising S_{1A} and S_{1B}, handle

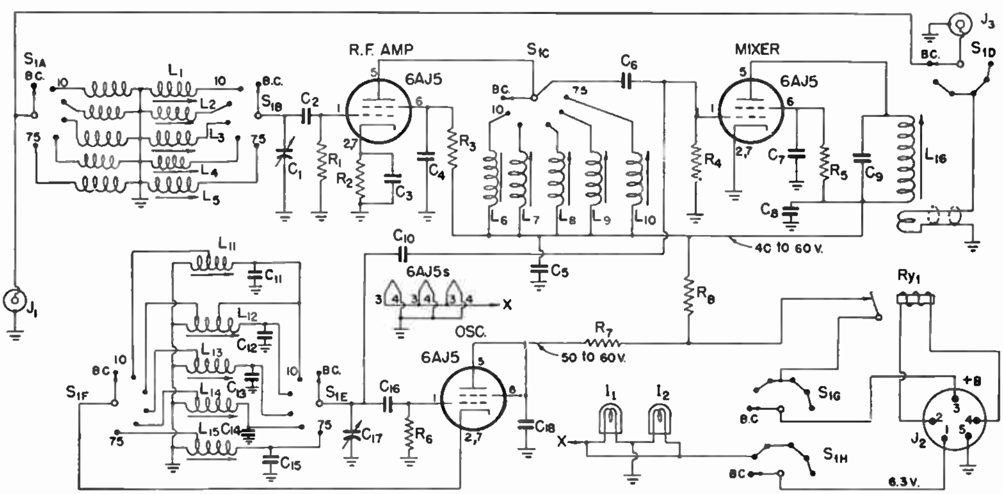


Fig. 20-6 — Circuit of the bandswitching converter.

C_1 — 50-μμfd. miniature variable.
C_2, C_6 — 50-μμfd. mica.
C_3 — 100-μμfd. mica.
C_4, C_5, C_7, C_8, C_{18} — 0.001-μfd. mica.
C_9 — 220-μμfd. mica.
C_{10} — 3 μμfd.
C_{11} — 51-μμfd. silvered mica.
C_{12}, C_{13} — 160-μμfd. silvered mica.
C_{14} — 150-μμfd. silvered mica.
C_{15} — 33-μμfd. silvered mica.
C_{16} — 33-μμfd. mica.
C_{17} — 15-μμfd. variable.
R_1, R_4, R_6 — 10,000 ohms, ½ watt.
R_2 — 180 ohms, ½ watt.
R_3, R_5 — 2000 ohms, ½ watt.
R_7, R_8 — Values dependent on supply voltage. Adjust for voltages marked.
I_1, I_2 — 12-volt dial lamp.
J_1, J_3 — Coaxial connector.
J_2 — 5-pin male power plug.
Ry_1 — 6-volt relay.
S_1 — Ceramic rotary switch — 4 wafers, 2 circuits per wafer, 6 positions per circuit, and 1 wafer, 1 circuit, 6 positions (1 below, 4 above chassis) (made from Centralab kit parts).

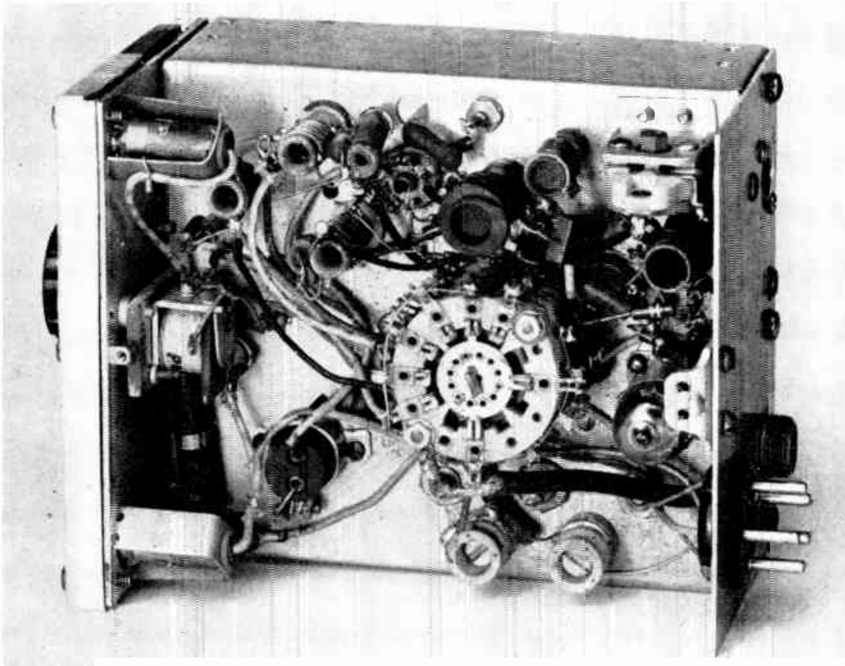


Fig. 20-7 — Top view of the bandswitching converter, showing oscillator and mixer coils grouped around the bandswitch. The relay mounted against the front edge of the chassis cuts the power to the converter during transmissions.

the r.f. input circuits. The other four wafers are mounted above and a clearance hole for the switch shaft is drilled in the chassis. Additional bracing against the action of the control lever is provided by adding a strap bracket across the index head at right angles to the assembly rods. This strap is fastened to holes in the index head and with long screws to the chassis.

Coil Table for Bandswitching Converter

Coil	*Band, Mc.*	*Lμh.*	*Turns*	*Wire Size*	*Length, Inches*	*Ant. Turns*	*Cath. Tap*
L_1*	29	0.85	12	28 d.s.c.	¾	3	—
L_2*	21	1.4	16	28 d.s.c.	¾	3	—
L_3*	14	3.1	24	28 d.s.c.	¾	4	—
L_4	7	11	52	28 d.s.c.	c-w.	7	—
L_5	3.5	38	92	34 d.s.c.	c-w.	15	—
L_6*	29	1.5	18	28 d.s.c.	¾	—	—
L_7*	21	2.8	27	28 d.s.c.	¾	—	—
L_8*	14	6.2	35	28 d.s.c.	c-w.	—	—
L_9	7	25	98	34 d.s.c.	c-w.	—	—
L_{10}	4	82	140	34 enam.	c-w.	—	—
L_{11}*	29	0.35	9	28 d.s.c.	¾	—	3
L_{12}*	21	0.26	7	28 d.s.c.	¾	—	3
L_{13}*	14	0.56	10	28 d.s.c.	¾	—	4
L_{14}*	7	1.9	18	28 d.s.c.	¾	—	6
L_{15}	4	15	60	28 d.s.c.	c-w.	—	20
L_{16}*	1.5	52	100	34 enam.	c-w.	25	—

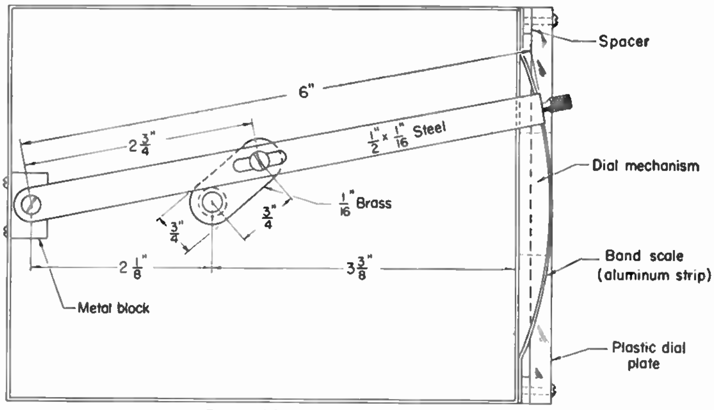
* Wound on National XR-91 iron-slug form, ⅝ in. diam., 1¼ in. long, as close as possible to end opposite slug screw; others same, but on XR-93 forms, 1¾ in. long. Antenna coils wound over ground end of r.f. grid coils with same size wire. Cathode-tap turns counted from ground end of oscillator coils.

A sketch of the switch operating mechanism is shown in Fig. 20-8. Dimensions can be adjusted to suit a variety of conditions. It is merely a matter of experimenting with a few pieces of cardboard and some thumbtacks to find dimensions that will fit each case. The short arm attached to the switch shaft should preferably be of brass so that the nut can be soldered fast. The set-screw collar to which the short arm is attached is a panel bearing. The threaded neck is cut and filed down so that it is a little longer than the thickness of the arm. The excess is then hammered down over the arm to make a firm joint. Solder flowed around the hole will add strength. The flange of the panel bearing should be drilled and tapped for two set screws. The bandswitch scale is a strip of thin aluminum. The positions for the various bands are marked with a scriber and the lines filled in with crayon.

The r.f. tube is the only one mounted top-side up. The mixer and oscillator tubes are upside down and have their connections and associated coils above the chassis. This arrangement permits better utilization of space and the chassis becomes a shield for the r.f. circuit.

Adjustment

Standard automobile receivers are designed for high-impedance antennas and transmission lines. Since the output of the converter is coupled to a low-impedance coax line, considerable mismatch results. Most b.c. receivers have enough gain so that the losses as a consequence can be tolerated. However, the gain



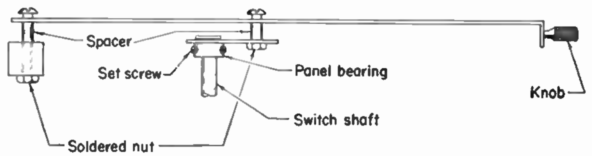
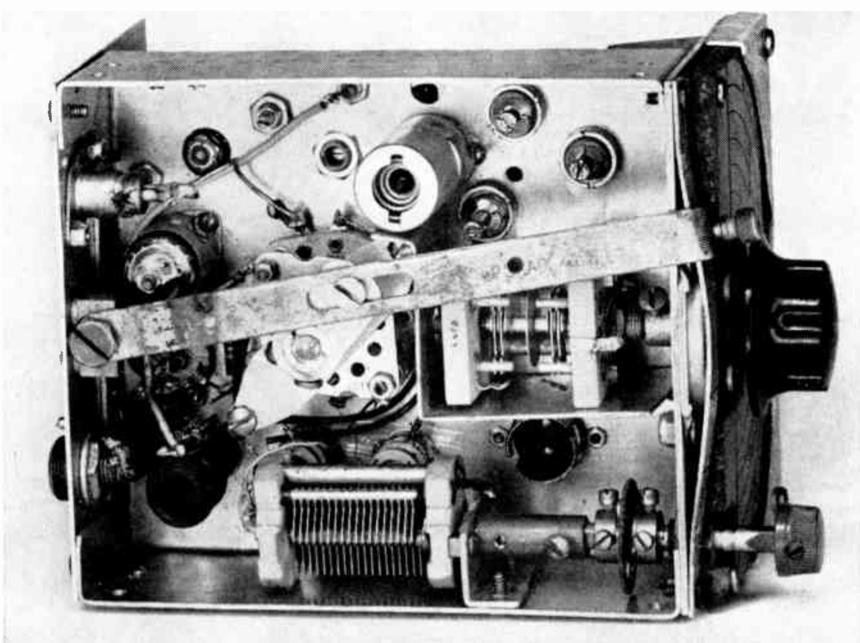


Fig. 20-8 — Sketches showing the construction and dimensions of the bandswitch mechanism for the multiband converter.

can be increased considerably by modifying the r.f. coil in the b.c. set. This is accomplished by winding a link of about 25 turns of No. 28 wire on the "cold" end of the antenna coil. This modification, however, will reduce the gain on the b.c. band. One compromise is to use one push button only for the converter and modify only the coil associated with that channel.

The entire converter was wired and aligned with a grid-dip meter before applying power. Depending on the forms used, some slight alteration in the number of turns shown in the coil table may be necessary.

Fig. 20-9 — Bottom view of the bandswitching converter showing the switch operating mechanism and inverted mounting of the h.f. oscillator and mixer tubes.



A Mobile Converter for 28 and 50 Mc.

The converter shown in Figs. 20-10 to 20-13 was designed for mobile reception on 6, 10, and 11 meters, but it may also be used in fixed-station work with good results. The intermediate frequency is 1500 kc., to permit its use with mobile broadcast receivers.

Circuit Details

The converter circuit diagram is shown in Fig. 20-11. A 6AK5 broadband r.f. amplifier is followed by a 6J6 mixer-oscillator. The oscillator circuit is the ultraudion type, operating 1500 kc. below the signal frequency. The need for gang-tuned circuits is eliminated by the broadband r.f. amplifier; thus only the oscillator tuning condenser, C_1, requires adjustment during normal tuning operation. Band

Fig. 20-10 — A bandswitching converter for 6, 10 and 11 meters. The pilot light at the lower right has an adjustable beam, for convenience in mobile work.

changing is accomplished with a 5-section selector switch, shown on the diagram as $S_{1A, B, C, D, E}$.

Seven commercially-available coils are used, six of them being identical except for the setting of the slugs. The wide inductance range of the slug-tuned units makes it possible to use similar coils for the r.f., mixer and oscillator coils for both ranges. Padder capacitance is added across the 10-meter r.f. and mixer coils, L_4 and L_6, and across both oscillator coils, L_7 and L_8. Varying the slug position takes care of the necessary differences in coil inductance for all these positions.

A single whip antenna may be used for both broadcast and amateur reception. A jumper connection between sections A and E of S_1 completes the circuit between the antenna and the broadcast receiver, with the switch in the position marked $B.C.$ in Fig. 20-11. A filament switch, S_2, is provided to remove the load of the converter tubes from the car battery when the receiver is being used for broadcast reception.

Broadbanding of the r.f. and mixer circuits is accomplished through the use of low-Q coils and tight coupling in the antenna circuit. The plate coil of the mixer is self-resonant at the intermediate frequency, giving a degree of broadness sufficient to permit tuning the receiver over a limited range near the high end of the broadcast band, providing a vernier effect.

Construction

All of the metal components are formed from 1/16-inch aluminum stock. The interior view, Fig. 20-12, shows the "L"-shaped section which serves as the front panel and the bottom plate of the unit. The panel and the bottom areas are each 5 inches square. Lips, 1/2 inch wide, are folded over along the top and side edges of the panel and also along the sides of the bottom section. The rolled-over edges are drilled and tapped to accommodate 6-32 machine screws.

A three-sided portion and a square top plate complete the converter cabinet. The sides are 5 inches square and the rear wall is 5 1/8 inches wide. All three sides are 5 inches high with 1/2-inch flanges folded over on the top edges and drilled and tapped for 6-32 screws. The sides and bottom edges of the case are drilled to clear machine screws; the holes should line up with the tapped holes of the panel-bottom assembly. A rectangular hole, 1 7/8 inches high and 2 inches wide, is cut at the bottom left-hand corner (as seen from the rear of the converter) of the rear wall, to provide clearance for the cable connectors. The top plate for the converter measures 5 by 5 inches. Holes, drilled along the edges, allow the cover to be fastened to the flanges at the top of the cabinet.

The physical shape of the converter chassis can best be visualized by study of the interior views. The chassis is 5 by 4 7/8 by 1 3/4 inches in size, with flanges 1/2 inch wide folded over along the front and the bottom edges to provide a means of mounting. A 2 1/4 × 3 3/4-inch cut-out at the center of the chassis allows clearance for the bandswitch. A large round hole located in the rear wall of the chassis simplifies the job of finding the oscillator padder condenser when this control requires adjustment.

A vertical partition used as the mounting surface for the oscillator tuning condenser, C_1, also serves as the shield between the plate and the grid circuits of the r.f. amplifier. It is 3 1/2 inches wide and 4 3/4 inches high, and is notched to clear the main chassis and the spacer bars and rotor arm of the bandswitch. The partition is held in place by a spade lug which passes through the chassis and by a mounting

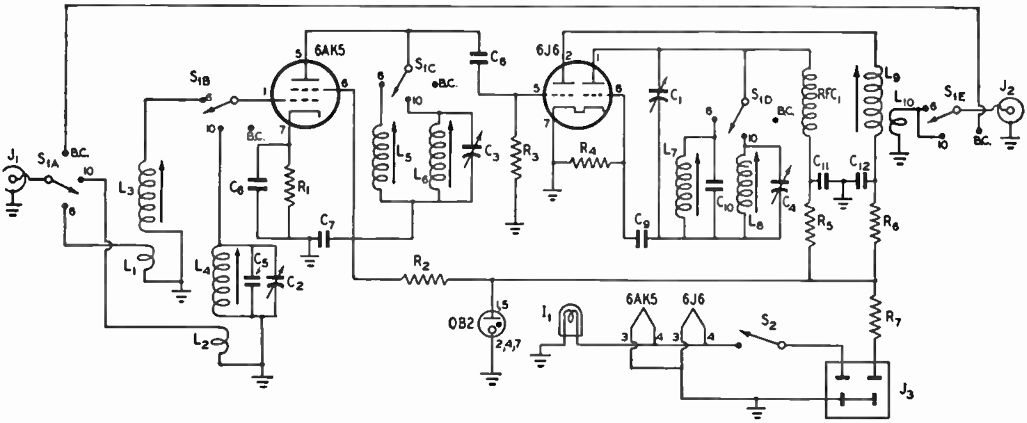


Fig. 20-11 — Circuit diagram of the bandswitching v.h.f. converter.

C_1 — 15-μμfd. variable reduced to one stator and 2 rotor plates (Millen 20015).
C_2, C_3, C_4 — 3–30-μμfd. mica trimmer (Millen 27030).
C_6, C_7 — 0.0015-μfd. ceramic (Centralab DA048002A).
C_8, C_9 — 100-μμfd. ceramic (Centralab CC32Z).
C_5, C_{10} — 10-μμfd. ceramic (Centralab CC20Z).
C_{11} — 500-μμfd. ceramic (Centralab D6501).
C_{12} — 0.01-μfd. ceramic (Centralab DA048003A).
R_1 — 220 ohms, ½ watt.
R_2, R_6 — 680 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 12,000 ohms, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_7 — 5000 ohms, 10 watts.
L_1, L_2 — 4 turns No. 28 d.s.c. close-wound over ground ends of L_3 and L_4.
L_3, L_4, L_5, L_6, L_7, L_8 — 6 turns No. 20 enameled wire close-wound on ⅜-inch diameter form; slug-tuned: inductance range 0.35 to 1.0 μh. (Cambridge Thermionic Corp. LS3-30 Mc.).
L_9 — Scramble-type winding on ⅜-inch slug-tuned form: inductance range 325 to 750 μh. (Cambridge Thermionic Corp. LS3-1 Mc.).
L_{10} — 20 turns No. 28 d.s.c. scramble-wound next to L_9.
I_1 — Adjustable-beam dial-light assembly.
J_1, J_2 — Coaxial-cable jacks (Amphenol 75-PC1M).
J_3 — 3-prong cable connector (Jones P-303AB).
RFC_1 — 300-μh. r.f. choke (Millen 34300).
$S_{1\,A,B,C,D,E}$ — 2-gang 6-circuit bandswitch (two Centralab SS sections).
S_2 — S.p.s.t. toggle switch.

lip which is screwed to the bottom side of the cabinet. It is located 3 inches in from the front edge of the chassis.

The heater switch and the pilot-light assembly are mounted at the lower left- and right-hand corners of the front panel with the bandswitch at the center, 1⅛ inches up from the bottom edge. The selector-switch index plate should have a rotor-shaft length of at least 3 inches, and the switch wafers should be mounted on the shaft with the first separated from the index plate by 1-inch spacers and with the second wafer separated from the first by 1⅝ inches.

The National MCN dial is centered above the bandswitch with the control shaft 3 inches above the bottom edge of the panel. It is wise to cut the large mounting hole suggested in the dial-mounting instruction sheet and then do the final fastening down of the dial after the tuning condenser and its mounting plate have been permanently secured in place.

The interior view of the completed converter shows the 6AK5 amplifier tube in front of the shield partition, with the grid inductances to

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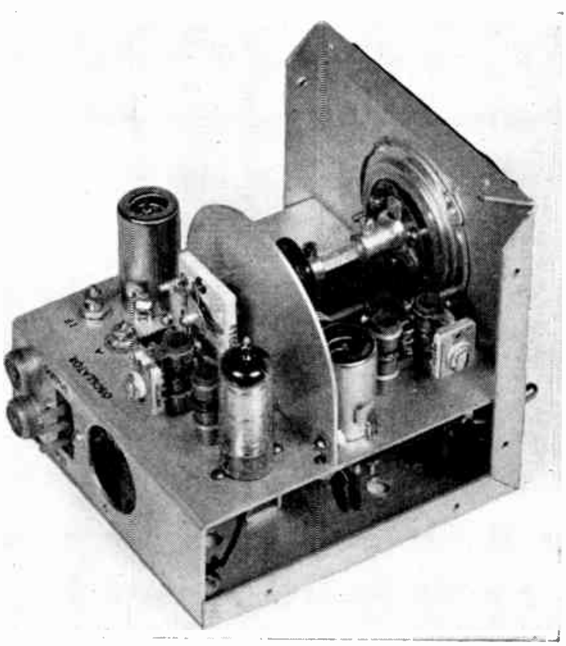


Fig. 20-12 — Interior view of the converter. Only the oscillator is tuned by the front-panel control, eliminating tracking problems.

the right of the tube. The padder condensers for 27 and 28 Mc. are mounted on the forward coil. From left to right across the rear of the chassis are the mixer-oscillator tube, five of the slug-tuned inductances, and the regulator tube. The i.f. output coil and the two oscillator coils are mounted below the chassis, as seen in the bottom view of the chassis subassembly. The r.f. plate coils are above the chassis to the left of the 0B2 regulator, the 28-Mc. coil being the one with the trimmer condenser mounted across the terminals.

Construction will be simpler if the builder uses coils as shown. The Type LS3 30-Mc. inductors will resonate at 50 Mc. with the tube and circuit capacitances, and only a small padder capacitance is required to tune them to 27 and 28 Mc.

Coaxial jacks for the antenna and i.f. output cables are at the rear of the chassis to the left of the power-cable jack. They are closely grouped so that the input and output cables may be taped together to form a common cable.

Wiring can be done readily if the subassembly method is employed. The bottom-view photograph of the chassis, Fig. 20-13, shows how the circuit components are closely grouped around the tube sockets, with wiring completed to the point of making connections to the band-switch. Twin-Lead of the 75-ohm type is used to make the connection between the antenna input jack and the bandswitch. The two wires enclosed in spaghetti at the right of the chassis in the bottom view are the 6.3-volt leads which go to the heater switch.

Testing

The heater requirements of the converter are 6.3 volts at 0.625 amp., and the plate supply should deliver 200 to 250 volts at 25 to 30 ma. These may be drawn from the receiver with which the converter is to be used, or a separate supply may be employed. With power turned on, the plate voltage of the mixer and r.f. amplifier should measure 105 volts and the 6AK5 cathode resistor should provide a drop of approximately 2 volts. The 6AK5 cathode current should be about 8.5 ma. The regulator-tube drain will be about 8 ma.

Alignment of the converter is made most simple if a calibrated signal generator is available, otherwise amateur transmitter signals of known frequency may be used. The r.f. and i.f. circuits can be peaked on background noise. The oscillator stage should be on the low side of the signal frequency. It is possible to vary the bandspread of the converter over a wide range. With a fairly low order of padder capacitance, and with the inductance increased by the tuning slug, the 10- and 11-meter bands can be covered with one swing of the tuning dial. Anyone not interested in 11 meters can increase the bandspread on the 10-meter range by adding more padder capacitance and by decreasing the inductance of L_8. The converter as shown has 13 divisions of bandspread at 11 meters and 52 divisions at 10 meters, with the logging of frequencies made on the B scale of the dial. Bandspread for the 50-Mc. band is 48 divisions on the A scale. This spread may be increased by the same method.

Some operators favor a selected group of frequencies within a band. A slight improvement in the performance of the converter can be made in this case by peaking the r.f. amplifier circuits at a favorite spot rather than at the center of a band. There may be a tendency toward regeneration in the 50-Mc. r.f. amplifier, however, if the input and plate circuits are peaked at precisely the same frequency, making stagger tuning desirable.

Reducing Spurious Responses

In localities where there are stations operating in the high FM band a converter or receiver having broadband r.f. stages will experience considerable interference on the 50-Mc. range. This can be corrected in several ways, the simplest being the insertion of a 100-Mc. trap in the antenna lead.

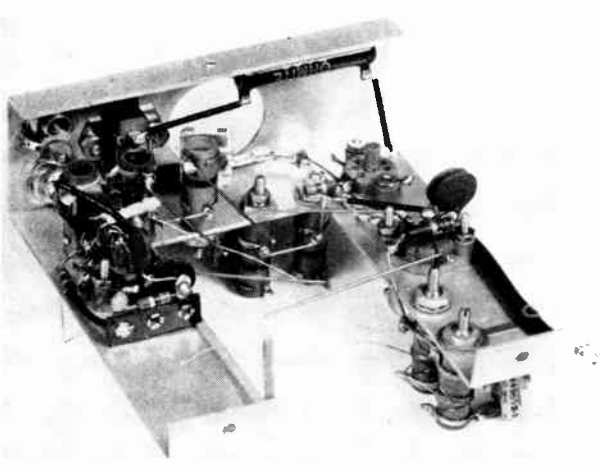


Fig. 20-13 — Construction of the converter is made easier if as much wiring as possible is done before the assembling is completed. This bottom view of the chassis subassembly shows the wiring completed to the point of connection to the bandswitch.

A Crystal-Controlled Converter for Two-Meter Mobile Reception

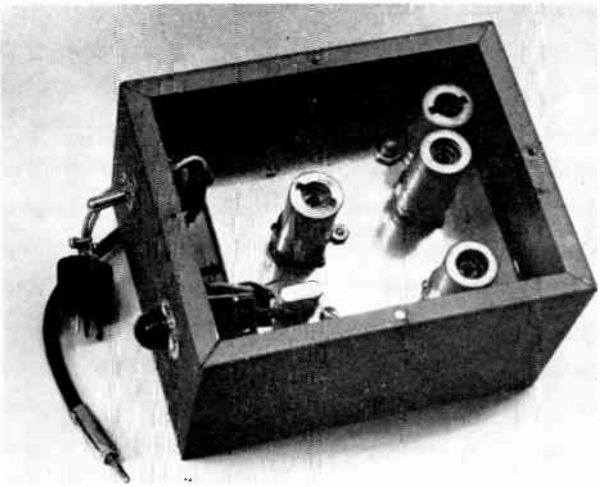


Fig. 20-14 — Top view of W2UTH's crystal-controlled converter for 2-meter mobile reception. The oscillator-multiplier tube and crystal are at the left. At the right are the r.f. amplifier, mixer and i.f. amplifier, looking up from the bottom. Because no external adjustments are needed, the converter may be built in almost any shape that will fit available space in the car.

The 144-Mc. mobile converter shown in Figs. 20-14 through 20-16 is designed primarily for mobile operation. Therefore to serve the aims of simplicity, compactness and low battery drain, some of the features that might be considered desirable in a home-station unit have been omitted. However, the cost is low and the performance of the system is entirely satisfactory, both as to stability and sensitivity.

Circuit

Since the tuning range of the usual car broadcast receiver is insufficient to permit coverage of the entire 2-meter band without changing crystals, this converter is designed to work into another converter which, in turn, works into the regular car receiver. This second converter is used as a tunable i.f. and should cover the range of 26 to 30 Mc. to provide the necessary 4-Mc. range to take care of the whole of the 2-meter band.

The r.f. stage uses a 6AK5, pentode connected. This results in a slight sacrifice in noise figure, compared to that obtainable with a triode, but with the other noises usually prevalent in mobile work, the ultimate in first-tube is not so important in practice. The mixer is a 6AB4 triode.

The oscillator is the simplest form of triode circuit, using a crystal at 39.33 Mc. in the first half of the 6J6, the second portion tripling to 118 Mc. Crystals such as the James Knights JK-H17 or H-173, the Bliley BH-6, or GE G64B, can be readily obtained for this frequency.

Where the mixer is a separate tube from the oscillator-multiplier, some injection coupling may be necessary, although the minimum required value should be used. The 1.5 μμfd. needed was obtained by connecting two 3-μμfd. units in series.

The i.f. stage, using a 6AK5, employs an output circuit that provides low-impedance coupling to the following converter.

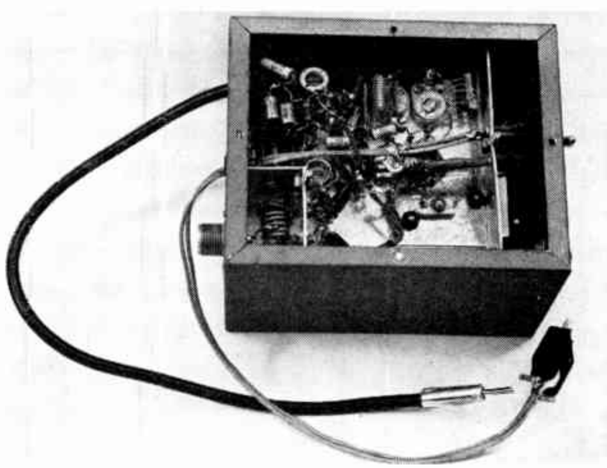


Fig. 20-15 — Bottom view of the 2-meter converter. The coil form at the upper left is the mixer plate circuit. Oscillator-multiplier components are at the upper right.

The converter is built on a 5 × 5-inch chassis that fits inside a standard utility box. Since there is no adjustment required during operation, the unit can be built in almost any shape that can be fitted into available space in the car. The coils and condensers are mounted under the chassis, and once the initial adjustment is made, they are left alone.

In order to isolate the input and output circuits, of the r.f. amplifier, a small right-angle shield is placed across the 6AK5 socket in such a way as to enclose the antenna coil. The shield may be seen in the lower left side in the bottom view of Fig. 20-15. The antenna is connected directly to the grid coil through coaxial cable.

The mixer output coil, L_4, is mounted between the 6AB4 and the i.f. amplifier tube, in the upper right-hand corner in the top view of Fig. 20-14.

At a supply voltage of 150, the converter drain will be about 15 ma. If a higher supply voltage is used, R_{15} should be increased accordingly. Adjustment is straightforward. The slug in L_5 is first adjusted for maximum background noise in the output of the system. Then L_4 is adjusted for maximum response on 2-meter signals in the most-used part of the band. L_1 can be peaked up by squeezing the turns together or spreading them apart slightly as needed.

With a 19-inch whip good signals have been obtained with this converter at distances up to 30 miles or more.

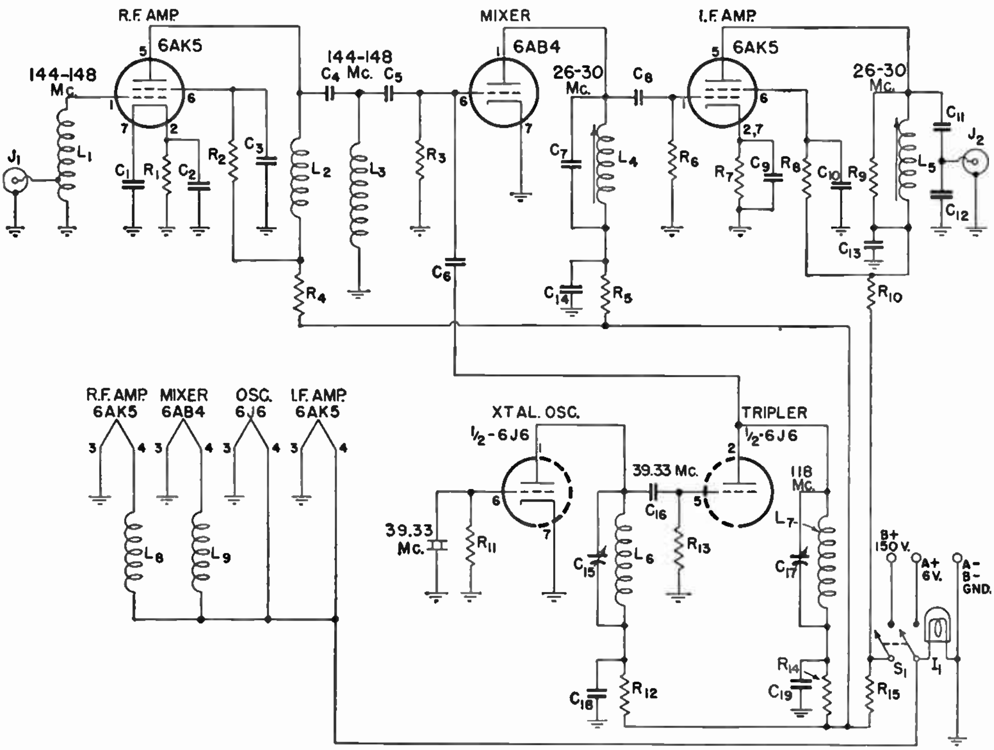


Fig. 20-16 — Schematic diagram and parts list for the crystal-controlled 2-meter converter. If crystals lower in frequency than 39 Mc. are to be used an overtone oscillator circuit can be substituted for the crystal circuit shown.

C_1, C_2, C_3, C_9, C_{10}, C_{13}, C_{14}, C_{18}, C_{19} — 0.001 μfd.
C_4, C_{11} — 5 $\mu\mu$fd.
C_5, C_8 — 50 $\mu\mu$fd.
C_6 — 1.5 $\mu\mu$fd. (two 3-$\mu\mu$fd. in series).
C_7 — 10 $\mu\mu$fd.
C_{12} — 30 $\mu\mu$fd.
C_{15}, C_{17} — 4-30-$\mu\mu$fd. ceramic trimmer.
C_{16} — 25 $\mu\mu$fd.
(All fixed capacitors ceramic.)
R_1 — 150 ohms.
R_2 — 10,000 ohms.
R_3 — 0.68 megohm.
R_4 — 1000 ohms.
R_5 — 3300 ohms.
R_6 — 0.1 megohm.
R_7 — 680 ohms.
R_8 — 39,000 ohms.
R_9 — 7000 ohms.
R_{10} — 1500 ohms.
R_{11} — 47,000 ohms.
R_{12}, R_{14} — 4700 ohms.
R_{13} — 0.22 megohm.
R_{15} — 5600 ohms, 1 watt. (All other resistors ½ watt.)
L_1 — 5 turns No. 16, ⅜-inch diam., ½ inch long, tapped at 1½ turns.
L_2 — ½-watt resistor wound full of No. 30 enameled wire.
L_3 — 3 turns No. 16, ⅜-inch diam., ¼ inch long.
L_4 — 10 turns No. 24 enam. on 13/32-inch diam. form (Millen 69041), brass slug.
L_5 — 10 turns No. 20 enam. on ½-inch slug-tuned form from BC-624 receiver. National XR-50 also usable.
L_6 — 11 turns No. 18, ½-inch diam. (B & W No. 3003 Miniductor).
L_7 — 3 turns No. 18, ½-inch diam.
L_8, L_9 — ½-watt resistor wound full of No. 18 enam.
J_1 — Coaxial fitting, female.
J_2 — Coaxial fitting, male.
S_1 — Double-pole single-throw toggle switch.

A Multiband Mobile Transmitter

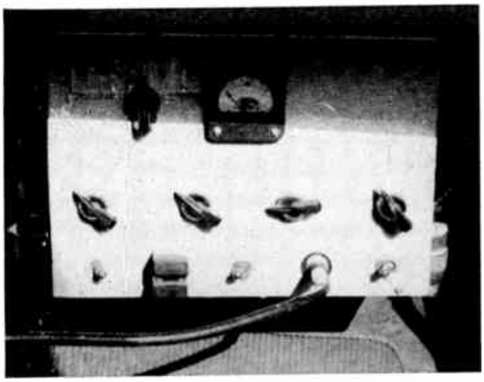


Fig. 20-17 — The bandswitching mobile transmitter installed under the dashboard of W2RPU's car.

The unit shown in Figs. 20-17 through 20-19 is a complete bandswitching mobile transmitter, including modulator and covering all bands from 4 to 29 Mc.

The circuit diagram is shown in Fig. 20-19. Either crystal control or VFO is available simply by snapping the toggle, S_1. A 6C4 is used in the VFO and this is the only indirectly-heated tube in the transmitter. All others are direct-heater types. The heater of the 6C4 operates from a separate circuit through S_2 so that it can be left on during receiving periods. This cuts down initial drift and eliminates waiting for the cathode to come up to temperature before each transmission. VFO output is taken from the cathode tap to minimize loading effects on frequency. The tuning range of the VFO is limited to 3500 to 4000 kc. This makes it necessary to use crystal control on 11 meters, unless it is desired to extend the VFO range. The plate voltage for the VFO is stabilized by an OB2 regulator tube.

The 5618 following the VFO may be used as an 80- or 40-meter crystal oscillator, or as an amplifier or doubler for the VFO, since the output circuit, C_9L_2, will tune to either band, one near maximum capacitance and the other near minimum.

The next stage, also using a 5618, may be operated as a doubler to 14 Mc., as a tripler to 21 Mc., or a quadrupler to 28 Mc., depending on the setting of C_{13} which covers all three bands. This stage is inserted or removed from the circuit by S_3. Thirty volts of fixed bias from the modulator-biasing battery practically cuts off plate current to the 5618 when this stage is not in use.

A 5516 is used in the final amplifier. This tube has the same power rating as the 2E25, but it is shorter physically so that it can be fitted into a smaller space. The use of an all-band tuner in the final-amplifier output circuit eliminates the necessity for plug-in coils or switching.

In the audio section, a carbon microphone drives a triode-connected 5618 which, in turn, drives two 2E30s in the Class AB_2 modulator. Microphone voltage is obtained from the car battery through the filter consisting of C_{20} and L_9.

The milliammeter, MA_1, can be switched to read current at the important points in the circuit. When switched to position E, it can be used to check plate voltage for the rig's final amplifier stage.

In the front-view photograph of Fig. 20-17, the control knobs across the panel are, from left to right, for VFO, first 5618, second 5618, and final amplifier. The meter switch is to the left of the meter. Along the bottom are the VFO-crystal switch, a dual crystal socket (one socket unwired for a spare crystal), the frequency-multiplier switch, S_3, microphone-control jack and the VFO heater switch.

In the rear-view photograph of Fig. 20-18, the four tuning condensers are lined up across the panel, just above the chassis level. C_{19} is a dual midget Hammarlund, originally of 140 μμfd. per section. To obtain the desired range, one rotor and two stator plates were removed from each section. The high-frequency coil, L_4, is mounted vertically at the rear of the condenser, while L_5 is placed at right angles alongside the condenser to minimize coupling between the two. Care should be taken to make sure, with a grid-dip meter, that the circuit when completed does not tune simultaneously to fundamental and harmonic frequencies. This can be controlled by altering the coils somewhat.

L_3 is mounted vertically behind the meter. L_2, at right angles, is fastened to C_9. L_1 is vertical behind C_1. The r.f. tubes are lined up across the center of the chassis. The 6C4 is hidden by the biasing battery to the right. The two 5618s are to the right of the 5516 final-amplifier tube. A baffle shield is placed between the tube and L_3 to the right. The audio components and the OB2

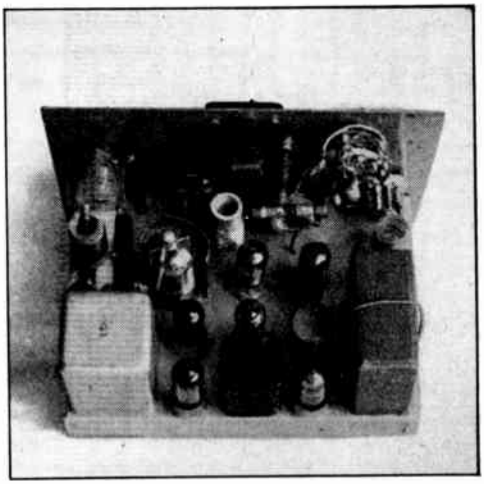


Fig. 20-18 — Rear interior view of W2RPU's mobile transmitter, showing the arrangement of components on the chassis.

occupy the rear portion of the chassis. All small components are mounted underneath. The chassis measures 8¼ inches long, 5⅞ inches from front to back and 1 inch deep.

Although this transmitter may be operated from a suitable dynamotor, there is an advantage in the use of two supplies. While the rest of the transmitter may be operated at 300 volts, a voltage of 250 is the maximum rated value for the 2E30 modulators. A separate supply for the Class AB_2 modulator with its varying plate current also improves the voltage regulation for the rest of the transmitter. Two 100-ma. vibrator-type power supplies, one delivering 300 volts and the other 250 volts, are recommended.

The two exciter tank circuits, C_9L_2 and $C_{13}L_3$, can be resonated to the desired bands by observing grid current to the following stage. A grid current of 2 to 3 ma. should be adequate for the multiplier stage and 3 to 5 ma. for the final.

The antenna should be of the center-loaded type. The RG-8/U coaxial cable feeding the base of the antenna is tapped on L_5 at a compromise point that serves for all bands. Some slight improvement can be gained by adjusting the tap for the band considered most important.

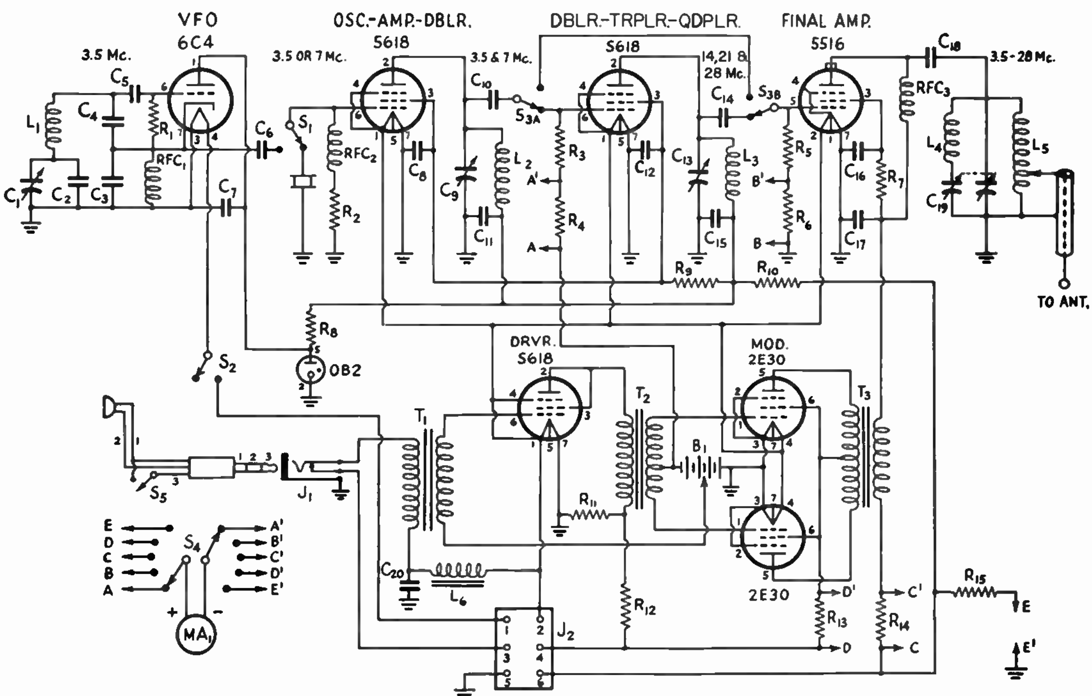


Fig. 20-19 — **Circuit diagram of the multiband mobile transmitter.**

C_1 — 50-μμfd. variable (National PSE-50).
C_2 — 100-μμfd. silvered mica.
C_3, C_4 — 0.001-μfd. silvered mica.
C_5, C_6 — 100-μμfd. mica.
C_7 — 0.01-μfd. mica.
C_8, C_{11}, C_{12}, C_{15} — 0.001-μfd. mica.
C_9, C_{13} — 100-μμfd. variable (National PSE-100 with ¾-inch shaft).
C_{10}, C_{14} — 47-μμfd. ceramic.
C_{16}, C_{17} — 0.001-μfd. 1000-volt mica.
C_{18} — 0.01-μfd. 1000-volt mica.
C_{19} — 110-μμfd.-per-section variable (Hammarlund HFD-140; see text).
C_{20} — 25-μfd. 25-volt electrolytic.
R_1, R_2 — 0.1 megohm, ½ watt.
R_3 — 56,000 ohms, ½ watt.
R_4, R_6 — 100 ohms, ½ watt.
R_5 — 27,000 ohms, 1 watt.
R_7 — 2500 ohms, 5 watts.
R_8 — 10,000 ohms, 2 watts.
R_9 — 27,000 ohms, 2 watts.
R_{10} — 2000 ohms, 2 watts.
R_{11} — 56,000 ohms, 2 watts.
R_{12} — 5000 ohms, 2 watts.
R_{13}, R_{14} — Meter shunts made of resistance wire to provide for full-scale meter reading of 100 ma.
R_{15} — 0.15 megohm, 1 watt (depends on meter used).
L_1 — 48 turns No. 26 enam., 1-inch diam., 1¼ inches long (may have to be slightly modified to provide proper bandspread).
L_2 — 28 turns No. 24 enam., 1-inch diam., ⅞ inch long.
L_3 — 9 turns No. 20 enam., ¾-inch diam., ⅞ inch long.
L_4 — 16 turns No. 20 enam., ¾-inch diam., ⅞ inch long.
L_5 — 19 turns No. 20 enam., 1¼-inch diam., 1¼ inches long, tapped 1½ turns.
L_6 — 10-hy. 30-ma. choke (filter).
B_1 — 30-volt battery with tap at 7½ volts.
J_1 — 3-contact open-circuit microphone jack (midget).
J_2 — 6-contact connector (Jones P-306-AB).
MA_1 — Milliammeter, 10-ma. scale.
RFC_1, RFC_2 — 2.5-mh. r.f. choke (National R-50).
RFC_3 — 2.5-mh. r.f. choke (National R-100U).
S_1 — S.p.d.t. toggle switch.
S_2 — S.p.d.t. toggle switch.
S_3 — D.p.d.t. toggle switch.
S_4 — 2-pole 5-position rotary switch.
S_5 — Push-to-talk switch.
T_1 — Midget output transformer: single plate to 200 ohms (mic. connected to 200 ohms).
T_2 — Single plate to p.p. grids for Class AB_2.
T_3 — Modulation transformer, Class AB_2.

NOTE: J_2 connections as follows: (1) VFO heater, (2) other heaters, (3) push-to-talk control to power supplies, (4) +h.v. audio, (5) ground, (6) +h.v. r.f.

A Band-Changing Mobile Transmitter for 50 and 144 Mc.

Figs 20-20 through 20-25 show circuits and constructional details of a compact transmitter covering the 6- and 2-meter bands. Band-changing is done entirely by the panel controls. The unit is only 3 inches deep, and therefore is suitable for instrument-panel mounting.

Output on either band may be obtained using crystals in the 8-, 12-, or 25-Mc. ranges. Although it is possible to operate the 2E26 output stage at higher voltage, the unit is designed primarily to work from a 300-volt 100-ma. supply. A single 200-ma. supply should take care of both this unit and a modulator in the latter case. Changing from one band to the other is accomplished through the use of wide-range tanks in the exciter, and a multicircuit tuner in the output. Metering circuits are included.

Circuit

The circuit of the unit is shown in Fig. 20-22. Type 5763s are used in the Tri-tet oscillator and the driver stage. The oscillator has a fixed cathode circuit resonant at approximately 15 Mc. C_5 has sufficient range to tune the oscillator output circuit from 24 through 36 Mc. This circuit is tuned to 25 Mc. for 50-Mc. output from the transmitter, and may be tuned to either 24 or 36 Mc. for final output at 144 Mc.

The multiplier output circuit, $C_{12}L_3$, covers the range of 48 to 72 Mc., and operates as a doubler to 50 Mc., or as either a doubler or tripler (depending on the oscillator output frequency) to 72 Mc. for final output at 144 Mc. The multiplier is capacity-coupled to the 2E26 amplifier grid. This stage operates straight through at 50 Mc., and as a doubler to 144 Mc. A combination of fixed bias and grid leak is used. The value of fixed bias is not critical — 22 to 45 volts. The 22K screen resistor gives proper screen voltage over a supply-voltage range of 300 to 400 volts.

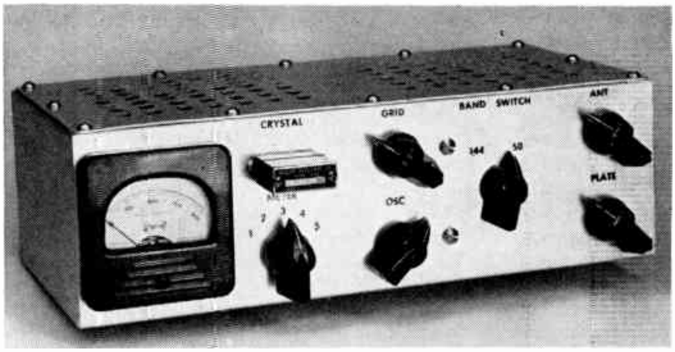


Fig. 20-20 — The crystal is mounted above the meter switch, to the left of the amplifier grid-tuning control. The tuning knob for the oscillator is at the lower left-hand side of the output switch, S_1. Controls for the output and amplifier plate circuits are at the right. The unit may be used vertically by orientating the meter. Ventilating holes should be drilled in the end used as the top.

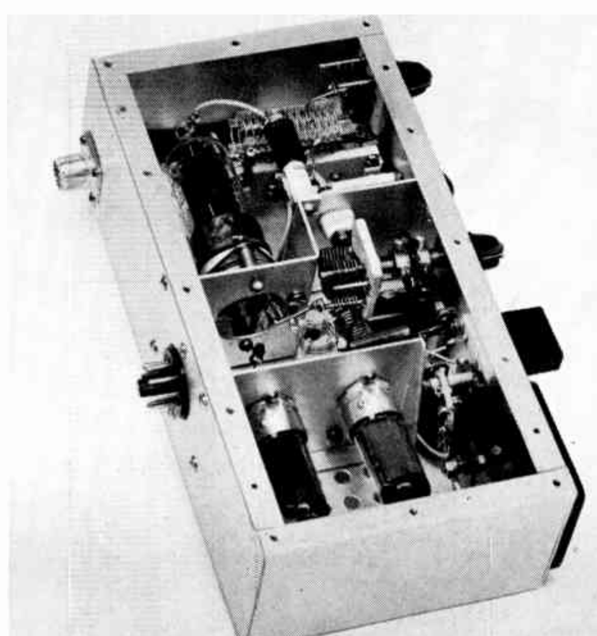


Fig. 20-21 — In this view the perforated top cover has been removed to show the completed transmitter. The input and output connectors are on the rear chassis wall and the 5763 subassembly is in the foreground, to the left of the meter switch. The Z-shaped partition supports C_{12}, RFC_4 and the 2E26. C_{12} is mounted on a feed-through bushing. The oscillator tuning capacitor, C_5, is panel-mounted directly below C_{12}. The output switch, S_1 is partially hidden by the Z-shaped plate. The multicircuit tuner is at the upper end of the chassis, just below the link tuning condenser, C_{18}.

The plate tuner for the amplifier consists of a capacitor, C_{17}, and inductors L_4 and L_5. Output from the amplifier is transferred to J_1 by a series-tuned circuit consisting of C_{18}, L_6 and S_1. L_6 is electrically subdivided by a tap which connects to C_{18}. That portion of L_6 above the tap provides output coupling at 50 Mc., and the lower section of the coil couples to L_5 when S_1 is set for 144-Mc. operation.

The metering circuit uses S_2, a 200-ma. d.c. milliammeter, and resistors R_4, R_8, R_{10}, R_{12} and R_{13}. R_{13} is connected to Terminals E and E_1 of the switch and, in turn, to Pins 7 and 8 of the power-input connector, J_2. The latter set of connections allows the plate current of an external modulator to be checked by the meter.

Provision for connecting either a single or a pair of supplies to the transmitter is made at J_2. If a single 300-volt pack is used for the entire unit, it is necessary to connect a jumper between Pins 3 and 5 of J_2. With separate supplies for exciter and final, connect the 300-volt supply to Pin 3 and the amplifier supply to Pin 5. When a modulator is connected to the transmitter, connect the secondary of the modulation transformer between Pins 5 and 8 of J_2, connect +h.v. to the 2E26 to Pin 8, and then return the +h.v. lead of the modulation-transformer primary to Pin 7.

Construction

A 3 × 5 × 10-inch aluminum chassis is used as the housing for the transmitter. The construction is made easier through the use of subassemblies. Fig. 20-24, along with the sketch of Fig. 20-23, identifies the components for the oscillator-multiplier section. The bracket supporting the components has ⅜-inch lips along the right and bottom edges for fastening to the chassis. The wire leader that later connects to C_5 should be about 3 inches long, while the five leads that will be joined to J_2 and S_2 can be about 5 inches long.

Fig. 20-21 shows a Z-shaped partition spanning the chassis. This can be made and installed most easily in two pieces overlapping and fastened together at the center. The height is made to fit the chassis depth. In Fig. 20-21, the segment lengths, from left to right, are 2½, 1⅛, and 2½ inches. Lips are bent at the ends and along the bottom for fastening to the chassis. A 1¼-inch hole is punched in the center of the segment on which the 2E26 is mounted, while a small feed-through bushing (Millen 32100) is set in the other segment. Position this bushing so that C_{12}, which is mounted on it, will be at the right level, and clear of the partition segment to the rear. The 2E26 socket is mounted on ⅝-inch spacers. Prongs 1, 2, 4, 6 and 8, and the screen by-pass, C_9, should be returned directly to ground on the *socket* side of the partition. A 2-terminal tie point to the rear of the socket supports the heater lead and the h.v. end of the screen resistor, R_{11}.

Mount the meter-shunt resistors across the terminals of S_2. Join Contacts A_1 and B_1, and connect 8-inch leads to the rotor-arm contacts and to Stationary Contacts C_1, D_1, E and E_1. A

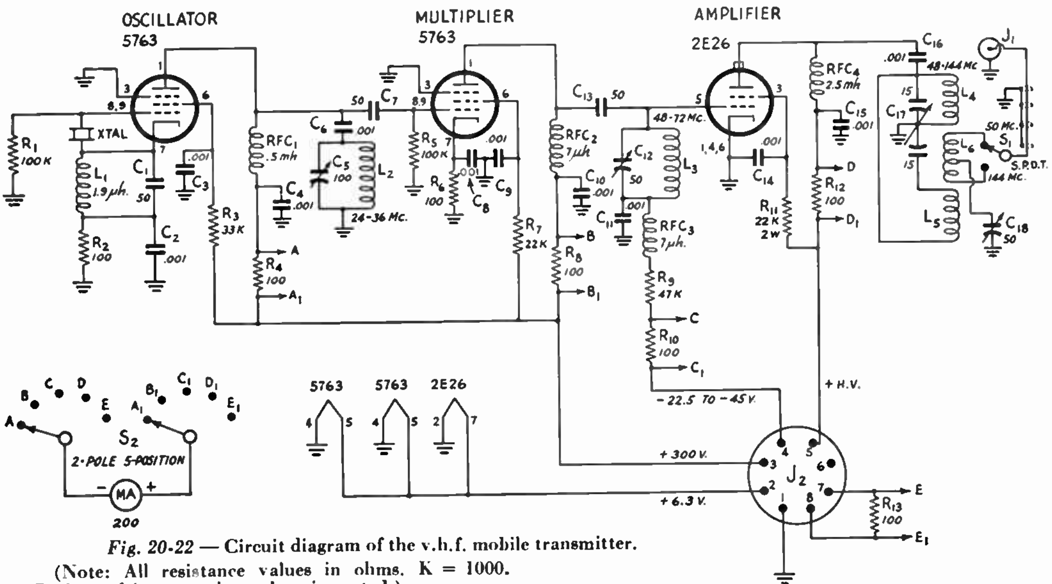


Fig. 20-22 — **Circuit diagram of the v.h.f. mobile transmitter.**

(Note: All resistance values in ohms. K = 1000. Resistors ½ watt, unless otherwise noted.)

C_5 — 100-μμf. variable (Hammarlund HF-100).
C_{12}, C_{18} — 50-μμf. variable (Hammarlund HF-50).
C_{17} — 15-μμf.-per-section variable (Hammarlund HFD-15-X).
L_1 — 1.9 μh., 34 turns No. 22 enam., ¼-inch diam., close-wound.
L_2 — 0.44 μh., 6 turns No. 20 tinned, ½-inch diam., ⅜ inch long (B & W 3003).
L_3 — 0.155 μh., 3 turns No. 18 tinned, ½-inch diam., ⅜ inch long (B & W 3002).
L_4 — 0.36 μh. (see text).
L_5 — 0.2 μh. (see text).
L_6 — See text.
J_1 — Amphenol coaxial connector.
J_2 — 8-prong male connector.
RFC_1 — National type R-50 r.f. choke.
RFC_2, RFC_3 — Ohmite type Z-50 r.f. choke.
RFC_4 — National type R-100S r.f. choke.
S_1, S_2 — 2-pole 6-position miniature selector switch, S_1 used as s.p.d.t. (Centralab PA-2003).

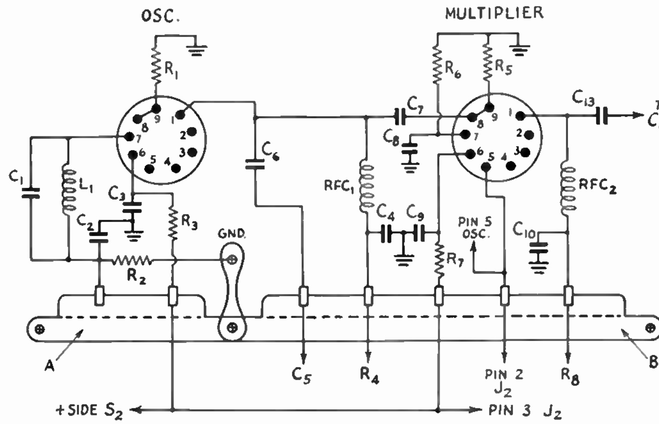


Fig. 20-23 — Drawing of the parts layout for the exciter subassembly. *A* and *B* are 2- and 5-terminal tie-point strips.

lead about 1 foot long should be soldered to Contact D.

In constructing the multicircuit tuner, first reduce the 3006 B & W Miniductor to a total of 14¼ turns. Without breaking the supporting bars, clip the winding at points that will leave 5 full turns at one end and 3¼ turns at the opposite end. The 6 turns left intact between end windings are used as the output coupling inductance, L_6. Short leads of No. 16 wire should now be soldered to the free ends of the three windings. Also, solder a short lead 1¼ turns in from the 144-Mc. end of the coupling coil. This should place the tap at the top of the coil when it is mounted.

To assemble the tuner, turn C_{17} with the insulated support bar facing toward the partition. Place the coil about ⅜ inch above the condenser, and bend the four leads from L_4 and L_5 into place. The *outside* ends of these sections go directly to the *rear* stator terminal of the condenser, while the inside lead of L_5 goes to the front stator terminal. The inside end of L_4 is grounded to the frame at the rear.

In mounting parts on the chassis, center J_2 on the rear wall 4¼ inches from the exciter end of the chassis, and J_1 in the lower corner of the amplifier end. On the panel side, the shafts for C_{17} and C_{18} are 1 inch from the right end. S_1 is centered 2⅞ inches from the right end, while the controls for C_5 and C_{12} are 4¾ inches in. A panel bearing is needed for C_{12}, which is fitted with an insulating shaft coupling. The remaining two controls are 6⅝ inches from the right-hand end. The meter is at the left-hand end.

The subassemblies may now be positioned while the mounting holes are marked. The bracket for the 5763s is placed 3¼ inches from the left-hand end of the chassis, while the rear end of the Z-shaped partition comes at 5⅛ inches from the same end.

Before fastening the subassemblies in place, proceed with the wiring. Connect S_1 to L_1 and J_1; solder the tap on L_6 to C_{18}; mount L_2 on the terminals of C_5; connect the rotor arms of S_2 to the meter.

Mount the exciter assembly and attach the proper loose leads to C_5, J_2 and S_2. Mount a tie point at the right-hand mounting screw of the crystal socket, and fasten R_9 between the tie point and Contact C of S_2. Run leads to the crystal

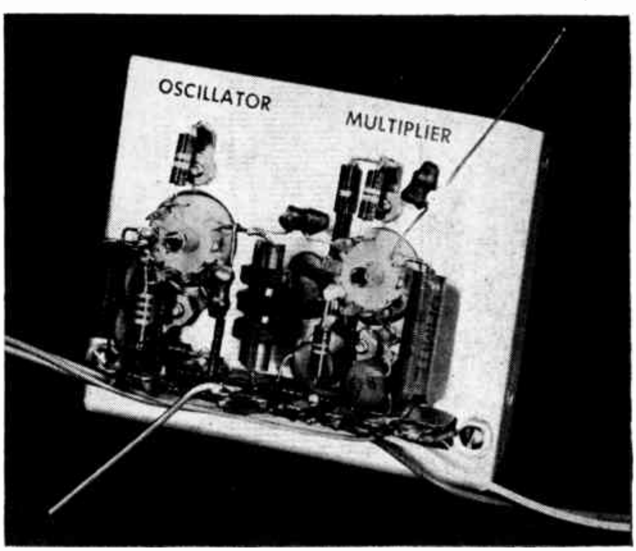


Fig. 20-24 — This subassembly measures 2 13/16 by 3½ inches and supports most of the components for the exciter stages. C_{13}, with one end floating free, is at the upper right-hand corner. The wire leaders at the bottom of the plate connect to the oscillator tank, meter switch and power connector, as shown by Fig. 20-22.

Voltage and Current Chart for the V.H.F. Mobile Transmitter

Oscillator				*Multiplier*				*Amplifier*				
Crystal Freq., Mc.	E_s	I_p, *Ma.*	*Plate Freq., Mc.*	E_g	E_s	I_p, *Ma.*	*Plate Freq., Mc.*	E_g	I_g, *Ma.*	E_s	I_p, *Ma.*	*Plate Freq., Mc.*
8.3	210	20	25	−80	240	25	50	−190	4	135	45	50
12.5	235	15	"	−120	245	27	"	−210	4.5	120	"	"
25.0	210	20	"	−60	240	25	"	−185	4	145	"	"
8.0	240	20	24	−85	250	25	72	−155	3.2	170	50	144
12.0	220	16	24	−140	255	27	"	−190	4	155	47	"
"	225	18	36	−115	245	"	"	−215	4.5	150	"	"
24.0	210	21	24	−65	250	"	"	−140	3	180	50	"

socket and then mount the Z-shaped partition in place.

Testing

For 50-Mc. operation, the crystal frequency must lie within one of the following ranges: 8.333 to 9.0 Mc.; 12.5 to 13.5 Mc.; 25.0 to 27.0 Mc. With a small B battery for fixed bias and a 300-volt supply connected to the exciter, but not the amplifier, tuning of the exciter at 50 Mc. requires only that C_5 and C_{12} be resonated at 25 and 50 Mc. respectively. The chart shows the approximate operating conditions for the 5763s.

Before testing the amplifier, turn the supply off and connect a jumper between Pins 3 and 5 of J_2, and connect a 115-volt 10-watt lamp to the output connector. S_1 should be set at the 50-Mc. position. Apply power and resonate C_{17}, indicated by a dip in plate current. This should come well toward minimum capacitance. Set C_{18} near full capacitance and retune C_{17} for resonance. (The amplifier data in the chart were taken with the dummy load. In operation, the currents will depend upon loading.) If biasing voltages are checked, use a v.t.v.m., or a general-purpose test instrument with a radio-frequency choke inductance of at least 1 mh. connected in series.

In tuning up for 144-Mc. output, work with the exciter stages only at first, using a crystal in any one of the following frequency ranges: 8.0 to 8.222 Mc.; 12.0 to 12.333 Mc.; 24 to 24.666 Mc. If a 12-Mc. crystal is selected, the oscillator may be tuned to either 24 or 36 Mc. In either case, the multiplier must be tuned to 72 Mc. by C_{12}. The oscillator is always tuned to 24 Mc. with crystals in the 8- and 24-Mc. ranges.

In checking amplifier operation at 144 Mc., S_1 must be in the 144-Mc. position. The plate current will show a relatively small dip at resonance on this band. For resonance, condensers C_{17} and C_{18} will be set well toward minimum capacitance.

Antenna

The tuned-link output circuit is designed for use with low-impedance antenna systems, so quarter-wave whips are recommended. A logical system for mobile work would make use of a two-section 50-Mc. whip that can be reduced to 144-Mc. dimensions by removing a top section.

Fig. 20-25 shows the circuit of an appropriate modulator.

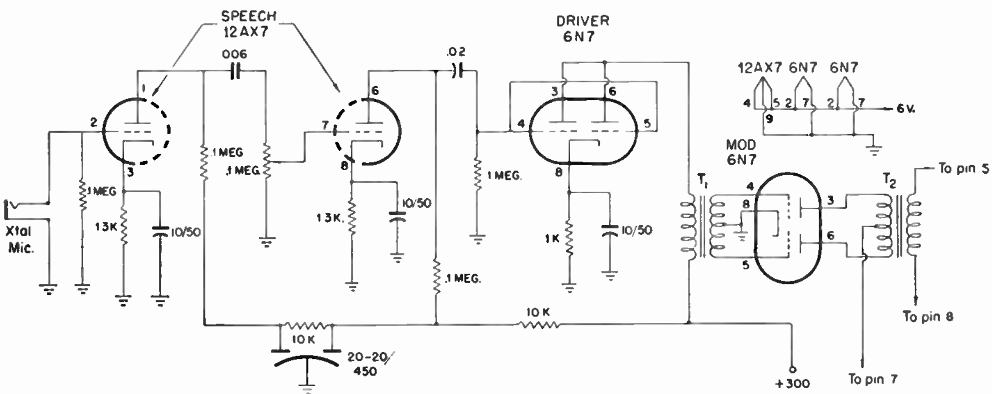


Fig. 20-25 — Circuit of a modulator for the 50- and 144-Mc. mobile transmitter. Pin numbers on modulation transformer leads refer to J_2 in Fig. 20-22.

T_1 — Driver transformer; parallel 6N7 to Class B 6N7 grids (Stancor A-4702).
T_2 — Class B modulation transformer (Stancor A-3845; 5000-ohm tap).

Mobile Power Supply

By far the majority of amateur mobile installations depend upon the car storage battery as the source of power. The tube types used in equipment are chosen so that the filaments or heaters may be operated directly from the battery. High voltage may be obtained from a supply of the vibrator-transformer-rectifier type or from a small motor-generator operating from the battery.

Filaments

Because tubes with directly-heated cathodes (filament-type tubes) have the advantage that they can be turned off during receiving periods and thereby reduce the average load on the battery, they are preferred by some for transmitter applications. However, the choice of types with direct heating is limited, especially among those for 6-volt operation, and the saving may not always be as great as anticipated, because directly-heated tubes may require greater filament power than those of equivalent rating with indirectly-heated cathodes. In most cases, the power required for transmitter filaments will be quite small compared to the total power consumed.

Plate Power

Under steady running conditions, the vibrator-transformer-rectifier system and the motor-generator-type plate supply operate with approximately the same efficiency. However, for the same power, the motor-generator's over-all efficiency may be somewhat lower because it draws a heavier starting current. On the other hand, the output of the generator requires less filtering and sometimes trouble is experienced in eliminating interference from the vibrator.

Converter units, both in the vibrator and rotating types, are also available. These operate at 6 or 12 volts d.c. and deliver 115 volts a.c. This permits operating standard a.c.-powered equipment in the car. Although these systems have the advantage of flexibility, they are less efficient than the previously-mentioned systems because of the additional losses introduced by the transformers used in the equipment.

Mobile Power Considerations

Since the car storage battery is a low-voltage source, this means that the current drawn from the battery for even a moderate amount of power will be large. Therefore, it is important that the resistance of the 6-volt circuit be held to a minimum by the use of heavy conductors, no longer than necessary, and good solid connections. A heavy-duty relay should be used in the line between the battery and the plate-power unit. An ordinary toggle switch, located in any convenient position, may then be used for the power control. A second relay may sometimes be advisable for switching the filaments. If the power unit must be located at some distance from the battery (in the trunk, for instance) the 6-volt cable should be of the heavy military type.

A complete mobile installation may draw 30 to 40 amperes or more from the 6-volt battery. This requires a considerably increased demand from the car's battery-charging generator. The voltage-regulator systems on cars of recent years will take care of a moderate increase in demand if the car is driven fair distances regularly at a speed great enough to insure maximum charging rate. However, if much of the driving is in urban areas at slow speed, or at night, it may be necessary to modify the charging system. Special communications-type generators, such as those used in police-car installations, are designed to charge at a high rate at slow engine speeds. The charging rate of the standard system can be increased within limits by tightening up slightly on the voltage-regulator and current-regulator springs. This should be done with caution, however, checking for excessive generator temperature or abnormal sparking at the commutator. The average car generator has a rating of 35 amperes, but it may be possible to adjust the regulator so that the generator will at least hold even with the transmitter, receiver, lights, heater, etc., all operating at the same time.

Another scheme that has been used to increase generator output at slow driving speeds is to decrease slightly the diameter of the generator pulley. This means, of course, that the generator will be running above normal at high driving speeds. Some generators will not stand the higher speed without damage.

If higher transmitter power is used, it may be necessary to install an a.c. charging system. In this system, the generator delivers a.c. and works into a rectifier. A charging rate of 75 amperes is easily obtained. Commutator trouble often experienced with d.c. generators at high current is avoided, but the cost of such a system is rather high.

Some mobile operators prefer to use a separate battery for the radio equipment. Such a system can be arranged with a switch that cuts the auxiliary battery in parallel with the car battery for charging at times when the car battery is lightly loaded. The auxiliary battery can also be charged at home when not in use.

A tip: many mobile operators make a habit of carrying a pair of heavy cables five or six feet long, fitted with clips to make a connection to the battery of another car in case the operator's battery has been allowed to run too far down for starting.

(See power-supply chapter, vibrator supplies.)

The Mobile Antenna

For mobile operation in the range between 1.8 and 30 Mc., the vertical whip antenna is almost universally used. Since longer whips present mechanical difficulties, the length is usually limited to a dimension that will resonate as a quarter-wave antenna in the 10-meter band. The car body serves as the ground connection. This antenna length is approximately 8 feet.

Fig. 20-26 — The quarter-wave whip at resonance will show a pure resistance at the feed point X.

With the whip length adjusted to resonance in the 10-meter band, the impedance at the feed point, X, Fig. 20-26, will appear as a pure resistance at the resonant frequency. This resistance will be composed almost entirely of radiation resistance (see index), and the efficiency will be high. However, at frequencies lower than the resonant frequency, the antenna will show an increasingly large capacitive reactance and a decreasingly small radiation resistance.

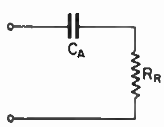


Fig. 20-27 — At frequencies below the resonant frequency, the whip antenna will show capacitive reactance as well as resistance. R_R is the radiation resistance, and C_A represents the capacitive reactance.

The equivalent circuit is shown in Fig. 20–27. For the average 8-ft. whip, the reactance of the capacitance, C_A, may range from about 150 ohms at 21 Mc. to as high as 8000 ohms at 1.8 Mc., while the radiation resistance, R_R, varies from about 15 ohms at 21 Mc. to as low as 0.1 ohm at 1.8 Mc. Since the resistance is low, considerable current must flow in the circuit if any appreciable power is to be dissipated as radiation in the resistance. Yet it is apparent that little current can be made to flow in the circuit so long as the comparatively high series reactance remains.

Fig. 20-28 — The capacitive reactance at frequencies lower than the resonant frequency of the whip can be canceled out by adding an equivalent inductive reactance in the form of a loading coil in series with the antenna.

Eliminating Reactance

The capacitive reactance can be canceled out by connecting an equivalent inductive reactance, L_L, in series, as shown in Fig. 20-28, thus tuning the system to resonance.

Unfortunately, all coils have resistance, and this resistance will be added in series, as indicated at R_C in Fig. 20-29. While a large coil may radiate some energy, thus adding to the radiation resistance, the latter will usually be negligible compared to the loss resistance introduced. However, adding the coil makes it possible to feed power to the circuit.

Ground Loss

Another element in the circuit dissipating power is the ground-loss resistance. Fundamentally, this is related to the nature of the soil in the area under the antenna. Little information

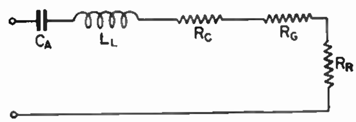


Fig. 20-29 — Equivalent circuit of a loaded whip antenna. C_A represents the capacitive reactance of the antenna, L_L an equivalent inductive reactance. R_C is the loading-coil resistance, R_G the ground-loss resistance, and R_R the radiation resistance.

is available on the values of resistance to be expected in practice, but some measurements have shown that it may amount to as much as 10 or 12 ohms at 4 Mc. At the lower frequencies, it may constitute the major resistance in the circuit.

Fig. 20-29 shows the circuit including all of the elements mentioned above. Assuming C_A lossless and the loss resistance of the coil to be represented by R_C, it is seen that the power output of the transmitter is divided among three resistances — R_C, the coil resistance; R_G, the ground-loss resistance; and R_R, the radiation resistance. Only the power dissipated in R_R is radiated. The power

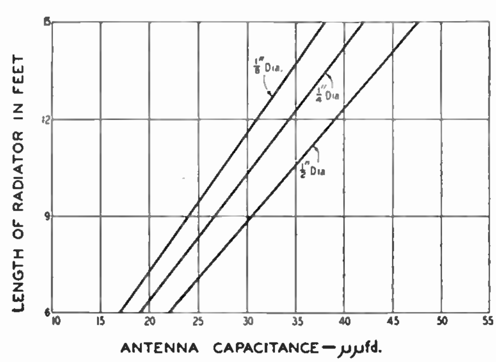


Fig. 20-30 — Graph showing the approximate capacitance of short vertical antennas for various diameters and lengths. These values should be approximately halved for a center-loaded antenna.

Approximate Values for 8-ft. Mobile Whip

f_{kc}	Loading $L_{\mu h}$	R_C (Q50) Ohms	R_C (Q300) Ohms	R_R Ohms	Feed R* Ohms	Matching $L_{\mu h}$*
Base Loading						
1800	345	77	13	0.1	23	3
3800	77	37	6.1	0.35	16	1.2
7200	20	18	3	1.35	15	0.6
14,200	4.5	7.7	1.3	5.7	12	0.28
21,250	1.25	3.4	0.5	14.8	16	0.28
29,000	....	....	....	....	36	0.23
Center Loading						
1800	700	158	23	0.2	34	3.7
3800	150	72	12	0.8	22	1.4
7200	40	36	6	3	19	0.7
14,200	8.6	15	2.5	11	19	0.35
21,250	2.5	6.6	1.1	27	29	0.29

R_C = Loading-coil resistance; R_R = Radiation resistance.

* Assuming loading coil Q = 300, and including estimated ground-loss resistance.

Suggested coil dimensions for the required loading inductances are shown in a following table.

developed in R_C and R_G is dissipated in heat. Therefore, it is important that the latter two resistances be minimized.

MINIMIZING LOSSES

There is little that can be done about the nature of the soil. However, poor electrical contact between large surfaces of the car body, and especially between the point where the feed line is grounded and the rest of the body, can add materially to the ground-loss resistance. For example, the feed line, which should be grounded as close to the base of the antenna as possible, may be connected to the bumper, while the bumper may have poor contact with the rest of the body because of rust or paint.

Loading Coils

The accompanying table shows the approximate loading-coil inductance required for the various bands. The graph of Fig. 20-30 shows the approximate capacitance of whip antennas of various average diameters and lengths. For 1.8, 4 and 7 Mc., the loading-coil inductance required (when the loading coil is at the base) will be approximately the inductance required to resonate in the desired band with the whip capacitance taken from the graph. For 14 and 21 Mc., this rough calculation will give more than the required inductance, but it will serve as a starting point for final experimental adjustment that must always be made.

Also shown in the table are approximate values of radiation resistance to be expected with an 8-ft. whip, and the resistances of loading coils — one group having a Q of 50, the other a Q of 300. A comparison of radiation and coil resistances will show the importance of reducing the coil resistance to a minimum, especially on the three lower-frequency bands.

To minimize loading-coil loss, the coil should have a high ratio of reactance to resistance, i.e., high Q. A 4-Mc. loading coil wound with small wire on a small-diameter solid form of poor quality, and enclosed in a metal protector, may have a Q as low as 50, with a resistance of 50 ohms or more. High-Q coils require a large conductor, "air-wound" construction, turns spaced, the best insulating material available, a diameter not less than half the length of the coil (not always mechanically feasible), and a minimum of metal in the field. Such a coil for 4 Mc. may show a Q of 300 or more, with a resistance of 12 ohms or less. This reduction in loading-coil resistance may be equivalent to increasing the transmitter power by 3 times or more. Most low-loss transmitter plug-in coils of the 100-watt size or larger, commercially produced, show a Q of this order. Where larger inductance values are required, lengths of low-loss space-wound coils are available (B & W).

Suggested Loading-Coil Dimensions

Req'd $L_{\mu h}$	Turns	Wire Size	Diam. In.	Length In.	Form or B & W Type
700	190	22	3	10	Polystyrene
345	135	18	3	10	Polystyrene
150	100	16	2½	10	Polystyrene
77	75	14	2½	10	Polystyrene
77	29	12	5	4¼	160T
40	28	16	2½	2	80B less 7 t.
40	34	12	2½	4¼	80T
20	17	16	2½	1¼	80B less 18 t.
20	22	12	2½	2¾	80T less 12 t.
8.6	16	14	2	2	40B less 4 t.
8.6	15	12	2½	3	40T less 5 t.
4.5	10	14	2	1¼	40B less 10 t.
4.5	12	12	2½	4	40T
2.5	8	12	2	2	15B
2.5	8	6	2⅜	4½	15T
1.25	6	12	1¾	2	10B
1.25	6	6	2⅜	4½	10T

Center Loading

The radiation resistance of the whip can be approximately doubled by placing the loading coil at the center of the whip, rather than at the base, as shown in Fig. 20-31. (The optimum position varies with ground resistance. The center is optimum for average ground resistance.) However, the inductance of the loading coil must be approximately doubled over the value required at the base to tune the system to resonance. For a coil of the same Q, the coil resistance will also be doubled. But, even if this is the case, center loading represents a gain in antenna efficiency, especially at the lower frequencies. This is because the ground-loss resistance remains the same, and the increased radiation resistance becomes a larger portion of the total circuit resistance, even though the coil resistance also increases. However, as turns are added to a loading coil (other factors being equal) the inductance (and therefore the reactance) increases at a greater rate than the resistance, and the larger coil will usually have a higher Q.

Fig. 20-31 — **Placing the loading coil at the center of the whip antenna, instead of at the base, increases the radiation resistance, although a larger coil must be used.**

Fig. 20-32 — **The top-loaded 4-Mc. antenna used by W6SCX. The loading coil is a B & W transmitting coil. The coil can be tuned by the variable link which is connected in series with the two halves of the coil.**

Top Loading Capacitance

Since the coil resistance varies with the inductance of the loading coil, the coil resistance can be reduced by reducing the number of turns. This can be done, while still maintaining resonance, by adding capacitance to the portion of the antenna above the coil. This capacitance can be provided by attaching a capacitive surface

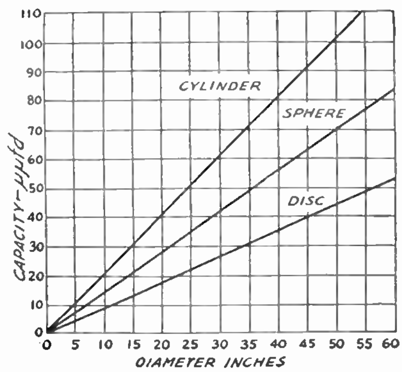


Fig. 20-33 — **Capacitances of spheres, disks and cylinders in free space. These values are approximately those to be expected when used with top-loaded whip antennas. The cylinder length is assumed to be equal to its diameter.**

as high up on the antenna as is mechanically feasible. Capacitive "hats," as they are usually called, may consist of a light-weight metal ball, cylinder, disk, or wheel structure as shown in Fig. 20-32. Fig. 20-33 shows the approximate added capacitance to be expected from top-loading devices of various forms and dimensions. This should be added to the capacitance of the whip above the loading coil (from Fig. 20-30) in determining the approximate inductance of the loading coil.

When center loading is used, the amount of capacitance to be added to permit the use of the same loading inductance required for base loading is not great, and should be seriously considered, since the total gain made by moving the coil to the center of the antenna may be quite marked.

Tuning the Band

Especially at the lower frequencies, where the resistance in the circuit is low compared to the coil reactance, the antenna will represent a very high-Q circuit, making it necessary to retune for relatively small changes in frequency. While many methods have been devised for tuning the whip over a band, one of the simplest and most efficient is shown in the sketches of Figs. 20-34 and 20-35, and the photograph of Fig. 20-37. In this case, a standard B & W plug-in coil is used as the loading coil. A length of large-diameter

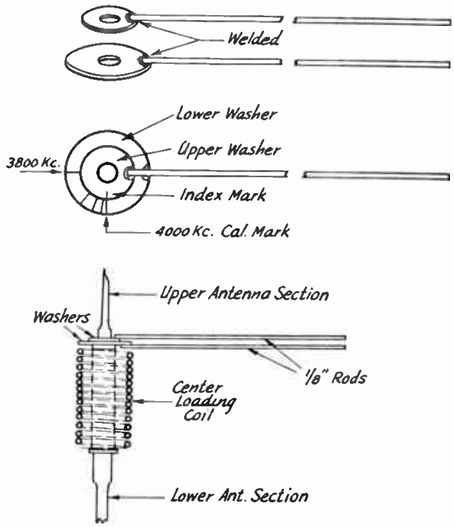


Fig. 20-34 — Details of rod construction. Dimensions can be varied to suit the whip diameter and the builder's convenience. Adjustment of rod lengths is described in the text.

polystyrene rod is drilled and tapped to fit between the upper and lower sections of the antenna. The assembly also serves to clamp a pair of metal brackets on each side of the polystyrene block that serve both as support and connections to the loading-coil jack bar.

A 1/8-inch steel rod, about 15 inches long, is brazed to each of two large-diameter washers with holes to pass the threaded end of the upper section. The rods form a loading capacitance that varies as the upper rod is swung away from the lower one, the latter being stationary. Enough variation in tuning can be obtained to cover the 80-meter band. Fig. 20-34 shows the top washer slightly smaller to facilitate marking a frequency scale on the stationary washer, after the upper

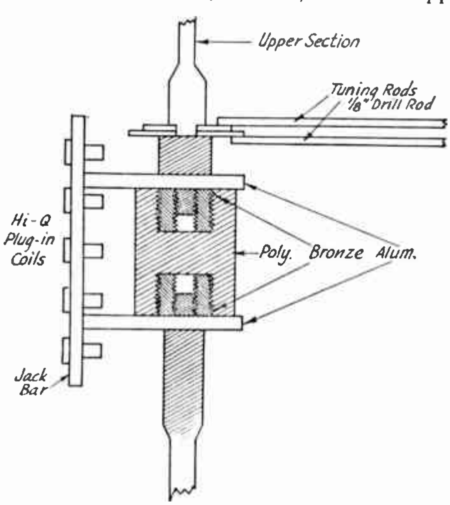


Fig. 20-35 — Construction details of the mounting for the rods and plug-in coil.

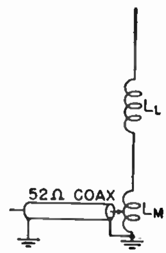
washer has been marked with an index. After the movable rod has been set, it is clamped in position by tightening up the upper antenna section. The plug-in mounting provides a convenient means of changing loading coils to go to another band.

FEEDING THE ANTENNA

It is usually found most convenient to feed the whip antenna with coax line. Unless very low-*Q* loading coils are used, the feed-point impedance will always be appreciably lower than 52 ohms — the characteristic impedance of the commonly-used coax line, RG-8/U or RG-58/U. Since the length of the transmission line will seldom exceed 10 ft., the losses involved will be negligible, even at 29 Mc., with a fairly-high s.w.r. However, unless a line of this length is made reasonably flat, difficulty may be encountered in obtaining sufficient coupling with a link to load the transmitter output stage.

One method of obtaining a match is shown in Fig. 20-36. A small inductance, L_M, is inserted at

Fig. 20-36 — A method of matching the loaded whip to 52-ohm coax cable. L_L is the loading coil and L_M the matching coil.



the base of the antenna, the loading-coil inductance being reduced correspondingly to maintain resonance. The line is then tapped on the coil at a point where the desired loading is obtained. The table shows the approximate inductance to be used between the line tap and ground. It is advisable to make the experimental matching coil larger than the value shown, so that there will be provision for varying either side of the proper position. The matching coil can also be of the plug-in type for changing bands.

Adjustment

For operation in the bands from 29 to 1.8 Mc., the whip should first be resonated at 29 Mc. with the matching coil inserted, but the line disconnected, using a grid-dip oscillator coupled to the matching coil. Then the line should be attached, and the tap varied to give proper loading, using a link at the transmitter end of the line whose reactance is approximately 52 ohms at the operating frequency, tightly coupled to the output tank circuit. After the proper position for the tap has been found, it may be necessary to readjust the antenna length slightly for resonance. This can be checked on a field-strength meter several feet away from the car.

The same procedure should be followed for each of the other bands, first resonating, with the g.d.o. coupled to the matching coil, by adjusting the loading coil.

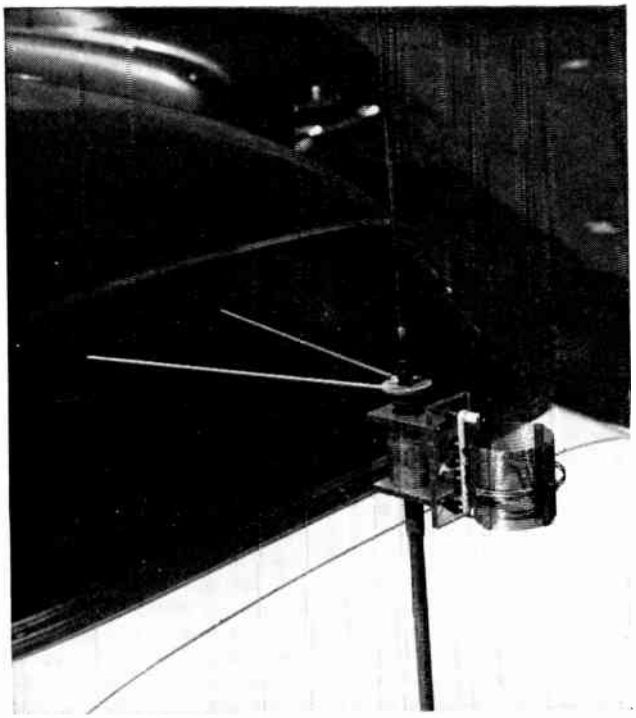


Fig. 20-37 — W8AUN's adjustable capacity hat for tuning the whip antenna over a band. The coil is a B & W type B 160-meter coil, with a turn or two removed. Spreading the rods apart increases the capacitance. This simple top loader has sufficient capacitance to permit the use of approximately the same loading-coil inductance at the center of the antenna as would normally be required for base loading.

After the position of the matching tap has been found, the size of the matching coil can be reduced to only that portion between the tap and ground, if desired. If turns are removed here, it will be necessary to reresonate with the loading coil.

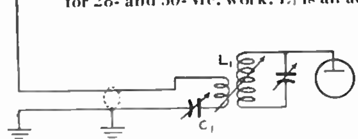
If an entirely flat line is desired, a s.w.r. indicator should be used while adjusting the line tap. With a good match, it should not be necessary to readjust for resonance after the line tap has been set.

It should be emphasized that the figures shown in the table are only approximate and may be altered considerably depending on the type of car on which the antenna is mounted and the spot at which the antenna is placed.

● ANTENNAS FOR 50 AND 144 MC.

A common type of antenna employed for mobile operation on 50 and 144 Mc. is the quarter-wave radiator which is fed with a coaxial line. The antenna, which may be a flexible telescoping "fish pole," is mounted in any of several places on the car. Quite a good match may be obtained by this method with the 50-ohm coaxial line now available; however, it is well to provide some means of tuning the system, so that all variables can be taken care of. The simplest tuning arrangement consists of a variable condenser connected between the low side of the transmitter coupling coil and ground, as shown in Fig. 20-38. This condenser should have a maximum capacitance of 75 to 100 μμfd. for 50 Mc., and should be adjusted for maximum loading with the least coupling to the transmitter. Some method of varying the coupling to the transmitter should be provided.

Fig. 20-38 — Method of feeding quarter-wave mobile antennas with coaxial line. C_1 should have a maximum capacitance of 75 to 100 μμfd. for 28- and 50-Mc. work. L_1 is an adjustable link.



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Measurements

It is practically impossible to operate an amateur station without making measurements at one time or another, even though the methods used may be quite crude. An example of a simple measurement is one that determines whether an amplifier stage in a transmitter is properly tuned; it can be done with no more elaborate equipment than a flashlight lamp and a piece of wire, but whatever the method used, a measurement is essential because the circuit itself gives no visible indication of the state of its tuning. The more refined the measuring equipment and methods, the more information can be obtained, and with more information at hand it becomes possible to adjust a piece of equipment for optimum performance more quickly and surely. Measuring and test equipment is especially valuable in building and in the initial adjustment of radio gear, and in locating and correcting breakdowns and faults.

The basic measurements are those of current, voltage, and frequency. Determination of the values of circuit elements — resistance, inductance and capacitance — are almost equally important. The inspection of waveform in audio-frequency circuits is highly useful. For these purposes there is available a wide assortment of instruments, both complete and in kit form; the latter, particularly, compare very favorably in cost with strictly home-built instruments and are frequently more satisfactory both in appearance and calibration. The instruments described in this chapter are ones having features of particular usefulness in amateur applications.

In using any instrument it should always be kept in mind that there is no such thing as an "absolute" measurement, and that measurements depend not only on the inherent accuracy of the instrument itself (which, in the case of commercially built units is usually within a few per cent, and in any event should be specified by the manufacturer) but also the conditions under which the measurement is made. Large errors can be introduced by failing to recognize the existence of conditions that affect the instrument readings. The instrument can only record what it sees — and what it sees may be something quite different from what the operator *thinks* it sees. This is particularly true in certain types of r.f. measurements, where there are many stray effects that are hard to eliminate.

D.C. Measurements

A direct-current instrument — voltmeter, ammeter, milliammeter or microammeter — is a device in which magnetic force is used to deflect a pointer over a calibrated scale in proportion to the current flowing. In the **D'Arsonval** type a coil of wire, to which the pointer is attached, is pivoted between the poles of a permanent magnet, and when current flows through the coil it causes a magnetic field that interacts with that of the magnet to cause the coil to turn. The turning force is exerted against a spiral spring attached to the coil and the pointer deflection is directly proportional to the current.

A less expensive type of instrument is the **moving-vane** type, in which a pivoted iron vane is pulled into a coil of wire by the magnetic field set up when current flows through the coil. The farther the vane extends into the coil the greater the magnetic force on it, for a given change in current, so this type of instrument does not have "linear" deflection — that is, the scale is cramped at the low-current end and spread out at the high-current end.

The same basic instrument is used for measuring either current or voltage. Good-quality instruments are made with fairly high sensitivity — that is, they give full-scale pointer deflection with very small currents — when intended to be used as voltmeters. The sensitivity of instruments intended for measuring large currents can be lower, but a highly sensitive instrument can be, and frequently is, used for measurement of currents much greater than needed for full-scale deflection.

● VOLTMETERS

Only a fraction of a volt is required for full-scale deflection of a sensitive instrument (1 milliampere or less full scale) so a high resistance is connected in series with it, Fig. 21-1, for measuring voltage. Knowing the current and the resistance, the voltage can easily be calculated from Ohm's Law. The meter is calibrated in terms of the voltage drop across the series resistor or **multiplier.** Practically any desired full-scale

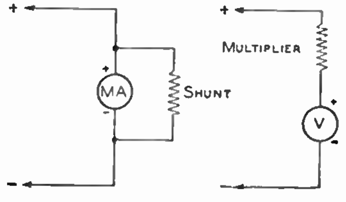


Fig. 21-1 — How voltmeter multipliers and milliammeter shunts are connected to extend the range of a d.c. meter.

voltage range can be selected by proper choice of multiplier resistance, and voltmeters frequently have several ranges selected by a switch.

The sensitivity of the voltmeter is usually expressed in "ohms per volt." A sensitivity of 1000 ohms per volt means that the resistance of the voltmeter is 1000 times the full-scale voltage, and by Ohm's Law the current required for full-scale deflection is 1 milliampere. A sensitivity of 20,000 ohms per volt, another commonly used value, means that the instrument is a 50-microampere meter. The higher the resistance of the voltmeter the more accurate the measurements in high-resistance circuits, because the current taken by the voltmeter may cause the voltage to differ from its value with the voltmeter disconnected. This is shown in Fig. 21-2.

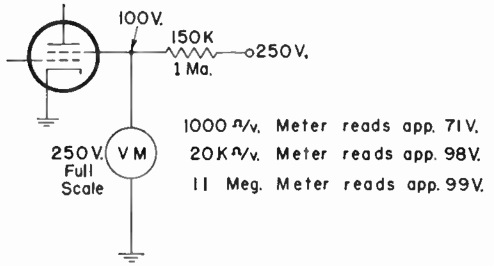


Fig. 21-2 — **Effect of voltmeter resistance on accuracy of readings. It is assumed that the d.c. resistance of the screen circuit is constant at 100 kilohms. The actual current and voltage without the voltmeter connected are 1 ma. and 100 volts. The voltmeter readings will differ because the different types of meters draw different amounts of current through the 150-kilohm resistor.**

The required multiplier resistance is found by dividing the desired full-scale voltage by the current, in amperes, required for full-scale deflection of the meter alone. Strictly, the internal resistance of the meter should be subtracted from the value so found, but this is seldom necessary (except perhaps for very low ranges) because the meter resistance will be negligibly small compared with the multiplier resistance. An exception is when the instrument is already provided with an internal multiplier, in which case the multiplier resistance required to extend the range is

$$R = R_m(n - 1)$$

where R is the multiplier resistance, R_m is the total resistance of the instrument itself, and n is the factor by which the scale is to be multiplied. For example, if a 1000-ohms-per-volt voltmeter having a calibrated range of 0–10 volts is to be extended to 1000 volts, R_m is $1000 \times 10 = 10{,}000$ ohms, n is $1000/10 = 100$, and $R = 10{,}000(100-1) = 990{,}000$ ohms.

If a milliammeter is to be used as a voltmeter, the value of series resistance can be found by Ohm's Law:

$$R = \frac{1000E}{I}$$

where E is the desired full-scale voltage and I the full-scale reading of the instrument in milliamperes.

The accuracy of a voltmeter depends on the calibration accuracy of the instrument itself and the accuracy of the multiplier resistors. Precision wire-wound resistors are used in high-quality instruments, but for most purposes standard ½- or 1-watt composition resistors will make an acceptable and economical substitute. Such resistors are supplied in tolerances of 5, 10 or 20 per cent ± the marked values. By obtaining matched pairs from the dealer's stock, one of which is, for example, 4 per cent low while the other is 4 per cent high, and using the pairs in parallel or series to obtain the required value of resistance, good accuracy can be obtained at small cost. High-voltage multipliers are preferably made up of several resistors in series; this not only raises the breakdown voltage but tends to average out errors in the individual resistors attributable to manufacturing tolerances.

MILLIAMMETERS AND AMMETERS

A microammeter or milliammeter can be used to measure currents larger than its full-scale reading by connecting a resistance **shunt** across its terminals as shown in Fig. 21-1. This diverts part of the current through the shunt, and the total current is the sum of that through the shunt and that through the meter. Knowing the meter resistance and the shunt resistance, the relative currents can easily be calculated.

The value of shunt resistance required for a given full-scale current range is given by

$$R = \frac{R_m}{n - 1}$$

where R is the shunt, R_m is the internal resistance of the meter, and n is the factor by which the original meter scale is to be multiplied. The internal resistance of a milliammeter is preferably determined from the manufacturer's catalog, but if this information is not available it can be determined by the method shown in Fig. 21-3. Do not use an ohmmeter to measure the internal resistance of a milliammeter; it may ruin the instrument.

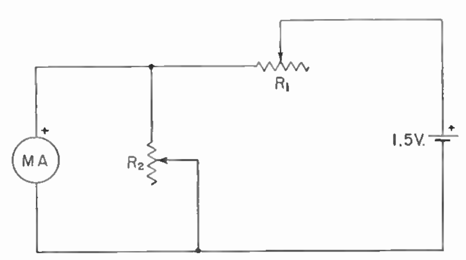


Fig. 21-3 — **Determining the internal resistance of a milliammeter or microammeter. R_1 is an adjustable resistor having a maximum value about twice that necessary for limiting the current to full scale with R_2 disconnected; adjust it for exactly full-scale reading. Then connect R_2 and adjust it for exactly half-scale reading. The resistance of R_2 is then equal to the internal resistance of the meter, and the resistor may be removed from the circuit and measured separately. Internal resistances vary from a few ohms to several hundred ohms, depending on the sensitivity of the instrument.**

Homemade milliammeter shunts can be constructed from any of the various special kinds of resistance wire, or from ordinary copper wire if no resistance wire is available. The Copper Wire Table in the data chapter gives the resistance per 1000 feet for various sizes of copper wire. After computing the resistance required, determine the smallest wire size that will carry the full-scale current (at 250 circular mils per ampere). Measure off enough wire (pulled tight but not stretched) to provide the required resistance. Accuracy can be checked by causing enough current to flow through the meter to make it read full scale without the shunt; connecting the shunt should then give the correct reading on the new full-scale range.

Any current-measuring instrument should have very low resistance compared with the resistance of the circuit being measured; otherwise, inserting the instrument will cause the current to differ from its value with the instrument out of the circuit. (This does not matter if the instrument is left permanently in the circuit.)

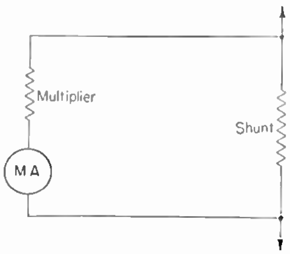


Fig. 21-4 — **Voltmeter method of measuring current. This method permits using relatively large values of resistance in the shunt, standard values of fixed resistors frequently being usable. If the multiplier resistance is 20 times the shunt resistance (or more) the error in assuming that all the current flows through the shunt will not be of consequence in most practical applications.**

However, the resistance of many circuits in radio equipment is quite high and the circuit operation is affected little, if at all, by adding as much as a few hundred ohms in series. In such cases the voltmeter method of measuring current, shown in Fig. 21-4, is frequently convenient. A voltmeter — or low-range milliammeter provided with a multiplier and operating as a voltmeter — having a full-scale voltage range of a few volts, is used to measure the voltage drop across a comparatively high resistance acting as a shunt. The formula above is used for finding the proper value of shunt resistance for a given scale-multiplying factor, R_m in this case being the multiplier resistance.

D.C. Power

Power in direct-current circuits is determined by measuring the current and voltage. When these are known, the power is equal to the voltage in volts multiplied by the current in amperes. If the current is measured with a milliammeter, the reading must be divided by 1000 to convert it to amperes.

● RESISTANCE MEASUREMENTS

Measurement of d.c. resistance is based on measuring the current through the resistance when a known voltage is applied, then using Ohm's Law. A simple circuit is shown in Fig. 21-5.

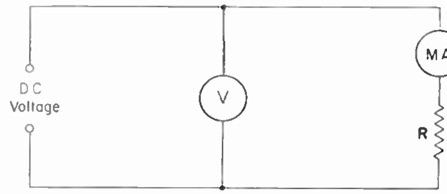


Fig. 21-5 — **Measuring resistance with a voltmeter and milliammeter. If the approximate resistance is known the voltage can be selected to cause the milliammeter, *MA*, to read about half scale. If not, additional resistance should be first connected in series with *R* to limit the current to a safe value for the milliammeter. The set-up then measures the total resistance, and the value of *R* can be found by subtracting the known additional resistance from the total.**

The internal resistance of the ammeter or milliammeter, *MA*, should be low compared with the resistance, *R*, being measured, since the voltage read by the voltmeter, *V*, is the voltage across *MA* and *R* in series. The instruments and the d.c. voltage should be chosen so that the readings are in the upper half of the scale, if possible, since the percentage error is less in this region.

An **ohmmeter** is an instrument consisting fundamentally of a voltmeter (or milliammeter, depending on the circuit used) and a small dry battery as a source of d.c. voltage, calibrated so the value of an unknown resistance can be read

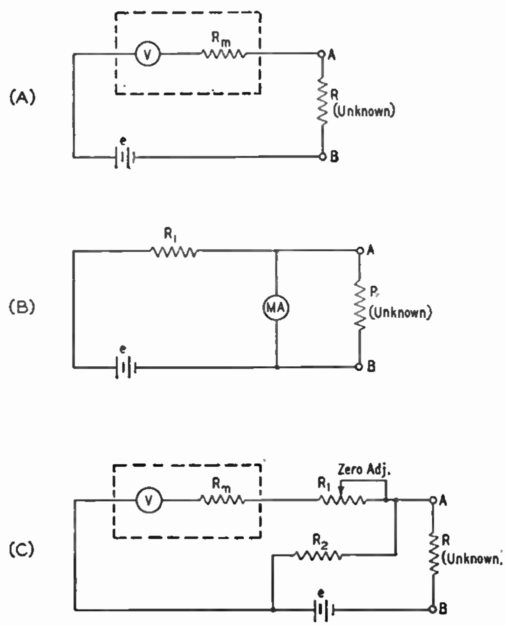


Fig. 21-6 — **Ohmmeter circuits. Values are discussed in the text.**

directly from the scale. Typical ohmmeter circuits are shown in Fig. 21-6. In the simplest type, shown in Fig. 21-6A, the meter and battery are connected in series with the unknown resistance. If a given deflection is obtained with terminals *A-B* shorted, inserting the resistance to be measured will cause the meter reading to decrease. When the resistance of the voltmeter is known, the following formula can be applied:

$$R = \frac{eR_m}{E} - R_m$$

where R is the resistance under measurement,
e is the voltage applied (*A-B* shorted),
E is the voltmeter reading with R connected, and
R_m is the resistance of the voltmeter.

The circuit of Fig. 21-6A is not suited to measuring low values of resistance (below a hundred ohms or so) with a high-resistance voltmeter. For such measurements the circuit of Fig. 21-6B can be used. The milliammeter should be a 0–1 ma. instrument, and R_1 should be equal to the battery voltage, e, multiplied by 1000. The unknown resistance is

$$R = \frac{I_2 R_m}{I_1 - I_2}$$

where R is the unknown,
R_m is the internal resistance of the milliammeter,
I_1 is the current in ma. with R disconnected from terminals *A-B*, and
I_2 is the current in ma. with R connected.

The formula is approximate, but the error will be negligible if e is at least 3 volts so that R_1 is at least 3000 ohms.

A third circuit for measuring resistance is shown in Fig. 21-6C. In this case a high-resistance voltmeter is used to measure the voltage drop across a reference resistor, R_2, when the unknown resistor is connected so that current flows through it, R_2 and the battery in series. By suitable choice of R_2 (low values for low resistance, high values for high-resistance unknowns) this circuit will give equally good results on all resistance values in the range from one ohm to several megohms, provided that the voltmeter resistance, R_m, is always very high (50 times or more) compared with the resistance of R_2. A 20,000-ohms-per-volt instrument (50-μamp. movement) is generally used. Assuming that the current through the voltmeter is negligible compared with the current through R_2, the formula for the unknown is

$$R = \frac{eR_2}{E} - R_2$$

where R and R_2 are as shown in Fig. 21-6C,
e is the voltmeter reading with *A-B* shorted, and
E is the voltmeter reading with R connected.

The "zero adjuster," R_1, is used to set the voltmeter reading exactly to full scale when the meter is calibrated in ohms. A 10,000-ohm variable resistor is suitable with a 20,000-ohms-per-volt meter. The battery voltage is usually 3 volts for ranges up to 100,000 ohms or so and 6 volts for higher ranges.

Combination Instruments

Since the same basic instrument is used for measuring current, voltage and resistance, the three functions can readily be combined in one unit using a single meter. Various models of the "VOM" (volt-ohm-milliammeter) are available commercially, the less expensive ones using a 0-1 milliammeter. A simple circuit based on such a meter is shown in Fig. 21-7. It has five current ranges, from 1 ma. to 1 ampere, three voltage ranges, 10 volts to 1000 volts, and two resistance ranges. Fig. 21-8 shows the ohmmeter calibration; the low-ohms curve is for a meter having an internal resistance of 55 ohms and should be calculated from the formula above (Fig. 21-6B) for instruments of different resistance.

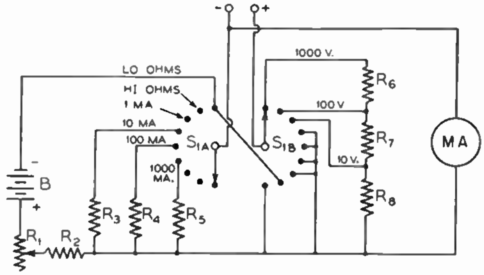


Fig. 21-7 — **Diagram of the volt-ohm-milliammeter.**
R_1 — 2000-ohm wire-wound variable.
R_2 — 3000 ohms, ½ watt.
R_3 — 10-ma. shunt, 6.11 ohms (see text).
R_4 — 100-ma. shunt, 0.555 ohm (see text).
R_5 — 1000-ma. shunt, 0.055 ohm (see text).
R_6 — 1000-volt multiplier, 0.9 megohm, ½ watt.
R_7 — 100-volt multiplier, 90,000 ohms, ½ watt.
R_8 — 10-volt multiplier, 10,000 ohms, ½ watt.
B — 4.5-volt dry battery.
S_{1A-B} — 9-point 2-pole selector switch.
MA — 0-1 milliammeter.

Ordinary carbon resistors can be used as voltmeter multipliers, connecting them in series or parallel to obtain a given value. The 10-, 100- and 1000-ma. shunts can be made of copper wire wound on small forms. The approximate lengths and sizes of the wire for the shunts are as follows: R_3, 9 feet No. 38 enameled; R_4, 5 feet No. 30 enameled; R_5, 8½ feet No. 18.

It is possible to buy special VOM scales to replace the 0–1 scale for certain types of milliammeters. In such case the circuit recommended for that scale should be used.

More expensive instruments use a 50-μamp. meter in the VOM, with large scales for easy reading. Such instruments frequently include a.c. scales as well, and in general are better purchased complete than made at home.

The VOM, even a very simple one, is among the most useful instruments for the amateur. Besides current and voltage measurements, it

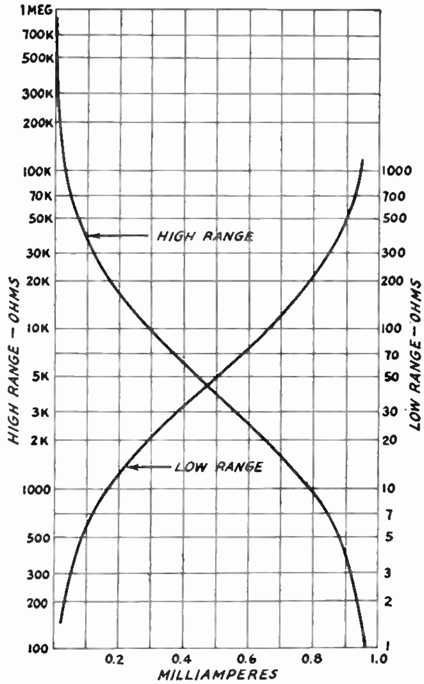


Fig. 21-8 — Calibration curve for the high- and low-resistance ranges of the volt-ohm-milliammeter.

can be used for checking continuity in circuits, for finding defective components before installation — shorted condensers, open or otherwise defective resistors, etc. — shorts or opens in wiring, and many other checks that, if applied during the construction of a piece of equipment, save much time and trouble. It is equally useful for servicing, when a component fails during regular operation.

THE VACUUM-TUBE VOLTMETER

The usefulness of the **vacuum-tube voltmeter (VTVM)** is based on the fact that a vacuum tube can amplify without taking power from the source of voltage applied to its grid. It is therefore possible to have a voltmeter of extremely high resistance, and thus take negligible current from the circuit under measurement, without using a d.c. instrument of exceptional sensitivity.

While there are several possible circuits, the one commonly used is shown in Fig. 21-9. A dual triode, V_1, is arranged so that, with no voltage applied to the left-hand grid, equal currents flow through both sections. Under this condition the two cathodes are at the same potential and no current flows through *M*. The currents can be adjusted to balance by potentiometer R_{11}, which takes care of variations in the tube sections and in the values of cathode resistors R_9 and R_{10}. When a voltage is applied to the left-hand grid the current through that tube section changes but the current through the other section remains unchanged, so the balance is upset and the meter indicates. The sensitivity of the meter is regulated by R_8, which serves to adjust the calibration. R_{12}, common to the cathodes of both tube sections, is a feed-back resistor that stabilizes the system and makes the readings linear. R_6 and C_1 form a filter for any a.c. component that may be present, and R_6 is balanced by R_7 connected to the grid of the second tube section.

To stay well within the linear range of operation the scale is limited to 3 volts or less in the average commercial instrument. Higher ranges are obtained by means of the voltage divider formed by R_1 to R_5, inclusive. As many ranges as desired can be used. Common practice is to use 1 megohm at R_1, and to make the sum of R_2 to R_5, inclusive, 10 megohms, thus giving a total resistance of 11 megohms, constant for all voltage ranges.

For measuring a.c. voltages the rectifier circuit shown at the lower left of Fig. 21-9 is used. One section of the double diode, V_2, is a half-wave rectifier and the second half acts as a balancing device, adjustable by R_{17}, to eliminate contact potential effects that would cause a constant d.c. voltage to appear at the VTVM grid. When measuring a.c., R_8 is usually set so that the r.m.s. a.c. calibration coincides with the d.c. calibration. A separate resistor is frequently switched in for the purpose.

Values to be used in the circuit depend consid-

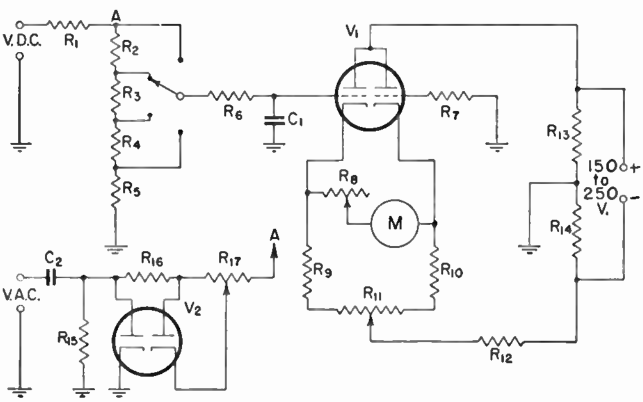


Fig. 21-9 — Vacuum-tube voltmeter circuit.

C_1 — 0.002- to 0.005-μfd. mica.
C_2 — 0.01 μfd., 1000 to 2000 volts, paper or mica.
R_1 — 1 megohm, ½ watt.
R_1 to R_5, inclusive — To give desired voltage ranges, totaling 10 megohms.
R_6, R_7 — 2 to 3 megohms.
R_8 — 10,000-ohm variable.
R_9, R_{10} — 2000 to 3000 ohms.
R_{11} — 5000- to 10,000-ohm potentiometer.
R_{12} — 10,000 to 50,000 ohms.
R_{13}, R_{14} — App. 25,000 ohms. A 50,000-ohm slider-type wire-wound can be used.
R_{15} — 10 megohms.
R_{16} — 3 megohms.
R_{17} — 10-megohm variable.
M — Microammeter, range from 0–200 μamp. to 0–1 ma.
V_1 — Dual triode, 6SN7 or 12AU7.
V_2 — Dual diode, 6H6 or 6AL5.

erably on the supply voltage and the sensitivity of the meter, *M*. R_{12}, and R_{13}–R_{14}, should be adjusted so that the voltmeter circuit can be brought to balance, and to give full-scale deflection on *M* with about 3 volts applied to the grid. The meter connections can be reversed to read voltages that are negative with respect to ground.

The VTVM has the disadvantage that it requires a source of power for its operation, as compared with a regular d.c. instrument. Also, it is susceptible to r.f. pick-up when working around an operating transmitter, unless well shielded and filtered. The fact that one of its terminals is grounded is also disadvantageous in some cases, since a.c. readings in particular may be inaccurate if an attempt is made to measure a circuit having both sides "hot" with respect to ground. Nevertheless, the high resistance of the VTVM more than compensates for these disadvantages, especially since in the majority of measurements they do not apply.

CALIBRATION

When extending the range of a d.c. instrument calibration usually is necessary, although resistors for voltmeter multipliers often can be purchased to close-enough tolerances so that the new range will be accurately known. However, in calibrating an instrument such as a VTVM a known voltage must be available to provide a starting point. Fresh dry cells have an open-circuit terminal voltage of approximately 1.6 volts, and one or more of them may be connected in series to provide several calibration points on the low range. Gas regulator tubes in a power supply, such as the 0C3, 0D3, etc., also provide a stable source of voltage whose value is known within a few per cent. Once a few such points are determined the voltmeter ranges may be extended readily by adding multipliers or a voltage divider as appropriate.

Shunts for a milliammeter may be adjusted by first using the meter alone in series with a source of voltage and a resistor selected to limit the current to full scale. For example, a 0-1 milliammeter may be connected in series with a dry cell and a 2000-ohm variable resistor, the latter being adjusted to allow exactly 1 milliampere to flow. Then the shunt is added across the meter and its resistance adjusted to reduce the meter reading by exactly the scale factor, *n*. If *n* is 5, the shunt would be adjusted to make the meter read 0.2 milliampere, so the full-scale current will be 5 ma. Using the new scale, the second shunt is added to give the next range, the same procedure being followed. This can be carried on for several ranges, but it is advisable to check the meter on the highest range against a separate meter used as a standard, since the errors in this process tend to be cumulative.

Measurement of Frequency and Wavelength

ABSORPTION FREQUENCY METERS

The simplest possible frequency-measuring device is a resonant circuit, tunable over the desired frequency range and having its tuning dial calibrated in terms of frequency. It operates by extracting a small amount of energy from the oscillating circuit to be measured, the frequency being determined by the tuning setting at which the energy absorption is maximum (Fig. 21-10).

Although such an instrument is not capable of very high accuracy, because the *Q* of the tuned circuit cannot be high enough to avoid uncertainty in the exact setting and because any two coupled circuits interact to some extent and change each others' tuning, the **absorption wavemeter** or frequency meter is nevertheless a highly useful instrument. It is compact, inexpensive, and requires no power supply. There is no ambiguity in its indications, as is frequently the case with the heterodyne-type instruments described later.

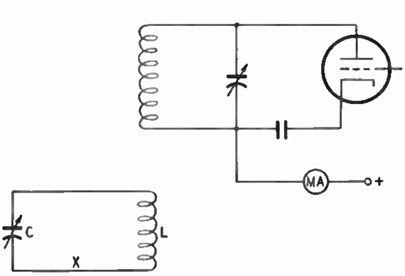


Fig. 21-10 — Absorption frequency meter and a typical application. The meter consists simply of a calibrated resonant circuit *LC*. When coupled to an amplifier or oscillator the tube plate current will rise when the frequency meter is tuned to resonance. A flashlight lamp may be connected in series at X to give a visual indication, but it decreases the selectivity of the instrument and makes it necessary to use rather close coupling to the circuit being measured.

When an absorption meter is used for checking a transmitter, the plate current of the tube connected to the circuit being checked can provide the necessary resonance indication. When the frequency meter is loosely coupled to the tank circuit the plate current will give a slight upward flicker as the meter is tuned through resonance. The accuracy is greatest when the loosest possible coupling is used.

A receiver oscillator may be checked by tuning in a steady signal and heterodyning it to give a beat note as in ordinary c.w. reception. When the frequency meter is coupled to the oscillator coil and tuned through resonance the beat note will change. Again, the coupling should be made loose enough so that a just-perceptible change in beat note is observed.

An approximate calibration for the wavemeter, adequate for most purposes, may be obtained by comparison with a calibrated receiver. The usual receiver dial calibration is sufficiently accurate. A simple oscillator circuit covering the same range as the frequency meter will be useful in calibration. Set the receiver to a given frequency, tune the oscillator to zero beat at the same frequency, and adjust the frequency meter to resonance with the oscillator as described above. This gives one calibration point. When a sufficient number of such points has been obtained a graph may be drawn to show frequency *vs.* dial settings on the frequency meter.

● INDICATING WAVEMETERS

The plain absorption meter requires fairly close coupling to the oscillating circuit to affect the plate current of a tube sufficiently to give visual indication. The sensitivity of the instrument can be increased, by adding a rectifier and d.c. microammeter or milliammeter, to the point where very loose coupling will suffice for a good reading. A typical circuit for this purpose is given in Fig. 21-11, and Figs. 21-12 and 21-13 show how such an instrument can be constructed. For convenience in use, the tuned circuit is mounted in a small metal box that can be held in one hand for close coupling to a circuit. The d.c. meter can be connected or not as desired, since it is separate (it can also be mounted in a small box) so the instrument can be used either as a plain absorption meter or as an indicating-type meter.

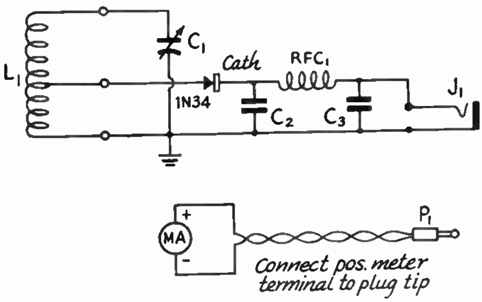


Fig. 21-11 — **Circuit diagram of indicating wavemeter. With the meter plug removed, it can be used as a compact absorption meter of the ordinary type.**

C_1 — 50-μμfd. variable (Hammarlund HF-50).
C_2, C_3 — 0.001-μfd. disc ceramic.
J_1 — Open-circuit jack.
MA — D.c. milliammeter, 0–1 or less.
P_1 — 'Phone plug.

Coil Data, L_1

Freq. Range	*Turns*	*Wire*	*Diameter*	*Turns/inch*	*Tap**
1.6–4.2 Mc.	139	32 enam.	¾ in.	Close-wound	32
3.6–10.5 Mc.	40	32 enam.	¾ in.	Close-wound	12
7.8–24.0 Mc.	40	24 tinned	½ in.	32	14
17.8–52.0 Mc.	15	20 tinned	½ in.	16	5
38 –117 Mc.	4	20 tinned	½ in.	16	1⅓
80 –270 Mc.	Hairpin of No. 14 wire, ⅝ in. spacing, 2 inches long including coil form pins. Tapped 1½ in. from ground end.				

* Turns from ground end.
Coil forms are Amphenol 24-5H, ¾-in. diameter.

The rectifier is a crystal diode, tapped down on the tuned-circuit coil to avoid excessive loading

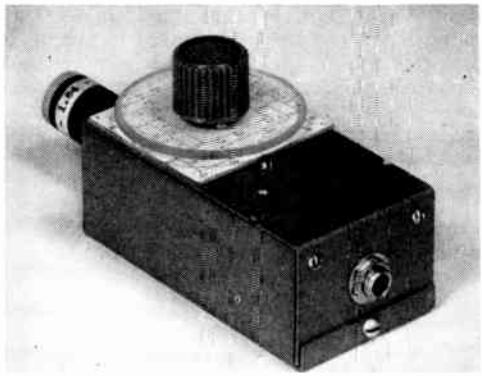


Fig. 21-12 — **A compact absorption wavemeter provided with a crystal rectifier and jack for an indicating meter. The meter can be mounted in a separate box, if desired. The dial is similar to that used on the grid-dip meter described later in this chapter.**

of the circuit which would broaden the tuning. Tapping down also improves the sensitivity, by providing an approximate impedance match between the tuned circuit and the crystal-circuit load. By plugging a headset into the output jack ('phones having 2000 ohms or greater resistance should be used for greatest sensitivity) the wavemeter can be used as a monitor for modulated transmissions.

It is of course possible to mount the d.c. meter in the same unit with the wavemeter proper, but this increases the bulk and weight. The separate units have the advantage, also, that a long line can be used to connect the two, since such a line carries only d.c., so the meter can be placed at a remote point to pick up r.f. while the indicator is placed at the spot where adjustments are being made. This is frequently useful in antenna work, for example.

Where connection to an a.c. line is convenient, a VTVM can be used instead of the milliammeter or microammeter, and because of its high resistance will considerably increase the sensitivity and selectivity of the wavemeter.

In addition to the uses mentioned above, a meter of this type may be used for final adjust-

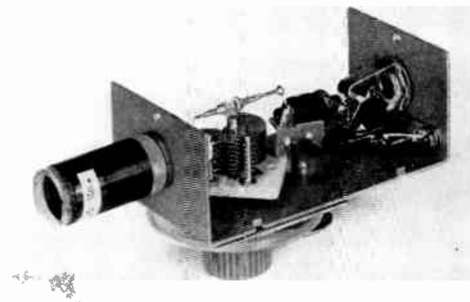


Fig. 21-13 — **Inside the indicating-type wavemeter. The tuning condenser should be mounted as close as possible to the coil socket so the leads will be of negligible length. The box is 1⅝ × 2¼ × 4 inches.**

ment of neutralization in r.f. amplifiers. For this purpose it may be loosely coupled to the plate tank coil. Alternatively, L_1 may be removed and the final-amplifier link output terminals connected to the coil socket. The latter method tends to ensure that the pick-up is from the final tank coil only.

LECHER WIRES

At very-high and ultrahigh frequencies it is possible to determine frequency by actually measuring the length of the waves generated. The measurement is made by observing standing waves on a two-wire parallel transmission line or **Lecher wires.** Such a line shows pronounced resonance effects, and it is possible to determine quite accurately the current loops (points of maximum current). The physical distance between two consecutive current loops is equal to one-half wavelength. Thus the wavelength can be read directly in meters (39.37 inches = 1 meter; 0.3937 inch = 1 cm.), or in centimeters for the very-short wavelengths.

The Lecher-wire line should be at least a wavelength long — that is, 7 feet or more on 144 Mc. — and should be entirely air-insulated except where it is supported at the ends. It may be made of copper tubing or of wires stretched tightly. The spacing between wires should not exceed about 2 per cent of the shortest wavelength to be measured. The positions of the current loops are found by means of a "shorting bar," which is simply a metal strip or knife edge which can be slid along the line to vary its effective length.

Making Measurements

For measuring the frequency of a transmitter, a convenient and fairly sensitive indicator can be made by soldering the ends of a one-turn loop of wire, of about the same diameter as the transmitter tank coil, to a low-current flashlight bulb. The loop should be coupled to the tank coil to give a moderately bright glow. A coupling loop should be connected to the ends of the Lecher wires and brought near the tank coil, as shown in Fig. 21-15. Then the shorting bar should be slid along the wires outward from the transmitter until the lamp gives a sharp dip in brightness. This point should be marked and the shorting bar moved out until a second dip is obtained. The distance between the two points will be equal to half the wavelength. If the measurement is made in inches, the frequency will be

$$F_{\text{Mc.}} = \frac{5905}{length \text{ (inches)}}.$$

If the length is measured in meters,

$$F_{\text{Mc.}} = \frac{150}{length \text{ (meters)}}.$$

In checking a superregenerative receiver, the Lecher wires may be similarly coupled to the receiver coil. In this case the resonance indication may be obtained by setting the receiver just to the point where the hiss is obtained, then as the bar is slid along the wires a spot will be found where the receiver goes out of oscillation. The distance between two such spots is equal to a half-wavelength.

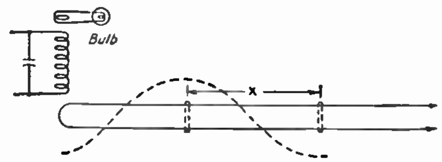


Fig. 21-15 — Coupling a Lecher wire system to a transmitter tank coil. Typical standing-wave distribution is shown by the dashed line. The distance X between the positions of the shorting bar at the current loops equals one-half wavelength.

The shorting bar must be kept at right angles to the two wires. A sharp edge on the bar is desirable, since it not only helps make good contact but also definitely locates the *point* of contact.

Accurate readings result when the loosest possible coupling is used between the line and the tank coil. Careful measurement of the exact distance between two current loops also is essential.

HETERODYNE METHODS

Heterodyne methods of frequency measurement make use of a stable oscillator generating either a known frequency or one that is variable over a known range. Measurement consists in comparing the unknown frequency with the known frequency of the oscillator, using an ordinary receiver for detecting both. This method is more accurate than others, because frequency

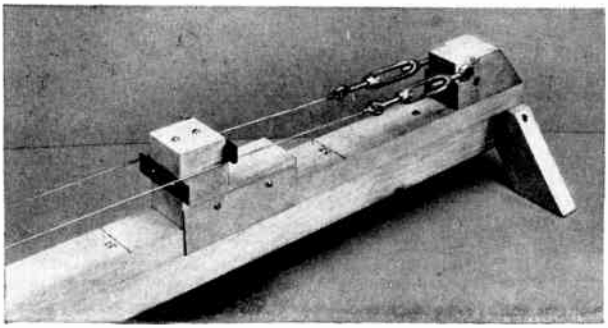


Fig. 21-14 — One end of a typical Lecher wire system. The wire is No. 16 bare solid-copper antenna wire (hard-drawn). The turnbuckles are held in place by a 3/16 × 2-inch bolt through the anchor block. The other end of the line, which is connected to the pick-up loop, should be insulated.

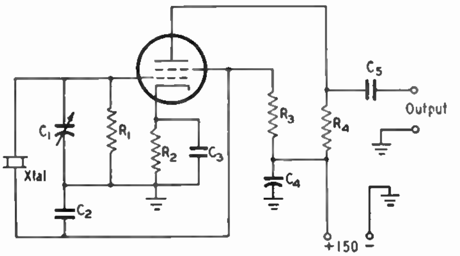


Fig. 21-16 — Circuit for crystal-controlled frequency standard. Tubes such as the 6SK7, 6SH7, 6AU6, etc., are suitable.

C_1 — 50-μμfd. variable.
C_2 — 150-μμfd. mica.
C_3 — 0.0022-μfd. mica.
C_4 — 0.01-μfd. paper.
C_5 — 22-μμfd. mica.
R_1 — 0.47 megohm, ½ watt.
R_2 — 1000 ohms, ½ watt.
R_3 — 0.1 megohm, ½ watt.
R_4 — 0.15 megohm, ½ watt.

differences of less than a cycle can be observed by aural (beat-note) methods, and the oscillator can be calibrated to practically any degree of precision by comparison with standard frequencies transmitted from WWV and WWVH.

Care must be used in heterodyne frequency measurement because in most cases harmonics are used and the measured frequency can be in error by a large factor if the wrong harmonic is picked. Also, a superheterodyne receiver will give many spurious responses in the presence of a strong signal and harmonics, so these must be recognized and ignored in making measurements. In general, heterodyne methods are most useful in measuring frequency to a high degree of accuracy *after* the frequency is known approximately from other methods. The absorption wavemeter is useful for making the first approximation and thus eliminating the possible gross errors.

Frequency Measurement with the Receiver

An ordinary receiver has the essential elements needed for frequency measurement. Its dial readings must be calibrated in terms of frequency, of course, before measurements can be made. Manufactured receivers are generally so calibrated; the accuracy of the calibration will vary with the receiver model, but if the receiver is well made and has good inherent stability, a bandspread dial calibration can be relied upon to within perhaps 0.2 per cent. For most accurate measurement, maximum response in the receiver should be determined by means of a carrier-operated tuning indicator (such as an S-meter), the receiver beat oscillator being turned off. If the receiver has a crystal filter, it should be set in a fairly "sharp" position to increase the accuracy.

When checking the frequency of your own transmitter, the receiving antenna should be disconnected so the signal will not overload or "block" the receiver. Also, the r.f. gain should be reduced as a further precaution against overloading. If the receiver still blocks without an antenna the frequency may be checked by turning off the power amplifier and tuning in the oscillator alone. It is difficult to avoid blocking under almost any conditions with a regenerative receiver, and so this type is not very suitable for checking the frequency of one's own transmitter.

THE HETERODYNE FREQUENCY METER

The **heterodyne frequency meter** is an oscillator with a precise frequency calibration. The oscillator must be so designed and constructed that it can be accurately calibrated and will retain its calibration over long periods of time.

The oscillator used in the frequency meter must be very stable. Mechanical considerations are most important in its construction. No matter how good the instrument may be electrically, its accuracy cannot be depended upon if the mechanical construction is flimsy. Frequency stability can be improved by avoiding the use of phenolic compounds and thermoplastics (bakelite, polystyrene, etc.) in the oscillator circuit, employing only high-grade ceramics instead. Plug-in coils ordinarily are not acceptable; instead, a solidly-built and firmly-mounted tuned circuit should be permanently installed. The oscillator panel and chassis should be as rigid as possible.

For amateur purposes the most useful type of meter is one covering the amateur bands only. The VFOs described in the chapter on transmitters are typical of the circuits and construc-

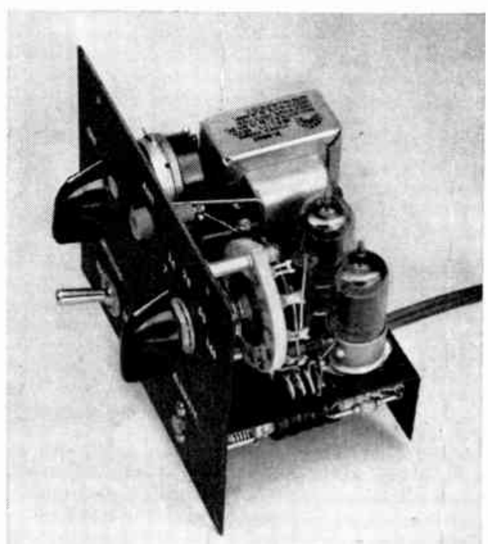
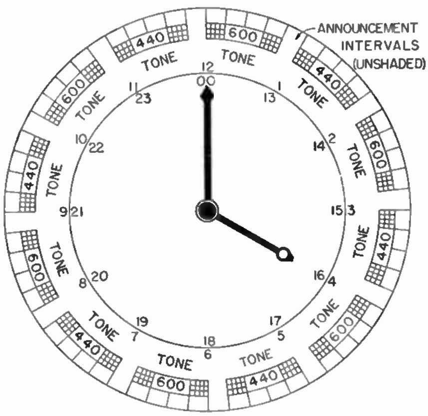


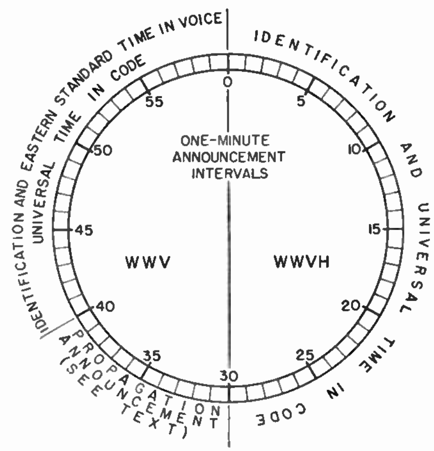
Fig. 21-17 — A compact frequency standard and harmonic amplifier for generating either 100- or 1000-kc. intervals throughout the spectrum to 150 Mc. It has a self-contained power supply using the transformer shown in the upper part of the photo. The output control is at the upper left, and the switch in the foreground is the harmonic-amplifier bandswitch. The dual crystal is between the bandswitch and output control. The toggle switch at the lower left corner of the panel selects either 1000- or 100-kc. intervals.

tion since they are designed with the same considerations in mind — i.e., to be highly stable both electrically and mechanically. Hence a good VFO, if accurately calibrated in frequency, is also a good heterodyne frequency meter.

Calibration must be done by comparing the oscillator frequency at various points in its range with signals of known frequency. The best method is to calibrate from a secondary frequency standard, described in the next section, at intervals of, say, 100 kc. and fill in the calibration curve by interpolation. The oscillator usually works over the approximate range 1750–2000 kc., harmonics being used for the higher amateur bands. If the calibration is done on the highest band — 28-32 Mc. — at intervals of 100 kc. it is equivalent to having calibration points at intervals of 100/16 = 6.25 kc. on the fundamental-frequency range.

STANDARD FREQUENCIES AND TIME SIGNALS





Standard radio and audio frequencies are broadcast continuously from WWV, operated by the Central Radio Propagation Laboratory, National Bureau of Standards, Washington, D. C. on the following frequencies:

Freq., Mc.	*Modulations (c.p.s.)*
2.5	1, 440 or 600
5	1, 440 or 600
10	1, 440 or 600
15	1, 440 or 600
20	1, 440 or 600
25	1, 440 or 600

Similar broadcasts are given from WWVH, Puunene, T.H., on the following frequencies:

Freq., Mc.	*Modulations (c.p.s.)*
5	1, 440 or 600
10	1, 440 or 600
15	1, 440 or 600

Transmissions are as given in the charts above, except that the WWVH broadcast is interrupted for 4 minutes following each hour and half hour and for periods of 40 minutes beginning at 0700 and 1900 universal time.

Time Signals

The 1-c.p.s. modulation is a 5-millisecond pulse at intervals of precisely one second, and is heard as a tick. Time intervals as transmitted are accurate to within 2 parts in 100 million + 1 microsecond. The tick on the 59th second is omitted.

Accuracy

Transmitted frequencies are accurate within 2 parts in 100 million.

Propagation Notices

During the announcement intervals at 20 minutes after and 10 minutes before the hour, propagation notices applying to transmission paths over the north Atlantic are transmitted from WWV on 2.5, 5, 10, 15, 20, and 25 Mc. These notices, in telegraphic code, consist of the letter N, W, or U followed by a number. The letter designations apply to propagation conditions as of the time of the broadcast, and have the following significance:

W — Ionospheric disturbance in progress or expected.
U — Unstable conditions, but communication possible with high power.
N — No warning.

The number designations apply to expected propagation conditions during the subsequent 12 hours and have the following significance:

Digit	*Forecast*
1	Impossible
2	Very Poor
3	Poor
4	Fair to Poor
5	Fair
6	Fair to Good
7	Good
8	Very Good
9	Excellent

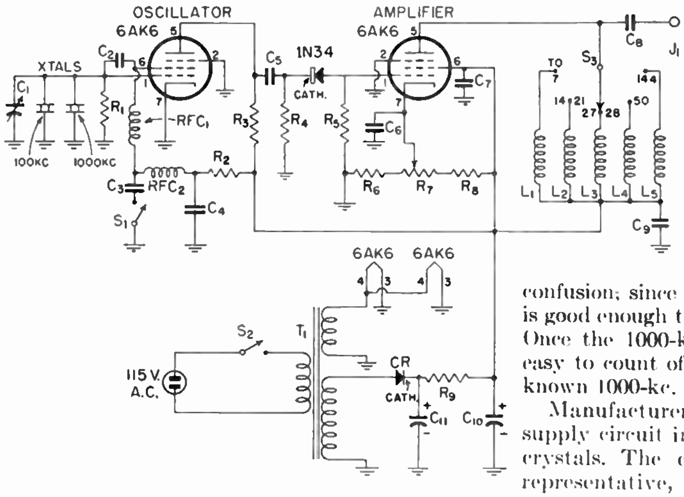


Fig. 21-18 — **Circuit diagram of the frequency standard and harmonic amplifier.**

C_1 — 25-μμfd. midget variable (Hammarlund MAPC-25).
C_2 — 3 μμfd. (2½ inches of 75-ohm Twin-Lead).
C_3, C_4 — 0.1-μfd. paper, 400 volts.
C_5 — 250-μμfd. ceramic.
C_6, C_7, C_9 — 0.001-μfd. disc ceramic.
C_8 — 100-μμfd. ceramic.
C_{10}, C_{11} — 20-μfd. electrolytic, 250 volts.
R_1 — 4.7 megohm, ½ watt.
R_2 — 22,000 ohms, ½ watt.
R_3, R_4, R_5 — 0.47 megohm, ½ watt.
R_6 — 470 ohms, ½ watt.
R_7 — 5000-ohm potentiometer.
R_8 — 47,000 ohms, 1 watt.
R_9 — 1000 ohms, 1 watt.
L_1 — 1-mh. r.f. choke (National R-50).
L_2 — 4-μh r.f. choke (National R-60).
L_3 — 2-μh r.f. choke (National R-60).
L_4 — 0.5 μh. (1-μh. r.f. choke, National R-33, with 10 turns removed).
L_5 — 3 turns No. 16, ¼-inch diam., ⅜ inch long.
CR — 65-ma. selenium rectifier.
J_1 — Tip jack.
RFC_1 — 0.5-mh. r.f. choke (National R-50).
RFC_2 — 5-mh. r.f. choke (National R-100S).
S_1 — S.p.s.t. toggle switch.
S_2 — S.p.s.t. toggle switch mounted on R_7.
S_3 — 1-pole 6-position selector switch; shorting type (Centralab 2500).
T_1 — Power transformer, 150 volts, 25 ma.; 6.3 volts, 0.5 amp. (Merit P-3046).
XTAL — 100–1000-kc. dual frequency crystal (Valpey DFS).

THE SECONDARY FREQUENCY STANDARD

The **secondary frequency standard** is a highly-stable oscillator generating a single frequency, usually 100 kc. It is nearly always crystal-controlled, and inexpensive 100-kc. crystals are available for the purpose. Since the harmonics are multiples of 100 kc. throughout the spectrum, some of them can be compared directly with the standard frequencies transmitted by WWV. The edges of most amateur bands also are exact multiples of 100 kc., so it becomes possible to determine the band edges very accurately. This is an important consideration in amateur frequency measurement, since the only regulatory requirement is that an amateur transmission be inside the assigned band and not on a specific frequency.

Intervals of 100 kc. are sometimes too close for accurate identification of a given harmonic, so special crystals that operate at both 1000 and 100 kc. are available. Intervals of 1000 kc. are sufficiently far apart to avoid confusion; since the average receiver calibration is good enough to provide positive identification. Once the 1000-kc. harmonics are spotted, it is easy to count off the 100-kc. intervals from the known 1000-kc. points.

Manufacturers of 100-kc. crystals usually supply circuit information for their particular crystals. The circuit given in Fig. 21-16 is representative, and will generate usable harmonics up to 30 Mc. or so. The variable condenser, C_1, provides a means for adjusting the frequency to exactly 100 kc. Harmonic output is taken from the circuit through a small condenser, C_5. There are no particular constructional points to be observed in building such a unit. Power for the tube heater and plate may be taken from the supply in the receiver with which the unit is to be used. The plate voltage is not critical, but it is recommended that it be taken from a VR-150 regulator if the receiver is equipped with one.

Sufficient signal strength usually will be secured if a wire is run between the output terminal connected to C_5 and the antenna post on the receiver. At the lower frequencies a metallic connection may not be necessary.

Figs. 21-17 through 21-19 show a compact standard, complete with power supply, that will give usable harmonics from both 100 and 1000 kc. up through the 144-Mc. band. It uses a dual crystal, either fundamental frequency being selected by a switch, and the output of the oscillator is fed to a crystal-diode rectifier to increase the amplitude of the high-order harmonics. These harmonics are then amplified in the second tube, a stage having broadly-tuned plate circuits centering in the higher-frequency amateur bands, switched in or out as required. A cathode gain control is provided in the amplifier circuit for regulating the output amplitude. The whole unit is constructed in a 5 × 3 × 4 box of the type having its own chassis, the small size being used so the unit can be squeezed into limited space on the operating table. It can be put on a larger chassis and box if desired, since the construction is not critical. Sufficient signal strength in the receiver should be secured by connecting a short piece of wire to the output terminal, but on very high frequencies it may be necessary to connect the wire to one antenna post on the receiver.

Adjusting to Frequency

In either Fig. 21-16 or 21-18 the frequency can be adjusted exactly to 100 kc. by making use of

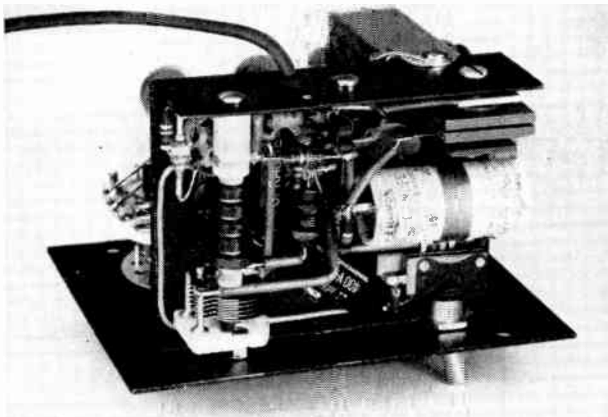


Fig. 21-19 — Below-chassis view of the frequency standard. The 1N34A harmonic generator is at the upper left. The variable condenser at the bottom is for adjustment of the oscillator frequency to exactly 100 kc. At the upper right, mounted on the rear lip of the chassis, is the selenium rectifier for the power supply. The filter condenser is just below it. Small resistors and condensers are grouped around the tube sockets.

the WWV transmissions tabulated in this chapter. Select the WWV frequency that gives a good signal at your location at the time of day most convenient. Tune it in with the receiver b.f.o. off and wait for the period during which the modulation is absent. Then switch on the 100-kc. oscillator and adjust its frequency, by means of C_1, until its harmonic is in zero beat with WWV. The exact setting is easily found by observing the slow pulsation in background noise as the harmonic comes close to zero beat, and adjusting to where the pulsation disappears or occurs at a very slow rate. The pulsations can be observed even more readily by switching on the receiver's b.f.o., after approximate zero beat has been secured, and observing the rise and fall in intensity (not frequency) of the beat tone. For best results the WWV signal and the signal from the 100-kc. oscillator should be about the same strength. It is advisable not to try to set the 100-kc. oscillator when the WWV signal is modulated, since it is difficult to tell whether the harmonic is being adjusted to zero beat with the carrier or with one of the sidebands.

Frequency Checking

The secondary standard provides signals of known frequency that can be tuned in on the station receiver. Determination of the frequency of a transmitter is then carried out by the method described earlier under "Frequency Measurement with the Receiver," using these points as positive identification of band edges. By using the known 100-kc. points the receiver calibration can be corrected so that, by interpolation, the frequency of a signal lying between the calibration points can be determined with good accuracy.

More Precise Methods

The methods described in this section are quite adequate for the primary purpose of amateur frequency measurements — that is, determining whether or not a transmitter is operating inside the limits of an amateur band, and the approximate frequency inside the band. For measurement of an unknown frequency to a high degree of accuracy more advanced methods can be used. Accurate signals at closer intervals can be obtained by using a multivibrator in conjunction with the 100-kc. standard, and thus obtaining signals at intervals of, say, 10 kc. or some other integral divisor of 100. Temperature control is frequently used on the 100-kc. oscillator to give a high order of stability (Collier, "What Price Precision?", *QST*, September and October, 1952). Also, the secondary standard can be used in conjunction with a variable-frequency interpolation oscillator to fill in the standard intervals (Woodward, "A Linear Beat-Frequency Oscillator for Frequency Measurement," *QST*, May, 1951). An interpolation oscillator and standard can be combined in one instrument, one application of this type having been described in *QST* for May, 1949 (Grammer, "The Additive Frequency Meter").

Test Oscillators

For many measurements and tests, it is necessary to have a source of signal at some desired frequency or range of frequencies. Although there is a wide variety of test oscillators capable of generating such signals, for most amateur work one or two simple types are quite adequate. A variable-frequency oscillator covering as much as possible of the r.f. spectrum, calibrated in frequency, has many useful applications. For 'phone work, an audio signal source is equally valuable in testing and adjustment of speech amplifiers, modulators and associated audio circuits and equipment. Both types can be built quite easily and at low cost.

THE GRID-DIP METER

The **grid-dip meter** is a simple vacuum-tube oscillator to which a low-range milliammeter or microammeter has been added to read the oscillator grid current. A 0–1 milliammeter is sensitive enough in most cases. The grid-dip meter is so called because when the oscillator is coupled to a tuned circuit, the grid current will show a decrease or "dip" when the oscillator is tuned through resonance with the unknown circuit. The reason for this is that the external circuit will absorb energy from the oscillator when both are tuned to the same frequency; the loss of energy from the oscillator circuit causes the feedback to decrease and this in turn is accompanied by a decrease in grid current. The dip in grid current is quite sharp when the circuit to which the oscillator is coupled has reasonably high Q.

The grid-dip meter is most useful when it covers a wide frequency range and is compactly constructed so that it can be coupled to circuits in hard-to-reach places such as in a transmitter or receiver chassis. It can thus be used to check tuning ranges and to find unwanted resonances of the type described in the chapter on TVI. Since it is its own source of r.f. energy it does not, like the absorption wavemeter, require the circuit being checked to be energized. In addition to resonance checks, the grid-dip meter also can be used as a signal source for receiver alignment and similar purposes and, as described later in this chapter, is useful in measurement of inductance and capacitance in the range of values used in r.f. circuits.

Figs. 21-20 to 21-22, inclusive, show a grid-dip meter of quite compact construction using plug-in coils to cover a continuous frequency range of 1600 kc. to 160 Mc., and thus useful in all amateur bands up through 144 Mc. as well as for checking for resonances in the low group of v.h.f. TV channels, the most important from the standpoint of harmonic TVI. It is small and light, and can be held and tuned with one hand since the dial extends slightly over the edges of the box so it can be operated with the thumb. The milliammeter is not contained in the oscillator itself but can be mounted separately in any convenient spot for viewing. Fig. 21-23 shows the milliammeter mounted in a standard meter case which also contains the power supply for the oscillator. The cable connecting the two units can be any desired length.

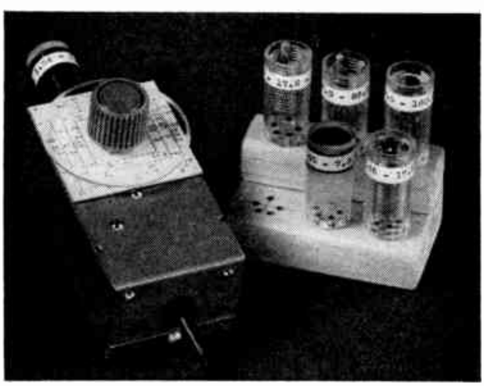


Fig. 21-20 — **A compact and light-weight grid-dip meter for one-hand operation. It is built in a 1⅝ × 2⅛ × 4-inch "Channel-lock" box and uses six plug-in coils to cover the range 1600 kc. to 160 Mc. The power supply and milliammeter for reading grid current are in a separate unit.**

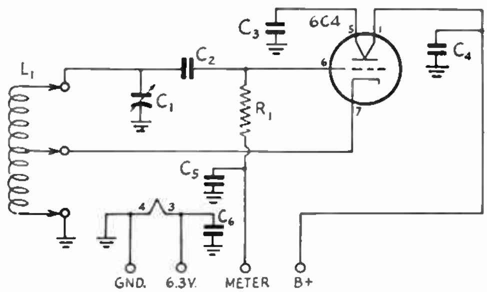


Fig. 21-21 — **Circuit diagram of the grid-dip meter.**
C_1 — 50-μμfd. midget variable (Hammarlund HF-50).
C_2 — 100-μμfd. ceramic.
C_3, C_4, C_6 — 0.001-μfd. disc ceramic.
C_5 — 0.01-μfd. disc ceramic.
R_1 — 22,000 ohms, ½ watt.

Coil Data, L_1

Freq. Range	*Turns*	*Wire*	*Diameter*	*Turns/inch*	*Tap**
1.59– 3.5 Mc.	139	32 enam.	¾ in.	Close-wound	32
3.45– 7.8 Mc.	40	32 enam.	¾ in.	Close-wound	12
7.55–17.5 Mc.	40	24 tinned	½ in.	32	14
17.2–40 Mc.	15	20 tinned	½ in.	16	5
37 –85 Mc.	4	20 tinned	½ in.	16	1⅓
78 –160 Mc.	Hairpin of No. 14 wire, ⅜ in. spacing, 2 inches long including coil form pins. Tapped 1½ in. from ground end.				

* Turns from ground end.
Coil forms are Amphenol 24-5H, ¾-in. diameter.

The oscillator circuit, shown in Fig. 21-21, is a grounded-plate Hartley, with the cathode tap adjusted for maximum sensitivity — that is, greatest change in grid current when tuning through resonance with a coupled circuit — rather than maximum grid current. For satisfactory operation at the highest frequency, the leads in the tuned circuit should be kept as short as possible, and the tuning condenser, C_1, is mounted so that its rotor and stator terminals are practically touching the corresponding pins on the coil socket. The tube socket is mounted on a bracket made from aluminum and placed at an angle so that the tube can be removed. The cathode connection between the tube socket and the coil socket is made of flat copper strip to reduce its inductance as much as possible.

Coils for the two low-frequency ranges are wound on the outsides of the forms in normal fashion, but with the exception of the highest range the remaining coils are lengths of B & W Miniductor mounted inside the forms. A hairpin-shaped coil is used for the highest range. As the coil forms are polystyrene, which softens at relatively low temperatures, care must be used in soldering to the pins. It is helpful to drill a metal plate, a few inches square and ⅟₁₆ inch or so thick,

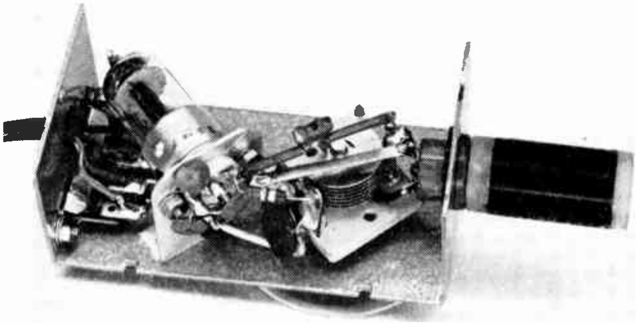


Fig. 21-22 — The grid-dip oscillator is built on the U-shaped portion of the box. C_3, C_4 and C_6 are grounded to a soldering lug at the left of the socket. Wires in the power and meter cable terminate at a 4-point terminal strip at the left.

so the coil pins will fit snugly; then if the plate is pressed firmly against the bottom of the form during soldering it will conduct the heat away from the polystyrene rapidly enough to prevent softening, if the soldering operation is not prolonged.

A transparent dial cut from a piece of ⅛-inch Plexiglas (obtainable at hobby stores) is used in preference to a solid dial so the calibration can be placed on top of the box, where there is more room for lettering. A hairline indicator is scratched on the dial, which is also provided with a standard small knob, fastened to it by small machine screws threaded in from the bottom.

The power supply shown in Fig. 21-23 uses a miniature power transformer with a selenium rectifier and a simple filter to give approximately 120 volts for the oscillator plate. The potentiometer shown in Fig. 21-24 is for adjustment of plate voltage. In any grid-dip meter the grid current will be different in different parts of the frequency range, with fixed plate voltage, so it is ordinarily necessary to choose a plate voltage that will keep the reading on scale in the part of the range where the grid current is highest. This usually results in rather low grid current at some other part of the range. With variable plate voltage this compromise is unnecessary.

The instrument may be calibrated by listening to its output with a calibrated receiver. The calibration should be as accurate as possible, although "frequency-meter accuracy" is not required in the applications for which a grid-dip meter is useful.

The grid-dip meter may be used as an indicating-type absorption wavemeter by shutting off the plate voltage and using the grid and cathode of the tube as a diode. However, this type of circuit is not as sensitive as the crystal-detector type shown earlier in this chapter, because of the high-resistance grid leak in series with the meter.

In using the grid-dip meter for checking the resonant frequency of a circuit the coupling should be kept to the point where the dip in grid current is just perceptible. This reduces interaction between the two circuits to a minimum and gives the highest accuracy. With too-close

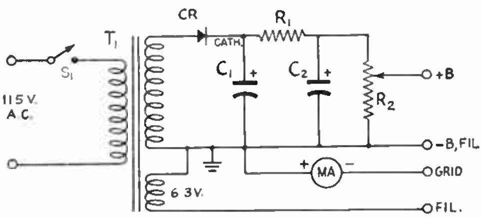


Fig. 21-24 — Circuit diagram of the power supply for the grid-dip meter.

C_1, C_2 — 16-μfd. electrolytic, 150 volts.
R_1 — 1000 ohms, ½ watt.
R_2 — 0.1-megohm potentiometer.
T_1 — Power transformer, 6.3 volts and 125 to 150 volts. (Merit P-3046 or equivalent.)
CR — 20-ma. selenium rectifier.
MA — 0-1 d.c. milliammeter.

coupling the oscillator frequency may be "pulled" by the circuit being checked, in which case different readings will be obtained when resonance is approached from the high side as compared with approaching from the low side.

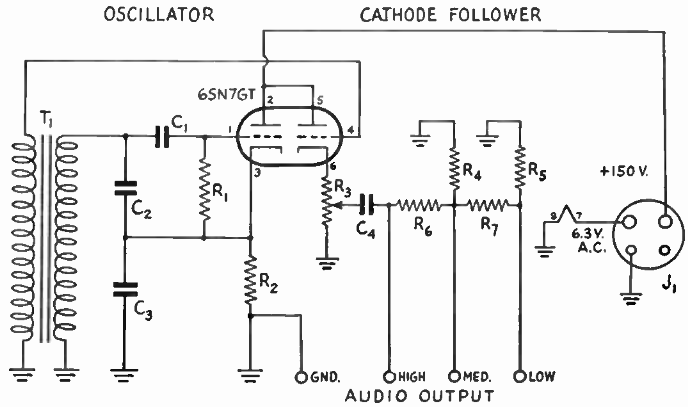
Fig. 21-23 — Power supply and milliammeter for the grid-dip meter are contained in a meter case. The control on top is for varying the plate voltage to maintain the grid current in the proper region.

AUDIO-FREQUENCY OSCILLATORS

A useful accessory for testing audio-frequency amplifiers and modulators is an audio-frequency signal generator or oscillator. Checks for distortion, gain, and the ordinary troubles that occur in such amplifiers do not require elaborate equipment; in most cases, a single audio frequency will suffice. The chief requirement is that the audio oscillator be able to generate a reasonably good sine wave.

Figs. 21-25 and 21-26 show a simple oscillator of a type entirely adequate for 'phone transmitter testing using the methods described in the chapter on amplitude modulation. It generates a fixed frequency of approximately 400

Fig. 21-25 — Circuit diagram of the simple audio oscillator.
C_1, C_4 — 0.1 μfd., 600-volt paper.
C_2 — 0.01 μfd., 600-volt paper (Sprague 6TM-S4).
C_3 — 0.03 μfd., 600-volt paper (Sprague 6TM-S3).
R_1 — 1 megohm, ½ watt.
R_2 — 10,000 ohms, ½ watt.
R_3 — 5000-ohm potentiometer
R_4, R_5 — 1700 ohms, ½ watt.
R_6, R_7 — 17,000 ohms, ½ watt.
J_1 — 4-prong chassis connector, male.
T_1 — Interstage audio transformer (Stancor A-4711).



cycles, and since it is provided with a step attenuator giving maximum outputs of approximately 1, 0.1, and 0.01 volts r.m.s., as well as continuously-variable output control, it can be used as a substitute for any type of microphone by proper choice of the high, medium, or low output.

The circuit diagram is given in Fig. 21-25. One section of a double triode is used as a Colpitts oscillator, with C_2, C_3 and the secondary winding of T_1 forming the tuned circuit. (With the transformer specified, the entire secondary winding is used.) The primary winding of T_1 is connected to the grid of the second triode section, which is used as a cathode follower. Variable output from the unit is taken from the arm of a potentiometer, R_3, connected as the cathode-follower load. The high output is taken directly from R_3, while the two lower outputs are taken from a ladder-type divider, R_4R_6 and R_5R_7. These points are brought out to tip jacks.

Molded paper condensers should be used at C_2 and C_3; cardboard-cased tubulars have been found to be unreliable in this circuit.

The power requirements are quite low — the total cathode current of the 6SN7GT is only 7.5 ma. and can be taken from any convenient source of about 150 volts. The 6SN7GT heater requires 0.6 amp. at 6.3 volts.

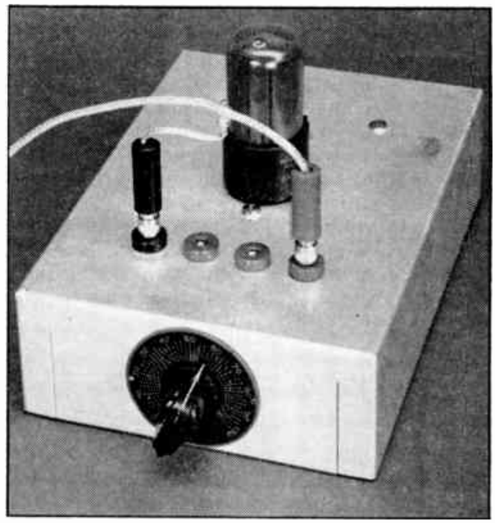


Fig. 21-26 — A simple and inexpensive audio oscillator for use in checking 'phone transmitter operation. It generates a good sine wave of fixed frequency and is provided with an attenuator so that the output level can be set at the proper value for substituting for any type of microphone.

R.F. Measurements

The measurement of fundamental quantities such as current, voltage and power at radio frequencies, and circuit elements such as inductance and capacitance, can be accomplished with equipment readily available to or easily constructed by the amateur. Measurements of this type at r.f. are equally as useful in building, testing, and operating equipment as their counterparts in d.c. circuits.

● R.F. CURRENT

R.f. current-measuring devices use a **thermocouple** in conjunction with an ordinary d.c. instrument. The thermocouple is made of two dissimilar metals which, when heated, generate a small d.c. voltage. The thermocouple is heated by a resistance wire through which the r.f. current flows, and since the d.c. voltage developed is proportional to the heating, which in turn is proportional to the power used by the heating element, the deflections of the d.c. instrument are proportional to power rather than to current. This causes the calibrated scale to be compressed at the low-current end and spread out at the high-current end. The useful range of such an instrument is about 3 or 4 to 1; that is, an r.f. ammeter having a full-scale reading of 1 ampere can be read with satisfactory accuracy down to about 0.3 ampere, one having a full scale of 5 amperes can be read down to about 1.5 amperes, and so on. No single instrument can be made to handle a wide range of currents. Neither can the r.f. ammeter be shunted satisfactorily, as can be done with d.c. instruments, because even a very small amount of reactance in the shunt will cause the readings to be highly dependent on frequency.

R.F. VOLTAGE

An **r.f. voltmeter** is a rectifier-type instrument, in which the r.f. is converted to d.c., which is then measured with a d.c. milliammeter. The best type of rectifier for most applications is a crystal diode, such as the 1N34 and similar types, because its capacitance is so low as to have little effect on the behavior of the r.f. circuit to which it is connected. The principal limitation of these rectifiers is their rather low value of safe inverse peak voltage. Vacuum-tube diodes are considerably better in this respect, but their size, shunt capacitance, and the fact that power is required for heating the cathode constitute serious disadvantages in many applications. Typical circuits for crystal-diode r.f. voltmeters are given in Fig. 21-27.

One of the principal uses for such voltmeters is as null indicators in r.f. bridges, as described later in this chapter. Another useful application is in measurement of the voltage between the conductors of a coaxial line, to show when a transmitter is adjusted for optimum output. In either case the voltmeter impedance should be high compared with that of the circuit under measurement, to avoid taking appreciable power, and the relationship between r.f. voltage and the reading of the d.c. instrument should be as linear as possible — that is, the d.c. indication should be directly proportional to the r.f. voltage at all points of the scale.

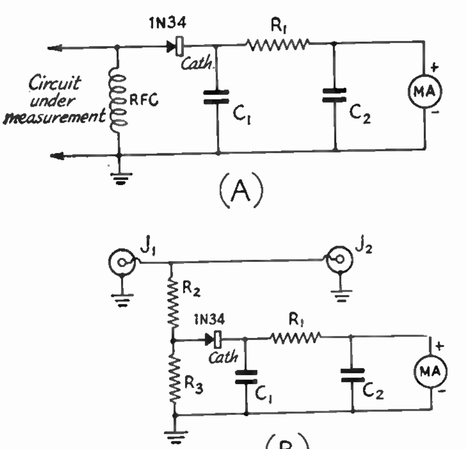


Fig. 21-27 — R.f. voltmeter circuits using a crystal rectifier and d.c. microammeter or 0–1 milliammeter. The circuit at A is suitable for measuring low voltages — up to about 35 volts maximum. B is for measuring the voltage between the conductors of a coaxial line. The total resistance of R_2 and R_3 should be of the order of 7500 ohms, with the ratio of R_2 to R_3 chosen to apply not more than 10 volts to the crystal circuit, based on the unmodulated carrier power in the line. In both circuits, R_1 should be not less than 10,000 ohms for a 0–1 milliammeter, and should be increased in proportion to the sensitivity of the meter (e.g., 20,000 ohms for a 0–500 microammeter, 100,000 ohms for a 0–100 microammeter). C_1 and C_2 should be 0.001 μfd. or more. In B, J_1 and J_2 represent coaxial connectors. The voltmeter is preferably built in a shielded box, the 2 × 4 × 4 size being large enough to contain the whole instrument.

Fig. 21-28 — R.f. ammeter mounted for connecting into a coaxial line for measuring power. A "2-inch" instrument will fit into a 2 × 4 × 4 metal box. The shunt capacitance of an ammeter mounted in this way has a negligible effect on the accuracy at frequencies as high as 30 Mc. if the instrument has a bakelite case. Metal-cased meters should be mounted on a bakelite panel which can in turn be mounted in a cut-out which clears the meter case by about ¼ inch.

All rectifiers show a variation in resistance with applied voltage, the resistance being highest when the applied voltage is small. These variations can be fairly well "swamped out" by using a high value of resistance in the d.c. circuit of the rectifier. A resistance of at least 10,000 ohms is necessary for reasonably good linearity, and higher values are beneficial. For this reason a fairly sensitive d.c. instrument should be used — if possible, a 0–100 microammeter, although a 0–1 milliammeter will serve quite well in many cases. A VTVM is ideal for the purpose since its extremely high input resistance exceeds anything that is practical with an ordinary microammeter. High resistance in the d.c. circuit also raises the impedance of the r.f. voltmeter and reduces its power consumption.

The basic voltmeter circuit is shown in Fig. 21-27A, and is simply a half-wave rectifier with a meter and a resistor, R_1, for improving the linearity. The time constant of C_1R_1 should be large compared with the period of the lowest radio frequency to be measured — a condition that can easily be met if R_1 is 10,000 ohms and C_1 is 0.001 μfd. or more — so C_1 will stay charged near the peak value of the r.f. voltage. The radio-frequency choke may be omitted if there is a low-resistance d.c. path through the circuit being measured. C_2 provides additional r.f. filtering for the d.c. circuit.

A practical arrangement for measuring the r.f. voltage in a coaxial line from a transmitter is shown at B. A voltage divider, R_2R_3, is connected across the line, the resistance values being chosen so the inverse peak voltage rating of the rectifier is not exceeded. This rating is in the vicinity of 50 volts, which limits the r.m.s. voltage that may be applied to the crystal to a maximum of 35 volts. If the approximate power carried by the line is known, the voltage can easily be calculated if the line is flat. A standing-wave ratio of 4 to 1 will cause the voltage to be twice the calculated value at a voltage loop, and 100 per cent modulation also doubles the voltage. Since it is unlikely that the s.w.r. will exceed 4 to

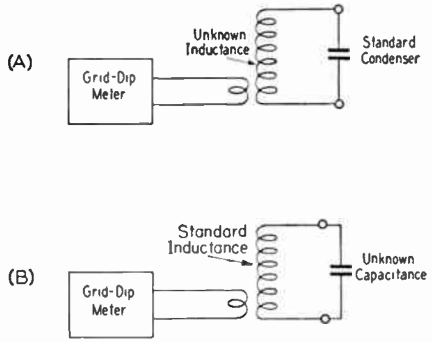


Fig. 21-29 — Set-ups for measuring inductance and capacitance with the grid-dip meter.

1 in a properly operated coax line, the safety factor will be adequate if the voltage divider is designed on the basis of applying one-fourth the rated value of voltage, or 8 to 10 volts, to the crystal. The total resistance in the divider should be about 100 times the line impedance so the power consumed by the voltmeter will not exceed 1 per cent of the power in the line. Composition resistors should be used, allowing 1 watt dissipation in R_2 (which usually dissipates practically all the voltmeter power) for each 100 watts in the line. The necessary dissipation can be built up by using resistors in series.

In constructing such a voltmeter care must be used to prevent stray coupling between the line and any part of the voltmeter, and also between the voltage divider and the crystal rectifier circuit. Also, the resistor or resistors comprising R_2 should be kept away from grounded metal in order to reduce stray capacitance.

Calibration

Calibration is not necessary for purely comparative measurements. A calibration in actual voltage requires a known resistive load and an r.f. ammeter. The set-up is the same as for r.f. power measurement as described later, and the voltage calibration is obtained by calculation from the known power and known load resistance, using Ohm's Law — $E = \sqrt{PR}$. As many points as possible should be obtained, by varying the power output of the transmitter, so that the linearity of the voltmeter can be checked.

Different voltage ranges may be secured, with a fixed voltage divider, by changing the value of R_1. It is advisable to calibrate on the lowest range and then, with a fixed value of power in the line, increase R_1 until the desired scale factor is obtained.

R.F. POWER

Measurement of r.f. power requires a resistive load of known value and either an r.f. ammeter or a calibrated r.f. voltmeter. The power is then either I^2R or E^2/R, where R is the load resistance in ohms.

The simplest method of obtaining a load of known resistance is to use an antenna system with coax-coupled matching circuit of the type described in the chapter on transmission lines. When the circuit is adjusted, by means of an s.w.r. bridge, to bring the s.w.r. down to 1 to 1 the load is resistive and of the value for which the bridge was designed (52 or 75 ohms). Fig. 21-28 shows a convenient way of mounting an r.f. ammeter for measuring current in a coaxial line. The instrument can be inserted in the line in place of the s.w.r. bridge after the matching has been completed, and the transmitter is then adjusted — without touching the matching circuit — for maximum current. The ammeter may be left in the line during regular operation if desired, but it should be kept in mind that a mismatch such as might be caused by an accident to the antenna system may result in damage to the instrument since under such conditions it is possible for the current to reach several times its normal value.

An r.f. voltmeter of the type described in the preceding section also can be used for power measurement in a similar set-up. It has the advantage that, because its scale is substantially linear, a much wider range of powers can be measured with a single instrument.

INDUCTANCE AND CAPACITANCE

The ability to measure the inductance of coils and the capacitance of condensers frequently saves time that might otherwise be spent in cut-and-try. A convenient instrument for this purpose is the grid-dip oscillator, described earlier in this chapter.

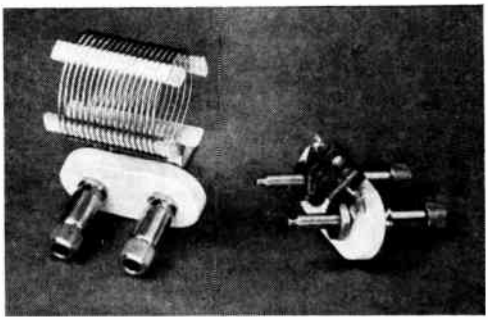


Fig. 21-30 — A convenient mounting, using binding-post plates, for L and C standards made from commercially-available parts. The condenser is a 100-μμfd. silver mica unit, mounted so the lead length is as nearly zero as possible. The inductance standard, 5 μh., is 17 turns of No. 3015 B & W Miniductor, 1-inch diameter, 16 turns per inch.

For measuring inductance, the coil is connected to a condenser of known capacitance as shown at A in Fig. 21-29. With the unknown coil connected to the standard condenser, the pick-up loop is coupled to the coil and the oscillator frequency adjusted for the grid-current dip, using the loosest coupling that gives a detectable indication. The inductance is then given by the formula

$$L_{\mu h.} = \frac{25{,}330}{C_{\mu\mu fd.}\, f_{Mc.}^{2}}$$

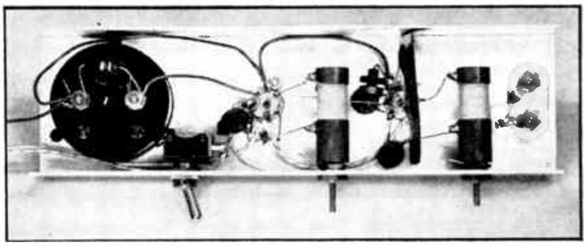


Fig. 21-34 — The logarithmic f.s. meter is constructed on a small aluminum channel. A small copper plate between the two coils is used for reducing the interstage coupling to the point where the r.f. amplifier is non-regenerative.

connecting a fairly large resistance in series with the milliammeter (or microammeter). About 10,000 ohms is required for good linearity. This considerably reduces the sensitivity of the meter, but the lower sensitivity can be compensated for by making the pick-up antenna sufficiently large.

A Sensitive Logarithmic F.S. Meter

For indicating the effect of antenna adjustments at a distant station, a logarithmic type of indicator is desirable in the field-strength meter since the meter readings with such an instrument are directly proportional to decibels. Figs. 21-32 to 21-34, inclusive, show a meter of this type. It makes use of the fact that the rectified d.c. output of a detector following a.v.c.-controlled r.f. stages tends to be logarithmic with respect to the r.f. voltage applied to the receiver.

As shown in Fig. 21-33, the circuit includes an r.f. amplifier, a detector, and a d.c. amplifier, using miniature battery tubes. The rectified r.f. voltage developed across R_1 in the diode circuit of the 1U5 is applied through the ground connection to the grid of the 1T4 r.f. amplifier and thus controls its gain. The 1½-volt "A" battery is not connected to ground but is allowed to "float," permitting the a.v.c. voltage to be effective on the grids.

In the unit shown in the photographs, slug-tuned coils are used because of their small size and because they eliminate the need for variable tuning condensers. However, ordinary condenser-tuned circuits can be substituted; the only requirement is that the circuits must be tunable to the frequency at which the antenna is being adjusted. The only critical point about the construction of such a meter is to lay out the tuned circuits so that the r.f. amplifier is stable; otherwise, any convenient layout may be used.

With the values shown in Fig. 21-33 the no-signal plate current should be very close to 0.5 milliampere. A less-sensitive d.c. instrument will require more "B" voltage. Whatever the type of meter, the current may be brought to exactly full scale, with no signal input, by shunting it with a variable resistor of suitable range, depending on the internal resistance.

Fig. 21-35 is a typical calibration curve. The readings are approximately logarithmic over about 70 per cent of the scale, with a range of about 20 db. Used with a folded-dipole pick-up antenna, the instrument is sensitive enough for use a few thousand feet away from a beam antenna fed with a few hundred watts.

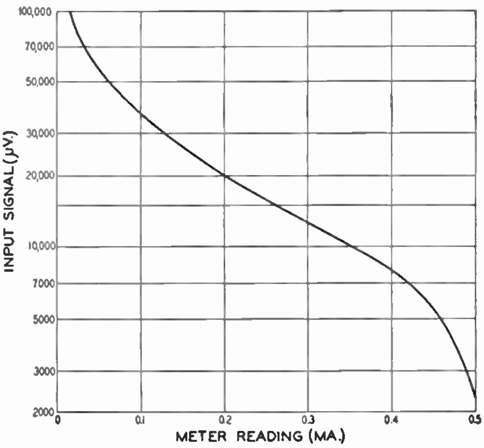


Fig. 21-35 — Typical calibration curve of the logarithmic field-strength meter. The curve is sufficiently logarithmic, for practical purposes, between about 0.05 and 0.45 ma. The way in which the readings vary with applied signal, and not the absolute value of the signal, is the important point, and since this will not change significantly so long as the same circuit is used, the curve above may be used with any similar instrument.

CHECKING STANDING WAVES

Standing waves on a transmission line can be measured if it is possible to measure the current at every point along the line, or the voltage between the two conductors at every point along the line. Rough checks on parallel-conductor lines can be made by going along the line with an absorption wavemeter having a crystal rectifier, taking care to keep the pick-up coil (or pick-up antenna) at the same distance from the line at every measurement. With such a device the maximum milliammeter reading usually will indicate current loops if a small pick-up coil is used, and voltage loops if a short pick-up antenna is used.

An alternative indicator, also useful with parallel-conductor lines, is a neon lamp. With moderate amounts of transmitter power, a low-wattage lamp will glow when the glass bulb is brought into contact with one line wire. As the lamp is moved along the line, a change in brightness indicates standing waves. If the glow is substantially the same all along the line the s.w.r. can be considered to be low enough for practical purposes.

Standing-Wave Ratio Indicators

Simple indicators such as those just mentioned are useful for checking the presence of

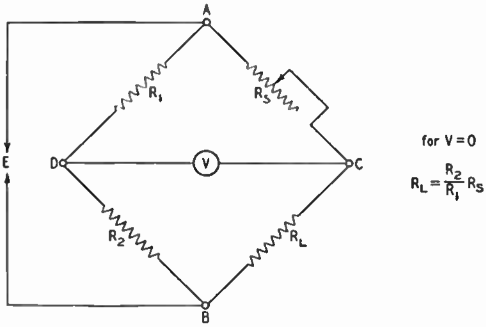


Fig. 21-36 — This fundamental bridge circuit is the basis for one type of device for measuring standing-wave ratio.

standing waves along a transmission line but are not adequate for actual measurement of the standing-wave ratio. Also, it is frequently inconvenient, and sometimes impossible, to move a current or voltage indicator along a transmission line for the distance required in checking standing waves.

An alternative method uses a bridge circuit to measure the standing-wave ratio. Fig. 21-36 will serve to illustrate the basic principles. R_1 and R_2 are fixed resistors having known values, and R_S is a calibrated variable resistor. The unknown resistance to be measured, R_L, is connected in series with R_S to form a voltage divider across the source of voltage, E. The resistance of the voltmeter, V, should be very much larger than any of the four resistance "arms" of the bridge for maximum accuracy. From Ohm's Law it is apparent that when R_1/R_2 equals R_S/R_L the voltage drops across R_1 and R_S are equal (this is also true of the voltage drops across R_2 and R_L) and there is no difference of potential between points C and D. Hence the voltmeter reading is zero ("null") and the bridge is said to be "balanced." Under any other conditions the potentials at C and D are not the same and the voltmeter reads the difference of potential.

The basis for s.w.r. measurements with a bridge is the fact that the input impedance of a properly-terminated transmission line is a pure resistance equal to the line's characteristic impedance. If a matched line is connected as the unknown arm of an appropriate bridge circuit the bridge can be balanced in the usual way and the indicating instrument will show a null. However, if the line is not properly terminated the voltage reflected back from the far end of the line will appear at the terminals of the bridge and will register on the voltmeter. The relationship between voltmeter reading (in percentage of full scale) and standing-wave ratio is shown in Fig. 21-37. This curve applies *only* when the voltmeter impedance is extremely high — 20 times or more — compared with the impedance for which the bridge is designed.

While other bridge circuits can be used for s.w.r. measurement, the resistance bridge is about the simplest and easiest to build. It lends itself well to construction for coaxial lines and when so designed can be used for measurement of open-wire lines as shown later in this chapter.

Bridge Construction

The voltmeter used in s.w.r. bridge circuits employs a crystal diode and is subject to the considerations described earlier in this chapter. In most cases, the bridge is used chiefly in the adjustment of an antenna matching system or in the adjustment of a coax-coupled matching network of the type described in the chapter on transmission lines. The object in such cases is to get the best possible match, as indicated by a null reading on the voltmeter, and not particularly to make accurate s.w.r. measurements. For this purpose the voltmeter requirements are not rigorous because it takes no current when the bridge is balanced, and a 0–1 milliammeter with a few thousand ohms resistance in series will serve very well. The circuit of Fig. 21-38 and the construction of Fig. 21-39 are quite satisfactory for a bridge intended primarily for impedance matching.

A principal point in the construction of an s.w.r. bridge is to avoid stray coupling between the resistors forming the bridge arms and between the arms and the voltmeter circuit. This can be done by keeping the resistance arms separated and at right angles to each other, and by placing the crystal and its connecting leads so that the loop so formed is not in inductive relationship with any loops formed by the bridge arms. Shielding between the bridge arms and the crystal circuit is helpful in reducing such couplings, although it is not always necessary. The two resistors forming the "ratio arms," R_1 and

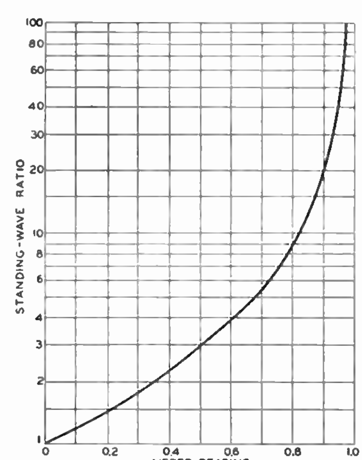


Fig. 21-37 — Standing-wave ratio in terms of meter reading (relative to full scale) after setting outgoing voltage to full scale. This graph is a plot of the formula

$$S.W.R. = \frac{V_o + V_r}{V_o - V_r}$$

where V_o and V_r are the outgoing and reflected components, respectively, of the voltage on the transmission line.

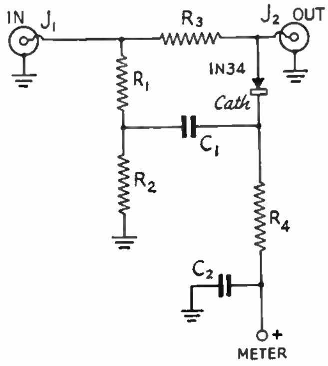


Fig. 21-38 — A simple bridge circuit useful for impedance-matching in coaxial lines.

C_1, C_2 — 0.005-μfd. disk ceramic.
R_1, R_2 — 47-ohm composition, ½ watt.
R_3 — 50- or 75-ohm (depending on line impedance) composition, ½ watt.
R_4 — 1000-ohm composition, ½ watt.
J_1, J_2 — Coaxial connector.

The meter may be a 0–1 milliammeter or d.c. voltmeter of any type having a sensitivity of 1000 ohms per volt or greater, and a full-scale range of 5 to 10 volts. Negative side of meter connects to ground.

R_2, should have identical relationships with metal parts, to keep the shunt capacitances equal, and also should have the same lead lengths so the inductances will balance. Leads should be kept as short as possible.

S.W.R. Measurement with a Bridge

For reasonably accurate measurement of s.w.r. the bridge must not only be well constructed, along the lines described above, but must have a voltmeter of very high impedance compared with the line impedance and must have provision for measuring the voltage applied to the bridge as well as the voltage developed between the arms. This is so the applied voltage can be kept constant (by regulating the transmitter output) both with and without the transmission line connected to the load terminals. If the input voltage is not maintained at a constant value the readings are unreliable. The same d.c. instrument can be used for both voltage measurements, but separate crystal rectifiers must be provided. Fig. 21-40 is the circuit of a bridge so equipped. Since the "input" voltmeter is simply used as a reference, its linearity is not important, nor does its reading have to bear any definite relationship to that of the "bridge" voltmeter, except that its range has to be at least twice that of the latter.

The resistance in the bridge voltmeter circuit should be of the order of 100 times the line impedance to avoid voltmeter errors; that is, R_4 plus the voltmeter resistance should be at least 50,000 ohms. This generally requires a sensitive d.c. instrument such as a 0–100 microammeter, a 20,000-ohms-per volt voltmeter, or, better, a VTVM.

Testing and Calibration

In a bridge intended for s.w.r. measurement rather than simple matching, the first check is to apply just enough r.f. voltage so that the bridge voltmeter reads full scale with the load terminals open. Measure the input voltage, then short-circuit the load terminals and readjust the input to the same voltage. The bridge voltmeter should again register full scale. If it does not, the ratio arms, R_1 and R_2, probably are not exactly equal. These two resistors should be carefully matched, although their actual value is not critical. This test should be made at the highest frequency to be used. If a similar test at a low frequency shows better balance, the probable cause is stray inductance or capacitance in one arm not balanced by equal strays in the other.

After the "short" and "open" readings have been equalized, the bridge should be checked for null balance with a "dummy" resistor equal to the line impedance connected to the load terminals. It is convenient to mount a half- or 1-watt resistor of the proper value in a coax connector, keeping it centered in the connector and using the minimum lead length. The bridge voltmeter should read zero at all frequencies. A reading above zero that remains constant at all frequencies indicates that the "dummy" resistor is

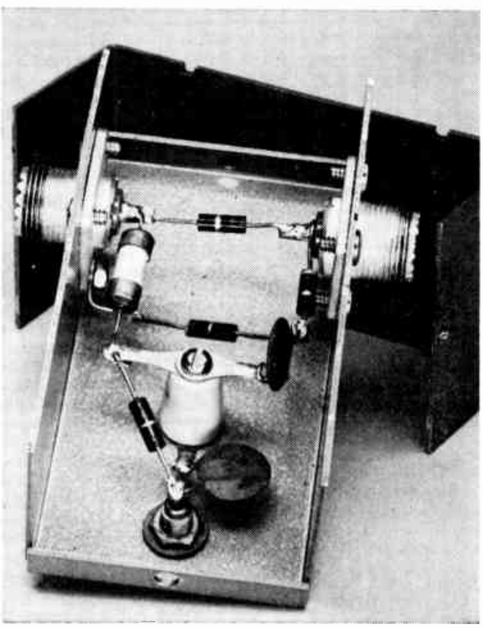


Fig. 21-39 — An inexpensive bridge for matching adjustments using the circuit of Fig. 21-38. It is built in a 1⅝ × 2⅛ × 4-inch "Channel-lock" box. The standard resistor, R_3, bridges the two coax connectors. A pin jack is provided for connection to the d.c. meter; the meter negative can be connected to the case or a coax fitting.

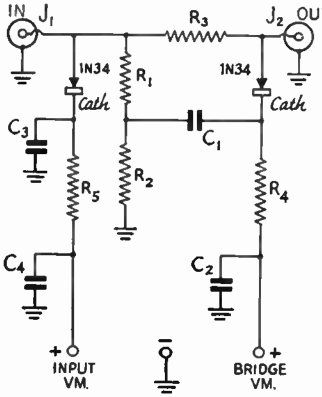


Fig. 21-40 — Bridge circuit for s.w.r. measurements. This circuit is intended for use with a d.c. voltmeter, range 5 to 10 volts, having a resistance of 10,000 ohms per volt or greater.

C_1, C_2, C_3, C_4 — 0.005-μfd. disk ceramic.
R_1, R_2 — 47-ohm composition, ½ or 1 watt.
R_3 — 50- or 75-ohm (depending on line impedance) composition, ½ or 1 watt.
R_4, R_5 — 10,000 ohms, ½ watt.
J_1, J_2 — Coaxial connectors.
Meter connects to either "input" or "bridge" position as required.

not matched to R_3, while readings that vary with frequency indicate stray reactive effects or stray coupling between parts of the bridge.

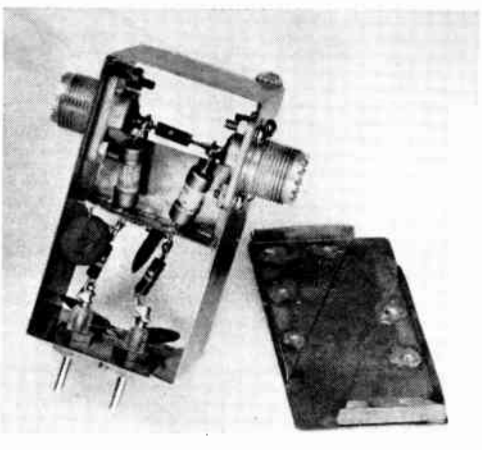
When the operation is satisfactory on the two points just described, the null should be checked with the dummy resistor connected to the bridge through several different lengths of transmission line, to ensure that R_3 actually matches the line impedance. If the null is not complete in this test both the dummy resistor and R_3 will have to be adjusted until a good match is obtained. With care, composition resistors can be filed down to raise the resistance, so it is best to start with resistors somewhat low in value. With each change in R_3, adjust the dummy resistor to give a good null when connected directly to the bridge, then try it at the end of several different lengths of line, continuing until the null is satisfactory under all conditions of line length and frequency. A discrepancy of a few per cent of the full-scale reading is tolerable.

With a high-impedance voltmeter, the s.w.r. readings will closely approximate the theoretical curve of Fig. 21-37. The calibration can be checked by using composition resistors as loads. Adjust the transmitter coupling so that the bridge voltmeter reads full scale with the output terminals open, and then check the input voltage. Connect various values of resistance across the output terminals, making sure that the input voltage is readjusted to be the same in each case, and note the reading with the meter in the bridge position. The s.w.r. is given by

$$S.W.R. = \frac{R_L}{R_0} \text{ or } \frac{R_0}{R_L}$$

where R_0 is the line impedance for which the bridge has been adjusted to null, and R_L is the resistance used as a load. Use the formula that places the larger of the two resistances in the numerator. If the readings do not correspond exactly for the same s.w.r. when appropriate resistors above and below the line impedance for which the bridge is designed are used, the current taken by the voltmeter is affecting the measurements.

Using a 0–100 microammeter, a 20,000-ohms-per-volt voltmeter on a 5-volt or higher range, or a VT voltmeter, the difference between "up" and "down" s.w.r. measurements should be negligible, provided the load resistors used for this test can be measured (at d.c.) with sufficient accuracy. Values over 1000 ohms or so should not be used at the higher frequencies.



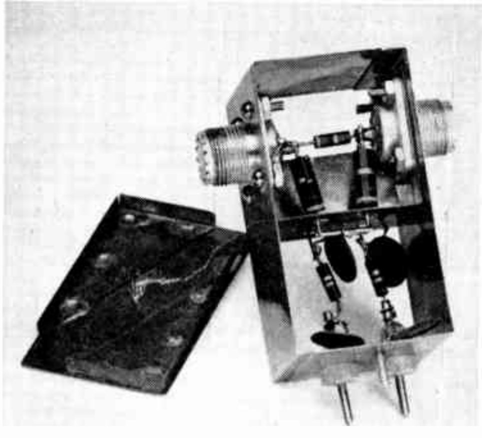


Fig. 21-41 — Top and bottom views of s.w.r. bridge using the circuit of Fig. 21-40. The box is constructed from flashing copper and measures 3 inches long, 1¾ deep and 1⅝ wide, the width being selected to be just great enough to permit connecting a 1-watt standard resistor, R_3, to the coax fittings with substantially no lead length. A small piece of copper shields the bridge arms from the crystal rectifiers. R_1 and R_2 are symmetrically placed with respect to R_3 and are at right angles to it to reduce stray coupling. The positive side of the d.c. meter connects to the feed-through bushings and the negative to the screw below them.

Using the Bridge

The procedure is the same whether the bridge is used for matching or for s.w.r. measurement. Apply power with the load terminals either open or shorted, and adjust the input until the bridge voltmeter reads full scale. Because the bridge operates a very low power level it may be necessary to couple it to a low-power driver stage rather than to the final amplifier. Alternatively, the plate voltage and excitation for the final amplifier may be reduced to the point where the power output is of the order of a few watts. Then connect the load and observe the voltmeter reading. For matching, adjust the matching network until the best possible null is obtained. For s.w.r. measurement, note the input voltage after adjusting for full-scale with the load terminals open or shorted, then connect the load and readjust the transmitter for the same input voltage. The bridge voltmeter then indicates the standing-wave ratio.

Parallel-Conductor Lines

Bridge measurements made directly on parallel-conductor lines are frequently subject to considerable error because of "antenna" currents flowing on such lines. These currents, which are either induced on the line by the field around the antenna or coupled into the line from the transmitter by stray capacitance, are in the same phase in both line wires and hence do not balance out like the true transmission-line currents. They will nevertheless actuate the bridge voltmeter, causing an indication that has no relationship to the standing-wave ratio.

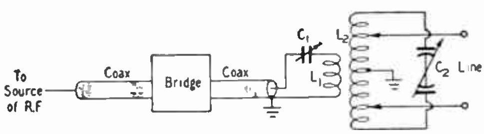


Fig. 21-42 — Circuit for using coaxial s.w.r. bridge for measurements on parallel-conductor lines. Values of circuit components are identical with those used for the similar "antenna-coupler" circuit discussed in the chapter on transmission lines.

The effect of "antenna" currents on s.w.r. measurements can be largely overcome by using a coaxial bridge and coupling it to the parallel-conductor line through a properly-designed impedance-matching circuit. A suitable circuit is given in Fig. 21-42. It closely resembles the common type of "antenna coupler," and in fact such a coupler can be used for the purpose. In the balanced tank circuit the "antenna" or parallel components on the line tend to balance out and so are not passed on to the s.w.r. bridge. It is essential that L_1 be coupled to a "cold" point on L_2 to minimize capacitive coupling, and also desirable that the center of L_2 be grounded to the chassis on which the circuit is mounted.

Values should be such that L_2C_2 can be tuned to the operating frequency and that L_1 provides sufficient coupling, as described in the transmission-line chapter. The measurement procedure is as follows:

Connect a noninductive (½- or 1-watt carbon) resistor, having the same value as the characteristic impedance of the parallel-conductor line, to the "line" terminals. Apply r.f. to the bridge, adjust the taps on L_2 (keeping them equidistant from the center), while varying the capacitance of C_1 and C_2, until the bridge shows a null. After the null is obtained, do not touch any of the circuit adjustments. Next, short-circuit the "line" terminals and adjust the r.f. input until the bridge voltmeter reads full scale. Remove the short-circuit and test resistor, and connect the regular transmission line. The bridge will then indicate the standing-wave ratio on the line.

The circuit requires rematching, with the test resistor, whenever the frequency is changed appreciably. It can, however, be used over a portion of an amateur band without readjustment, with negligible error.

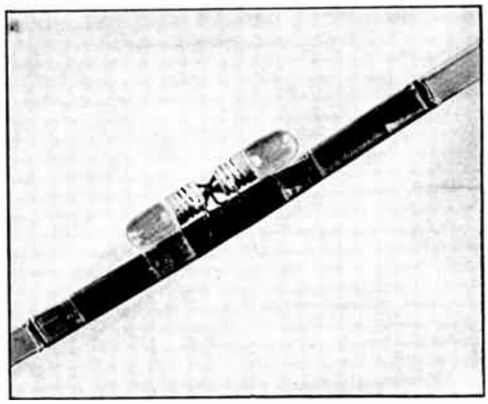


Fig. 21-43 — The "twin-lamp" standing-wave indicator mounted on 300-ohm Twin-Lead. Scotch tape is used for fastening.

The "Twin-Lamp"

A simple and inexpensive standing-wave indicator for 300-ohm line is shown in Fig. 21-43. It consists only of two flashlight lamps and a short piece of 300-ohm line. When laid flat against the line to be checked, the combination of inductive and capacitive coupling is such that outgoing power on the line causes the lamp nearest to the transmitter to light, while reflected power lights the lamp nearest the load. The power input to the line should be adjusted to make the lamp nearest the transmitter light to full brilliance. If the line is properly matched and the reflected power is very low, the lamp toward the antenna will be dark. If the s.w.r. is high, the two lamps will glow with practically equal brilliance.

The length of the piece of 300-ohm line needed in the twin-lamp will depend on the transmitter power and the operating frequency. A few inches will suffice with high power at high frequencies, while a foot or two may be needed with low power and at low frequencies.

In constructing the twin-lamp, cut one wire in the exact center of the piece and peel the ends back on either side just far enough to provide leads to the flashlight lamps. Remove about 1/4

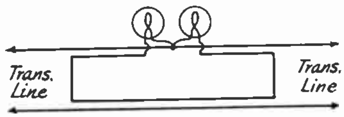


Fig. 21-44 — Wiring diagram of the "twin-lamp" standing-wave indicator.

inch of insulation from one wire of the main transmission line at some convenient point. Use the lowest-current flashlight bulbs or dial lamps available. Solder the tips of the bulbs together and connect them to the bare point in the transmission line, then solder the ends of the cut portion of the short piece to the shells of the bulbs. Figs. 21-43 and 21-44 should make the construction clear.

Installing the twin-lamp on a line introduces a discontinuity in the line impedance which causes the s.w.r. from the twin-lamp back to the transmitter to differ from the s.w.r. existing between the antenna and twin-lamp. For this reason it is desirable to remove it after s.w.r. checks have been made. It is convenient to mount the twin-lamp on a short length of line fitted to a 300-ohm plug at one end and a mating socket at the other. If similar plugs and sockets are used on the transmitter and regular transmission line, the whole test unit can be inserted and taken out at will.

The twin-lamp will respond to "antenna" currents on the transmission line in much the same way as the bridge circuits discussed earlier. There is therefore always a possibility of error in its indications, unless it has been determined by other means that "antenna" currents are inconsequential compared with the true transmission-line current.

The Oscilloscope

The **cathode-ray oscilloscope** gives a visual representation of signals at both audio and radio frequencies and can therefore be used for many types of measurements that are not possible with instruments of the types discussed earlier in this chapter. In amateur work, one of the principal uses of the 'scope is for displaying an amplitude-modulated signal so a 'phone transmitter can be adjusted for proper modulation and continuously monitored to keep the modulation percentage within proper limits. For this purpose a very simple circuit will suffice, and an oscilloscope designed expressly for this purpose is described in this section.

The versatility of the 'scope can be greatly increased by adding amplifiers and linear deflection circuits, but the design and adjustment of such circuits tends to be complicated if optimum performance is to be secured, and is somewhat outside the field of this chapter. Special components are generally required. Oscilloscope kits for home assembly are available from a number of suppliers, and since their cost compares very favorably with that of a home-built instrument of comparable design, they are recommended for serious consideration by those who have need for or are interested in the wide range of measurements that is possible with a fully-equipped 'scope.

CATHODE-RAY TUBES

The heart of the oscilloscope is the **cathode-ray tube,** a vacuum tube in which the electrons emitted from a hot cathode are first accelerated to give them considerable velocity, then formed into a beam, and finally allowed to strike a special translucent screen which *fluoresces,* or gives off light at the point where the beam strikes. A beam of moving electrons can be moved laterally, or **deflected,** by electric or magnetic fields, and since its weight and inertia are negligibly small, it can be made to follow instantly the variations in periodically-changing fields at both audio and radio frequencies.

The electrode arrangement that forms the electrons into a beam is called the **electron gun.** In the simple tube structure shown in Fig. 21-45, the gun consists of the cathode, grid, and anodes Nos. 1 and 2. The intensity of the electron beam is regulated by the grid in the same way as in an ordinary tube. Anode No. 1 is operated at a positive potential with respect

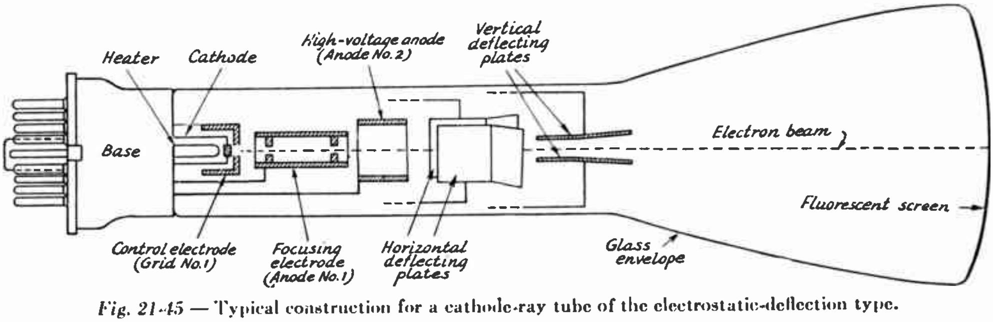


Fig. 21-45 — Typical construction for a cathode-ray tube of the electrostatic-deflection type.

to the cathode, thus accelerating the electrons that pass through the grid, and is provided with small apertures through which the electron stream passes. On emerging from the apertures the electrons are traveling in practically parallel straight-line paths. The electrostatic fields set up by the potentials on anode No. 1 and anode No. 2 form an **electron lens** system which makes the electron paths converge or focus to a point at the fluorescent screen. The potential on anode No. 2 is usually fixed, while that on anode No. 1 is varied to bring the beam into focus. Anode No. 1 is, therefore, called the **focusing electrode.**

Electrostatic deflection, the type generally used in the smaller tubes, is produced by **deflecting plates.** Two sets of plates are placed at right angles to each other, as indicated in Fig. 21-45. The fields are created by applying suitable voltages between the two plates of each pair. Usually one plate of each pair is connected to anode No. 2, to establish the polarities of the vertical and horizontal fields with respect to the beam and to each other.

Formation of Patterns

When periodically-varying voltages are applied to the two sets of deflecting plates, the path traced by the fluorescent spot forms a **pattern** that is stationary so long as the amplitude and phase relationships of the voltages remain unchanged. Fig. 21-46 shows how such patterns are formed. The horizontal sweep voltage is assumed to have the "sawtooth" waveshape indicated. With no voltage applied to the vertical plates the trace simply sweeps from left to right across the screen along the horizontal axis $X-X'$ until the instant H is reached, when it reverses direction and returns to the starting point. The sine-wave voltage applied to the vertical plates similarly would trace a line along the axis $Y-Y'$ in the absence of any deflecting voltage on the horizontal plates. However, when both voltages are present the position of the spot at any instant depends upon the voltages on both sets of plates at that instant. Thus at time B the horizontal voltage has moved the spot a short distance to the right and the vertical voltage has similarly moved it upward, so that it reaches the actual position B' on the screen. The resulting trace is easily followed from the other indicated positions, which are taken at equal time intervals.

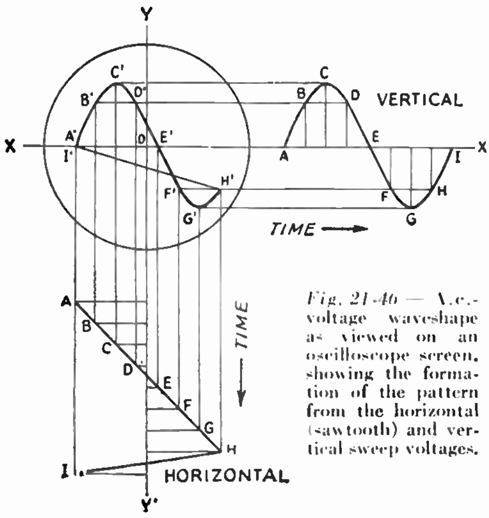


Fig. 21-46 — A.c.-voltage waveshape as viewed on an oscilloscope screen, showing the formation of the pattern from the horizontal (sawtooth) and vertical sweep voltages.

Types of Sweeps

A sawtooth sweep-voltage waveshape, such as is shown in Fig. 21-46, is called a **linear sweep,** because the deflection in the horizontal direction is directly proportional to time. If the sweep were perfect the **fly-back** time, or time taken for the spot to return from the end (H) to the beginning (I or A) of the horizontal trace, would be zero, so that the line HI would be perpendicular to the axis $Y-Y'$. Although the fly-back time cannot be made zero in practicable sweep-voltage generators it can be made quite small in comparison to the time of the desired trace AH, at least at most frequencies within the audio range. The line $H'I'$ is called the **return trace;** with a linear sweep it is less brilliant than the pattern, because the spot is moving much more rapidly during the fly-back time than during the time of the main trace.

The linear sweep shows the shape of the wave in the same way that it is usually represented graphically. If the period of the a.c. voltage applied to the vertical plates is considerably less than the time taken to sweep horizontally across the screen, several cycles of the vertical or "signal" voltage will appear in the pattern.

The shape of the pattern obtained, with a given signal waveshape on the vertical plates, obviously will depend upon the shape of the horizontal sweep voltage. If the horizontal sweep is sinusoidal, the main and return sweeps each occupy the same time and the spot moves faster horizontally in the center of the pattern than it does at the ends. When two sinusoidal voltages of the same frequency are applied to both sets of plates, the pattern may be a straight line, an ellipse, or a circle, depending upon the amplitudes and phase relationships of the two voltages.

For many amateur purposes a satisfactory horizontal sweep is simply a 60-cycle voltage of adjustable amplitude. In modulation monitoring (described in the chapter on amplitude modulation) audio-frequency voltage can be taken from the modulator to supply the horizontal sweep. For examination of audio-frequency waveforms, the linear sweep is essential. Its frequency should be adjustable over the entire range of audio frequencies to be inspected on the oscilloscope.

Lissajous Figures

When sinusoidal a.c. voltages are applied to the two sets of deflecting plates in the oscilloscope the resultant pattern depends on the relative amplitudes, frequencies and phase of the two voltages. If the ratio between the two frequencies is constant and can be expressed in intergers a stationary pattern will be produced. This makes it possible to use the oscilloscope for

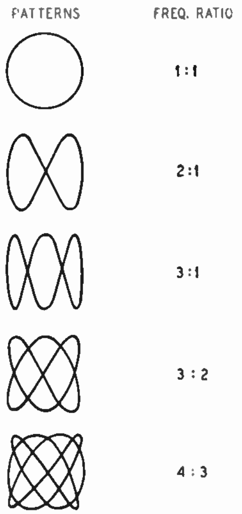


Fig. 21-47 — Lissajous figures and corresponding frequency ratios for a 90-degree phase relationship between the voltages applied to the two sets of deflecting plates.

determining an unknown frequency, provided a variable frequency standard is available, or for determining calibration points for a variable-frequency oscillator if a few known frequencies are available for comparison.

The stationary patterns obtained in this way are called **Lissajous figures.** Examples of some of the simpler Lissajous figures are given in Fig. 21-47. The frequency ratio is found by counting the number of loops along two adjacent edges. Thus in the third figure from the top there are three loops along a horizontal edge and only one along the vertical, so the ratio of the vertical frequency to the horizontal frequency is 3 to 1. Similarly, in the fifth figure from the top there are four loops along the horizontal edge and three along the vertical edge, giving a ratio of 4 to 3. Assuming that the known frequency is applied to the horizontal plates, the unknown frequency is

$$f_2 = \frac{n_2}{n_1} f_1$$

where f_1 = known frequency applied to horizontal plates,

f_2 = unknown frequency applied to vertical plates,

n_1 = number of loops along a vertical edge, and

n_2 = number of loops along a horizontal edge.

An important application of Lissajous figures is in the calibration of audio-frequency signal generators. For very low frequencies the 60-cycle power-line frequency is held accurately enough to be used as a standard in most localities. The medium audio-frequency range can be covered by comparison with the 440- and 600-cycle modulation on the WWV transmissions. An oscilloscope having both horizontal and vertical amplifiers is desirable, since it is convenient to have a means for adjusting the voltages applied to the deflection plates to secure a suitable pattern size. It is possible to calibrate over a 10-to-1 range, both upwards and downwards, from each of the latter frequencies and thus cover the audio range useful for voice communication.

A Simple Oscilloscope

Figs. 21-48 through 21-50 show the circuit and constructional details of a simple 2-inch oscilloscope suitable for the r.f. measurements described in the chapter on amplitude modulation. The compact assembly, with everything sup-

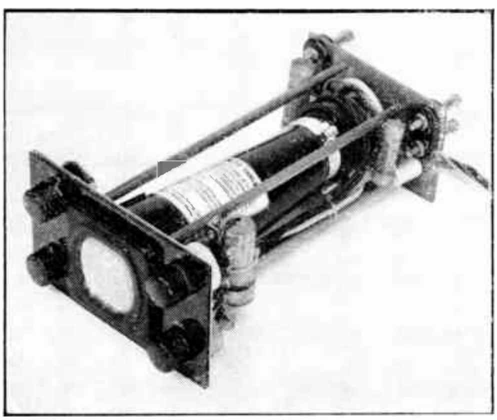


Fig. 21-48 — A 2-inch oscilloscope of compact construction, suitable for modulation measurements and monitoring. It is designed around the 2BP1 cathode-ray tube and can be mounted either in the transmitter itself or in a separate cabinet. (*Built by W1BHD and W1NUQ.*)

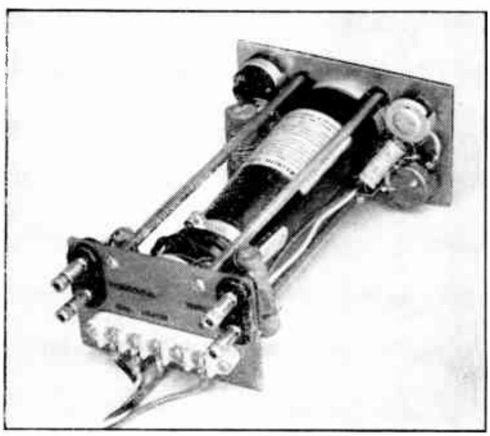


Fig. 21-49 — Rear view of the 2-inch oscilloscope. The 2BP1 is supported by the strap at the end of the shield, which clamps around the tube base. The tube socket floats, with short flexible leads running to the terminal board.

ported by the 3¼ by 5¼-inch panel, makes it possible to mount it right in a transmitter unit, if desired. In such case the heater power and high voltage for the 2BP1 tube may be taken from the transmitter power supply. The heater of the tube requires 6.3 volts at 0.6 ampere. The high voltage may be anything between 500 and 1000 volts, the maximum current being about 600 microamperes.

Fig. 21-50 is the circuit diagram of the unit. Four controls are provided, for adjusting the focus and brightness and for centering the pattern both horizontally and vertically. The horizontal and vertical signal input terminals are isolated from the c.r.t. deflection plates for d.c. by blocking condensers C_1 and C_2. These condensers should be rated to stand the maximum voltage applied

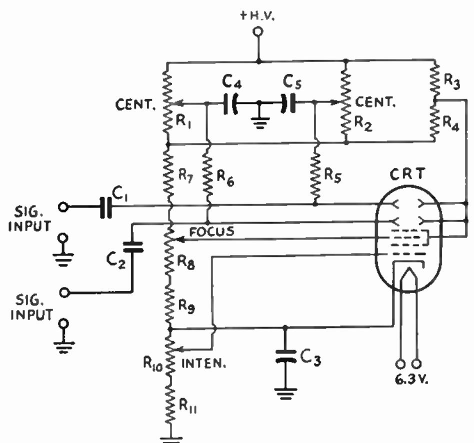


Fig. 21-50 — **Circuit diagram of the 2-inch oscilloscope. The high voltage may be between 500 and 1000 volts, according to the voltage available.**

C_1, C_2, C_4, C_5 — 0.01-μfd., 1000-volt rating.
C_3 — 0.5 μfd., 500 volts.
R_1, R_2 — 3-megohm volume control.
R_3, R_4 — 82,000 ohms, ½ watt.
R_5, R_6 — 2.2 megohms, ½ watt.
R_7 — 0.75 megohm, 1 watt.
R_8, R_{10} — 0.25-megohm volume control.
R_9 — 0.1 megohm, 1 watt.
R_{11} — 0.27 megohm, 1 watt.

to the tube plus the peak signal voltage. The signal voltage required for full deflection depends on the high voltage used, and for 500-volt operation is 65 volts per inch horizontally and 40 volts per inch vertically. At 1000 volts the corresponding figures are 130 volts per inch horizontally and 80 volts per inch vertically.

As shown in Figs. 21-48 and 21-49, the four control potentiometers are mounted in pairs each side of the c.r.t. face on the panel. Quarter-inch brass rods support a small bakelite panel at the rear. Power connections are made by means of a

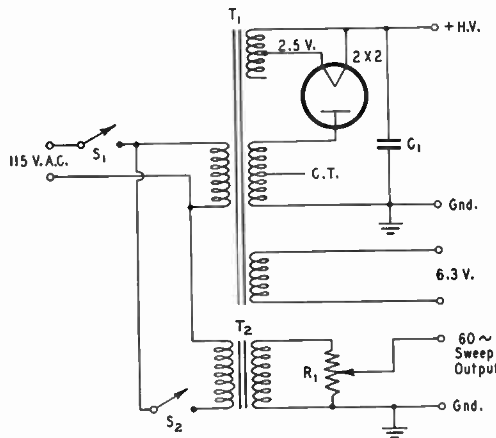


Fig. 21-51 — **Suggested power supply for the 2-inch oscilloscope if power is not supplied by the transmitter. A 60-cycle sweep circuit is included.**

C_1 — 0.25 to 1 μfd., 1000 volts.
R_1 — 0.5-megohm volume control.
S_1, S_2 — S.p.s.t. toggle.
T_1 — Small replacement transformer, 250 to 350 volts each side c.t., current rating unimportant. The 2X2 rectifier filament is supplied by one-half of the 5-volt rectifier winding. Filament secondary 6.3 volts, current required 0.6 amp.
T_2 — Audio transformer, 1 to 1 ratio suitable.

terminal strip, and double binding-post assemblies are used for the signal inputs. The brass rod supports are drilled and tapped at the ends, and at the front are assembled to the same holes that mount the bezel (Millen 80072) and the tube shield (Millen 80042). The latter is used to protect the tube from both low-frequency a.c. and r.f. fields that act on the beam and distort the pattern.

Connections and use of an oscilloscope of this type for modulation checking are described in the chapter on amplitude modulation. For the trapezoidal pattern some of the audio voltage from the modulator should be applied to the horizontal plates through a voltage divider as described in that chapter. For continuous monitoring of modulation a 60-cycle sweep can be used on the horizontal plates. The 60-cycle voltage can be obtained through a small audio transformer from the power line, as indicated in Fig. 21-51, with a potentiometer for setting it to the proper value to give a pattern of the desired size.

The unit can of course be mounted in a standard utility box or cabinet, if desired, in which event it is convenient to include a power supply. A suitable diagram is given in Fig. 21-51. Any small replacement transformer can be used since the power required is extremely small.

Assembling a Station

An amateur station is generally far better known by its signal and good operation than by its physical appearance. Good operating and a clean signal will build a reputation faster than thousands of dollars invested in special equipment and an elaborate "shack," and it is this very fact that makes amateur radio the democratic hobby that it is. However, most amateurs take pride in the arrangement of their stations, in the same way that they are careful of the appearance and arrangement of anything else that is part of the household. An antenna installation is the only external indication of the amateur station, and the degree of neatness required is generally determined by the district where the amateur lives and the attitude of the neighbors. However, with the advent of all different kinds of television receiving antennas, neighbors are in a much less favorable position to complain about the appearance of an amateur antenna system in the vicinity. TVI is something else, however!

The actual location inside the house of the "shack" — the room where the transmitter and receiver are located — depends, of course, on the free space available for amateur activities. Fortunate indeed is the amateur with a separate room that he can devote to his amateur station, or the few who can have a special small building separate from the main house. However, most amateurs must share a room with other domestic activities, and amateur stations will be found tucked away in a corner of the living room, a bedroom, a large closet, or even under the kitchen stove! A spot in the cellar or the attic can almost be classed as a separate room, although it may lack the "finish" of a normal room.

Regardless of the location of the station, however, it should be designed for maximum operating convenience and safety. It is foolish to have the station arranged so that the throwing of several switches is required to go from "receive" to "transmit," just as it is silly to have the equipment arranged so that the operator is in an uncomfortable and cramped position during his operating hours. The reasons for building the station as safe as possible are obvious, if you are interested in spending a number of years with your hobby!

CONVENIENCE

The first consideration in any amateur station is the operating position, which includes the operator's table and chair and the pieces of equipment that are in constant use (the receiver, send-receive switch, and key or microphone). The table should be as large as possible, to allow sufficient room for the receiver or receivers, frequency-measuring equipment, monitoring equipment, control switches, and keys and microphones, with enough space left over for the logbook, a pad and pencil, and perhaps a *large* ash tray. Suitable space should be included for radiogram blanks and a call book, if these accessories are in frequent use. If the table is small, or the number of pieces of equipment is large, it is often necessary to build a shelf or rack for the auxiliary equipment, or to mount it in some less convenient location in or under the table. If one has the facilities, a semicircular "console" can be built of wood, or a simpler solution is to use two small wooden cabinets to support a table top of wood or Masonite. Home-built tables or consoles can be finished in any of the available oil stains, varnishes, paints or lacquers. Many operators use a large piece of plate glass over part of their table, since it furnishes a good writing surface and can cover miscellaneous charts and tables,

This compact station is arranged for clean-cut c.w. operation, with no frills or extras. The homemade modern-style table provides adequate operating space, a cubbyhole for log and Call Book, and drawers for QSL cards and spare parts. (*W9NN, Des Plaines, Ill.*)

prefix lists, operating aids, calendar, and similar accessories.

If the major interests never require frequent band changing, or frequency changing within a band, the transmitter can be located some distance from the operator, in a location where the meters can be observed from time to time (and the color of the tube plates noted!). If frequent band or frequency changes are a part of the usual operating procedure, the transmitter should be mounted close to the operator, either along one side or above the receiver, so that the controls are easily accessible without the need for leaving the operating position.

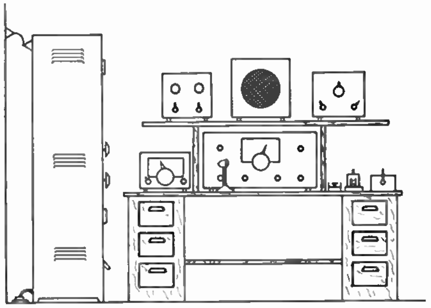


Fig. 22-1 — In a station assembled for maximum ease in frequency or band changing, the transmitter should be located next to the operating position, as shown above. On the operating table, the receiver is in front of the operator and VFO or crystal-switching oscillator on the left. (The VFO or crystal oscillator could be part of the transmitter proper, but most operators seem to prefer a separate VFO.)

The frequency standard and other auxiliary equipment can be mounted on a shelf above the receiver. The operating table can be an old desk, or a top supported by two small wooden cabinets. The "send-receive" switch is to the right of the telegraph keys — other switches are on the transmitter or the individual units.

The above arrangement can be made to look cleaner by arranging all of the equipment on the table behind a single panel or a set of panels. In this case, provision must be made for getting behind the panel for servicing the units.

A compromise arrangement would place the VFO or crystal-switched oscillator at the operating position and the transmitter in some convenient location not adjacent to the operator. Since it is usually possible to operate over a portion of a band without retuning the transmitter stages, an operating position of this type is an advantage over one in which the operator must leave his position to make a change in frequency.

Controls

The operator has an excellent chance to exercise his ingenuity in the location of the operating controls. The most important controls in the station are the receiver tuning dial and the send-receive switch. The receiver tuning dial should be located four to eight inches above the operating table, and if this requires mounting the receiver off the table, a small shelf or bracket will do the trick. With the

One of the most convenient station arrangements is to build a semicircular operating table as shown here. All operating controls are readily available, and considerably more equipment can be grouped around the operator than when an ordinary desk is used. (*W2SAI, Riverton, N. J.*)

single exception of the amateur whose work is almost entirely in traffic or rag-chew nets, which require little or no attention to the receiver, it will be found that the operator's hand is on the receiver tuning dial most of the time. If the tuning knob is too high or too low, the hand gets cramped after an extended period of operating, hence the importance of a properly-located receiver. The majority of c.w. operators tune with the left hand, preferring to leave the right hand free for copying messages and handling the key, and so the receiver should be mounted where the knob can be reached by the left hand. 'Phone operators aren't tied down this way, and tune the communications receiver with the hand that is more convenient.

The hand key should be fastened securely to the table, in a line just outside the right shoulder and far enough back from the front edge of the table so that the elbow can rest on the table. A good location for the semiauto-

In this arrangement, the two receivers (with separate loudspeakers) and the transmitter VFO are all within easy reach of the operator, while the monitoring oscilloscope on the left-hand transmitter rack can be easily seen from the operating position. (*W7JU, Boulder City, Nev.*)

matic or "bug" key is right next to the hand-key, although some operators prefer to mount the automatic key in front of them on the left, so that the right forearm rests on the table parallel to the front edge.

The best location for the microphone is directly in front of the operator, so that he doesn't have to shout across the table into it, or run up the speech-amplifier gain so high that all manner of external sounds are picked up. If the microphone is supported by a boom or by a flexible "goose neck," it can be placed in front of the operator without its base taking up valuable table space.

In any amateur station worthy of the name, it should be necessary to throw no more than one switch to go from the "receive" to the "transmit" condition. In 'phone stations, this switch should be located where it can be easily reached by the hand that isn't on the receiver. In the case of c.w. operation, this switch is most conveniently located to the right or left of the key, although some operators prefer to have it mounted on the left-hand side of the operating position and work it with the left hand while the right hand is on the key. Either location is satisfactory, of course, and the choice depends upon personal preference. Some operators use a foot-controlled switch, which is a convenience but doesn't allow too much freedom of position during long operating periods.

If the microphone is hand-held during 'phone operation, a "push-to-talk" switch on the microphone is convenient, but hand-held microphones tie up the use of one hand and

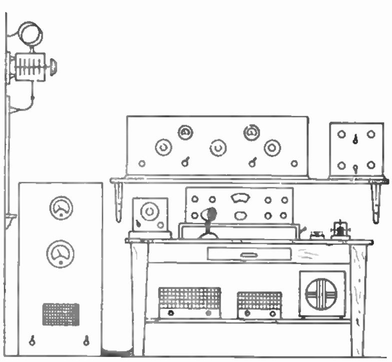
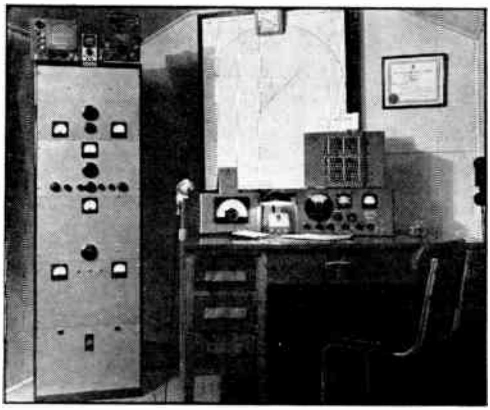


Fig. 22-2 — When little space is available for the amateur station, the equipment has to be spotted where it will fit. In the above arrangement, the transmitter, modulator and power supplies (separate units) are sandwiched in alongside the operating table and on a shelf above the table. The antenna tuning unit is mounted over the feed-through insulators that bring the antenna line into the "shack," and loudspeaker and small power supplies are mounted under the table. The operating position is clean, however, with the VFO, receiver and keys at table level. The tuning knob of this receiver would be uncomfortably low if the receiver weren't raised by the wooden arch, and the "send-receive" switch is mounted on the right-hand side of this arch, next to the hand key. Interconnecting leads should be cabled along the back of the table and table legs, to keep them inconspicuous.



This illustrates how concealing all interconnecting wires and eliminating gear not necessary to communication results in an extremely neat station. (*VE3AUJ, Woodstock, Ont.*)

are not too desirable, although they are widely used in mobile and portable work.

The location of other switches, such as those used to control power supplies, filaments, 'phone/c.w. change-over and the like, is of no particular importance, and they can be located on the unit with which they are associated. This is not strictly true in the case of the 'phone/c.w. DX man, who sometimes has need to change in a hurry from c.w. to 'phone. In this case, the change-over switch should be at the operating table, although the actual change-over should be done by a relay controlled by the switch.

If a rotary beam is used the control of the beam should be convenient to the operator. The direction indicator, however, can be located anywhere within sight of the operator, and does not have to be located on the operating table unless it is included with the control.

When several fixed beams are used, the selection of any one should be possible from the operating position, to minimize the time required to select the proper one. This generally means using a series of antenna relays or a stepping switch.

Frequency Spotting

In a station where a VFO is used, or where a number of crystals is available, the operator should be able to turn on only the oscillator of his transmitter, so that he can spot accurately his location in the band with respect to other stations. This allows him to see if he has anything like a clear channel (if such a thing exists in the amateur bands!), or to see what his frequency is with respect to another station. Such a provision can be part of the "send-receive" switch. Switches are available with a center "off" position, a "hold" position on one side, for turning on the oscillator only, and a "lock" position on the other side for turning on the transmitter and antenna relays. If oscillator keying is used, the key serves the same pur-

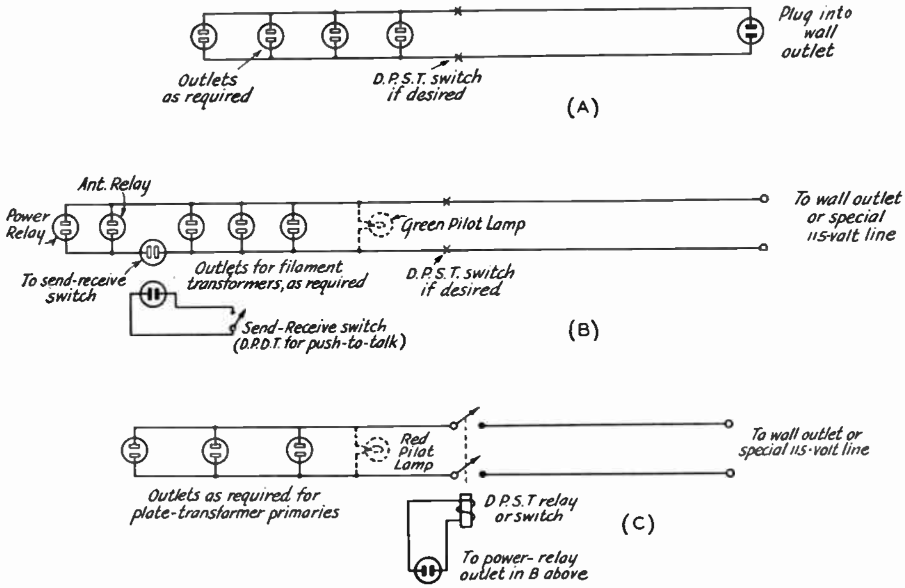


Fig. 22-3 — Power circuits for a high-power station. A shows the outlets for the receiver, monitoring equipment, speech amplifier and the like. The outlets should be mounted inconspicuously on the operating table. B shows the transmitter filament circuits and control-relay circuits, if the latter are used. C shows the plate-transformer primary circuits, controlled by the power relay. A heavy-duty switch can be used instead of the relay, in which case the antenna relay would be connected in circuit C.

If 115-volt pilot lamps are used, they can be connected as shown. Lower-voltage lamps must be connected across suitable windings on transformers.

With "push-to-talk" operation, the "send-receive" switch can be a d.p.d.t. affair, with the second pole controlling the "on-off" circuit of the receiver.

pose, provided a "send-receive" switch is available to turn off the high-voltage supplies and prevent a signal going out on the air during adjustment of the oscillator frequency.

For 'phone operation, the telegraph key or an auxiliary switch can control the transmitter oscillator, and the "send-receive" switch can then be wired into the control system so as to control the oscillator as well as the other circuits.

Comfort

Of prime importance is the comfort of the operator. If you find yourself getting tired after a short period of operating, examine your station to find what causes the fatigue. It may be that the chair is too soft or hasn't a straight back or is the wrong height for you. The key or receiver may be located so that you assume an uncomfortable position while using them. If you get sleepy fast, the ventilation may be at fault. (Or you may need sleep!)

● POWER CONNECTIONS AND CONTROL

Following a few simple rules in wiring your power supplies and control circuits will make it an easy job to change units in the station. If the station is planned in this way from the start, or if the rules are recalled when you are rebuilding, you will find it a simple matter to revise your station from time to time without a major rewiring job.

It is neater and safer to run a single pair of wires from the outlet over to the operating table or some central point, rather than to use a number of adapters at the wall outlet.

Interconnections

The wiring of any station will entail two or three common circuits, as shown in Fig. 22-3. The circuit for the receiver, monitoring equipment and the like, assuming it to be taken from a wall outlet, should be run from the wall to an inconspicuous point on the operating table, where it terminates in a multiple outlet large enough to handle the required number of plugs. A single switch between the wall outlet and the receptacle will then turn on all of this equipment at one time.

The second common circuit in the station is that supplying voltage to rectifier- and transmitter-tube filaments, bias supplies, and anything else that is not switched on and off during transmit and receive periods. The coil power for control relays should also be obtained from this circuit. The power for this circuit can come from a wall outlet or from the transmitter line, if a special one is used.

The third circuit is the one that furnishes power to the plate-supply transformers for the r.f. stages and for the modulator. (See chapter on Power Supplies for high-power considerations. When it is opened, the transmitter is disabled except for the filaments, and the transmitter should be safe to work on. However, one always feels safer when working on the transmitter if he has turned off every power supply pertaining to the transmitter.

With these three circuits established, it becomes a simple matter to arrange the station for different conditions and with new units. Anything on the operating table that runs all the time ties into the first circuit. Any new power supply or r.f. unit gets its filament power from the second circuit. Since the third circuit is controlled by the send-receive switch (or relay), any power-supply primary that is to be switched on and off for send and receive connects to circuit No. 3.

Break-In and Push-To-Talk

In c.w. operation, "break-in" is any system that allows the transmitting operator to hear the other station's signal during the "key-up" periods between characters and letters. This allows the sending station to be "broken" by the receiving station at any time, to shorten calls, ask for "fills" in messages, and speed up operation in general. With present techniques, it requires the use of a separate receiving antenna and, with high power, some means for protecting the receiver from the transmitter when the key is "down." Several methods, applicable to high-power stations, are described in Chapter Eight. If the transmitter is low-powered (50 watts or so), no special equipment is required except the separate receiving antenna and a receiver that "recovers" fast. Where break-in operation is used, there should be a switch on the operating table to turn off the plate supplies when adjusting the oscillator to a new frequency, although during all break-in work this switch will be closed.

"Push-to-talk" is an expression derived from the "push" switch on some microphones, and it means a 'phone station with a single control for all change-over functions. Strictly speaking, it should apply only to a station where this single send-receive switch must be held in place during transmission periods, but any fast-acting switch will give practically the same effect. A control switch with a center "off" position, and one "hold" and one "lock" position, will give more flexibility than a straight "push" switch. The one switch must control the antenna change-over relay, the transmitter power supplies, and the receiver "on-off" circuit. This latter is necessary to disable the receiver during transmit periods, to avoid acoustic feed-back.

Switches and Relays

It is dangerous to use an overloaded switch in the power circuits. After it has been used for some time, it may fail, leaving the power on the circuit even after the switch is thrown to the "off" position For this reason, large switches, or relays with adequate ratings, should be used to control the plate power. Relays are rated by coil voltages (for their control circuits) and by their contact current ratings.

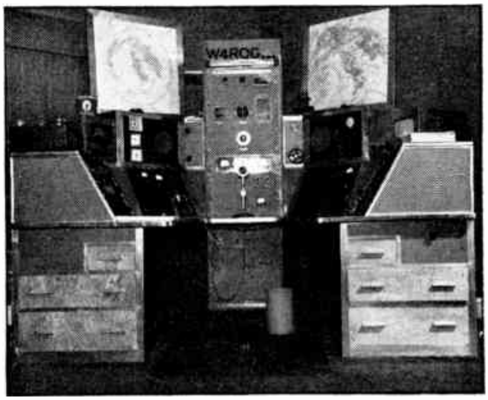
When relays are used, the send-receive switch closes the circuit to their coils, thus closing the relay contacts. The relay contacts are in the power circuit being controlled, and thus the switch handles only the relay-coil current.

SAFETY

Of prime importance in the layout of the station is the personal safety of the operator and of visitors, invited or otherwise, during normal operating practice. If there are small children in the house, every step must be taken to prevent their accidental contact with power leads of any voltage. A locked room is a fine idea, if it is possible, otherwise housing the transmitter and power supplies in metal cabinets is an excellent, although expensive, solution. Lacking a metal cabinet, a wooden cabinet or a wooden framework covered with wire screen is the next-best solution. Many stations have the power supplies housed in metal cabinets in the operating room or in a closet or basement, and this cabinet or entry is kept locked — with the key out of reach of everyone but the operator. The power leads are run through conduit to the transmitter, using ignition cable for the high-voltage leads. If the power supplies and transmitter are in the same cabinet, a lock-type main switch for the incoming line power is a good precaution.

A simple substitute for a lock-type main switch is an ordinary line plug with a short connecting wire between the two pins. By wiring a female receptacle in series with the main power line in the transmitter, the shorting plug will act as the main safety lock. When the plug is removed and hidden, it will be impossible to energize the transmitter, and a stranger or child isn't likely to spot or suspect the open receptacle.

An essential adjunct to any station is a **shorting stick** for discharging any high voltage to ground before any work or coil changing is done in the transmitter. Even if interlocks and power-supply bleeders are used, the failure of



This example of a "console" shows how it is possible to find room for a receiver and multiband kilowatt transmitter (plus power supplies and modulator), together with a wide variety of accessories including a 7-inch TV receiver, tape recorder and panoramic adapter. (*W4RQG, Winston-Salem, N. C.*)

one or more of these components may leave the transmitter in a dangerous condition. The shorting stick is made by mounting a small metal hook, of wire or rod, on one end of a dry stick or bakelite rod. A piece of ignition cable or other well-insulated wire is then run from the hook on the stick to the chassis or common ground of the transmitter, and the stick is hung alongside the transmitter. Whenever the power is turned off in the transmitter to work on the rig, or to change coils, the shorting stick is first used to touch the several high-voltage leads (tank condenser, filter condenser, tube plate connection, etc.) to insure that there is no high voltage at any of these points. This simple device has saved many a life. Use it!

Fusing

A minor hazard in the amateur station is the possibility of fire through the failure of a component. If the failure is complete and the component is large, the house fuses will generally blow. However, it is unwise and inconvenient to depend upon the house fuses to protect the lines running to the radio equipment, and every power supply should have its primary circuit individually fused, at about 150 to 200 per cent of the maximum rating of the supply. Circuit breakers can be used instead of fuses if desired.

Wiring

Control-circuit wires running between the operating position and a transmitter in another part of the room should be hidden, if possible. This can be done by running the wires under the floor or behind the base molding, bringing the wires out to terminal boxes or regular wall fixtures. Such construction, however, is generally only possible in elaborate installations, and the average amateur must content himself with trying to make the wires as inconspicuous as possible. If several pairs of leads must be run from the operating table to the transmitter, as is generally the case, a single piece of rubber- or vinyl-covered multiconductor cable will always look neater than several pieces of rubber-covered lamp cord.

The antenna wires always present a problem, unless coaxial-line feed is used. Open-wire line from the point of entry of the antenna line should always be arranged neatly, and it is generally best to support it at several points. Many operators prefer to mount their antenna-tuning assemblies right at the point of entry of the feedline, together with an antenna change-over relay (if one is used), and then the link from the tuning assembly to the transmitter can be made of inconspicuous coaxial line or Twin-Lead. If the transmitter is mounted near the point of entry of the line, it simplifies the problem of "What to do with the feeders?"

General

You can check your station arrangement by asking yourself the following questions. If all of your answers are an honest "Yes," your station will be one of which you can be proud.

1) Is your station safe, under normal operating conditions, both for the operator and the visitor?

2) Is the operating position comfortable, even after several hours of operating?

3) Do you throw not more than one switch to go from "receive" to "transmit"?

4) Does it take only a short time to explain to another amateur how to work your station?

5) Do you show your station to visiting amateurs or laymen without apologizing for its appearance?

This complete half-kilowatt station is an excellent example of how all of the equipment can be concealed inconspicuously in the living room. The cabinet is built of ¾-inch plywood finished in blonde mahogany and, with the doors closed, looks like a custom-built radio-phonograph combination. (*W9MQK, Madison, Wisc.*)

BCI and TVI

Every amateur has the obligation to make sure that the operation of his station does not, because of any shortcomings in equipment, cause interference with other radio services. It is unfortunately true that much interference is directly the fault of broadcast and TV receiver construction. Nevertheless, the amateur can and should help to alleviate interference even though the responsibility for it does not lie with him.

Successful handling of interference cases requires winning the listener's coöperation. Here are a few pointers on how to go about it.

Clean House First

The first step obviously is to make sure that the transmitter has no radiations outside the bands assigned for amateur use. The best check on this is your own AM or TV receiver. It is always convincing if you can say — and demonstrate — that you do not interfere with reception in your own home.

Don't Hide Your Identity

Whenever you make equipment changes — or shift to a hitherto unused band or type of emission — that might be expected to change the interference situation, check with your neighbors. If no one is experiencing interference, so much the better: it does no harm to keep the neighborhood aware of the fact that you are operating without bothering anyone.

Should you change location, announce your presence and conduct occasional tests on the air, requesting anyone whose reception is being spoiled to let you know about it so steps may be taken to eliminate the trouble.

Act Promptly

The average person will tolerate a limited amount of interference, but no one can be expected to put up with frequent and extended interruptions to programs. The sooner you take steps to eliminate the interference, the more agreeable the listener will be; the longer he has to wait for you, the less willing he will be to coöperate.

Present Your Story Tactfully

When you interfere, it is natural for the complainant to assume that your transmitter is at fault. If you are certain that the trouble is not caused by harmonics or other spurious emissions from your transmitter, explain to the listener that it is simply the presence of your strong signal on his receiving antenna that causes the difficulty, and that some modifications will have to be made in the receiver if he is to expect interference-free reception.

Arrange for Tests

Most listeners are not very competent observers of the various aspects of interference. If at all possible, enlist the help of another amateur and have him operate your transmitter while you see what happens at the affected receiver. You can then determine for yourself where the trouble is most likely to be.

Avoid Working on the Receiver

If your tests show that the fault has to be remedied in the receiver itself, *do not offer to work on the receiver*. It is not your fault that the receiver design is defective. Recommend that the work be done by a reliable serviceman, and offer to advise the latter as to the cause and cure if necessary.

In General

In this "public relations" phase of the problem a great deal depends on your own attitude. Most people will be willing to meet you half way, particularly when the interference is not of long standing, if you as a person make a good impression. Your personal appearance is important. So is what you say about the receiver — no one takes kindly to hearing his possessions derided. If you discuss your interference problems on the air, do it in a constructive way — one calculated to increase listener coöperation, not destroy it.

Causes and Cure of BCI

Interference with AM broadcasting usually falls into one or more rather well-defined categories. A knowledge of the general types of interference and the methods required to eliminate it will lead to a rapid appraisal of the situation and will avoid much cut-and-try in finding a cure.

Transmitter Defects

Out-of-band radiation is something that must be cured at the transmitter. Parasitic oscillations are a frequently unsuspected source of such radiations, and no transmitter can be considered satisfactory until it has been thoroughly checked for both low- and high-frequency parasitics. Very often parasitics show up only as transients, causing key clicks in c.w. transmitters and "splashes" or "burps" on modulation peaks in AM transmitters. Methods for detecting and eliminating parasitics are discussed in the transmitter chapter.

In c.w. transmitters the sharp make and break that occurs with unfiltered keying causes transients that, in theory, contain frequency components through the entire radio spectrum. Practically, they are often strong enough in the immediate vicinity of the transmitter to cause serious interference to broadcast reception. Key clicks can be eliminated by the methods detailed in the chapter on keying.

A distinction must be made between clicks generated in the transmitter itself and those set up by the mere opening and closing of the key contacts when current is flowing. The latter are of the same nature as the clicks heard in a receiver when a wall switch is thrown to turn a light on or off, and may be more troublesome nearby than the clicks that actually go out on the signal. A filter for eliminating them usually has to be installed as close as possible to the key contacts.

Overmodulation in AM 'phone transmitters generates transients similar to key clicks. It can be prevented either by using automatic systems for limiting the modulation to 100 per cent, or by continuously monitoring the modulation. Methods for both are described in the chapter on amplitude modulation. In this connection, the term "overmodulation" means any type of nonlinear modulation that results from overloading or inadequate design. This can occur even though the actual modulation percentage is less than 100.

BCI is frequently made worse by radiation from the transmitter, power wiring, or the r.f. transmission line. This is because the signal causing the interference, in such cases, is radiated from wiring that is nearer the broadcast receiver than the antenna itself. In such cases much depends on the method used to couple the transmitter to the antenna, a subject that is discussed in the chapters on transmission lines and antennas. If it is at all possible the antenna itself should be placed so that it is not in close proximity to house wiring, telephone and power lines, and similar conductors.

Image and Oscillator-Harmonic Responses

Relatively few superhet broadcast receivers have any r.f. amplification preceding the mixer, so that the selectivity at the signal frequency is not especially high. The result is that strong signals from near-by transmitters, even though the transmitting frequency is far removed from the broadcast band, can force themselves to the mixer grid. They will normally be eliminated by the i.f. selectivity, except in cases where the transmitter frequency is the image of the broadcast signal to which the receiver is tuned, or when the transmitter frequency is so related to a harmonic of the broadcast receiver's local oscillator as to produce a beat at the intermediate frequency.

These image and oscillator-harmonic responses tune in and out on the broadcast receiver dial just like a broadcast signal, except that in the case of harmonic response the tuning rate is more rapid. Since most receivers use an intermediate frequency in the neighborhood of 450 kc., the interference is a true image only when the amateur transmitting frequency is in the 1750-kc. band. Oscillator-harmonic responses occur from 3.5- and 7-Mc. transmissions, and sometimes even from higher frequencies.

The problem is to reduce the amplitude of the amateur signal in the front end of the b.c. receiver. If the receiver uses an external antenna a wavetrap at the receiver antenna terminals may help. It may also be helpful to reduce the length of the receiving antenna — and particularly to avoid a length that might be near resonance at the transmitter frequency — or to change its direction with respect to the transmitting antenna. If the signal is being picked up by the antenna it will disappear when the antenna is disconnected. If it is still present under these circumstances the pick-up is in the set wiring or the power circuits. A line filter may be tried for the latter. Pick-up on the set wiring can only be cured by installing some shielding around the r.f. circuits. Copper window screening cut and fitted to size will usually do the trick.

Since images and harmonic responses occur at definite frequencies on the receiver dial, it is always possible to choose an operating frequency that will not give such a response on top of the broadcast stations that are favored in the vicinity. While your signal may still be heard when the receiver is tuned off the local stations, it will at least not interfere with program reception.

Cross-Modulation

With 'phone transmitters, there are occasionally cases where the voice is heard whenever the broadcast receiver is tuned to a b.c. station, but there is no interference when tuning between stations. This is cross-modulation, a result of rectification in one of the early stages of the receiver. Receivers that are susceptible to this trouble usually also get a similar type of interference from regular broadcasting if there is a strong local b.c. station and the receiver is tuned to some *other* station.

The remedy for cross-modulation in the receiver is the same as for images and oscillator-harmonic responses — reduce the strength of the amateur signal at the receiver by means of a wave-trap, line filter, or shielding, as required. The trouble is not always in the receiver, however, since cross modulation can occur in any rectifying circuit — such as a poor contact in water or steam piping, gutter pipes, and other conductors in the strong field of the transmitting antenna.

Audio-Circuit Rectification

The most frequent cause of interference from operation at the higher frequencies is from rectification of a signal that by one means or another gets into the audio system of the receiver. In the milder cases an amplitude-modulated signal will be heard with reasonably good quality, but is not tunable — that is, it is present no matter what the frequency to which the receiver dial is set. An unmodulated carrier may have no observable effect in such cases beyond causing a little hum. However, if the signal is very strong there will be a reduction of the audio output level of the receiver whenever the carrier is thrown on. This causes an annoying "jumping" of the program when the interfering signal is keyed. With 'phone transmission the change in audio level is not so objectionable because it occurs at less frequent intervals. Also, ordinary rectification gives no audio output from a frequency-modulated signal, so the interference can be made almost completely unnoticeable if FM or PM is used instead of AM.

Interference of this type is most prevalent in a.c.-d.c. receivers. The pick-up may occur in the audio-circuit wiring or the interfering signal may get into the audio circuits by way of the line cord. Power-line pick-up can be treated by means of line filters, but pick-up in the receiver wiring requires individual attention. Remedies that have been found successful are described in the sections following.

CHECKING AND CURING BCI

When a case of broadcast interference comes to your attention, set a definite time to conduct tests and then prepare to do the job as expeditiously as possible. As suggested before, get another amateur to operate your transmitter while you do the actual observing and testing at the listener's receiver. If you have a small broadcast receiver of your own that does not show interference, take it with you to demonstrate to the listener that the trouble is not in your transmitter but in his receiver. The procedure outlined below will save time in getting at the source of the trouble and eliminating it.

1) Determine whether the interference is tunable or not. This will usually indicate the methods required for elimination of the trouble, as it will show which of the general types of interference discussed above is present.

2) If the set has an external antenna, disconnect it and turn the volume control up full. If the interference is no longer present, it is merely necessary to prevent the r.f. appearing on the antenna from entering the set. If wavetraps reduce the amplitude of the interfering signal but do not eliminate it entirely, try a short piece of wire as a receiving antenna. Alternatively, the antenna may be relocated. It should be placed as far as possible from the transmitting antenna, and should run at right angles to it to minimize coupling.

3) If the interference persists after the antenna is disconnected, check for r.f. on the power line by using a sensitive wavemeter such as that described in the chapter on measurements to probe along the a.c. cord that connects the set to the power source. (This test also should be made with receivers using built-in loops.) Checks should be made at the transmitter frequency, and also at harmonic frequencies. If r.f. is detected in the line, by-pass both sides of the a.c. line to ground with 0.005-μfd. ceramic condensers at the point where the line cord enters the set. (A simple plug-and-socket adapter can be made up for this purpose.) If this does not completely eliminate the interference, try a line filter designed for the operating frequency.

4) If it is evident that the interference is being picked up on the receiver wiring, explain the situation to the owner and tell him that the exact cause cannot be determined without removing the chassis from the cabinet, and that, in any event, the receiver will have to be modified if the interference is to be eliminated. Recommend that the actual work be done by a radio serviceman. Offer to check into the cause yourself, if he will allow you to take the set to your shop (with the understanding that you will not make any changes in the receiver without his express permission) so the serviceman can be told what needs to be done.

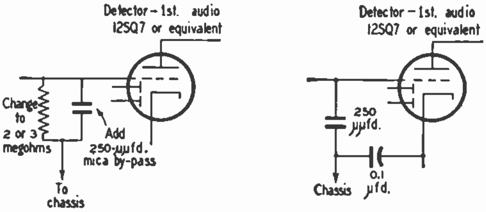


Fig. 23-1 — Two methods of eliminating r.f. from the grid of a combined detector/first-audio stage. At A, the value of the grid leak is reduced to 2 or 3 megohms, and a mica by-pass condenser is added. At B, both grid and cathode are by-passed.

5) In the event that the owner allows you to take the receiver, set it up near your transmitter and check to see if the amplitude of the interfering signal is changed by various settings of the receiver volume control. If it is, the r.f. is entering the set *ahead* of the volume control. If it is unaffected by the volume control, it is getting into the audio stages at a point following the volume control.

6) Pin the source down, if it is ahead of the volume control, by removing one tube at a time until one is found that kills the interference when it is removed. In sets using series-connected filaments, this will be possible only if a tube of equal heater rating, and with all but the heater pins clipped off, is *substituted* for the tube.

7) Determine which element (or elements) of the tube is picking up the interference by touching each tube pin with a test lead about three feet long. The lead, acting as an antenna, will cause the interference to increase when it is placed on a tube pin that is contributing to the interference. Once the sensitive points have been determined, the trouble can be eliminated by shielding the leads connected to the tube element that is affected, and by shielding the tube itself. Grid leads are the principal offenders, especially the long leads that run

from a tube cap to a tuning condenser terminal.

8) If the pick-up is found to be in the audio system — as is the case in many sets, especially when the transmitter is operating at 28 Mc. or higher — it can be eliminated by one or another of the methods shown in Figs. 23-1 and 23-2. Fig. 23-1A is a method that has proved successful with many a.c.-d.c. receivers. The value of the grid leak in the combined detector/first-audio tube (usually a 12SQ7 or its equivalent) is reduced to 2 or 3 megohms. The grid is then by-passed for r.f. with a 250-μμfd. mica condenser. Fig. 23-1B is a similar method. A third method that has worked in a.c.-d.c. receivers requires only that the heater of the detector/first-audio stage be by-passed to ground with a 0.001-μfd. condenser. The method shown in Fig. 23-2 uses a 75,000-ohm ½-watt resistor to form, with the tube capacitance, a low-pass filter. The resistor is connected between the grid pin of the tube and *all other* wires connected to the grid. In all cases, both sides of the a.c. line should be by-passed to chassis with 0.001- to 0.01-μfd. condensers.

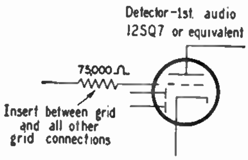


Fig. 23-2 — Using a 75,000-ohm resistor to form a low-pass filter with the tube capacitance. The resistor must be mounted at the tube pin, between the grid and all other grid connections.

Wavetraps and A.C. Line Filters

A wavetrap consists of a parallel-tuned circuit that is connected in series with the broadcast antenna and the antenna post of the receiver. It should be designed to resonate at the frequency of the interfering signal. The circuit of a simple trap is shown in Fig. 23-3. If interference results from operation in more than one amateur band several traps may be connected in series, each tuned to the center of one of the bands in which operation is contemplated. To adjust the wavetrap, have another licensed amateur operate the transmitter while you tune the trap for maximum attenuation of the interference.

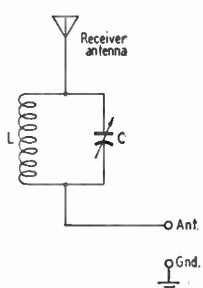


Fig. 23-3 — A simple wavetrap circuit. *L* and *C* must resonate at the frequency of the interfering signal. Suitable constants are tabulated below.

Band	*C*	*L*			
3.5	140 μμfd.	16 μh.,	32 turns	#22, 1″ diam.,	1″ long
7	100 μμfd.	6	19	#22, 1″	1″
14	50 μμfd.	3.5	14	#18, 1″	1″
21	35 μμfd.	2.2	12	#18, 1″	1″
28	25 μμfd.	1.5	9	#18, 1″	1″

A common form of a.c. line filter is shown in Fig. 23-4. This type of filter will usually do some good if the signal is being picked up on the house wiring and transferred to the set by way of the line cord. The values used for the coils and condensers are in general not critical. The effectiveness of the filter will depend considerably on the ground connection used, and it may be necessary to try grounding to several different possible ground connections to secure the best results. A filter of this type will usually not be very helpful if the signal is being picked up on the line cord itself, which may be the case when the transmitter is on v.h.f. In such a case it should be installed inside the receiver chassis and grounded to the chassis at the point where the line cord enters.

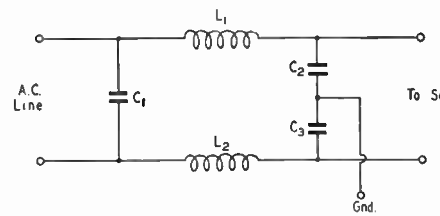


Fig. 23-4 — A.c. line filter for receivers. The values of C_1, C_2 and C_3 are not generally critical; capacitances from 0.001 to 0.01 μfd. can be used. L_1 and L_2 can be a 2-inch winding of No. 18 enameled wire on a half-inch diameter form.

The tuned filter shown in Fig. 23-5 is often more effective than the untuned type when only one frequency needs to be eliminated. After installation, the condenser is simply adjusted to reduce the interference to the greatest possible extent. It is advisable to mount either type of filter in a small shield box, to prevent pick-up in the filter and to make it less conspicuous.

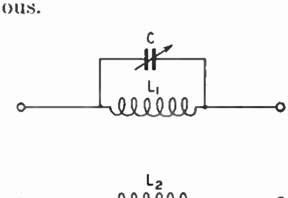


Fig. 23-5 — Resonant filter for the a.c. line. A single condenser tunes both L_1 and L_2, which are unity-coupled, one wound on top of the other. Constants for amateur bands are tabulated below.

Band	*C*	L_1 - L_2
3.5	110 – 150 (fixed)	25 t. No. 18, 1¼″ dia. × 2⅜″ long
7	110 μμfd.	18 t. No. 18, 1¼″ dia. × 2⅜″ long
14	100 μμfd.	12 t. No. 18, 1¼″ dia. × 2⅜″ long
21	50 μμfd.	10 t. No. 18, 1¼″ dia. × 2⅜″ long
28	25 μμfd.	9 t. No. 18, 1½″ dia. × 2⅜″ long

D.c.c. wire is recommended for all coils.

Interference with Television

Interference with the reception of television signals usually presents a more difficult problem than interference with AM broadcasting. In BCI cases the interference almost always can be attributed to deficient selectivity or spurious responses in the BC receiver. While similar deficiencies exist in many television receivers, it is also true that amateur transmitters generate harmonics that fall inside many or all television channels. These spurious radiations cause interference that ordinarily cannot be eliminated by anything that may be done at the receiver, so must be prevented at the transmitter itself.

The over-all situation is further complicated by the fact that television broadcasting is in three distinct bands, two in the v.h.f. region and one in the u.h.f.

V.H.F. Television

For the amateur who does most of his transmitting on frequencies below 30 Mc. the TV band of principal interest is the low v.h.f. band between 54 and 88 Mc. If harmonic radiation can be reduced to the point where no interference is caused to Channels 2 to 6, inclusive, it is almost certain that any harmonic troubles with channels above 174 Mc. will disappear also.

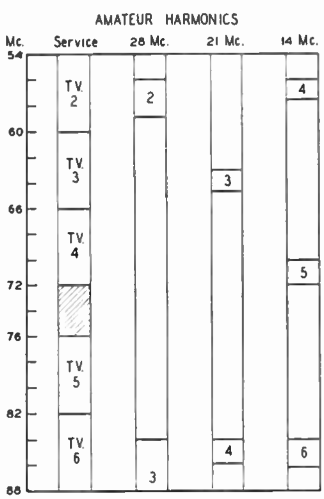
The relationship between the v.h.f. television channels and harmonics of amateur bands from 14 through 28 Mc. is shown in Fig. 23-6. Harmonics of the 7- and 3.5-Mc. bands are not shown because they fall in every television channel. However, the harmonics above 54 Mc. from these bands are of such high order that they are usually rather low in amplitude, although they may be strong enough to interfere if the television receiver is quite close to the amateur transmitter. Low-order harmonics — up to about the sixth — are usually the most difficult to eliminate.

Of the amateur v.h.f. bands, only 50 Mc. will have harmonics falling in a v.h.f. television channel (Channels 11, 12 and 13). However, a transmitter for any amateur v.h.f. band may cause interference if it has multiplier stages either tuned to or having harmonics in one or more of the v.h.f. TV channels. The r.f. energy on such frequencies can be radiated directly from the transmitting circuits or coupled by stray means to the transmitting antenna.

Frequency Effects

The degree to which transmitter harmonics or other undesired radiation actually in the TV channel must be suppressed depends principally on two factors, the strength of the TV signal on the channel or channels affected, and the relationship between the frequency of the spurious radiation and the frequencies of the TV picture and sound carriers within the channel. If the TV signal is very strong, interference can be eliminated by comparatively simple methods. However, if the TV signal is very weak, as in "fringe" areas where the received picture is visibly degraded by the appearance of set noise or "snow" on the screen, it may be necessary to go to extreme measures.

In either case the intensity of the interference depends very greatly on the exact frequency of the interfering signal. Fig. 23-7 shows the placement of the picture and sound carriers in the standard TV channel. In Channel 2, for example, the picture carrier frequency is 54 + 1.25 = 55.25 Mc. and the sound carrier frequency is 60 − 0.25 = 59.75 Mc. The second harmonic of 28,010 kc. (56,020 kc. or 56.02 Mc.) falls 56.02 − 54 = 2.02 Mc. above the low edge of the channel and is in the region marked "Severe" in Fig. 23-7. On the other hand, the second harmonic of 29,500 kc. (59,000 kc. or 59 Mc.) is 59 − 54 = 5 Mc. from the low edge of the channel and falls in the region marked "Mild." Interference at this frequency has to be about 100 times as strong as at 56,020 kc. to cause effects of equal intensity.



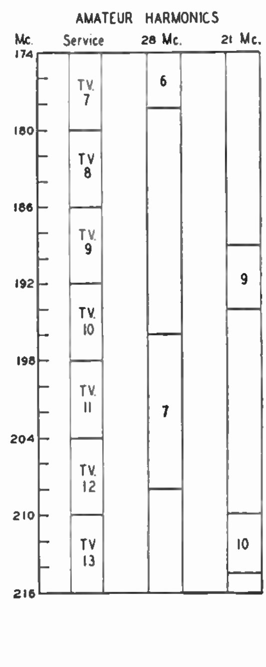


Fig. 23-6 — **Relationship of amateur-band harmonics to v.h.f. TV channels. Harmonic interference from transmitters operating below 30 Mc. is most likely to be serious in the low-channel group (54 to 88 Mc.).**

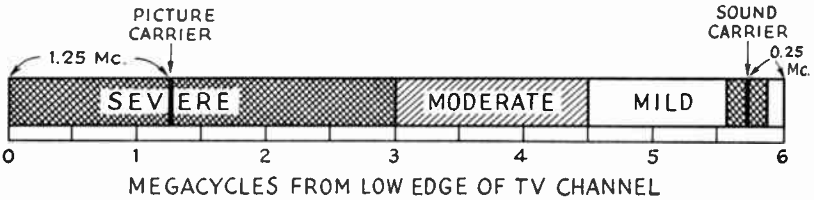


Fig. 23-7 — Location of picture and sound carriers in a monochrome television channel, and relative intensity of interference as the location of the interfering signal within the channel is varied without changing its strength. The three regions are not actually sharply defined as shown in this drawing, but merge into one another gradually.

Thus an operating frequency that puts a harmonic near the picture carrier requires about 40 db. more harmonic suppression in order to avoid interference, as compared with an operating frequency that puts the harmonic near the upper edge of the channel.

For a region of 100 kc. or so either side of the sound carrier there is another "Severe" region where a spurious radiation will interfere with reception of the sound program, and this region also should be avoided. In general, a signal of intensity equal to that of the picture carrier will not cause noticeable interference if its frequency is in the "Mild" region shown in Fig. 23-7, but the same intensity in the "Severe" region will utterly destroy the picture.

Interference Patterns

The visible effects of interference vary with the type and intensity of interference. Complete "blackout," where the picture and sound disappear completely, leaving the screen dark, occurs only when the transmitter and receiver are quite close together. Strong interference ordinarily causes the picture to be broken up, leaving a jumble of light and dark lines, or turns the picture "negative" — the normally white parts of the picture turn black and the normally black parts turn white. "Cross-hatching" — diagonal bars or lines in the picture — accompanies the latter, usually, and also represents the most common type of less-severe interference. The bars are the result of the beat between the harmonic frequency and the picture carrier frequency. They are broad and relatively few in number if the beat frequency is comparatively low — near the picture carrier — and are numerous and very fine if the beat frequency is very high — toward the upper end of the channel. Typical cross-hatching is shown in Fig. 23-8. If the frequency falls in the "Mild" region in Fig. 23-7 the cross-hatching may be so fine as to be visible only on close inspection of the picture, in which case it may simply cause the apparent brightness of the screen to change when the transmitter carrier is thrown on and off.

Fig. 23-8 — "Cross-hatching," caused by the beat between the picture carrier and an interfering signal inside the TV channel.

Whether or not cross-hatching is visible, an amplitude-modulated transmitter may cause "sound bars" in the picture. These look about as shown in Fig. 23-9. They result from the variations in the intensity of the interfering signal when modulated. Under most circumstances modulation bars will not occur if the amateur transmitter is frequency- or phase-modulated. With these types of modulation the cross-hatching will "wiggle" from side to side with the modulation.

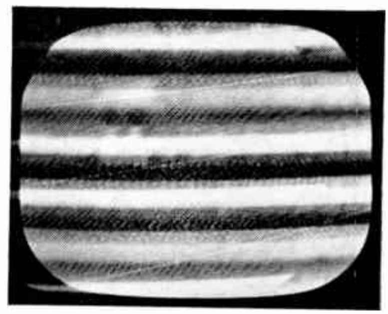


Fig. 23-9 — "Sound bars" or "modulation bars" accompanying amplitude modulation of an interfering signal. In this case the interfering carrier is strong enough to destroy the picture, but in mild cases the picture is visible through the horizontal bars. Sound bars may accompany modulation even though the unmodulated carrier gives no visible cross-hatching.

Except in the more severe cases, there is seldom any effect on the sound reception when interference shows in the picture, unless the frequency is quite close to the sound carrier. In the latter event the sound may be interfered with even though the picture is clean.

Reference to Fig. 23-6 will show whether or not harmonics of the frequency in use will fall in any television channels that can be received in the locality. It should be kept in mind that not only harmonics of the final frequency may interfere, but also harmonics of any frequencies that may be present in buffer or frequency-multiplier stages. In the case of 144-Mc. transmitters, frequency-multiplying combinations that require a doubler or tripler stage to operate on a frequency actually in a low-band v.h.f. channel in use in the locality should be avoided.

Harmonic Suppression

Effective harmonic suppression has three separate phases:

1) Reducing the amplitude of harmonics generated in the transmitter. This is a matter of circuit design and operating conditions.

2) Preventing stray radiation from the transmitter and from associated wiring. This requires adequate shielding and filtering of all circuits and leads from which radiation can take place.

3) Preventing harmonics from being fed into the antenna.

It is impossible to build a transmitter that will not generate *some* harmonics, but it is obviously advantageous to reduce their strength, by circuit design and choice of operating conditions, by as large a factor as possible before attempting to prevent them from being radiated. Harmonic radiation from the transmitter itself or from its associated wiring obviously will cause interference just as readily as radiation from the antenna, so measures taken to prevent harmonics from reaching the antenna will not reduce TVI if the transmitter itself is radiating harmonics. But once it has been found that the transmitter itself is free from harmonic radiation, devices for preventing harmonics from reaching the antenna can be expected to produce results.

REDUCING HARMONIC GENERATION

Since reasonably-efficient operation of r.f. power amplifiers always is accompanied by harmonic generation, good judgment calls for operating all frequency-multiplier stages at a very low power level — plate voltages not exceeding 250 or 300. When the final output frequency is reached, it is desirable to use as few stages as possible in reaching the output power level, and to use tubes that require a minimum of driving power.

Circuit Design and Layout

Harmonic currents of considerable amplitude flow in both the grid and plate circuits of r.f. power amplifiers, but they will do relatively little harm if they can be effectively by-passed to the cathode of the tube. Fig. 23-10A shows the paths followed by harmonic currents in an amplifier circuit; because of the high reactance of the tank coil there is little harmonic current in it, so the harmonic currents simply flow through the tank condenser, the plate (or grid) blocking condenser, and the tube capacitances. The lengths of the leads forming these paths is of great importance, since the inductance in this circuit will resonate with the tube capacitance at some frequency in the v.h.f. range (the tank and blocking capacitances usually are so large compared with the tube capacitance that they have little effect on the resonant frequency). If such a resonance happens to occur at or near the same frequency as one of the transmitter harmonics, the effect is just the same as though a harmonic tank circuit

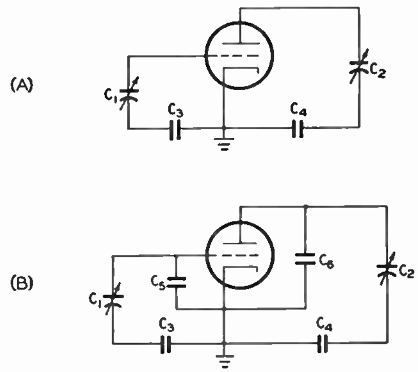


Fig. 23-10 — (A) A v.h.f. resonant circuit is formed by the tube capacitance and the leads through the tank and blocking condensers. Regular tank coils are not shown, since they have little effect on such resonances. (B) Using low-inductance condensers shunting the tube elements to lower the resonance point below the TV channels. C_5 and C_6 usually are 15 to 50 μμfd. and either of vacuum or tubular construction.

had been deliberately introduced; the harmonic at that frequency will be tremendously increased in amplitude.

Such resonances are unavoidable, but by keeping the path from plate to cathode and from grid to cathode as short as is physically possible, the resonant frequency usually can be raised above 100 Mc. in amplifiers of medium power. This puts it between the two groups of television channels.

In low-frequency transmitters where physically-short return paths from plate or grid to cathode are difficult because of the shape and size of tubes and tank condensers, the arrangement shown in Fig. 23-10B is frequently helpful. Condensers C_5 and C_6 should be of the vacuum or tubular type and should be mounted as close as possible to the tube connections. They form resonant circuits in themselves with the tube capacitance, but generally at a sufficiently high frequency so that no harm is done. At lower frequencies than this self-resonance, they effectively add to the tube capacitance and thus tune the inductance of the leads through the regular tank and blocking condensers to a considerably lower frequency than the tube alone. The resonance therefore can be shifted to a frequency *below* 54 Mc. and again is outside the TV range. This method is most useful at 3.5 and 7 Mc. because it increases the tank capacitance to the point where there may be very little tank coil left, at the higher frequencies.

It is easier to place grid-circuit v.h.f. resonances where they will do no harm when the amplifier is link-coupled to the driver stage, since this generally permits shorter leads and more favorable conditions for by-passing the harmonics than is the case with capacitive coupling. Link coupling also reduces the coupling between the driver and amplifier at harmonic frequencies, thus preventing driver harmonics from being amplified.

The inductance of leads from the tube to the tank condenser can be reduced not only by shortening but by using flat strip instead of wire conductors. It is also better to use the chassis as the return from the blocking condenser to cathode, since a chassis path will have less inductance than almost any other form of connection.

The v.h.f. resonance points in amplifier tank circuits can be found by coupling a grid-dip meter covering the 50–250 Mc. range to the grid and plate leads. If a resonance is found in or near a TV channel, methods such as those described above should be used to move it well out of the TV range. The grid-dip meter also should be used to check for v.h.f. resonances in the tank coils, because coils made for 14 Mc. and below usually will show such resonances. In making the check, disconnect the coil entirely from the transmitter and move the grid-dip meter coil along it while exploring for a dip in the 54–88 Mc. band. If a resonance falls in a TV channel that is in use in the locality, changing the number of turns will move it to a frequency where it will not be troublesome.

In many r.f. amplifiers the cathode connection of the tube is below chassis while the plate (and sometimes the grid) connection frequently is above. In such a case the blocking condenser should be mounted *below* chassis. If the ground return is made to the top, the r.f. current has to flow over the top and either through the hole for the tube socket or else entirely over the chassis surface before it reaches the cathode. This condition is highly undesirable not only because of v.h.f. resonances but because such chassis currents frequently cause instability in the amplifier.

Operating Conditions

Grid bias and grid current have an important effect on the harmonic content of the r.f. currents in both the grid and plate circuits. In general, harmonic output increases as the grid bias and grid current are increased, but this is not necessarily true of a *particular* harmonic. The third and higher harmonics, especially, will go through fluctuations in amplitude as the grid current is increased, and sometimes a rather high value of grid current will minimize one harmonic as compared with a low value of grid current. This characteristic can be used to advantage where a particular harmonic is causing interference, keeping in mind that the operating conditions that minimize one harmonic may greatly increase another.

For equal operating conditions, there is little or no difference between single-ended and push-pull amplifiers in respect to harmonic generation. Push-pull amplifiers are frequently trouble-makers on even harmonics because with such amplifiers the even-harmonic voltages are in phase at the ends of the tank circuit and hence appear with equal amplitude across the whole tank coil, if the center of the coil is not grounded. Under such circumstances the even harmonics can be coupled to the output circuit through stray capacitance between the tank and coupling coils. This does not occur in a single-ended amplifier if the coupling coil is placed at the cold end of the tank.

Harmonic Traps

If a harmonic in only one TV channel is particularly bothersome — frequently the case when the transmitter operates on 28 Mc. — a trap tuned to the harmonic frequency may be installed in the plate lead as shown in Fig. 23-11. At the harmonic frequency the trap represents a very high impedance and hence reduces the amplitude of the harmonic current flowing through the tank circuit. In the push-pull circuit both traps have the same constants. The *L/C* ratio is not critical but a high-*C* circuit usually will have least effect on the performance of the plate circuit at the normal operating frequency.

Since there is a considerable harmonic voltage across the trap, it may radiate unless the transmitter is well shielded. Traps should be placed so that there is no coupling between them and the amplifier tank circuit.

A trap is a highly-selective device and so is useful only over a small range of frequencies. A second- or third-harmonic trap on a 28-Mc. tank circuit usually will not be effective over more than 50 kc. or so at the fundamental frequency, depending on how serious the interference is without the trap. Because they are critical of adjust-

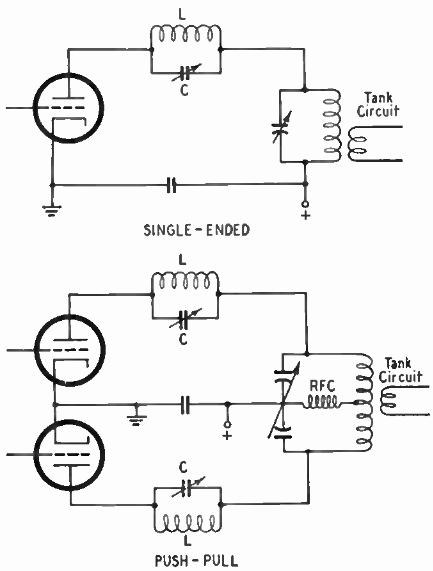


Fig. 23-11 — Harmonic traps in an amplifier plate circuit. *L* and *C* should resonate at the frequency of the harmonic to be suppressed. *C* may be a 25- to 50-μμfd. midget, and *L* usually consists of 3 to 6 turns about ½ inch in diameter for Channels 2 through 6. The inductance should be adjusted so that the trap resonates at about half capacity of *C* before being installed in the transmitter. It may be checked with a grid-dip meter. When in place, it is adjusted for minimum interference to the TV picture.

ment, it is better to prevent TVI by other means, if possible, and use traps only as a last resort.

PREVENTING RADIATION FROM THE TRANSMITTER

The extent to which interference will be caused by direct radiation of spurious signals depends on the operating frequency, the transmitter power level, the strength of the television signal, and the distance between the transmitter and TV receiver. Transmitter radiation can be a very serious problem if the TV signal is weak, if the TV receiver and amateur transmitter are close together, and if the transmitter is operated with high power.

Shielding

Direct radiation from the transmitter circuits and components can be prevented by proper shielding. To be effective, a shield must completely enclose the circuits and parts and must have no openings that will permit r.f. energy to escape. Unfortunately, ordinary metal boxes and cabinets do not provide good shielding, since such openings as louvers, lids, holes for running in connections, and so on, allow far too much leakage.

A primary requisite for good shielding is that all joints must make a good electrical connection along their entire length. A small slit or crack will let out a surprising amount of r.f. energy; so will ventilating louvers and large holes such as those used for mounting meters. On the other hand, small holes do not impair the shielding very greatly, and a limited number of ventilating holes may be used if they are small — not over ¼ inch in diameter. Also, wire screen makes quite effective shielding if the wires make good electrical connection where they cross over, so the leakage through large openings can be very much reduced by covering such openings with screening, well bonded to all edges of the opening.

The intensity of r.f. fields about coils, condensers, tubes and wiring decreases very rapidly with distance, so shielding is more effective, from a practical standpoint, if the components and wiring are not too close to it. It is advisable to have a separation of several inches, if possible, between "hot" points in the circuit and the nearest shielding.

For a given thickness of metal, the greater the conductivity the better the shielding. Copper is best, with aluminum, brass and steel following in that order. However, if the thickness is adequate for structural purposes (over 0.02 inch) and the shield and a "hot" point in the circuit are not in close proximity, any of these metals will be satisfactory. Greater separation should be used with steel shielding than with the other materials not only because it is considerably poorer as a shield but also because it will cause greater losses in near-by circuits than would copper or aluminum at the same distance. Wire screen used as a shield should also be kept at some distance from high-voltage or high-current r.f. points, since there is considerably more leakage through the mesh than through solid metal.

Where two pieces of metal join, as in forming a corner, they should overlap at least a half inch and be fastened together firmly with screws or bolts spaced at close-enough intervals to maintain firm contact all along the joint. The contact surfaces should be clean before joining, and should be checked occasionally — especially steel, which is almost certain to rust after a period of time.

The leakage through a given size of aperture in shielding increases with frequency, so such points as good continuous contact, screening of holes, and so on, become even more important when the radiation to be suppressed is in the high band — 174–216 Mc. — than in the low TV band. Hence 50- and 144-Mc. transmitters, which in general will have frequency-multiplier harmonics of relatively high intensity in this region, require special

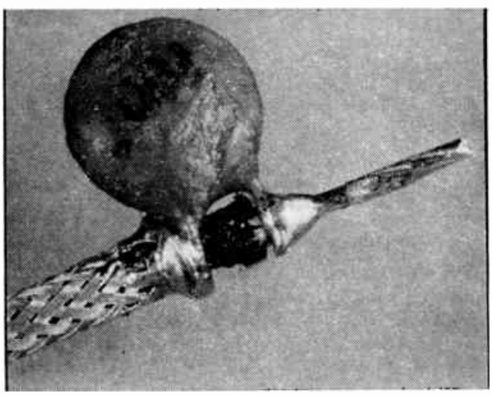


Fig. 23-12 — Proper method of by-passing the end of a shielded lead using disk ceramic condenser. The 0.001 μfd. size should be used for 1600 volts or less; 500 μμfd. at higher voltages. The leads are wrapped around the inner and outer conductors and soldered, so that the lead length is negligible. This photograph is about four times actual size.

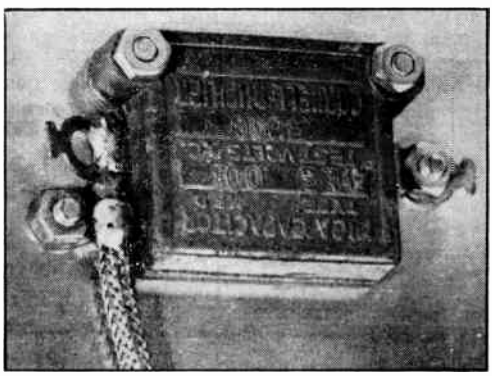


Fig. 23-13 — By-passing the end of a high-voltage lead. The end of the shield braid is soldered to a lug fastened to the chassis directly underneath. The other terminal of the condenser is similarly bolted directly to the chassis. When the by-pass is used at a terminal connection block the "hot" lead should be soldered directly to the terminal, if possible, but in any event connected to it by a very short lead.

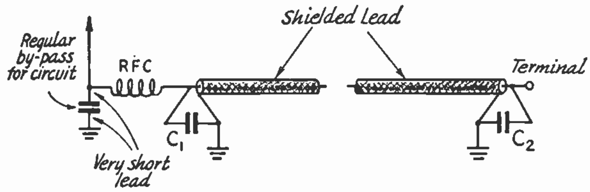


Fig. 23-14 — Additional r.f. filtering of supply leads may be required in regions where the TV signal is very weak. The r.f. choke should be physically small, and may consist of a 1-inch winding of No. 26 enameled wire on a ¼-inch form, close-wound. Manufactured single-layer chokes having an inductance of a few microhenrys also may be used.

attention in this respect if the possibility of interfering with a channel received locally exists.

Lead Treatment

Even very good shielding can be made completely useless when connections are run from external power supplies and other equipment to the circuits inside the shield. Every conductor so introduced into the shielding forms a path for the escape of r.f., which is then radiated by the connecting wires. Hence a step that is essential in every case is to prevent harmonic currents from flowing on the leads leaving the shielded enclosure.

Harmonic currents always flow on the d.c. or a.c. leads connecting to the tube circuits. A very effective means of preventing such currents from being coupled into other wiring, and one that provides desirable by-passing as well, is to use shielded wire for all such leads, maintaining the shielding from the point where the lead connects to the tube or r.f. circuit right through to the point where it is about to leave the chassis. The shield braid should be grounded to the chassis at both ends and at frequent intervals along the path.

Good by-passing of shielded leads also is essential. Bearing in mind that the shield braid about the conductor confines the harmonic currents to the *inside* of the shielded wire, the object of by-passing is to prevent their escape. Figs. 23-12 and 23-13 show the proper way to by-pass. The small-type 0.001-μfd. ceramic disk condenser, when mounted on the end of the shielded wire as shown in Fig. 23-12, actually forms a series-resonant circuit in the 54–88-Mc. range and thus represents practically a short-circuit for low-band TV harmonics. The exposed wire to the connection terminal should be kept as short as is physically possible, to prevent any possible harmonic pick-up exterior to the shielded wiring. Disk condensers of this capacitance are available in several voltage ratings up to 1600 volts. For higher voltages, the maximum capacitance available is approximately 500 μμfd., which is large enough for good by-passing of harmonics. Alternatively, mica condensers may be used as shown in Fig. 23-13, mounting the condenser flat against the chassis and grounding the end of the shield braid directly to chassis, keeping the exposed part as short as possible. Either 0.001-μfd. or 470-μμfd. (500 μμfd.) condensers should be used. The larger capacitance is series-resonant in Channel 2 and the smaller in Channel 6.

These by-passes are essential at the connection-block terminals, and desirable at the tube ends of the leads also. Installed as shown with shielded wiring, they have been found to be so effective that there is usually no need for further harmonic filtering. However, if a test shows that additional filtering is required, the arrangement shown in Fig. 23-14 may be used. Such an r.f. filter should be installed at the tube end of the shielded lead, and if more than one circuit is filtered care should be taken to keep the r.f. chokes separated from each other and so oriented as to minimize coupling between them. This is necessary for preventing harmonics present in one circuit from being coupled into another.

In difficult cases involving Channels 7 to 13 — i.e., close proximity between the transmitter and receiver, and a weak TV signal — additional lead-filtering measures may be needed to prevent radiation of interfering signals by 50- and 144-Mc. transmitters. A recommended method is shown in Fig. 23-15. It uses a shielded lead by-passed with a ceramic disk as described above, with the addition of a low-inductance feed-through type condenser and a small r.f. choke, the condenser being used as a terminal for the external connection. For voltages above 400, a condenser of compact construction (as indicated in the caption) should be used, mounted so that there is a very minimum of exposed lead, inside the chassis, from the condenser to the connection terminal.

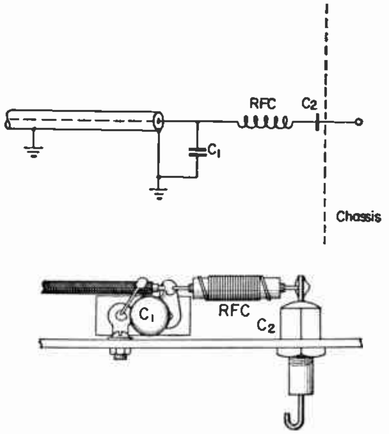


Fig. 23-15 — Additional lead filtering for harmonics or other spurious frequencies in the high v.h.f. TV band (174–216 Mc.).

C_1 — 0.001-μfd. disk ceramic.

C_2 — 0.001-μfd. feed-through by-pass (Erie Style 326). (For 500-2000-volt lead, substitute Plasticon Glass mike, LSG — 251, for C_2.)

RFC — 14 inches No. 26 enamel close-wound on 3/16 inch diam. form or resistor.

As an alternative to the series-resonant by-passing described above, feed-through type condensers such as the Sprague "Hypass" type may be used as terminals for external connections. The ideal method of installation is to mount them so they protrude through the chassis, with thorough bonding to the chassis all around the hole in which the condenser is mounted. The principle is illustrated in Fig. 23-16.

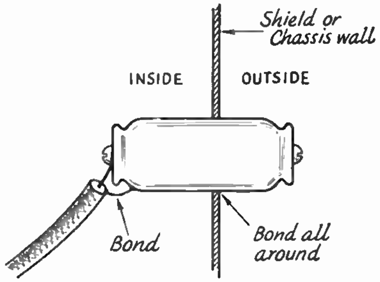


Fig. 23-16 — The best method of using the "Hypass" type feed-through condenser. Capacitances of 0.01 to 0.1 μfd. are satisfactory. Condensers of this type are useful for high-current circuits, such as filament and 115-volt leads, as a substitute for the r.f. choke shown in Fig. 23-14, in cases where additional lead filtering is needed.

Meters that are mounted in an r.f. unit should be enclosed in shielding covers, the connections being made with shielded wire with each lead by-passed as described above. The shield braid should be grounded to the panel or chassis immediately outside the meter shield, as indicated in Fig. 23-17. A by-pass may also be connected across the meter terminals, principally to prevent any fundamental current that may be present from flowing through the meter itself. As an alternative to individual meter shielding the meters may be mounted entirely behind the panel, and the panel holes needed for observation may be covered with wire screen that is carefully bonded to the panel all around the hole.

Care should be used in the selection of shielded wire for transmitter use. Not only should the insulation be conservatively rated for the d.c. voltage in use, but the insulation should be of material that will not easily deteriorate in soldering. The r.f. characteristics of the wire are not especially important, except that the attenuation of harmonics in the wire itself will be greater if the insulating material has high losses at radio frequencies; in other words, wire intended for use at d.c. and low frequencies is preferable to cables designed expressly for carrying r.f. The attenuation also will increase with the length of the wire; in general, it is better to make the leads as long as circumstances permit rather than to follow the more usual practice of using no more lead than is actually necessary. Where the wiring crosses or runs parallel, the shields should be spot-soldered together and connected to the chassis. For high voltages, automobile ignition cable covered with shielding braid is recommended.

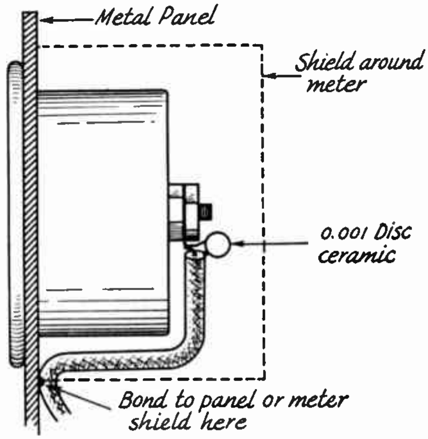


Fig. 23-17 — Meter shielding and by-passing. It is essential to shield the meter mounting hole since the meter will carry r.f. through it to be radiated. Suitable shields can be made from 2½- or 3-inch diameter shield cans of the type made for enclosing coils.

Proper shielding of the transmitter requires that the r.f. circuits be shielded entirely from the external connecting leads. A situation such as is shown in Fig. 23-18, where the leads in the r.f. chassis have been shielded and properly filtered

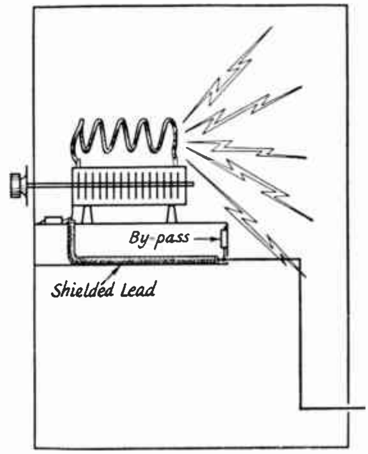


Fig. 23-18 — A metal cabinet can be an adequate shield, but there will still be radiation if the leads inside can pick up r.f. from the transmitting circuits.

but the chassis is mounted in a large shield, simply invites the harmonic currents to travel over the chassis and on out over the leads *outside* the chassis. The shielding about the r.f. circuits should make complete contact with the chassis on which the parts are mounted.

Checking Transmitter Radiation

A check for transmitter radiation always should be made before attempting to use low-pass filters or other devices for preventing harmonics from reaching the antenna system. The only really satisfactory indicating instrument is a television receiver. In regions where the TV signal is strong an indicating wavemeter such as one having a crystal or tube detector may be useful; if it is possible to get any indication at all on harmonics either on supply leads or around the transmitter itself, the harmonics are probably strong enough to cause interference. However, the absence of any such indication does not mean that harmonic interference will not be caused. If the techniques of shielding and lead filtering described in the preceding section are followed, the harmonic intensity on any external leads should be far below what any such instruments can detect.

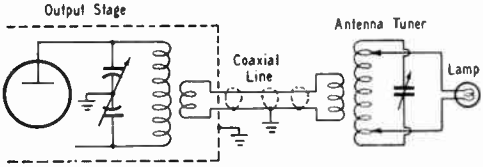


Fig. 23-19 **— Dummy-antenna circuit for checking harmonic radiation from the transmitter and leads. The matching circuit helps prevent harmonics in the output of the transmitter from flowing back over the transmitter itself, which may occur if the lamp load is simply connected to the output coil of the final amplifier. See transmission-line chapter for details of the matching circuit. Tuning must be adjusted by cut-and-try, as the bridge method described in the transmission-line chapter will not work with lamp loads because of the change in resistance when the lamps are hot.**

Radiation checks should be made with the transmitter delivering full power into a dummy antenna, such as an incandescent lamp of suitable power rating, preferably installed inside the shielded enclosure. If the dummy must be external, it is desirable to connect it through a coax-matching circuit such as is shown in Fig. 23-19. Shielding the dummy antenna circuit is also desirable, although it is not always necessary.

Make the radiation test on all frequencies that are to be used in transmitting, and note whether or not interference patterns show in the received picture. (These tests must be made while a TV signal is being received, since the beat patterns will not be formed if the TV picture carrier is not present.) If interference exists, its source can be detected by grasping the various external leads (by the insulation, not the live wire!) or bringing the hand near meter faces, louvers, and other possible points where harmonic energy might escape from the transmitter. If any of these tests cause a *change* — not necessarily an *increase* — in the intensity of the interference, the presence of harmonics at that point is indicated. The location of such "hot" spots usually will point the way to the remedy. If the TV receiver and the transmitter can be operated side-by-side, a length of wire connected to one antenna terminal on the receiver can be used as a probe to go over the transmitter enclosure and external leads. This device will very quickly expose the spots from which serious leakage is taking place.

As a final test, connect the transmitting antenna or its transmission line terminals to the outside of the transmitter shielding. Interference created when this test is applied indicates that weak currents are on the outside of the shield and can be conducted to the antenna when the normal antenna connections are used. Currents of this nature represent interference that can be conducted *over* low-pass filters, etc., and which therefore cannot be eliminated by such filters.

PREVENTING HARMONICS FROM REACHING THE ANTENNA

The third and last step in reducing harmonic TVI is to keep the spurious energy generated in or passed through the final stage from traveling over the transmission line to the antenna. It is seldom worthwhile even to attempt this until the radiation from the transmitter and its connecting leads has been reduced to the point where, with the transmitter delivering full power into a dummy antenna, it has been determined by actual testing with a television receiver that the radiation is below the level that can cause interference. If the dummy antenna test shows enough radiation to be seen in a TV picture, it is a practical certainty that harmonics will be coupled to the antenna system no matter what preventive measures are taken.

In inductively-coupled output systems, some harmonic energy will be transferred from the final amplifier through the mutual inductance between the tank coil and the output coupling coil. Harmonics of the output frequency transferred in this way can be greatly reduced by providing sufficient selectivity between the final tank and the transmission line. A good deal of selectivity, amounting to 20 to 30 db. reduction of the second harmonic and much higher reduction of higher-order harmonics, is furnished by a matching circuit of the type shown in Fig. 23-19 and described in the chapter on transmission lines. An "antenna coupler" is therefore a worthwhile addition to the transmitter.

In 50- and 144-Mc. transmitters, particularly, harmonics not directly associated with the output frequency — such as those generated in low-frequency early stages of the transmitter — may get coupled to the antenna by stray means. For example, a 144-Mc. transmitter might have an oscillator or frequency multiplier at 48 Mc., followed by a tripler to 144 Mc. Some of the 48-Mc. energy will appear in the plate circuit of the tripler, and if passed on to the grid of the final amplifier will appear as a 48-Mc. modulation on the 144-Mc. signal. This will cause a spurious signal at 192 Mc., which is in the high TV band, and the selectivity of the tank circuits may not be sufficient to prevent its being coupled to the antenna. Spurious signals of this type can be reduced by using link coupling between the driver stage and final amplifier (and between earlier stages as well) in addition to the suppression afforded by using an antenna coupler.

Capacitive Coupling

Harmonics and other spurious signals transferred from the tank by stray capacitance are not suppressed by an antenna coupler to the same extent as those transferred by pure inductive coupling. The upper drawing in Fig. 23-20 shows the link-coupled system as it might be used to couple into a parallel-conductor line. Inasmuch as a coil is a sizable metallic object, there is capacitance between the final tank coil and its associated link coil, and between the antenna tank

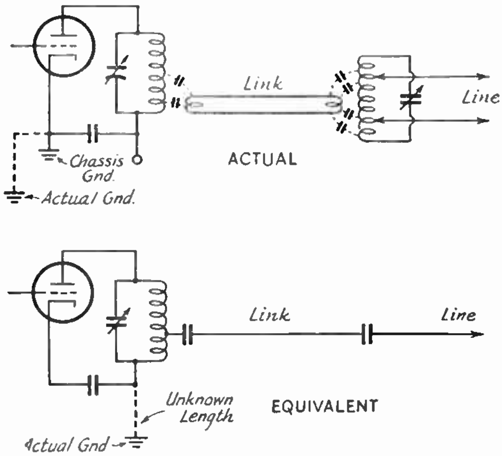


Fig. 23-20 — The stray capacitive coupling between coils in the upper circuit leads to the equivalent circuit shown below, for v.h.f. harmonics.

coil and its link. Energy coupled through these capacitances travels over the link circuit and the transmission line as though these were merely single conductors. The tuned circuits simply act as masses of metal and offer no selectivity at all for capacity-coupled energy. Although the actual capacitances are small, they offer a very good coupling medium for frequencies in the v.h.f. range.

Capacitive coupling can be reduced by coupling to a "cold" point on the tank coil — the end connected to ground or cathode in a single-ended stage. In push-pull circuits having a split-stator condenser with the rotor grounded for r.f., all parts of the tank coil are "hot" at even harmonics, but the center of the coil is "cold" at the fundamental and odd harmonics. If the center of the tank coil, rather than the rotor of the tank condenser, is grounded through a by-pass condenser the center of the coil is "cold" at all frequencies, but this arrangement is not very desirable because it causes the harmonic currents to flow through the coil rather than the tank condenser and this increases the harmonic transfer by pure inductive coupling.

With either single-ended or balanced tank circuits the coupling coil should be grounded to the chassis by a short, direct connection as shown in Fig. 23-21. If the coil feeds a balanced line or link, it is preferable to ground its center, but if it feeds a coax line or link one side may be grounded. Coaxial output is much preferable to balanced output, because the harmonics have to stay *inside* a properly installed coax system and tend to be attenuated by the cable before reaching the antenna coupler.

At high frequencies — and possibly as low as 14 Mc. — capacitive coupling can be greatly reduced by using a shielded coupling coil as shown in Fig. 23-22. The inner conductor of a length of coaxial cable is used to form a one-turn coupling coil. The outer conductor serves as an open-circuited shield around the turn, the shield being grounded to the chassis. The shielding has no effect on the inductive coupling. Because this construction is suitable only for one turn, the coil is not well adapted for use on the lower frequencies where many turns are required for good coupling. Shielded coupling coils having a larger number of turns are available commercially. A shielded coil is particularly useful with push-pull amplifiers when the suppression of even harmonics is important.

A shielded coupling coil or coaxial output will not prevent stray capacitive coupling to the antenna if harmonic currents can flow over the *outside* of the coax line. In Fig. 23-23, the arrangement at either A or C will allow r.f. to flow over the outside of the cable to the antenna system. The proper way to use coaxial cable is to shield the transmitter completely, as shown at B, and make sure that the outer conductor of the cable is a continuation of the transmitter shielding. This prevents r.f. inside the transmitter from getting out by any path except the *inside* of the cable. Harmonics flowing *through* a coax line can be stopped from reaching the antenna system by an antenna coupler or by a low-pass filter installed in the line.

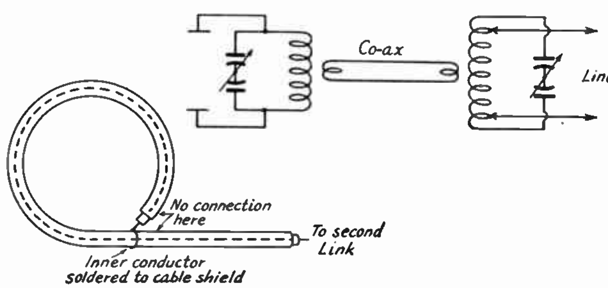
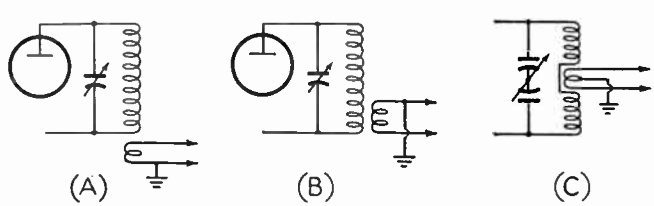


Fig. 23-22 — Shielded coupling coil constructed from coaxial cable. The smaller sizes of cable such as RG-59/U are most convenient when the coil diameter is 3 inches or less, because of greater flexibility. For larger coils RG-8/U or RG-11/U can be used.

Fig. 23-21 — Methods of coupling and grounding link circuits to reduce capacitive coupling between the tank and link coils. Where the link is wound over one end of the tank coil the side toward the hot end of the tank should be grounded, as shown at B.



(A)

(B)

(C)

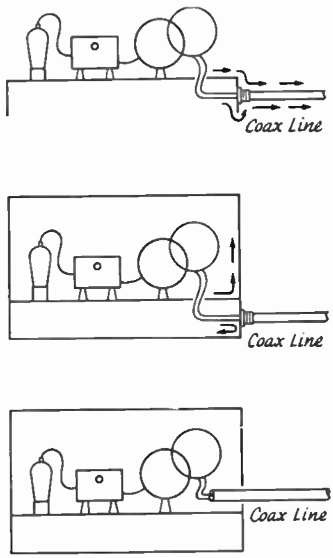


Fig. 23-23 — Right (B) and wrong (A and C) ways to connect a coaxial line to the transmitter. In either A or C, harmonic energy coupled by stray capacitance to the outside of the cable will flow without hindrance to the antenna system. In B the energy cannot leave the shield and hence can flow out only through, not over, the cable.

Low-Pass Filters

A low-pass filter properly installed in a coaxial line, feeding either a matching circuit (antenna coupler) or feeding the antenna directly, will provide very great attenuation of harmonics. When the main transmission line is of the parallel-conductor type, the coax-coupled matching-circuit arrangement is highly recommended as a means for using a coax low-pass filter.

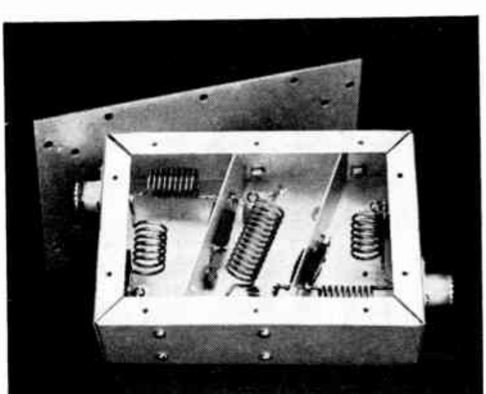


Fig. 23-24 — An inexpensive low-pass filter using silver-mica postage-stamp condensers. The box is a 2 by 4 by 6 aluminum chassis. Aluminum shields, bent and folded at the sides and bottom for fastening to the chassis, form shields between the filter sections. The diagonal arrangement of the shields provides extra room for the coils and makes it easier to fit the shields in the box, since bending to exact dimensions is not essential. The bottom plate, made from sheet aluminum, extends a half inch beyond the ends of the chassis and is provided with mounting holes in the extensions. It is held on the chassis with sheet-metal screws.

A properly-designed low-pass filter will not introduce appreciable power loss at the fundamental frequency if the coaxial line in which it is inserted is terminated so that the s.w.r. is low. (The s.w.r. can easily be measured by means of a simple bridge as described in the chapters on measurements and transmission lines.) Such a filter has the property of passing without loss all frequencies below its "cut-off" frequency, but simultaneously has large attenuation for all frequencies above the cut-off frequency.

Low-pass filters of simple and inexpensive construction for use with transmitters operating below 30 Mc. are shown in Figs. 23-24 and 23-26. These are designed to use mica condensers of readily-available capacitance values, for compactness and low cost. Both use the same circuit, Fig. 23-25, the only difference being in the L and C values. Technically, they are three-section filters having two full constant-k sections and two m-derived terminating half-sections, and their attenuation in the 54–88-Mc. range varies from over 50 to nearly 70 db., depending

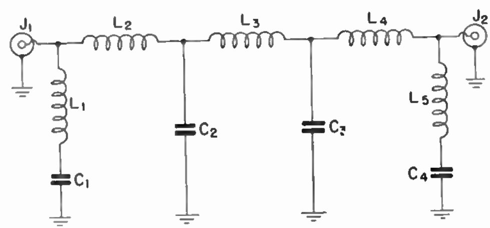


Fig. 23-25 — Low-pass filter circuit for attenuating harmonics in the TV bands. J_1 and J_2 are chassis-type coaxial connectors. In the table below the letters refer to the following:

A — Constructed as in Fig. 23-24, using 100- and 70-μμfd. 500-volt silver mica condensers in parallel for C_2 and C_3.

B — Same as *A* but with 70- and 50-μμfd. silver mica condensers in parallel for C_2 and C_3.

C — Constructed as in Fig. 23-26, using 100- and 50-μμfd. mica condensers, 1200-volt (case-style CM-45) in parallel for C_2 and C_3.

D and E — Constructed with variable condensers, 500- to 1000-volt rating, adjusted to values given.

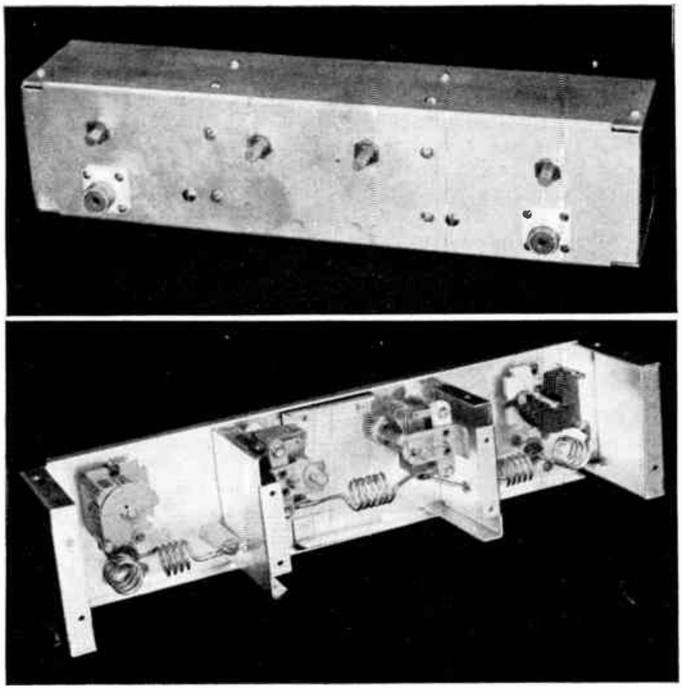
	A	B	C	D	E	
Z_o	52	75	52	52	75	ohms
f_c	36	35.5	41	40	40	Mc.
f_∞	44.4	47	54	50	50	Mc.
f_1	25.5	25.2	29	28.3	28.3	Mc.
f_2	32.5	31.8	37.5	36.1	36.1	Mc.
C_1, C_4	50	40	50	46	32	μμfd.
C_2, C_3	170	120	150	154	106	μμfd.
L_1, L_5	5½	6	4	5	6½	turns*
L_2, L_4	8	11[1]	7	7	9½	turns*
L_3	9	13	8	8½	11½	turns*

* No. 12 or No. 14 wire, ½ inch inside diameter, 8 turns per inch.

[1] A 9-turn coil with closer turn spacing to give the same inductance is shown in Fig. 23-24.

on the frequency and the particular set of values used. Above 174 Mc. the theoretical attenuation is better than 85 db., but will depend somewhat on internal resonant conditions associated principally with the lead lengths to the condensers. These leads should be kept as short as is physically possible.

The power that these filters can handle safely is determined by the voltage and current limitations of the mica condensers. These limitations are such that the power capacity is least at the highest frequency. The unit using postage-stamp silver mica condensers is capable of handling approximately 50 watts in the 28-Mc. band, when working into a properly-matched line, but is good for about 150 watts at 21 Mc. and 300 watts at 14 Mc. and lower frequencies. The unit with the larger mica condensers (case-type CM-45) will carry about 250 watts safely at 28 Mc., this rating increasing to 500 watts at 21 Mc. and a kilowatt at 14 Mc. and lower. If there is an appreciable mismatch between either filter and the line into which it works, these ratings will be considerably decreased, so in order to avoid condenser failure it is highly essential that the line on the output side of the filter be carefully matched by its load. This can be done with an s.w.r. bridge, and the matching is easy to control if the line from the filter terminates in a matching circuit of the type described in the chapter on transmission lines.



Figs. 23-27 — Low-pass filter for use with 50-Mc. transmitters and 52-ohm line. It uses variable air condensers adjusted to the proper capacitance values and is suited to powers up to a kilowatt.

The power capacity of these filters can be increased considerably by substituting r.f. type fixed condensers (such as the Centralab 850 series) or variable air condensers, in which event the power capability will be such as to handle the maximum amateur power on any band. The construction can be modified to accommodate either of the latter types of condenser, using a similar layout in a larger box.

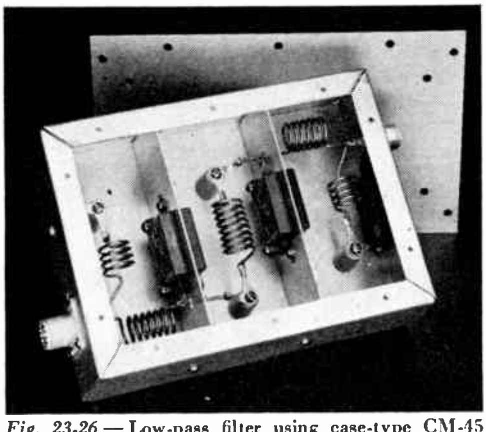


Fig. 23-26 — Low-pass filter using case-type CM-45 condensers. The box is a 2 by 5 by 7 aluminum chassis, fitted with a bottom plate of similar construction to the one used in Fig. 23-24.

Using condensers of standard tolerances, there should be little difficulty in getting proper filter operation. A grid-dip meter with an accurate calibration should be used for adjustment of the coils. First, wire up the filter without L_2 and L_4. Short-circuit J_1 at its inside end with a screwdriver or similar conductor, couple the grid-dip meter to L_1 and adjust the inductance of L_1, by varying the turn spacing, until the circuit resonates at f_∞ as given in the table. Do the same thing at the other end of the filter with L_5. Then couple the meter to the circuit formed by L_3, C_2 and C_3, and adjust L_3 to resonate at the frequency f_1 as given by the table. Then remove L_3, install L_2 and L_4 and adjust L_2 to make the circuit formed by L_1, L_2, C_1 and C_2 (without the short across J_1) resonate at f_2 as given in the table. Do the same with L_4 for the circuit formed by L_4, L_5, C_3 and C_4. Then replace L_3 and check with the grid-dip meter at any coil in the filter; a distinct resonance should be found at or very close to the cut-off frequency, f_c. The filter is then ready for use.

The filter constants suggested at D and E in Fig. 23-25 are based on the optimum design for good impedance characteristics — that is, with $m = 0.6$ in the end sections — and a cut-off frequency below the RETMA standard i.f. for television receivers (sound carrier at 41.25 Mc.; picture carrier at 45.75 Mc.). This is to avoid possible harmonic interference from 21 Mc. and below to the receiver's intermediate amplifier. The other designs similarly cut off at 41 Mc. or below, but m in these cases is necessarily based on the capacitances available in standard fixed condensers.

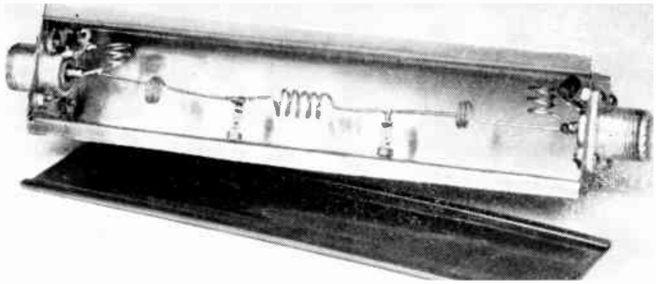


Fig. 23-29 — A 52-ohm low-pass filter for 144-Mc. transmitters.

Filters for 50- and 144-Mc. Transmitters

Since a low-pass filter must have a cut-off frequency above the frequency on which the transmitter operates, a filter for a v.h.f. transmitter cannot be designed for attenuation in all television channels. This is no handicap for v.h.f. work but means that the filter will not be effective when used with lower-frequency transmitters, unless it happens that no TV channels in use in the locality fall inside the pass-band of the filter.

Fig. 23-27 shows a filter for 52-ohm coax suitable for a 50-Mc. transmitter of any power up to the authorized limit. The circuit diagram is given in Fig. 23-28. If the values of inductance and capacitance can be measured (see chapter on measurements) the components can be preset and assembled without further adjustment. Alternatively, the grid-dip meter method described earlier may be used. The resonant frequencies are:

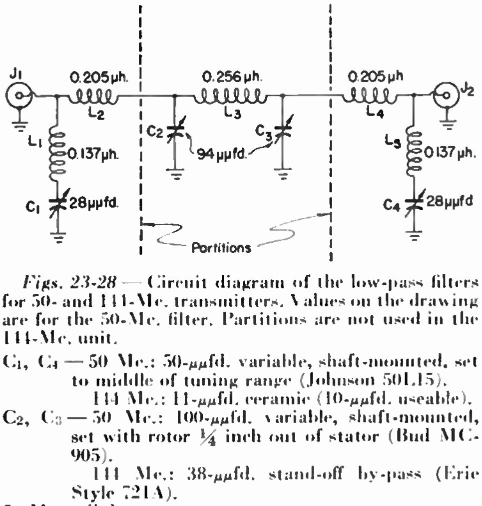
L_1C_1 (J_1 shorted), L_5C_4 (J_2 shorted)	81.5 Mc.
$L_3C_2C_3$ (L_2 and L_4 disconnected)	46 Mc.
$L_1L_2C_1C_2$ (L_3 disconnected), $L_4L_5C_3C_4$ (L_3 disconnected)	58.5 Mc.

The cut-off frequency is approximately 65 Mc.

The case for the 50-Mc. filter is a standard box (ICA Slip-cover, No. 29100) measuring 3⅛ by 13 by 2⅝ inches. The two end condensers, C_1 and C_4, are mounted with their two stator posts toward the ends of the filter. The two larger units are mounted in the center compartment with their rotor shafts toward the middle. The top leads from coils L_1 and L_5 are wrapped around the stator terminals of C_1 and C_4, and the bottom leads fit directly into the coaxial input and output fittings. The outer ends of coils L_2 and L_4 are soldered to the coaxial fitting terminals, and their inner ends are soldered to lugs supported on one-inch ceramic stand-off insulators. Leads from the stand-offs go through holes in the partitions to the bottom stator lugs on C_2 and C_3. L_3 is soldered to the two upper lugs on these two capacitors, thus completing the filter circuit. Lead lengths for the coils given in the parts list are the total lengths to be left when the winding is completed, including the portions that will be used in soldering operations.

This filter will give high attenuation in Channels 4–6 and all the high-band channels, and thus will take care of most of the spurious signals generated in a 50-Mc. transmitter.

A filter for low-power 144-Mc. transmitters is shown in Fig. 23-29. It is designed for maximum attenuation in the 190–215 Mc. region to suppress the spurious radiations in that range that frequently occur with 144-Mc. transmitters, but also has good attenuation for all frequencies above 170 Mc. Optimum capacitance values are given in Fig. 23-28. If possible, several units of the nearest standard values available should be measured and those having values closest to the optimum used. The inductance values are too small to be measured with sufficient accuracy, so the filter should be adjusted by the following method:



Figs. 23-28 — Circuit diagram of the low-pass filters for 50- and 144-Mc. transmitters. Values on the drawing are for the 50-Mc. filter. Partitions are not used in the 144-Mc. unit.

C_1, C_4 — 50 Mc.: 50-μμfd. variable, shaft-mounted, set to middle of tuning range (Johnson 50L15).
144 Mc.: 11-μμfd. ceramic (10-μμfd. useable).

C_2, C_3 — 50 Mc.: 100-μμfd. variable, shaft-mounted, set with rotor ¼ inch out of stator (Bud MC-905).
144 Mc.: 38-μμfd. stand-off by-pass (Erie Style 721A).

50-Mc. coil data:

L_1, L_5 — 3½ turns ⅝ inch long. Top leads ¾ inch, bottom leads ¼ inch long.

L_2, L_4 — 1½ turns ⅝ inch long. Leads ½ inch long each end.

L_3 — 5½ turns ⅞ inch long. Leads 1 inch long each.
All 50-Mc. coils No. 12 tinned, ½-inch diam., coil length measured between right-angle bends where leads begin.

144-Mc. coil data:

L_1, L_5 — 3 turns ¼ inch long. Leads ¼ inch long each end.

L_2, L_4 — 2 turns ⅛ inch long. Leads 1 inch long each end.

L_3 — 5 turns ¾ inch long. Leads ⅝ inch long each end.
All 144-Mc. coils No. 18 tinned, ¼-inch diam., lengths measured as for 50-Mc. coils.

J_1, J_2 — Coaxial fitting.

First, mount L_1 and C_1, short J_1 temporarily at its inner terminals, and adjust L_1 until the combination resonates at 200 Mc. as shown by grid-dip meter. Next, remove the short from J_1 and connect L_2 and C_2, adjusting L_2 until the circuit formed by $L_1L_2C_1C_2$ resonates at 144 Mc. Then disconnect L_2 and mount L_3 between C_2 and C_3. Adjust L_3 until the circuit $L_3C_2C_3$ resonates at 112 Mc. Next, disconnect L_3 and follow a similar procedure starting from the other end with L_5 and C_4. Finally, reconnect all coils and a check at any point in the filter should show resonance at 160 Mc., the approximate cut-off frequency.

The case for the 144-Mc. filter is made from flashing copper and is 1¼ inches square by 7⅛ inches long. The main portion of the case is cut from a single piece with the end tabs folded down and soldered to the sides. Flanges are folded over at the bottom, and a cover is made to slip over these.

Filter Installation

In order to give the harmonic attenuation of which it is capable, a low-pass filter must be installed in such a way that *all* the output of the transmitter flows through it. If harmonic currents are permitted to flow on the outside of the connecting coaxial cables, they will simply flow over the filter and on up to the antenna, and the filter does not have an opportunity to stop them. That is why it is so important to reduce the radiation from the transmitter and its leads to negligible proportions.

Fig. 23-30 shows the proper way to install a filter between a shielded transmitter and a matching circuit. Note that the coax, together with the shields about the transmitter and filter, forms a continuous shield to keep all the r.f. inside. It is thus forced to flow through the filter and the harmonics are attenuated. If there is no harmonic energy left after passing through the filter, shielding from that point on is not necessary; consequently, the matching circuit or antenna coupler does not need to be shielded. However, the antenna-coupler chassis arrangement shown in Fig. 23-30 is desirable because it will tend to prevent fundamental-frequency energy from flowing from the matching circuit back over the transmitter; this helps eliminate feed-back troubles in audio systems.

If the antenna is driven through coaxial line the matching circuit shown in Fig. 23-30 may be omitted. In that case the line goes directly from the filter to the antenna.

When a filter does not seem to give the harmonic attenuation of which it should be capable, the probable reason is that harmonics are by-passing it because of improper installation and inadequate transmitter shielding, including lead filtering. However, occasionally there are cases where the circuits formed by the cables and the apparatus to which they connect become resonant at a harmonic frequency. This greatly increases the harmonic output at that frequency. Such troubles can be completely overcome by substituting a slightly different cable length. The most critical length is that connecting the transmitter to the filter. Checking with a grid-dip meter at the final amplifier output coil usually will show whether an unfavorable resonance of this type exists.

SUMMARY

The methods of harmonic elimination outlined in this chapter have been proved beyond doubt to be effective even under highly unfavorable conditions. It must be emphasized once more, however, that the problem must be solved one step at a time, and the procedure must be in logical order. It cannot be done properly without two items of simple equipment: a grid-dip meter and wavemeter covering the TV bands, and a dummy antenna.

The proper procedure may be summarized as follows:

1) Take a critical look at the transmitter on the basis of the design considerations outlined under "Reducing Harmonic Generation".

2) Check all circuits, particularly those connected with the final amplifier, with the grid-dip meter to determine whether there are any resonances in the TV bands. If so, rearrange the circuits so the resonances are moved out of the critical frequency region.

3) Connect the transmitter to the dummy antenna and check with the wavemeter for the presence of harmonics on leads and around the transmitter enclosure. Seal off the weak spots in the shielding and filter the leads until the wavemeter shows no indication at any harmonic frequency.

4) At this stage, check for interference with a TV receiver. If there is interference, determine the cause by the methods described previously and apply the recommended remedies until the interference disappears.

5) When the transmitter is completely clean on the dummy antenna, connect it to the regular antenna and check for interference on the TV receiver. If the interference is not bad, an antenna coupler or matching circuit installed as previously described should clear it up. Alternatively, a low-pass filter may be used. If neither the antenna coupler nor filter makes any difference in the interference, the evidence is strong that the interference, at least in part, is being caused by receiver overloading because of the strong funda-

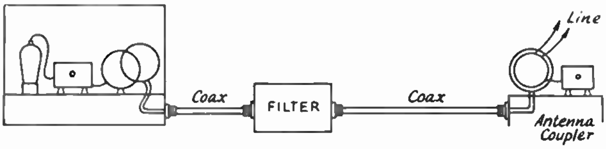


Fig. 23-30 — The proper method of installing a low-pass filter between the transmitter and antenna coupler or matching circuit. If the antenna is fed through coax the matching circuit may be omitted but the same construction should be used between the transmitter and filter. The filter should be thoroughly shielded.

mental-frequency field about the TV antenna and receiver. (See later section for identification of fundamental-frequency interference.) A coupler and, or filter, installed as described above, will invariably make a difference in the intensity of the interference if the interference is caused by transmitter harmonics alone.

6) If there is still interference after installing the coupler and/or filter, and the evidence shows that it is probably caused by a harmonic, more attenuation is needed. A more elaborate filter may be necessary. However, it is well at this stage to assume that part of the interference may be caused by receiver overloading, and take steps to alleviate such a condition before trying highly-elaborate filters, traps, etc., on the transmitter.

HARMONICS BY RECTIFICATION

Even though the transmitter is completely free from harmonic output it is still possible for interference to occur because of harmonics generated outside the transmitter. These result from rectification of fundamental-frequency currents induced in conductors in the vicinity of the transmitting antenna. Rectification can take place at any point where two conductors are in poor electrical contact, a condition that frequently exists in plumbing, downspouting, BX cables crossing each other, and numerous other places in the ordinary residence. It also can occur in any exposed vacuum tubes in the station, in power supplies, speech equipment, etc., that may not be enclosed in the shielding about the r.f. circuits. Poor joints anywhere in the antenna system are especially bad, and rectification also may take place in the contacts of antenna changeover relays. Another common cause is overloading the front end of the communications receiver when it is used with a separate antenna (which will radiate the harmonics generated in the first tube) for break-in.

Rectification of this sort will not only cause harmonic interference but also is frequently responsible for cross-modulation effects. It can be detected in greater or less degree in most locations, but fortunately the harmonics thus generated are not usually of high amplitude. However, they can cause considerable interference in the immediate vicinity in fringe areas, especially when operation is in the 28-Mc. band. The amplitude decreases rapidly with the order of the harmonic, the second and third being the worst. It is ordinarily found that even in cases where destructive interference results from 28-Mc. operation the interference is comparatively mild from 14 Mc., and is negligible at still lower frequencies.

There is nothing that can be done at either the transmitter or receiver when rectification occurs. The remedy is to find the source and eliminate the poor contact either by separating the conductors or bonding them together. A crystal wavemeter (tuned to the fundamental frequency) is useful for hunting the source, by showing which conductors are carrying r.f. and, comparatively, how much.

Interference of this kind is frequently intermittent, since the rectification efficiency will vary with vibration, the weather, and so on. The possibility of corroded contacts in the TV receiving antenna should not be overlooked, especially if it has been up a year or more.

TV RECEIVER DEFICIENCIES

Front-End Overloading

When a television receiver is quite close to the transmitter, the intense r.f. signal from the transmitter's fundamental may overload one or more of the receiver circuits to produce spurious responses that cause interference.

If the overload is moderate, the interference is of the same nature as harmonic interference; it is caused by harmonics generated in the early stages of the receiver and, since it occurs only on channels harmonically related to the transmitting frequency, is difficult to distinguish from harmonics actually radiated by the transmitter. In such cases additional harmonic suppression at the transmitter will do no good, but any means taken at the receiver to reduce the amateur fundamental strength fed to the first tube will effect an improvement. With more severe overloading interference also will occur on channels *not* harmonically related to the transmitting frequency, so such cases are easily identified.

Cross-Modulation

Under some circumstances overloading will result in cross-modulation or mixing of the amateur signal and that from a local FM or TV station. For example, a 14-Mc. signal can mix with a 92-Mc. FM station to produce a beat at 78 Mc. and cause interference in Channel 5, or with a TV station on Channel 5 to cause interference in Channel 3. Neither of the channels interfered with is in harmonic relationship to 14 Mc. Both signals have to be on the air for the interference to occur, and eliminating either at the TV receiver will eliminate the interference.

There are many combinations of this type, depending on the band in use and the local frequency assignments to FM and TV stations. The interfering frequency is equal to the amateur fundamental frequency either added to or subtracted from the frequency of some local station, and when interference occurs in a TV channel that is not harmonically related to the amateur transmitting frequency the possibilities in such frequency combination should be investigated.

I. F. Interference

Some TV receivers do not have sufficient selectivity to prevent strong signals in the intermediate-frequency range from forcing their way through the front end and getting into the i.f. amplifier. The once-standard intermediate frequency of, roughly, 21 to 27 Mc., is subject to interference from the fundamental-frequency output of transmitters operating in either the 21-

and 27-Mc. bands. Transmitters on 28 Mc. sometimes will cause this type of interference as well.

A form of i.f. interference peculiar to 50-Mc. operation near the low edge of the band occurs with some receivers having the standard "41-Mc." i.f., which has the sound carrier at 41.25 Mc. and the picture carrier at 45.75 Mc. A 50-Mc. signal that forces its way into the i.f. system of the receiver will cause a beat with the i.f. picture carrier that falls on or near the i.f. sound carrier, even though the interfering signal is not actually in the nominal pass-band of the i.f. amplifier.

There is a type of i.f. interference unique to the 144-Mc. band in localities where certain u.h.f. TV channels are in operation, affecting only those TV receivers in which double-conversion type plug-in u.h.f. tuning strips are used. The design of these strips involves a first intermediate frequency that varies with the TV channel to be received and, depending on the particular strip design, this first i.f. may be in or close to the 144-Mc. amateur band. Since there is comparatively little selectivity in the TV signal-frequency circuits ahead of the first i.f., a signal from a 144-Mc. transmitter will "ride into" the i.f., even when the receiver is at a considerable distance from the transmitter. The channels that can be affected by this type of i.f. interference are as follows:

Receivers with 21-Mc. second i.f.	*Receivers with 41-Mc. second i.f.*
Channels 14–18, inc.	Channels 20–25, inc.
Channels 41–48, inc.	Channels 51–58, inc.
Channels 69–77, inc.	Channels 82 and 83.

If the receiver is not close to the transmitter, a trap of the type shown in Fig. 23-33 will be effective. However, if the separation is small the 144-Mc. signal will be picked up directly on the receiver circuits and the best solution is to readjust the strip oscillator so that the first i.f. is moved to a frequency not in the vicinity of the 144-Mc. band. This has to be done by a competent technician.

I.f. interference is easily identified since it occurs on all channels — although sometimes the intensity varies from channel to channel — and the cross-hatch pattern it causes will rotate when the receiver's fine-tuning control is varied. When the interference is caused by a harmonic, overloading, or cross modulation, the structure of the interference pattern does not change as the fine-tuning control is varied, although its intensity may change.

High-Pass Filters

In all the above cases the interference can be eliminated if the fundamental signal strength can be reduced to a level that the receiver can handle. To accomplish this with signals on bands below 30 Mc., the most satisfactory device is a high-pass filter having a cut-off frequency between 30 and 50 Mc., installed at the tuner input terminals

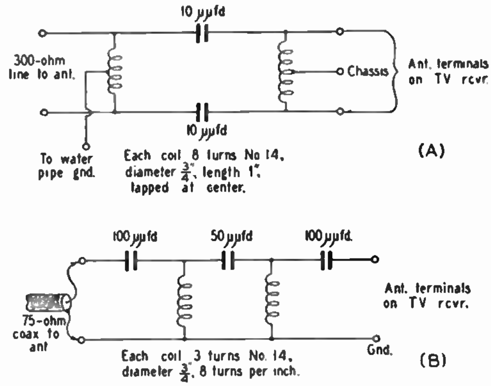


Fig. 23-31 — High-pass filters for installation at the TV receiver antenna terminals. A — balanced filter for 300-ohm line, B — for 75-ohm coaxial line. *Important:* Do not use a direct ground on an a.c.-d.c. chassis. Ground through a 0.001-μfd. mica condenser.

of the receiver. Circuits that have proved effective are shown in Figs. 23-31 and 23-32. Fig. 23-32 has one more section than the filters of Fig. 23-31 and as a consequence has somewhat better cut-off characteristics. All the circuits given are designed to have little or no effect on the TV signals but will attenuate all signals lower in frequency than about 40 Mc. These filters preferably should be constructed in some sort of shielding container, although shielding is not always necessary. The dashed lines in Fig. 23-32 show how individual filter coils can be shielded from each other. The condensers can be

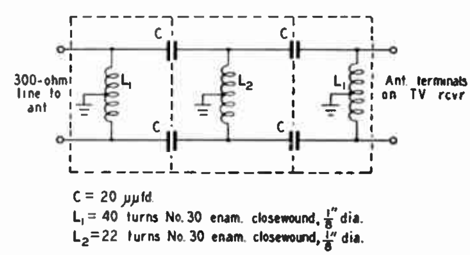


Fig. 23-32 — Another type of high-pass filter for 300-ohm line. The coils may be wound on 1/8-inch diameter plastic knitting needles. *Important:* Do not use a direct ground on an a.c.-d.c. chassis. Ground through a 0.001-μfd. mica condenser.

tubular ceramic units centered in holes in the partitions that separate the coils.

High-pass filters cannot be applied successfully in the case of 50-Mc. transmissions, because a filter having a sufficiently-sharp cut-off characteristic to give both good attenuation at 50–54 Mc. and no attenuation above 54 Mc. cannot be built, practically. A high-*Q* wavetrap, tuned to the transmitting frequency, is about the only practical solution. A successful design, using quarter-wave sections of Twin-Lead, is shown in Fig. 23-33. These "suck-out" traps absorb energy at the frequency to be eliminated, when carefully tuned. The assembly should be slid along the TV antenna lead-in until the most

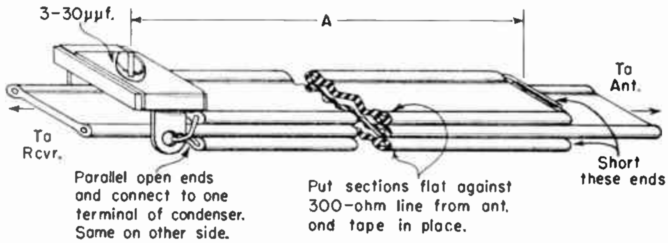


Fig. 23-33 — Absorption-type wavetrap using sections of 300-ohm line tuned to have an electrical length of ¼ wavelength at the transmitter frequency. Approximate physical lengths (dimension *A*) are 40 inches for 50 Mc. and 14 inches for 144 Mc., allowing for the loading effect of the capacitance at the open end. Two traps are used in parallel, one on each side of the line to the receiver. Install close to receiver antenna terminals.

effective position is found, and then fastened in place with Scotch Tape. An insulated tuning tool should be used for adjustment of the condenser, since it is at a "hot" point and will show considerable body-capacity effect.

High-pass filters are available commercially at moderate prices. In this connection, it should be understood by all parties concerned that while an amateur is responsible for *harmonic* radiation from his transmitter, it is no part of his responsibility to pay for or install filters, wavetraps, etc., that may be required at the receiver to prevent interference caused by his *fundamental* frequency. The set owner should be advised to get in touch with the organization from which he purchased the receiver or which services it, to make arrangements for proper installation. Proper installation usually requires that the filter be installed right at the input terminals of the r.f. tuner of the TV set and not merely at the antenna terminals, which may be at a considerable distance from the tuner. The question of cost is one to be settled between the set owner and the organization with which he deals. Some of the larger manufacturers of TV receivers have instituted arrangements for coöperating with the set dealer in installing high-pass filters at no cost to the receiver owner. FCC-sponsored TVI Committees, now operating in many cities, have all the information necessary for effectuating such arrangements.

If the fundamental signal is getting into the receiver by way of the line cord a line filter such as that shown in Fig. 23-4 may help. To be most effective it should be installed inside the receiver chassis at the point where the cord enters, making the ground connections directly to chassis at this point. It may not be so helpful if placed between the line plug and the wall socket unless the r.f. is actually picked up on the house wiring rather than on the line cord itself

Antenna Installation

Many television receivers will respond strongly to parallel currents on the receiving transmission line. Usually, the transmission line picks up a great deal more energy from a near-by transmitter than the television receiving antenna itself, causing parallel currents that should be, but are not, rejected by the receiver's input circuit. This situation can be improved by using shielded transmission line — coax or, in the balanced form, "twinax" — on the receiving installation. For best results the line should terminate in a coax fitting on the receiver chassis, but if this is not possible the shield should be grounded to the chassis right at the antenna terminals.

The use of shielded transmission line for the receiver also will be helpful in reducing response to harmonics actually being radiated from the transmitter or transmitting antenna. In most receiving installations the transmission line is very much longer than the antenna itself, and is consequently far more exposed to the harmonic fields from the transmitter. Much of the harmonic pick-up, therefore, is on the receiving transmission line when the transmitter and receiver are quite close together. Shielded line, plus relocation of either the transmitting or receiving antenna to take advantage of directive effects, often will result in reducing overloading, as well as harmonic pick-up, to a level that does not interfere with reception.

U.H.F. Television

Although u.h.f. television is comparatively new, experience so far indicates that harmonic TVI is far less troublesome in this band than in the v.h.f. TV band. Harmonics from transmitters operating below 30 Mc. are of such high order that they would normally be expected to be quite weak; in addition, the components, circuit conditions and construction of low-frequency transmitters are such as to tend to prevent very strong harmonics from being generated in this region. However, this is not true of v.h.f. transmitters, particularly those working in the 144-Mc. and higher bands. Here the problem is quite similar to the problem of the low v.h.f. TV band with respect to transmitters operating below 30 Mc.

There is one highly favorable factor in u.h.f. TV that does not exist in the most of the v.h.f. TV band: If harmonics are radiated, it is possible to move the transmitter frequency sufficiently (within the amateur band being used) to avoid interfering with a channel that may be in use in the locality. By restricting operation to a portion of the band that will not result in harmonic interference, it is possible to avoid the necessity for taking extraordinary precautions to prevent harmonic radiation.

TABLE 23-I

Harmonic Relationship—Amateur V.H.F. Bands and U.H.F. TV Channels

Amateur Band	*Harmonic*	*Fundamental Freq. Range*	*U.H.F. TV Channel Affected*
144 Mc.	4th	144.0–144.5	31
		144.5–146.0	32
		146.0–147.5	33
		147.5–148.0	34
	5th	144.0–144.4	55
		144.4–145.6	56
		145.6–146.8	57
		146.8–148	58
	6th	144–144.33	79
		144.33–145.33	80
		145.33–147.33	81
		147.33–148	82
220 Mc.	3rd	220–220.67	45
		220.67–222.67	46
		222.67–224.67	47
		224.67–225	48
	4th	220–221	82
		221–222.5	83
420 Mc	2nd	420–421	75
		421–424	76
		424–427	77
		427–430	78
		430–433	79
		433–436	80
		436–439	81
		439–442	82
		442–448	83

The frequency assignment for u.h.f. television consists of seventy 6-megacycle channels (Nos. 14 to 83, inclusive) beginning at 470 Mc. and ending at 890 Mc. The harmonics from amateur bands above 50 Mc. span the u.h.f. channels as shown in Table 23-I. Since the assignment plan calls for a minimum separation of six channels between any two stations in one locality, there is ample opportunity to choose a fundamental frequency that will move a harmonic out of range of a local TV frequency.

The basic methods of preventing harmonic radiation already described will apply equally well in the u.h.f. TV region. However, there is comparatively little information as yet on the best types of components for these frequencies. It is anticipated also that modifications of the techniques will be developed to give improved harmonic suppression as experience is acquired.

Color Television

It is probable that a number of stations will begin transmitting television programs in color during 1954, but it is not expected that any very considerable number of receivers capable of color reproduction will be in use immediately. The color TV system now proposed uses a subcarrier frequency spaced 3.58 megacycles from the regular picture carrier (or 4.83 Mc. from the low edge of the channel) for transmitting the color information. Harmonics which fall in the color subcarrier region can be expected to cause break-up of color in the received picture. This modifies the chart of Fig. 23-7 to introduce another "severe" region centering around 4.8 Mc. measured from the low-frequency edge of the channel. Hence with color television reception there is less opportunity to avoid harmonic interference by choice of operating frequency.

Construction Practices

TOOLS AND MATERIALS

While an easier, and perhaps a better, job can be done with a greater variety of tools available, by taking a little thought and care it is possible to turn out a fine piece of equipment with only a few of the common hand tools. A list of tools which will be indispensable in the construction of radio equipment will be found on this page. With these tools it should be possible to perform any of the required operations in preparing panels and metal chassis for assembly and wiring. It is an excellent idea for the amateur who does constructional work to add to his supply of tools from time to time as finances permit.

Several of the pieces of light woodworking machinery, often sold in hardware stores and mail-order retail stores, are ideal for amateur radio work, especially the drill press, grinding head, band and circular saws, and joiner. Although not essential, they are desirable should you be in a position to acquire them.

INDISPENSABLE TOOLS

Long-nose pliers, 6-inch.
Diagonal cutting pliers, 6-inch.
Wire stripper.
Screwdriver, 6- to 7-inch, ¼-inch blade.
Screwdriver, 4- to 5-inch, ⅛-inch blade.
Scratch awl or scriber for marking lines.
Combination square, 12-inch, for laying out work.
Hand drill, ¼-inch chuck or larger, 2-speed type preferable.
Electric soldering iron, 100 watts, ¼-in. tip.
Hack saw, 12-inch blades.
Center punch for marking hole centers.
Hammer, ball-peen, 1-lb. head.
Heavy knife.
Yardstick or other straightedge.
Carpenter's brace with adjustable hole cutter or socket-hole punches (see text).
Large, coarse, flat file.
Large round or rat-tail file, ½-inch diameter.
Three or four small and medium files—flat, round, half-round, triangular.
Drills, particularly ⅛-inch and Nos. 18, 28, 33, 42 and 50.
Combination oil stone for sharpening tools.
Solder and soldering paste (noncorroding).
Medium-weight machine oil.

ADDITIONAL TOOLS

Bench vise, 4-inch jaws.
Tin shears, 10-inch, for cutting thin sheet metal.
Taper reamer, ½-inch, for enlarging small holes.
Taper reamer, 1-inch, for enlarging holes.
Countersink for brace.
Carpenter's plane, 8- to 12-inch, for woodworking.
Carpenter's saw, crosscut.
Motor-driven emery wheel for grinding.
Long-shank screwdriver with screw-holding clip for tight places.
Set of "Spintite" socket wrenches for hex nuts.
Set of small, flat, open-end wrenches for hex nuts.
Wood chisel, ½-inch.
Cold chisel, ½-inch.
Wing dividers, 8-inch, for scribing circles.
Set of machine-screw taps and dies.
Dusting brush.
Socket punches, esp. 1⅛" and 1¼".

Twist Drills

Twist drills are made of either high-speed steel or carbon steel. The latter type is more common and will usually be supplied unless specific request is made for high-speed drills. The carbon drill will suffice for most ordinary equipment construction work and costs less than the high-speed type.

While twist drills are available in a number of sizes those listed in bold-faced type in Table 24-I will be most commonly used in construction of amateur equipment. It is usually desirable to purchase several of each of the commonly-used sizes rather than a standard set, most of which will be used infrequently, if at all.

Care of Tools

The proper care of tools is not alone a matter of pride to a good workman. He also realizes the energy which may be saved and the annoyance which may be avoided by the possession of a full kit of well-kept sharp-edged tools.

Drills should be sharpened at frequent intervals so that grinding is kept at a minimum each time. This makes it easier to maintain the rather critical surface angles required for best cutting with least wear. Occasional oilstoning of the cutting edges of a drill or reamer will extend the time between grindings.

The soldering iron can be kept in good condition by keeping the tip well tinned with solder and not allowing it to run at full voltage for long periods when it is not being used. After each period of use, the tip should be removed and cleaned of any scale which may have accumulated. An oxidized tip may be cleaned by dipping it in sal ammoniac while hot and then wiping it clean with a rag. If the tip becomes pitted it should be filed until smooth and bright, and then tinned immediately by dipping it in solder.

Useful Materials

Small stocks of various miscellaneous materials will be required in constructing radio apparatus, most of which are available from hardware or radio-supply stores. A representative list follows:

- Sheet aluminum, 16 or 18 gauge for brackets and shielding.
- ½ × ½-inch aluminum angle stock.
- ¼-inch diameter round brass or aluminum rod for shaft extensions.
- Machine screws: Round-head and flat-head, with nuts to fit. Most useful sizes: 4-36, 6-32 and 8-32, in lengths from ¼ inch to 1½ inches. (Nickel-plated iron will be found satisfactory except in strong r.f. fields, where brass should be used.)
- Bakelite, lucite and polystyrene scraps.
- Soldering lugs, panel bearings, rubber grommets, terminal-lug wiring strips, varnished-cambric insulating tubing.
- Shielded and unshielded wire.
- Tinned bare wire, Nos. 22, 14 and 12.

Machine screws, nuts, washers, soldering lugs, etc., are most reasonably purchased in quantities of a gross.

TABLE 24-I
Numbered Drill Sizes

Number	*Diameter (mils)*	*Will Clear Screw*	*Drilled for Tapping Iron, Steel or Brass**
1	228.0	—	—
2	221.0	12–24	—
3	213.0	—	14–24
4	209.0	12–20	—
5	205.0	—	—
6	204.0	—	—
7	201.0	—	—
8	199.0	—	—
9	196.0	—	—
10	193.5	10–32	—
11	191.0	10–24	—
12	189.0	—	—
13	185.0	—	—
14	182.0	—	—
15	180.0	—	—
16	177.0	—	12–24
17	173.0	—	—
18	**169.5**	**8–32**	—
19	166.0	—	12–20
20	161.0	—	—
21	159.0	—	10–32
22	157.0	—	—
23	154.0	—	—
24	152.0	—	—
25	149.5	—	10–24
26	147.0	—	—
27	144.0	—	—
28	**140.0**	**6–32**	—
29	**136.0**	—	**8–32**
30	128.5	—	—
31	120.0	—	—
32	116.0	—	—
33	**113.0**	**4–36, 4–40**	—
34	111.0	—	—
35	**110.0**	—	**6–32**
36	106.5	—	—
37	104.0	—	—
38	101.5	—	—
39	099.5	3–48	—
40	098.0	—	—
41	096.0	—	—
42	**093.5**	—	**4–36, 4–40**
43	089.0	2–56	—
44	086.0	—	—
45	082.0	—	3–48
46	081.0	—	—
47	078.5	—	—
48	076.0	—	—
49	073.0	—	2–56
50	**070.0**	—	—
51	067.0	—	—
52	063.5	—	—
53	059.5	—	—
54	055.0	—	—

*Use one size larger for tapping bakelite and hard rubber.

CHASSIS WORKING

With a few essential tools and proper procedure, it will be found that building radio gear on a metal chassis is no more of a chore than building with wood, and a more satisfactory job results. Aluminum is to be preferred to steel, not only because it is a superior shielding material, but because it is much easier to work and to provide good chassis contacts.

The placing of components on the chassis is shown quite clearly in the photographs in this *Handbook*. Aside from certain essential dimensions, which usually are given in the text, exact duplication is not necessary.

Much trouble and energy can be saved by spending sufficient time in planning the job. When all details are worked out beforehand the actual construction is greatly simplified.

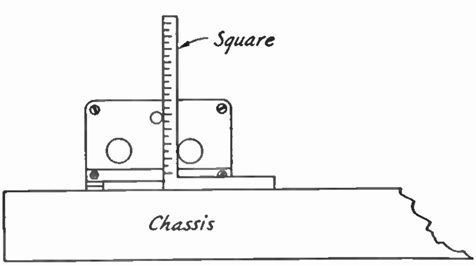


Fig. 24-1 — Method of measuring the heights of condenser shafts, etc. If the square is adjustable, the end of the scale should be set flush with the face of the head.

Cover the top of the chassis with a piece of wrapping paper or, preferably, cross-section paper, folding the edges down over the sides of the chassis and fastening with adhesive tape. Then assemble the parts to be mounted on top of the chassis and move them about until a satisfactory arrangement has been found, keeping in mind any parts which are to be mounted underneath, so that interferences in mounting may be avoided. Place condensers and other parts with shafts extending through the panel first, and arrange them so that the controls will form the desired pattern on the panel. Be sure to line up the shafts squarely with the chassis front. Locate any partition shields and panel brackets next, and then the tube sockets and any other parts, marking the mounting-hole centers of each accurately on the paper. Watch out for condensers whose shafts are off center and do not line up with the mounting holes. Do not forget to mark the centers of socket holes and holes for leads under i.f. transformers. etc., as well as holes for wiring leads. The small holes for socket-mounting screws are best located and center-punched, using the socket itself as a template, after the main center hole has been cut.

By means of the square, lines indicating accurately the centers of shafts should be extended to the front of the chassis and marked on the panel at the chassis line, the panel being fastened on temporarily. The hole centers may then be punched in the chassis with the center punch. After drilling, the parts which require mounting underneath may be located and the mounting holes drilled, making sure by trial that no interferences exist with parts mounted on top. Mounting holes along the front edge

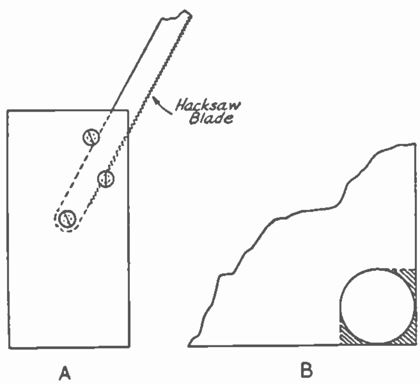


Fig. 24-2 — To cut rectangular holes in a chassis corner, holes may be filed out as shown in the shaded portion of B, making it possible to start the hack-saw blade along the cutting line. A shows how a single-ended handle may be constructed for a hack-saw blade.

of the chassis should be transferred to the panel, by once again fastening the panel to the chassis and marking it from the rear.

Next, mount on the chassis the condensers and any other parts with shafts extending to the panel, and measure accurately the height of the center of each shaft above the chassis, as illustrated in Fig. 24-1. The horizontal displacement of shafts having already been marked on the chassis line on the panel, the vertical displacement can be measured from this line. The shaft centers may now be marked on the back of the panel, and the holes drilled. Holes for any other panel equipment coming above the chassis line may then be marked and drilled, and the remainder of the apparatus mounted. Holes for terminals etc., in the rear edge of the chassis should be marked and drilled at the same time that they are done for the top.

Drilling and Cutting Holes

When drilling holes in metal with a hand drill it is important that the centers first be located with a center punch, so that the drill point will not "walk" away from the center when starting the hole. When the drill starts to break through, special care must be used. Often it is an advantage to shift a two-speed drill to low gear at this point. Holes more than ¼ inch in diameter may be started with a smaller drill and reamed out with the larger drill.

The chuck on the usual type of hand drill is limited to ¼-inch drills. Although it is rather tedious, the ¼-inch hole may be filed out to larger diameters with round files. Another method possible with limited tools is to drill a series of small holes with the hand drill along the inside of the diameter of the large hole, placing the holes as close together as possible. The center may then be knocked out with a cold chisel and the edges smoothed up with a file. Taper reamers which fit into the carpenter's brace will make the job easier. A large rattail file clamped in the brace makes a very good reamer for holes up to the diameter of the file, if the file is revolved counterclockwise.

For socket holes and other large round holes, an adjustable cutter designed for the purpose may be used in the brace. Occasional application of machine oil in the cutting groove will help. The cutter first should be tried out on a block of wood, to make sure that it is set for the correct diameter. The most convenient device for cutting socket holes is the socket-hole punch. The best type is that which works by turning a take-up screw with a wrench.

Rectangular Holes

Square or rectangular holes may be cut out by making a row of small holes as previously described, but is more easily done by drilling a ½-inch hole inside each corner, as illustrated in Fig. 24-2, and using these holes for starting and turning the hack saw. The socket-hole punch and the square punches which are now available also may be of considerable assistance in cutting out large rectangular openings. The burrs or rough edges which usually result after drilling or cutting holes may be removed with a file, or sometimes more conveniently with a sharp knife or chisel. It is a good idea to keep an old wood chisel sharpened and available for this purpose. A burr reamer will also be useful.

CONSTRUCTION NOTES

If a control shaft must be extended or insulated, a flexible shaft coupling with adequate insulation should be used. Satisfactory support for the shaft extension can be provided by means of a *metal* panel bearing made for the purpose. Never use panel bearings of the non-metal type unless the condenser shaft is grounded. *The metal bearing should be connected to the chassis with a wire or grounding strip.*

This prevents any possible danger of shock.

The use of fiber washers between ceramic insulation and metal brackets, screws or nuts will prevent the ceramic parts from breaking.

Cutting and Bending Sheet Metal

If a sheet of metal is too large to be cut conveniently with a hack saw, it may be marked with scratches as deep as possible along the line of the cut on both sides of the sheet and then clamped in a vise and worked back and forth until the sheet breaks at the line. Do not carry the bending too far until the break begins to weaken; otherwise the edge of the sheet may become bent. A pair of iron bars or pieces of heavy angle stock, as long or longer than the width of the sheet, to hold it in the vise will make the job easier. "C"-clamps may be used to keep the bars from spreading at the ends. The rough edges may be smoothed up with a file or by placing a large piece of emery cloth or sandpaper on a flat surface and running the edge of the metal back and forth over the sheet.

Bends may be made similarly. The sheet should be scratched on both sides, but not so deeply as to cause it to break.

Finishing Aluminum

Aluminum chassis, panels and parts may be given a sheen finish by treating them in a caustic bath. An enamelled container, such as a dishpan or infant's bathtub, should be used for the solution. Dissolve ordinary household lye in cold water in a proportion of ¼ to ½ can of lye per gallon of water. The stronger solution will do the job more rapidly. Stir the solution with a stick of wood until the lye crystals are complete dissolved. Be very careful to avoid any skin contact with the solution. It is also harmful to clothing. Sufficient solution should be prepared to cover the piece completely. When the aluminum is immersed, a very pronounced bubbling takes place and ventilation should be provided to disperse the escaping gas. A half hour to two hours in the solution should be sufficient, depending upon the strength of the solution and the desired surface.

Remove the aluminum from the solution with sticks and rinse thoroughly in cold water while swabbing with a rag to remove the black deposit. Then wipe off with a rag soaked in vinegar to remove any stubborn stains or fingerprints. (See May, 1950, *QST* for a method of coloring and anodizing aluminum.)

DECIMAL EQUIVALENTS OF FRACTIONS

1/32	.03125	17/32	.53125
1/16	.0625	9/16	.5625
3/32	.09375	19/32	.59375
1/8	.125	5/8	.625
5/32	.15625	21/32	.65625
3/16	.1875	11/16	.6875
7/32	.21875	23/32	.71875
1/4	.25	3/4	.75
9/32	.28125	25/32	.78125
5/16	.3125	13/16	.8125
11/32	.34375	27/32	.84375
3/8	.375	7/8	.875
13/32	.40625	29/32	.90625
7/16	.4375	15/16	.9375
15/32	.46875	31/32	.96875
1/2	.5	1	1.0

Soldering

The secret of good soldering is in allowing time for the *joint*, as well as the solder, to attain sufficient temperature. Enough heat should be applied so that the solder will melt when it comes in contact with the wires being joined, without touching the solder to the iron. Always use rosin-core solder, never acid-core. Except where absolutely necessary, solder should never be depended upon for the mechanical strength of the joint; the wire should be wrapped around the terminals or clamped with soldering terminals.

When soldering crystal diodes or carbon resistors in place, especially if the leads have been cut short and the resistor is of the small ½-watt size, the resistor lead should be gripped with a pair of pliers up close to the resistor so that the heat will be conducted away from the resistor. Overheating of the resistor while soldering can cause a permanent resistance change of as much as 20 per cent. Also, mechanical stress will have a similar effect, so that a small resistor should be mounted so that there is no appreciable mechanical strain on the leads.

Trouble is sometimes experienced in soldering to the pins of coil-forms or male cable plugs. It helps first to tin the inside of the pins by applying soldering paste to the hole, and then flowing solder into the pin. Then immediately clear the solder from the hot pin by a whipping motion or by blowing through the pin from the inside of the form or plug. Before inserting the wire in the pin, file the nickel plate from the tip. After soldering, round the solder tip off with a file.

When soldering to sockets, it is a good idea to have the tube or coil form inserted to prevent solder running down into the socket prongs. It also helps to conduct the heat away when soldering to polystyrene sockets, which often soften under the heat of the iron.

Wiring

The wire used in connecting up amateur equipment should be selected considering both the maximum current it will be called upon to handle and the voltage its insulation must stand without breakdown. Also, from the consideration of TVI, the power wiring of all transmitters should be done with wire that has a braided shielding cover. Receiver and audio circuits may also require the use of shielded wire at some points for stability, or the elimination of hum.

No. 20 stranded wire is commonly used for most receiver wiring (except for the high-frequency circuits) where the current does not exceed 2 or 3 amperes. For higher-current heater circuits, No. 18 is available. Wire with cellulose acetate insulation is good for voltages up to about 500. For higher voltages, thermoplastic-insulated wire should be used. Inexpensive wire strippers that make the removal of insulation from hook-up

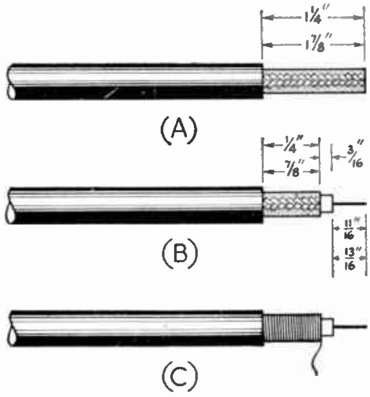


Fig. 24-3 — **Cable-stripping dimensions for Jones Type P-101 plugs. Smaller dimensions are for ¼-inch plugs, the larger dimensions for ½-inch plugs. As indicated in C, the remaining copper braid is wound with bare or tinned wire to make a snug fit in the sleeve of the plug.**

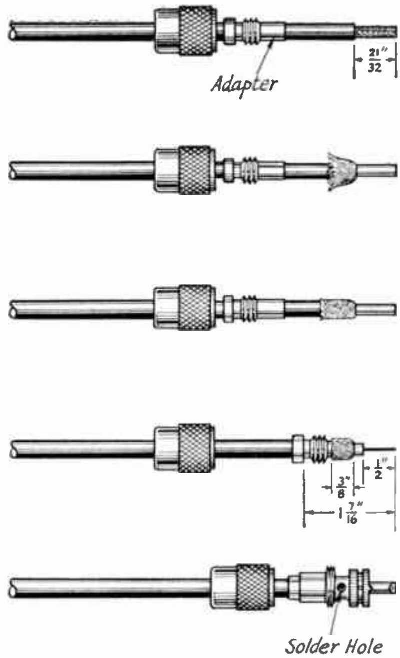


Fig. 24-5 — **Method of assembling ¼-inch cable, Amphenol Type 83-1SP (PL-259) plug and adapter.**

wire an easy job are available on the market.

In cases where power leads have several branches in the chassis, it is convenient to use fiber-insulated tie points or "lug strips" as anchorages or junction points. Strips of this type are also useful as insulated supports for resistors, r.f. chokes and condensers. High-voltage wiring should have exposed points held to a minimum, and those which cannot be avoided should be rendered as inaccessible as possible to accidental contact or short-circuit.

Where shielded wire is called for and capacitance to ground is not a factor, Belden type 8885 shielded grid wire may be used. If capacitance must be minimized, it may be necessary to use a piece of car-radio low-capacitance lead-in wire, or coaxial cable.

For wiring high-frequency circuits, rigid wire is often used. Bare soft-drawn tinned wire, sizes 22 to 12 (depending on mechanical requirements), is suitable. Kinks can be removed by stretching a piece 10 or 15 feet long and then cutting into short lengths that can be handled conveniently. R.f. wiring should be run directly from point to point with a minimum of sharp bends and the wire kept well spaced from the chassis or other grounded metal surfaces. Where the wiring must pass through the chassis or a partition, a clearance hole should be cut and lined with a rubber grommet. In case insulation becomes necessary, varnished cambric tubing (spaghetti) can be slipped over the wire.

In transmitters where the peak voltage does not exceed 2500 volts, the shielded grid wire mentioned above should be satisfactory for power circuits. For higher voltages, Belden type 8656, Birnbach type 1820, or shielded ignition cable can be used. In the case of filament circuits carrying heavy current, it may be necessary to use No. 10 or 12 bare or enameled wire, slipped through spaghetti, and then covered with copper braid pulled tightly over the spaghetti. The chapter

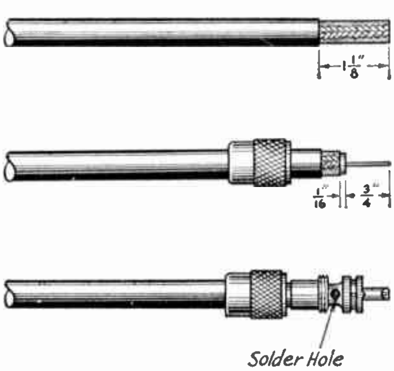


Fig. 24-4 — **Dimensions for stripping ½-inch cable to fit Amphenol Type 83-1SP (PL-259) plug.**

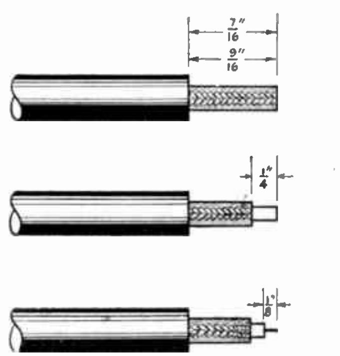


Fig. 24-6 — **Stripping dimensions for Amphenol 82-830 and 82-832 plug-in connectors. The longer exposed braid is for the first type.**

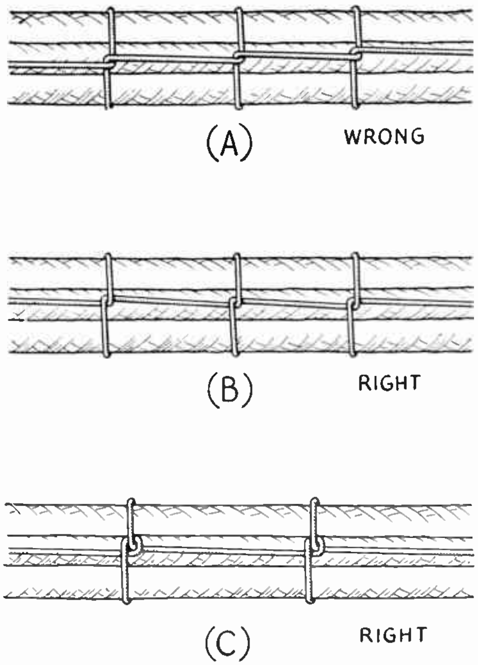


Fig. 24-7 — Methods of lacing cables. The method shown at C is more secure, but takes more time than the method of B. The latter is usually adequate for most amateur requirements.

on TVI shows the manner in which shielded wire should be applied. If the shielding is simply slid back over the insulation and solder flowed into the end of the braid, the braid usually will stay in place without the necessity for cutting it back or binding it in place. The braid should be burnished with sandpaper or a knife so that solder will take with a minimum of heat to protect the insulation underneath.

R.f. wiring in transmitters usually follows the method described above for receivers with due respect to the voltages involved.

Power and control wiring external to the transmitter chassis preferably should be of shielded wire bound into a cable. Fig. 24-7 shows the correct methods of lacing cables.

Coaxial Plug Connections

Considerable time and trouble can be saved in making cable connections to coaxial plugs by starting out with the correct stripping dimensions. Fig. 24-3 shows how the end of the cable should be prepared for connecting to Jones Type P-101 plugs. After the exposed braid has been wound, it should be carefully tinned, applying no more heat than is necessary, to avoid melting the inner insulation. A small amount of solder also should be flowed into the sleeve of the plug. Then, when the cable is inserted in the sleeve, the connection can be made secure by holding the iron against the sleeve until the solder inside melts. While joining the two, the plug may be held by inserting it in a hole drilled in a board. Figs. 24-4, 24-5 and 24-6 show details of connections to different types of Amphenol plugs and adapters. In Fig. 24-4, it is easiest to cut through to the wire with a sharp knife at a distance of 13/16 inch from the end of the wire and remove the insulation and shielding in one piece. Then slice off a 1/16-inch piece of polyethylene which may be slid back onto the wire.

After the braid in Fig. 24-5 has been frayed back, it will be necessary to file the braid down as much as possible to make it fit the plug.

COMPONENT VALUES

Values of composition resistors and small condensers (mica and ceramic) are specified throughout this *Handbook* in terms of "preferred values." In the preferred-number system, all values represent (approximately) a constant-percentage increase over the next lower value. The base of the system is the number 10. Only two significant figures are used. Table 24-II shows the preferred values based on tolerance steps of 20, 10 and 5 per cent. All other values are expressed by multiplying or dividing the base figures given in the table by the appropriate power of 10. (For example, resistor values of 33,000 ohms, 6800 ohms, and 150 ohms are obtained by multiplying the base figures by 1000, 100, and 10, respectively.)

"Tolerance" means that a variation of plus or minus the percentage given is considered satisfactory. For example, the actual resistance of a "4700-ohm" 20-per-cent resistor can lie anywhere between 3700 and 5600 ohms, approximately. The permissible variation in the same resistance value with 5-per-cent tolerance

TABLE 24-II

Standard Component Values

20% Tolerance	*10% Tolerance*	*5% Tolerance*
10	10	10
		11
	12	12
		13
15	15	15
		16
	18	18
		20
22	22	22
		24
	27	27
		30
33	33	33
		36
	39	39
		43
47	47	47
		51
	56	56
		62
68	68	68
		75
	82	82
		91
100	100	100

would be in the range from 4500 to 4900 ohms, approximately.

Only those values shown in the first column of Table 24-II are available in 20-per-cent tolerance. Additional values, as shown in the second column, are available in 10-per-cent tolerance; still more values can be obtained in 5-per-cent tolerance.

In the component specifications in this *Handbook*, it is to be understood that when no tolerance is specified the *largest* tolerance available in that value will be satisfactory.

Values that do not fit into the preferred-number system (such as 500, 25,000, etc.) easily can be substituted. It is obvious, for example, that a 5000-ohm resistor falls well within the tolerance range of the 4700-ohm 20-per-cent resistor used in the example above. It would not, however, be usable if the tolerance were specified as 5 per cent.

COLOR CODES

Standardized color codes are used to mark values on small components such as composition resistors and mica condensers, and to identify leads from transformers, etc. The resistor-condenser number color code is given in Table 24-III.

Fixed Condensers

The methods of marking "postage-stamp" mica condensers, molded paper condensers, and tubular ceramic condensers are shown in Fig. 24-8. Condensers made to American War Standards or Joint Army-Navy specifications are marked with the 6-dot code shown at the top. Practically all surplus condensers are in this category. The 3-dot RETMA code is used for condensers having a rating of 500 volts and ±20% tolerance only; other ratings and tolerances are covered by the 6-dot RETMA code.

Examples: A condenser with a 6-dot code has the following markings: Top row, left to right, black, yellow, violet; bottom row, right to left, brown, silver, red. Since the first color in the top row is black (significant figure zero) this is the AWS code and the condenser has mica dielectric. The significant figures are 4 and 7, the decimal multiplier 10 (brown, at right of second row), so the capacitance is 470 μμfd. The tolerance is ± 10%. The final color, the characteristic, deals with temperature coefficients and methods of testing, and may be ignored.

A condenser with a 3-dot code has the following colors, left to right: brown, black, red. The significant figures are 1, 0 (10) and the multiplier is 100. The capacitance is therefore 1000 μμfd.

A condenser with a 6-dot code has the following markings: Top row, left to right, brown, black, black; bottom row, right to left, black, gold, blue. Since the first color in the top row is neither black nor silver, this is the RETMA code. The significant figures are 1, 0, 0 (100) and the decimal multiplier is 1 (black). The capacitance is therefore 100 μμfd. The gold dot shows that the tolerance is ± 5% and the blue dot indicates 600-volt rating.

Ceramic Condensers

Conventional markings for ceramic condensers are shown in the lower drawing of Fig. 24-8. The colors have the meanings indicated in Table 24-IV. In practice, dots may be used instead of the *narrow* bands indicated in Fig. 24-8.

Example: A ceramic condenser has the following markings: Broad band, violet; narrow bands or dots, green, brown, black, green. The significant figures are 5, 1 (51) and the decimal multiplier is 1, so the capacitance is 51 μμfd. The temperature coefficient is − 750 parts per million per degree C., as given by the broad band, and the capacitance tolerance is ± 5%.

Fixed Composition Resistors

Composition resistors (including small wire-wound units molded in cases identical with the composition type) are color-coded as shown in

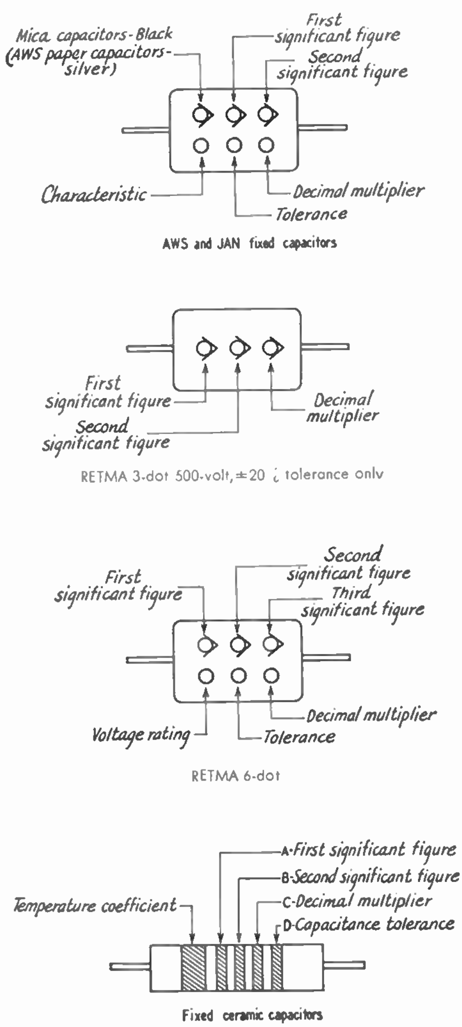
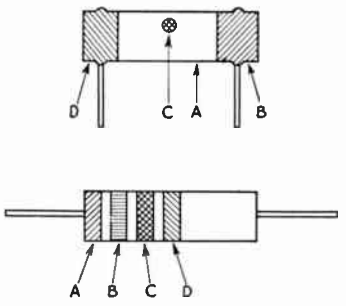


Fig. 24-8 — Color coding of fixed mica, molded paper, and tubular ceramic condensers. The color code for mica and molded paper condensers is given in Table 24-III. Table 24-IV gives the color code for tubular ceramic condensers.



Fixed composition resistors

Fig. 24-9 — Color coding of fixed composition resistors. The color code is given in Table 24-III. The colored areas have the following significance:

A — First significant figure of resistance in ohms
B — Second significant figure.
C — Decimal multiplier.
D — Resistance tolerance in per cent. If no color is shown, the tolerance is ±20%.

Fig. 24-9. Colored bands are used on resistors having axial leads; on radial-lead resistors the colors are placed as shown in the drawing. When bands are used for color coding the body color has no significance.

Examples: A resistor of the type shown in the lower drawing of Fig. 24-9 has the following color bands: A, red; B, red; C, orange; D, no color. The significant figures are 2, 2 (22) and the decimal multiplier is 1000. The value of resistance is therefore 22,000 ohms and the tolerance is ±20%.

A resistor of the type shown in the upper drawing has the following colors: body (A), blue; end (B), gray; dot, red; end (D), gold. The significant figures are 6, 8 (68) and the decimal multiplier is 100, so the resistance is 6800 ohms. The tolerance is ±5%.

I.F. Transformers

Blue — plate lead.
Red — "B" + lead.
Green — grid (or diode) lead.
Black — grid (or diode) return.

Note: If the secondary of the i.f.t. is center-tapped, the second diode plate lead is green-and-black striped, and black is used for the center-tap lead.

TABLE 24-III
Resistor-Condenser Color Code

Color	*Significant Figure*	*Decimal Multiplier*	*Tolerance (%)*	*Voltage Rating**
Black	0	1	–	—
Brown	1	10	1*	100
Red	2	100	2*	200
Orange	3	1000	3*	300
Yellow	4	10,000	4*	400
Green	5	100,000	5*	500
Blue	6	1,000,000	6*	600
Violet	7	10,000,000	7*	700
Gray	8	100,000,000	8*	800
White	9	1,000,000,000	9*	900
Gold	–	0.1	5	1000
Silver	–	0.01	10	2000
No color	–	—	20	500

* Applies to condensers only.

TABLE 24-IV
Color Code for Ceramic Condensers

Color	*Significant Figure*	*Decimal Multiplier*	*Capacitance Tolerance*		*Temp. Coeff. p.p.m./deg. C.*
			More than 10 μμfd. (in %)	*Less than 10 μμfd. (in μμfd.)*	
Black	0	1	± 20	2.0	0
Brown	1	10	± 1		−30
Red	2	100	± 2		−80
Orange	3	1000			−150
Yellow	4				−220
Green	5		± 5	0.5	−330
Blue	6				−470
Violet	7				−750
Gray	8	0.01		0.25	30
White	9	0.1	± 10	1.0	500

A.F. Transformers

Blue — plate (finish) lead of primary.
Red — "B" + lead (this applies whether the primary is plain or center-tapped).
Brown — plate (start) lead on center-tapped primaries. (Blue may be used for this lead if polarity is not important.)
Green — grid (finish) lead to secondary.
Black — grid return (this applies whether the secondary is plain or center-tapped).
Yellow — grid (start) lead on center-tapped secondaries. (Green may be used for this lead if polarity is not important.)

Note: These markings apply also to line-to-grid and tube-to-line transformers.

Loudspeaker Voice Coils

Green — finish
Black — start.

Loudspeaker Field Coils

Black and Red — start.
Yellow and Red — finish.
Slate and Red — tap (if any).

Power Transformers

1) Primary Leads. . . . . *Black*
   If tapped:
   Common. . . . . *Black*
   Tap. . . . . *Black and Yellow Striped*
   Finish. . . . . *Black and Red Striped*
2) High-Voltage Plate Winding. . . . . *Red*
   Center-Tap. . . *Red and Yellow Striped*
3) Rectifier Filament Winding. . . . . . . *Yellow*
   Center-Tap. . *Yellow and Blue Striped*
4) Filament Winding No. 1. . . . . *Green*
   Center-Tap. . *Green and Yellow Striped*
5) Filament Winding No. 2. . . . . *Brown*
   Center-Tap. *Brown and Yellow Striped*
6) Filament Winding No. 3. . . . . *Slate*
   Center-Tap. . . *Slate and Yellow Striped*

CHAPTER 25

Operating a Station

The enjoyment of our hobby usually comes from the operation of our station once we have finished its construction. Upon the *station* and its *operation* depend the communication records that are made.

An operator with a slow, steady, clean-cut method of sending has a big advantage over the poor operator. Good sending is partly a matter of practice but patience and judgment are just as important qualities of an operator as a good "fist." The technique of speaking in connected thoughts and phrases is equally important for the operator who uses voice.

OPERATING COURTESY AND TOLERANCE

Normal operating interests in amateur radio vary considerably. Some prefer to rag-chew, others handle traffic, others work DX, others concentrate on working certain areas, countries or states and still others get on for an occasional contact only to check a new transmitter or antenna.

Interference is one of the things we amateurs have to live with. However, we can conduct our operating in a way designed to alleviate it as much as possible. *Before putting the transmitter on the air, listen on your own frequency.* If you hear stations engaged in communication on that frequency, stand by until you are sure no interference will be caused by your operations, or *shift to another frequency.* No amateur or any group of amateurs has any *exclusive* claim to any frequency in any band. We must work together, each respecting the rights of others. Remember, those other chaps can cause you as much interference as you cause them, sometimes more! Where a VFO is used it is not necessary to stick to a single operating frequency though it is well to have one or two preferred and alternate frequencies. It has become general operating procedure these days to work stations on or near your own frequency. This practice will automatically assist in reducing interference.

C.W. PROCEDURE

The best operators, *both* those using voice and c.w., observe certain operating procedures developed from experience and regarded as "standard practice."

1) *Calls.* Calling stations may call efficiently by transmitting the call signal of the station called three times, the letters DE, followed by one's own station call sent three times. (Short calls with frequent "breaks" to listen have proved to be the best method.) Repeating the call of the station called four or five times and signing not more than two or three times has proved excellent practice, thus: WØBY WØBY WØBY WØBY WØBY DE W1AW W1AW $\overline{\text{AR}}$.

CQ. The general-inquiry call (CQ) should be sent not more than five times without interspersing one's station identification. The length of repeated calls is carefully limited in intelligent amateur operating. (CQ is not to be used when testing or when the sender is not expecting or looking for an answer. Never send a CQ "blind." Always be sure to listen on the transmitting frequency first.)

The directional CQ: To reduce the number of useless answers and lessen QRM, every CQ call should be made informative when possible.

> *Examples:* A United States station looking for any Hawaiian amateur calls: CQ KH6 CQ KH6 CQ KH6 DE W4IA W4IA W4IA K. A Western station with traffic for the East Coast when looking for an intermediate relay station calls: CQ EAST CQ EAST CQ EAST DE W5IGW W5IGW W5IGW K. A station with messages for points in Massachusetts calls: CQ MASS CQ MASS CQ MASS DE W7CZY W7CZY W7CZY K.

Hams who do not raise stations readily may find that their sending is poor, their calls ill-timed or judgment in error. When conditions are right to bring in signals from the desired locality, you can call them. Reasonably short calls, with appropriate and brief breaks to listen, will raise stations with minimum time and trouble.

2) *Answering a Call:* Call three times (or less); send DE; sign three times (or less); after contact is established decrease the use of the call signals of both stations to *once or twice.* When a station receives a call but does not receive the call letters of the station calling, QRZ? may be used. It means "By whom am I being called?" QRZ should not be used in place of CQ.

3) *Ending Signals and Sign-Off:* The proper use of $\overline{\text{AR}}$, K, $\overline{\text{KN}}$, $\overline{\text{SK}}$ and CL ending signals is as follows:

$\overline{\text{AR}}$ — End of transmission. Recommended after call to a specific station before contact has been established.

> *Example:* W6ABC W6ABC W6ABC W6ABC W6ABC DE W9LMN W9LMN $\overline{\text{AR}}$. Also at the end of transmission of a radiogram, immediately following the signature, preceding identification.

K — Go ahead (any station). Recommended after CQ and at the end of each transmission during QSO when there is no objection to others breaking in.

Example: CQ CQ CQ DE W1ABC W1ABC K or W9XYZ DE W1ABC K.

$\overline{\text{KN}}$ — Go ahead (specific station), all others keep out. Recommended at the end of each transmission during a QSO, or after a call, when calls from other stations are not desired and will not be answered.

Example: W4FGH DE XU6GRL $\overline{\text{KN}}$.

$\overline{\text{SK}}$ — End of QSO. Recommended before signing *last* transmission at end of a QSO.

Example: .... $\overline{\text{SK}}$ W8LMN DE W5BCD.

CL — I am closing station. Recommended when a station is going off the air, to indicate that it will not listen for any further calls.

Example: .... $\overline{\text{SK}}$ W7HIJ DE W2JKL CL.

4) *Test signals* to permit another station to adjust receiving equipment may consist of a series of Vs with the call signal of the transmitting station at frequent intervals. Remember that a test signal can be a totally unwarranted cause of QRM, and *always listen first* to find a clear spot if possible.

5) *Receipting* for conversation or traffic: Never send acknowledgment until the transmission has been entirely received. "R" means "All right, OK, I understand *completely*." Use R *only* when *all* is received correctly.

6) *Repeats.* When most of a transmission is lost, a call should be followed by correct abbreviations to ask for repeats. When a few words on the end of a transmission are lost, the *last word received correctly* is given after ?AA, meaning "all after." When a few words at the beginning of a transmission are lost, ?AB for "all before" a stated word should be used. The quickest way to ask for a fill in the middle of a transmission is to send the last word received correctly, a question mark, then the next word received correctly. Another way is to send "?BN [word] and [word]."

Do not send words twice (QSZ) unless it is requested. Send single. Do not fall into the bad habit of sending double *without a request* from fellows you work. Don't say "QRM" or "QRN" when you mean "QRS." Don't CQ unless there is definite reason for so doing. When sending CQ, use judgment.

General Practices

When a station has receiving trouble, the operator asks the transmitting station to "QSV." The letter "R" is often used in place of a decimal point (e.g., "3R5 Mc.") or the colon in time designation (e.g., "2R30 PM"). A long dash is sometimes sent for "zero."

The law concerning superfluous signals should be noted. If you *must* test, disconnect the antenna system and use an equivalent "dummy" antenna. Send your call frequently when operating. Pick a time for adjusting the station apparatus when few stations will be bothered.

The up-to-date amateur station uses "break-in." For best results send at a medium speed. Send evenly with proper spacing. The standard-type telegraph key is best for all-round use. Regular daily practice periods, two or three periods a day, are best to acquire real familiarity and proficiency with code.

No excuse can be made for "garbled" copy. Operators should copy what is sent and refuse to acknowledge a whole transmission until every word has been received correctly. *Good operators do not guess.* "Swing" in a fist is *not* the mark of a good operator. Unusual words are sent twice, the word repeated following the transmission of "?". If not *sure*, a good operator systematically asks for a fill or repeat. Sign your call frequently, interspersed with calls, and at the end of all transmissions.

On Good Sending

Assuming that an operator has learned sending properly, and comes up with a precision "fist" — not fast, but clean, steady, making well-formed rhythmical characters and spacing beautiful to listen to — he then becomes subject to outside pressures to his own possible detriment in everyday operating. He will want to "speed it up" because the operator at the other end is going faster, and so he begins, unconsciously, to run his words together or develops a "swing."

Perhaps one of the easiest ways to get into bad habits is to do too much playing around with special keys. Too many operators spend only enough time with a straight key to acquire "passable" sending, then subject their newly-developed "fists" to the entirely different movements of bugs, side-swipers, electronic keys, or what-have-you. All too often, this results in the ruination of what may have become a very good "fist."

Think about your sending a little. Are you satisfied with it? You should not be — ever. Nobody's sending is perfect, and therefore *every* operator should continually strive for improvement. Do you ever run letters together — like Q for MA, or P for AN — especially when you are in a hurry? Practically everybody does at one time or another. Do you have a "swing"? Any recognizable "swing" is a deviation from perfection. Strive to send like tape sending; copy a W1AW Bulletin and try to send it with the same spacing using a local oscillator on a subsequent transmission.

Check your spacing in characters, between characters and between words occasionally by making a recording of your fist on an inked tape recorder. This will show up your faults as nothing else will. Practice the correction of faults.

USING A BREAK-IN SYSTEM

Break-in avoids unnecessarily long calls, prevents QRM, gives more communication per hour of operating. Brief calls with frequent short pauses for reply can approach (but not equal) break-in efficiency.

A separate receiving antenna facilitates break-in operation. It is only necessary with break-in to pause just a moment with the key up (or to cut the carrier momentarily and pause in a 'phone conversation) to listen for the other station. The click when the carrier is cut off is as effective as the word "break."

C.w. telegraph break-in is usually simple to arrange. With break-in, ideas and messages to be transmitted can be pulled right through the holes in the QRM. Snappy, efficient amateur work with break-in usually requires a separate receiving antenna and arrangement of the transmitter and receiver to eliminate the necessity for throwing switches between transmissions.

In calling, the transmitting operator sends the letters "BK" at frequent intervals during his call so that stations hearing the call may know that break-in is in use and take advantage of the fact. *He pauses at intervals* during his call, to listen for a moment for a reply. If the station being called does not answer, the call can be continued.

With a tap of the key, the man on the receiving end can interrupt (if a word is missed). The other operator is constantly monitoring, awaiting just such directions. It is not necessary that *you* have perfect facilities to take advantage of break-in when the stations you work are break-in-equipped. After any invitation to *break* is given (and at each pause) press your key — and contact can start *immediately*.

VOICE OPERATING

The use of proper procedure to get best results is just as important as in using code. In telegraphy words must be spelled out letter by letter. It is therefore but natural that abbreviations and shortcuts should have come into widespread use. In voice work, however, abbreviations are not necessary, and should have less importance in our operating procedure.

The letter "K" has been agreed to in telegraphic practice so that the operator will not have to pound out the separate letters that spell the words "go ahead." The voice operator can *say* the words "go ahead" or "over," or "come in please."

One laughs on c.w. by spelling out HI. On 'phone *use* a laugh when one is called for. Be natural as you would with your family and friends.

The matter of reporting *readability* and *strength* is as important to 'phone operators as to those using code. With telegraph nomenclature, it is necessary to spell out words to describe signals or use the abbreviated signal reporting system (RST . . . see Chapter Twenty-Six). Using voice, we have the ability to "say it with words." "Readability four, Strength eight" is the best way to give a quantitative report. Reporting can be done so much more meaningfully with ordinary words: "You are weak but you are in the clear and I can understand you, so go ahead," or "Your signal is strong but you are buried under local interference." Why not say it with words?

Voice-Operating Hints

1) Listen before calling.

2) Make short calls with breaks to listen. Avoid long CQs; do not answer any.

3) Use push-to-talk. Give essential data concisely in first transmission.

4) Make reports honest. Use definitions of strength and readability for reference. Make your reports informative and useful. Honest reports and *full* word description of signals save amateur operators from FCC trouble.

5) Limit transmission length. Two minutes or less will convey much information. When three or more stations converse in round tables, brevity is essential.

6) Display sportsmanship and courtesy. Bands are congested . . . make transmissions meaningful . . . give others a break.

7) Check transmitter adjustment . . . avoid AM overmodulation and splatter. Do not radiate when moving VFO frequency or checking NFM swing. Use receiver b.f.o. to check stability of signal. Complete testing before busy hours!

Voice Equivalents to Code Procedure

Voice	*Code*	*Meaning*
Go ahead; over	K	Self-explanatory
Wait; stand by	$\overline{AS}$, QRX	Self-explanatory
Okay	R	Receipt for a correctly-transcribed message or for "solid" transmission with no missing portions

'Phone-Operating Practice

Efficient voice communication, like good c.w. communication, demands good operating. Adherence to certain points "on getting results" will go a long way toward improving our 'phone-band operating conditions.

Use push-to-talk technique. Where possible arrange on-off switches or controls for fast back-and-forth exchanges that emulate the practicality of the wire telephone. This will help reduce the length of transmissions and keep brother amateurs from calling you a "monologuist" — a guy who likes to hear himself talk!

Listen with care. Keep noise and "backgrounds" out of your operating room to facilitate good listening. It is natural to answer the strongest signal, but take time to listen and give some consideration to the *best* signals, regardless of strength. Every amateur cannot run a kilowatt, but there is no reason why every amateur cannot have a signal of good quality, and utilize uniform operating practices to aid in the understandability and ease of his own communications.

Interpose your call regularly and at frequent intervals. Three short calls are better than one long one. In calling CQ, one's call should certainly appear at least once for every five or six CQs. Calls with frequent breaks to listen will save time and be most productive of results. In identifying, always transmit your *own* call *last. Don't* say "This is W1ABC standing by for W2DEF"; say "W2DEF, this is W1ABC, over." FCC regulations show the call of the transmitting station sent *last.*

Include country prefix before call. It is not correct to say "9RRX, this is 1BDI." Correct and legal use is "W9RRX, this is W1BDI." FCC regulations require proper use of calls; stations have been cited for failure to comply with this requirement.

Monitor your own frequency. This helps in timing calls and transmissions. Send when there is a chance of being copied successfully — not when you are merely "more QRM." Timing transmissions is an art to cultivate.

Keep modulation constant. By turning the gain "wide open" you are subjecting anyone listening to the diversion of whatever noises are present in or near your operating room, to say nothing of the possibility of feed-back, echo due to poor acoustics, and modulation excesses due to sudden loud noises. Speak near the microphone, and don't let your gaze wander all over the station causing sharply-varying input to your speech amplifier; at the same time, keep far enough from the microphone so your signal is not modulated by your breathing. Change distance or gain only as necessary to insure uniform transmitter performance without overmodulation, splatter or distortion.

Make connected thoughts and phrases. Don't mix disconnected subjects. Ask questions consistently. Pause and get answers.

Have a pad of paper handy. It is convenient and desirable to jot down questions as they come in the course of discussion in order not to miss any. It will help you to make intelligent to-the-point replies.

Steer clear of inanities and soap-opera stuff. Our amateur radio and also our personal reputation as a serious communications worker depend on us.

Avoid repetition. Don't repeat back what the other fellow has just said. Too often we hear a conversation like this: "Okay on your new antenna there, okay on the trouble you're having with your receiver, okay on the company who just came in with some ice cream, okay . . . [etc.]." Just *say* you received everything OK. Don't try to prove it.

Use phonetics only as required. When clarifying genuinely doubtful expressions and in getting your call identified positively we suggest use of the ARRL Phonetic List. Limit such use to really-necessary clarification.

The speed of radiotelephone transmission (with perfect accuracy) depends almost entirely upon the skill of the two operators involved. One must learn to speak at a rate allowing perfect understanding as well as permitting the receiving operator to copy down the message text, if that is necessary. Because of the similarity of many English speech sounds, the use of alphabetical word lists has been found necessary. All voice-operated stations should use a *standard* list as needed to identify call signals or unfamiliar expressions.

ARRL Word List for Radiotelephony

ADAM	JOHN	SUSAN
BAKER	KING	THOMAS
CHARLIE	LEWIS	UNION
DAVID	MARY	VICTOR
EDWARD	NANCY	WILLIAM
FRANK	OTTO	X-RAY
GEORGE	PETER	YOUNG
HENRY	QUEEN	ZEBRA
IDA	ROBERT	

Example: W1AW . . . W 1 ADAM WILLIAM.

Round Tables. The round table has many advantages if run properly. It clears frequencies of interference, especially if all stations involved are on the same frequency, while the enjoyment value remains the same, if not greater. By use of push-to-talk, the conversation can be kept lively and interesting, giving each station operator ample opportunity to participate without waiting overlong for his turn.

Round tables can become very unpopular if they are not conducted properly. The monologuist, off on a long spiel about nothing in particular, cannot be interrupted; *make your transmissions short and to the point.* "Butting in" is discourteous and unsportsmanlike; *don't enter a round table, or any contact between two other amateurs, unless you are invited.* It is bad enough trying to understand voice through prevailing interference without the added difficulty of poor quality; *check your transmitter adjustments frequently.* In general, follow the precepts as hereinbefore outlined for the most enjoyment in round tables as well as any other form of radiotelephone communication.

● WORKING DX

Most amateurs at one time or another make "working DX" a major aim. As in every other phase of amateur work, there are right and wrong ways to go about getting best results in working foreign stations, and it is the intention of this section to outline a few of them.

The ham who has trouble raising DX stations readily may find that poor transmitter efficiency is not the reason. He may find that his sending is poor, or his calls ill-timed, or his judgment in error. When conditions are right to bring in the DX, and the receiver sensitive enough to bring in several stations from the desired locality, the way to work DX is to use the appropriate frequency and timing and *call these stations,* as against the common practice of calling "CQ DX."

The call CQ DX means slightly different things to amateurs in different bands:

a) On v.h.f., CQ DX is a general call ordinarily used only when the band is open, under favorable "skip" conditions. For v.h.f. work such a call is used for looking for new states and countries, also for distances beyond the customary "line-of-sight" range on most v.h.f. bands.

b) CQ DX on our 7-, 14- and 28-Mc. bands may be taken to mean "General call to any foreign station." The term "foreign station" usually refers to any station in a foreign continent. (*Experienced* amateurs in the U. S. A. and Canada do *not* use this call, but *answer* such calls made by foreign stations.)

DX OPERATING CODE
(For W/VE Amateurs)

Some amateurs interested in DX work have caused considerable confusion and QRM in their efforts to work DX stations. The points below, if observed by all W/VE amateurs, will go a long way toward making DX more enjoyable for everybody.

1. Call DX only after he calls CQ, QRZ?, signs $\overline{SK}$, or 'phone equivalents thereof.
2. Do *not* call a DX station:
   a. On the frequency of the station he is working until you are *sure* the QSO is over. This is indicated by the ending signal $\overline{SK}$ on c.w. and any indication that the operator is listening, on 'phone.
   b. Because you hear someone else calling him.
   c. When he signs $\overline{KN}$, $\overline{AR}$, CL, or 'phone equivalents.
   d. Exactly on his frequency.
   e. After he calls a directional CQ, unless of course you are in the right direction or area.
3. Keep within frequency-band limits. Some DX stations operate outside. Perhaps they can get away with it, but you cannot.
4. Observe calling instructions of DX stations. "10U" means call ten kc. *up* from his frequency, "15D" means 15 kc. *down*, etc.
5. Give honest reports. Many foreign stations *depend* on W and VE reports for adjustment of station and equipment.
6. Keep your signal clean. Key clicks, chirps, hum or splatter give you a bad reputation and may get you a citation from FCC.
7. *Listen* for and *call* the station you want. Calling CQ DX is not the best assurance that the *rare* DX will reply.
8. When there are several W or VE stations waiting to work a DX station, avoid asking him to "listen for a friend." Let your friend take his chances with the rest. Also avoid engaging DX stations in rag-chews against their wishes

c) CQ DX used on 3.5 Mc. under winter-night conditions may be used in this same manner. At other times, under average 3.5-Mc. propagation conditions, the call may be used in domestic work when looking for new states or countries in one's own continent, usually applying to stations located over 1000 miles distant from you.

The way to work DX is not to use a CQ call at *all* (in our continent). Instead, use your best tuning skill — and listen — and listen — and *listen. You have to hear them before you can work them.* Hear the desired stations first; time your calls well. Use your utmost skill. A sensitive receiver is often more important than the power input in working foreign stations. If you can hear stations in a particular country or area, chances are that you will be able to work someone there.

One of the most effective ways to work DX is to know the operating habits of the DX stations sought. Doing too much transmitting on the DX bands is not the way to do this. Again, *listening* is effective. Once you know the operating habits of the DX station you are after you will know when and where to call, and when to remain silent waiting your chance.

Some DX stations use the signals HM, MH, LM and ML to indicate where they are tuning for replies The meanings of these signals are as follows:

HM — Will start to listen at *high*-frequency end of band and tune toward *middle* of band.
MH — Will start to listen in the *middle* of the band and tune toward the *high*-frequency end.
LM — Will start to listen at *low*-frequency end of band and tune toward *middle* of band.
ML — Will start to listen in the *middle* of the band and tune toward the *low*-frequency end.

Example: If the procedure will be to tune from the middle of the band to the high end, a CQ call goes: CQ DE G5BY MH K.

ARRL has recommended some operating procedures to DX stations aimed at controlling some of the thoughtless operating practices sometimes used by W/VE amateurs. A copy of these recommendations (Operating Aid No. 5) can be obtained free of charge from ARRL Headquarters.

In any band, particularly at line-of-sight frequencies, when directional antennas are used, the directional CQ such as CQ W5, CQ north, etc., is the preferable type of call. Mature amateurs agree that CQ DX is a wishful rather than a practical type of call for most stations in the North Americas looking for contacts in foreign countries. Ordinarily, it is a cause of unnecessary QRM.

Conditions in the transmission medium make all field strengths from a given region more nearly equal at a distance, irrespective of power used. In general, the higher the frequency band, the less important power considerations become. This accounts in part for the relative popularity of the 14- and 28-Mc. bands among amateurs who like to work DX.

DATE TIME	STATION CALLED	CALLED BY	HIS FREQ. OR DIAL	HIS SIGNALS RST	MY SIGNALS RST	FREQ. MC.	EMISSION TYPE	POWER INPUT WATTS	TIME OF ENDING QSO	OTHER DATA
11-16-53										
6:15PM	WØTQD	×	3.65	589	569X	3.5	A1	250	6:43	Tfc – rec'd 6, sent 10
7:20	CQ	×				7				
7:21	×	W4TWI	7.16	369	579				7:32	Vy heavy QRM on me
9:25	W8UKS	×	3.83	59	47	3.9	A3	100	10:05	Sam
11-18-53										
7:05AM	VK4EL	×	14.03			14	A1	250		Answered a W6
7:09	ZL2ACV	×	14.07	339	559X				7:20	
7:21	×	KA2KW	14.07	469X	349				7:33	First KA
7:36	CQ	×								
7:37	×	W6TI	14.01	589	589C				8:12	

KEEP AN ACCURATE AND COMPLETE STATION LOG AT ALL TIMES! F.C.C. REQUIRES IT.

A page from the official ARRL log is shown above, answering every Government requirement in respect to station records. Bound logs made up in accord with the above form can be obtained from Headquarters for a nominal sum or you can prepare your own, in which case we offer this form as a suggestion. The ARRL log has a special wire binding and lies perfectly flat on the table.

KEEPING AN AMATEUR STATION LOG

The FCC requires every amateur to keep a complete station operating record. It may also contain records of experimental tests and adjustment data. A stenographer's notebook can be ruled with vertical lines in any form to suit the user. The Federal Communications Commission requirements are that a log be maintained that shows (1) the date and time of *each* transmission, (2) *all calls* and transmissions made (whether two-way contacts resulted or not), (3) the input power to the last stage of the transmitter, (4) the frequency band used, (5) the time of *ending* each QSO and the operator's identifying signature for responsibility for each session of operating. Messages may be written in the log or separate records kept — but record must be made for one year as required by the FCC. For the convenience of amateur station operators ARRL stocks both logbooks and message blanks, and if one uses the official log he is sure to comply fully with the Government requirements if the precautions and suggestions included in the log are followed.

Message Handling

Amateur operators in the United States and a few other countries enjoy a privilege not available to amateurs in most countries — that of handling third-party message traffic. In the early history of amateur radio in this country, some amateurs who were among the first to take advantage of this privilege formed an extensive relay organization which became known as the American Radio Relay League.

Thus, amateur message-handling has had a long and honorable history and, like most services, has gone through many periods of development and change. Those amateurs who handled traffic in 1914 would hardly recognize it the way some of us do it today, just as equipment in those days was far different from that in use now. Progress has been made and new methods have been developed in step with advancement in communication techniques of all kinds. Amateurs who handled a lot of traffic found that organized operating schedules were more effective than random relays, and as techniques advanced and messages increased in number, trunk lines were organized, spot frequencies began to be used, and there sprang into existence a number of traffic nets in which many stations operated on the same frequency to effect wider coverage in less time with fewer relays; but the old methods are still available to the amateur who handles only an occasional message.

Although message handling is as old an art as is amateur radio itself, there are many amateurs who do not know how to handle a message and have never done so. As each amateur grows older and gains experience in the amateur service, there is bound to come a time when he will be called upon to handle a written message, during a communications emergency, in casual contact with one of his many acquaintances on the air, or as a result of a request from a non-amateur friend. Regardless of the occasion, if it comes to you, you will want to rise to it! Considerable embarrassment is likely to be experienced by the amateur who finds he not only does not know the form in which the message should be prepared, but does not know what to do with the message once it has been filed or received in his station.

Traffic work need not be a complicated or time-consuming activity for the casual or occasional message-handler. Amateurs may participate in traffic work to whatever extent they wish, from an occasional message now and then to becoming a part of organized traffic systems.

This chapter explains some principles so the reader may know where to find out more about the subject and may exercise the message-handling privilege to best effect as the spirit and opportunity arise.

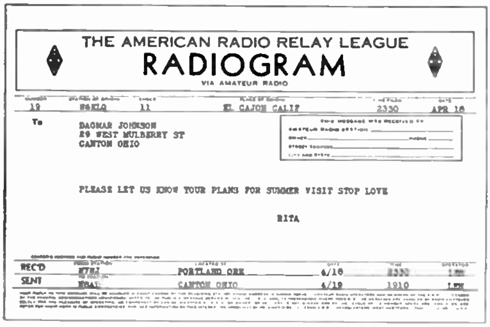
Responsibility

Amateurs who originate messages for transmission or who receive messages for relay or delivery should first consider that in doing so they are accepting the responsibility of clearing the message from their station on its way to its destination in the shortest possible time. Forty-eight hours after filing or receipt is the generally-accepted rule among traffic-handling amateurs, but it is obvious that if every amateur who relayed the message allowed it to remain in his station this long it might be a long time reaching its destination. Traffic should be relayed or delivered as quickly as possible.

Message Form

Once this responsibility is realized and accepted, handling the message becomes a matter of following generally-accepted standards of form and transmission. For this purpose, each message is divided into four parts: the preamble, the address, the text and the signature. Some of these parts themselves are subdivided. It is necessary in preparing the message for transmission and in actually transmitting it to know not only what each part is and what it is for, but to know in what *order* it should be transmitted, and to know the various procedure signals used with it when sent by c.w. If you are going to send a message, you may as well send it right.

Standardization is important! There is a great deal of room for expressing originality and individuality in amateur radio, but there are also times and places where such expression can only cause confusion and inefficiency. Recognizing the need for standardization in message form and message transmitting procedures, ARRL has long since recommended such standards, and most traffic-interested amateurs have followed them. In general, these recommendations, and the various changes they have undergone from year to year, have been at the request of amateurs participating in this activity, and they are completely outlined and explained in *Operating an Amateur Radio Station*, a copy of which is available upon request or by use of the coupon at the end of this chapter.


THE AMERICAN RADIO RELAY LEAGUE
RADIOGRAM
VIA AMATEUR RADIO

19 11 EL CAJON CALIF 2330 APR 18

To DAGMAR JOHNSON
29 WEST MULBERRY ST
CANTON OHIO

PLEASE LET US KNOW YOUR PLANS FOR SUMMER VISIT STOP LOVE

RITA

RECD PORTLAND ORE 4/18
SENT CANTON OHIO 4/19 1910

Here is an example of a plain-language message in correct ARRL form. The preamble is always sent as shown: number, station of origin, check, place of origin, time filed, date.

Clearing a Message

Amateurs not experienced in message handling should depend on the experienced message-handler to get a message through, if it is important; but the average amateur can enjoy operating with a message to be handled either through a local traffic net or by free-lancing. The latter may be accomplished by careful listening for an amateur station at desired points, directional CQs, use of the General Calling frequencies, or by making and keeping a schedule with another amateur for regular work between specified points. He may well aim at learning and enjoying through doing. The joy and accomplishment in thus developing one's operating skill to top perfection has a reward all its own.

The best way to clear a message is to put it into one of the many organized traffic networks, or to give it to a station who can do so. There are many amateurs who make the handling of traffic their principal operating activity, and many more still who participate in this activity to a greater or lesser extent. The result is a system of traffic nets which spreads to all corners of the United States and covers most U. S. possessions and Canada. Once a message gets into one of these nets, regardless of the net's size or coverage, it is systematically routed toward its destination in the shortest possible time.

If you decide to "take the bull by the horns" and put the message into a traffic net yourself (and more power to you if you do!), you will need to know something about how traffic nets operate, and the special Q signals and procedure they use to dispatch all traffic with a maximum of efficiency. Reference to net lists in *QST* (usually in the November and January issues) will give you the frequency and operating time of the net in your section, or other net into which your message can go. Listening for a few minutes at the time and frequency indicated should acquaint you with enough fundamentals to enable you to report into the net and indicate your traffic. From that time on you follow the instructions of the net control station, who will tell you when and to whom (and on what frequency, if different from the net frequency) to send your message. Since most nets use the special "QN" signals, it is usually very helpful to have a list of these before you (list available from ARRL Hq.).

Network Operation

About this time, you may find that you are enjoying this type of operating activity and want to know more about it, and to increase your proficiency. Many amateurs are happily "addicted" to traffic handling after only one or two brief exposures to it. Most traffic nets are at present being conducted by c.w., since this mode of communication seems to be more popular for record purposes — but this does not mean that high code speed is a necessary prerequisite to working in traffic networks. There are many nets organized specifically for the slow-speed amateur, and most of the so-called "fast" nets are usually glad to slow down to accommodate slower operators, especially those nets at state or section level.

The significant facet of net operation, however, is that code speed alone does *not* make for efficiency — sometimes quite the contrary! A high-speed operator who does not know net procedure can "foul up" a net much more completely and more quickly than can a slow operator. It is a proven fact that a bunch of high-speed operators who are not "savvy" in net operation cannot accomplish as much during a specified period as an equal number of slow operators who *know* net procedure. Don't let your code speed deter you from getting into traffic work. Given a little time, your speed will reach the point where you can compete with the best of them. Concentrate first on learning net procedure, for most traffic nowadays is handled on nets.

Much traffic is also being handled on 'phone nowadays. This mode is exceptionally well suited to short-range traffic work and requires knowledge of phonetics and procedure peculiar to voice operation. Procedure is of paramount importance on 'phone, since the public may be listening. The major problem, of course, is QRM.

Teamwork is the theme of net operation. The net which functions most efficiently is the net in which all participants are thoroughly familiar with the procedure used, and in which operators refrain from transmitting except at the direction of the net control station, and do not occupy time with extraneous comments, even exchange of pleasantries. There is a time and place for everything. When a net is in session it should concentrate on handling traffic until all traffic is cleared. Before or after the net is the time for rag-chewing and discussion. Some details of net operation are included in *Operating an Amateur Radio Station*, mentioned earlier, but the whole story cannot be told. There is no substitute for actual participation.

The National Traffic System

To facilitate and speed the movement of message traffic, there is in existence an integrated national system by means of which originated traffic will normally reach its destination area the same day the message is originated. This system uses the local section net as a basis. Each section net sends a representative to a "regional" net (normally covering a call area) and each "regional" net sends a representative to an "area" net (normally covering a time zone). After the area net has cleared all its traffic, its members then go back to their respective regional nets, where they clear traffic to the various section net representatives. By means of connecting schedules between the area nets, traffic can flow both ways so that traffic originated on the West Coast reaches the East Coast with a maximum of dispatch, and vice versa. In general local section nets function at 1900, regional nets at 1945, area nets at 2030 and the same or different regional personnel again at 2130. Some section nets conduct a late session at 2200 to effect traffic delivery the same night. Local time is referred to in each case.

The NTS plan somewhat spreads traffic opportunity so that casual traffic may be reported into nets for efficient handling one or two nights per week, early or late; or the ardent traffic man can operate in *both* early and late groups and in be-between to roll up impressive totals and speed traffic reliably to its destination. Old-time traffic men who prefer a high degree of organization and teamwork have returned to the traffic game as a result of the new system. Beginners have shown more interest in becoming part of a system nationwide in scope, in which *anyone* can participate. The National Traffic System has vast and intriguing possibilities as an amateur service. It is open to any amateur who wishes to participate.

The above is but the briefest résumé of what is of necessity a rather complicated arrangement of nets and schedules. Complete details of the System and its operation are available to anyone interested. Just drop a line to ARRL Headquarters.

Emergency Communication

One of the most important ways in which the amateur serves the public, thus making his existence a national asset, is by his preparation for and his participation in communications emergencies. Every amateur, regardless of the extent of his normal operating activities, should give some thought to the possibility of his being the only means of communication should his community be cut off from the outside world. It has happened many times, often in the most unlikely places; it has happened without warning, finding some amateurs totally unprepared; it can happen to *you*. Are you ready?

There are two principal ways in which any amateur can prepare himself for such an eventuality. One is to provide himself with equipment capable of operating on any type of emergency power (i.e., either a.c. or d.c.), and equipment which can readily be transported to the scene of disaster. Mobile equipment is especially desirable in most emergency situations.

Such equipment, regardless of its elaborateness or modernness, is of little use, however, if it is not used properly and at the right times; and so another way for an amateur to prepare himself for emergencies, by no means less important than the first, is to *learn to operate efficiently*. There are many amateurs who feel that they know how to operate efficiently who find themselves considerably handicapped at the crucial time by not knowing proper procedure, by being unable due to years of casual amateur operation to adapt themselves to snappy, abbreviated transmissions, and by being unfamiliar with message form and routing procedures. It is dangerous to overrate your ability in this respect; it is far better to assume that you have much to learn.

In general it can be said that there is more emergency equipment available than there are operators who know properly how to operate during emergency conditions, for such conditions require clipped, terse procedure with complete break-in on c.w. and fast push-to-talk on 'phone. The casual rag-chewing aspect of amateur radio, however enjoyable and worth-while in its place, must be forgotten at such times in favor of the business at hand. There is only one way to gain experience in this type of operation, and that is by practicing it. During an emergency is no time for practice; it should be done beforehand, as often as possible, on a regular basis.

This leads up to the necessity for emergency organization and preparedness. ARRL has long recognized this necessity and has provided for it. The Section Communications Manager (whose address appears on page 6 of any recent issue of *QST*) is empowered to appoint certain qualified amateurs in his section for the purpose of coördinating emergency communication organization and preparedness in specified areas or communities. This appointee is known as an Emergency Coördinator for the city or town. One is specified for each community. For coördination and promotion at section level a Section Emergency Coördinator arranges for and recommends the appointments of various Emergency Coördinators at activity points throughout the section. Emergency Coördinators organize amateurs in their communities according to local needs for emergency communication facilities.

The community amateurs taking part in the local organization are members of the Amateur Radio Emergency Corps (AREC). *All* amateurs are invited to register in the AREC, whether they are able to play an active part in their local organization or only a supporting rôle. Application blanks are available from your EC, SEC, SCM or direct from ARRL Headquarters. In the event that inquiry reveals no Emergency Coördinator appointed for your community, your SCM would welcome a recommendation either from yourself or from a radio club of which you are a member. By holding an amateur operator license, you have the responsibility both to your community and to amateur radio to uphold the traditions of the service.

Among the League's publications is a booklet entitled *Emergency Communications*. This booklet, while small in size, contains a wealth of information on AREC organization and functions and is invaluable to any amateur participating in emergency or civil defense work. It is free to

Before Emergency

PREPARE yourself by providing a transmitter-receiver set-up together with an emergency power source upon which you can depend.

TEST both the dependability of your emergency equipment and your own operating ability in the annual ARRL Simulated Emergency Test and the several annual on-the-air contests, especially Field Day.

REGISTER your facilities and your availability with your local ARRL Emergency Coördinator. If your community has no EC, contact your local civic and relief agencies and explain to them what the Amateur Service offers the community in time of disaster.

In Emergency

LISTEN before you transmit. Never violate this principle.

REPORT at once to your Emergency Coördinator so that he will have up-to-the-minute data on the facilities available to him. Work with local civic and relief agencies as the EC suggests, offer these agencies your services directly in the absence of an EC.

RESTRICT all on-the-air work in accordance with FCC regulations, Sec. 12.156, whenever FCC "declares" a state of communications emergency.

QRRR is the official ARRL "land SOS," a distress call for emergency only. It is for use *only* by a station seeking assistance.

RESPECT the fact that the success of the amateur effort in emergency depends largely on circuit discipline. The established Net Control Station should be the supreme authority for priority and traffic routing.

CO-OPERATE with those we serve. Be ready to help, but stay off the air unless there is a specific job to be done that you can handle more efficiently than any other station.

COPY all bulletins from W1AW. During time of emergency special bulletins will keep you posted on the latest developments.

After Emergency

REPORT to ARRL Headquarters as soon as possible and as fully as possible so that the Amateur Service can receive full credit. Amateur Radio has won glowing public tribute in many major disasters since 1919. Maintain this record.

AREC members and should be in every amateur's shack. Drop a line to the ARRL Communications Department if you want a copy, or use the coupon at the end of this chapter.

The Radio Amateur Civil Emergency Service

In order to be prepared for any eventuality, FCC and the Federal Civil Defense Administration (FCDA), in collaboration with ARRL, have promulgated the Radio Amateur Civil Emergency Service. RACES is a temporary peacetime service, intended primarily to serve civil defense and to continue operation during any extreme national emergency, such as war. It shares certain segments of frequencies with the regular Amateur Service on a non-exclusive basis. Its regulations have been made a sub-part of the familiar amateur regulations; that is, the present regulations have become sub-part A, the new RACES regulations being added as sub-part B. Copies of both parts are included in the latest edition of the ARRL *License Manual.*

If *every* amateur participated, we would still be far short of the total operating personnel required properly to implement RACES. As the service which bears the responsibility for the successful implementation of this important new function, we face not only the task of installing (and in some cases building) the necessary equipment, but also of the training of thousands of additional people. This can and should be a function of the local unit of the Amateur Radio Emergency Corps under its EC and his assistants, working in close collaboration with the local civil defense organization.

The first step in organizing RACES locally is the appointment of a Radio Officer by the local civil defense director, possibly on the recommendation of his communications officer. A complete and detailed communications plan must be approved successively by local, state and FCDA regional directors, by the FCDA Washington office, and by FCC. Once this has been accomplished, applications for station authorizations under this plan can be submitted direct to FCC. *QST* will carry further information from time to time, and ARRL will keep its field officials fully informed by bulletins as the situation requires. A series of three articles in *QST* for March, April and May, 1953, makes a useful reference and sets the stage for RACES.

In the event of war, civil defense will place great reliance on RACES for radio communications. RACES is an Amateur Service. Its implementation is logically a function of the Amateur Radio Emergency Corps — an *additional* function in peacetime, but probably an exclusive function in wartime. Therefore, your best opportunity to be of service will be to register with your local EC, and to participate *actively* in the local AREC/RACES program.

ARRL Operating Organization

Amateur operation must have point and constructive purpose to win public respect. Each individual amateur is the ambassador of the entire fraternity in his public relations and attitude toward his hobby. ARRL field organization adds point and purpose to amateur operating.

The Communications Department of the League is concerned with the practical operation of stations in all branches of amateur activity. Appointments or awards are available for rag-chewer, traffic enthusiast, 'phone operator, DX man and experimenter.

There are seventy-three ARRL Sections in the League's field organization, which embraces the United States, Canada and certain other territory. Operating affairs in each Section are supervised by a Section Communications Manager elected by members in that section for a two-year term of office. Organization appointments are made by the section managers, elected as provided in the Rules and Regulations of the Communications Department, which accompany the League's By-Laws and Articles of Association. Section communications managers' addresses for all sections are given in full in each issue of *QST*. SCMs welcome monthly activity reports from all amateur stations in their jurisdiction.

Whether your activity embraces 'phone or telegraphy, or both, there is a place for you in League organization.

● LEADERSHIP POSTS

To advance each type of station work and group interest in amateur radio, and to develop practical communications plans with the greatest success, appointments of leaders and organizers in particular single-interest fields are made by SCMs. Each leadership post is important. Each provides activities and assistance for appointee groups and individual members along the lines of natural interest. Some posts further the general ability of amateurs to communicate efficiently at all times, by pointing activity toward networks and round tables, others are aimed specifically at establishment of provisions for organizing the amateur service as a stand-by communications group to serve the public in disaster, civil defense need or emergency of any sort. The SCM appoints the following in accordance with section needs and individual qualifications:

PAM	'Phone Activities Manager. Organizes activities for OPSs and voice operators in his section. Promotes 'phone nets and recruits OPSs.
RM	Route Manager. Organizes and coördinates c.w. traffic activities. Supervises and promotes nets and recruits ORSs.
SEC	Section Emergency Coördinator. Promotes and administers section emergency radio organization.
EC	Emergency Coördinator. Organizes amateurs of a community or other area for emergency radio service; maintains liaison with officials and agencies served; also with other local communication facilities.

STATION APPOINTMENTS

ARRL's field organization has a place for every active amateur who has a station. The Communications Department organization exists to increase individual enjoyment and station effectiveness in amateur radio work, and we extend a cordial invitation to every amateur to participate fully in the activities and to apply to the SCM for one of the following station appointments. ARRL Membership and the General Class license or VE equivalent is prerequisite to appointments, except OES is available to Novice/Technician grades.

OPS	Official 'Phone Station. Sets high voice operating standards and procedures, furthers 'phone nets and traffic.
ORS	Official Relay Station. Traffic service, operates c.w. nets; noted for 15 w.p.m. and procedure ability.
OBS	Official Bulletin Station. Transmits ARRL and FCC bulletin information to amateurs.
OES	Official Experimental Station. Experimental operating, collects and reports v.h.f.-u.h.f.-s.h.f. propagation data, may engage in facsimile, TT, TV, etc., experiments working on 50 Mc. and/or above.
OO	Official Observer. Sends coöperative notices to amateurs to assist in frequency observance, insures high-quality signals, and prevents FCC trouble.

Emblem Colors

Members wear the emblem with black-enamel background. A red background for an emblem will indicate that the wearer is SCM. SEC's, ECs, RMs, PAMs may wear the emblem with green background. Observers and all *station* appointees are entitled to wear blue emblems.

SECTION NETS

Amateurs can add much experience and pleasure to their own amateur lives, and substance and accomplishment to the credit of all of amateur radio, when organized into effective interconnection of cities and towns.

The successful operation of a net depends a lot on the Net Control Station. This station should be chosen carefully and be one that will not hesitate to enforce each and every net rule and set the example in his own operation.

A progressive net grows, obtaining new members both directly and through other net members. Bulletins may be issued at intervals to keep in direct contact with the members regarding general net activity, to keep tab on net procedure, make suggestions for improvement, keep track of active members and weed out inactive ones.

A National Traffic System is sponsored by ARRL to facilitate the over-all expeditious relay and delivery of message traffic. The system recognizes the need for handling traffic beyond the section-level networks that have the popular support of both 'phone and c.w. groups (OPS and ORS) throughout the League's field organization. Area and regional provisions for NTS are furthered by Headquarters correspondence. The ARRL Net Directory, revised in December each year, includes the frequencies and times of operation of the hundreds of different nets operating on amateur band frequencies.

Radio Club Affiliation

ARRL is pleased to grant affiliation to any amateur society having (1) at least 51% of the voting club membership as full members of the League, and (2) at least 51% of society government-licensed radio amateurs. In high school radio clubs bearing the school name, the first above requirement is modified to require one full member, ARRL, in the club. Where a society has common aims and wishes to add strength to that of other club groups and strengthen amateur radio by affiliation with the national amateur organization, a request addressed to the Communications Manager will bring the necessary forms and information to initiate the application for affiliation. Such clubs receive field-organization bulletins and special information at intervals for posting on club bulletin boards or for relay to their memberships. A travel plan providing communications, technical and secretarial contact from the Headquarters is worked out seasonally to give maximum benefits to as many as possible of the several hundred active *affiliated* radio clubs. Papers on club work, suggestions for organizing, for constitutions, for radio courses of study, etc., are available on request.

Club Training Aids

One section of the ARRL Communications Department handles the Training Aids Program. This program is a service to ARRL affiliated clubs. Material is aimed at education, training and entertainment of club members. Interesting quiz material is available.

Training Aids include such items as motion-picture films, film strips, slides, and lecture outlines. Also, code-proficiency training equipment such as recorders, tape transmitters and tapes will be loaned when such items are available.

All Training Aids materials are loaned free (except for shipping charges) to ARRL affiliated clubs. Numerous groups use this ARRL service to good advantage. If your club is affiliated but has not yet taken advantage of this service, you are missing a good chance to add the available features to your meeting programs and general club activities. Watch club bulletins and *QST* or write the ARRL Communications Department for full details.

W1AW

The Maxim Memorial Station, W1AW, is dedicated to fraternity and service. Operated by the League headquarters, W1AW is located about four miles south of the Headquarters offices on a seven-acre site. The station is on the air daily, except holidays, and available time is divided between different bands and modes.

Telegraph and 'phone transmitters are provided for all bands from 1.8 to 144 Mc. The normal frequencies in each band for c.w. and voice transmissions are as follows: 1885, 3555, 3950, 7125, 7255, 14,100, 14,280, 21,020, 21,350, 28,060, 28,768, 52,000 and 145,600 kc. Operating-visiting hours and the station schedule are listed every other month in *QST*.

All amateurs are invited to visit W1AW, as well as to work the station from their own shacks. The station was established to be a living memorial to Hiram Percy Maxim and to carry on the work and traditions of the amateur fraternity.

OPERATING ACTIVITIES

Within the ARRL field organization there are several special activities. The first Saturday and Sunday of each month is set aside for all ARRL officials, officers and directors to get together over the air from their own stations. This activity is known to the gang as the LO party. For all appointees, other quarterly tests are scheduled to develop operating ability and a spirit of fraternalism.

In addition to these special activities for appointees and members, ARRL sponsors various other activities open to all amateurs. The DX-minded amateur may participate in the Annual ARRL International DX Competition during February and March. This popular contest may bring you the thrill of working new countries. Then there is the ever-popular Sweepstakes in November. Of domestic scope, the SS affords the opportunity to work new states for that WAS award. A Novice activity is planned annually. The interests of v.h.f. enthusiasts are also provided for in special activities planned by ARRL.

As in all our operating, the idea of having a good time is combined in the Annual Field Day with the more serious thought of preparing ourselves to render public service in times of emergency. A premium is placed on the use of equipment without connection to commercial power sources. Clubs and individual groups always have a good time in the "FD," learn much about the requirements for operating under knockabout conditions afield.

ARRL contest activities are diversified to appeal to all operating interests, and will be found announced in detail in issues of *QST* preceding the different events.

AWARDS

The League-sponsored operating activities heretofore mentioned have useful objectives and provide much enjoyment for members of the fraternity. Achievement in amateur radio is recognized by various certificates offered through the League and detailed below.

WAS Award

WAS means "Worked All States." This award is available regardless of affiliation or nonaffiliation with any organization. Here are the rules to follow in applying for WAS:

1) Two-way communication must be established on the amateur bands with each of the states; any and all amateur bands may be used. A card from the District of Columbia may be submitted in lieu of one from Maryland.

2) Contacts with all states must be made from the same location. Within a given community one location may be defined as from places no two of which are more than 25 miles apart.

3) Contacts may be made over any period of years, and may have been made any number of years ago, provided only that all contacts are from the same location.

4) QSL cards, or other written communications from stations worked confirming the necessary two-way contacts, must be submitted by the applicant to ARRL headquarters.

5) Sufficient postage must be sent with the confirmations to finance their return. No correspondence will be returned unless sufficient postage is furnished.

6) The WAS award is available to all amateurs.

7) Address all applications and confirmations to the Communications Department, ARRL, 38 La Salle Road, West Hartford, Conn.

DX Century Club Award

Here are the rules under which the DX Century Club Award will be issued to amateurs who have worked and confirmed contact with 100 countries in the postwar period. If you worked fewer than 100 countries before the war and have since worked and confirmed a sufficient number to make the 100 mark, the DXCC is still available to you under the rules detailed on page 74 of June, 1946, *QST*.

1) The Century Club Award Certificate for confirmed contacts with 100 or more countries is available to all amateurs everywhere in the world.

2) Confirmations must be submitted direct to ARRL headquarters for all countries claimed. Claims for a total of 100 countries must be included with first application. Confirmation from foreign contest logs may be requested in the case of the ARRL International DX Competition only, subject to the following conditions:

a) Sufficient confirmations of other types must be submitted so that these, plus the DX Contest confirmations, will total 100. In every case, Contest confirmations must not be requested for any countries from which the applicant has regular confirmations. That is, contest confirmations will be granted only in the case of countries from which applicants have no regular confirmations.

b) Look up the contest results as published in *QST* to see if your man is listed in the foreign scores. If he isn't, he did not send in a log and no confirmation is possible.

c) Give year of contest, date and time of QSO.

d) In future DX Contests do not request confirmations until after the final results have been published, usually in one of the early fall issues. Requests before this time must be ignored.

3) The ARRL Countries List, printed periodically in *QST*, will be used in determining what constitutes a "country." The Miscellaneous Data chapter of this *Handbook* contains the Postwar Countries List.

4) Confirmations must be accompanied by a list of claimed countries and stations to aid in checking and for future reference.

5) Confirmations from additional countries may be submitted for credit each time ten additional confirmations are available. Endorsements for affixing to certificates and showing the new confirmed total (110, 120, 130, etc.) will be awarded as additional credits are granted. ARRL DX Competition logs from foreign stations may be utilized for these endorsements, subject to conditions stated under (2).

6) All contacts must be made with amateur stations working in the authorized amateur bands or with other stations licensed to work amateurs.

7) In cases of countries where amateurs are licensed in the normal manner, credit may be claimed only for stations using regular government-assigned call letters. No credit may be claimed for contacts with stations in any countries in which amateurs have been temporarily closed down by special government edict where amateur licenses were formerly issued in the normal manner.

8) All stations contacted must be "land stations" . . . contacts with ships, anchored or otherwise, and aircraft, cannot be counted.

9) All stations must be contacted from the same call area, where such areas exist, or from the same country in cases where there are no call areas. One exception is allowed to this rule: where a station is moved from one call area to another, or from one country to another, all contacts must be made from within a radius of 150 miles of the initial location.

10) Contacts may be made over any period of years from November 15, 1945, provided only that all contacts be made under the provisions of Rule 9, and by the same station licensee; contacts may have been made under different call letters in the same area (or country), if the licensee for all was the same.

11) All confirmations must be submitted exactly as received from the stations worked. Any altered or forged confirmations submitted for CC credit will result in disqualification of the applicant. The eligibility of any DXCC applicant who was ever barred from DXCC to reapply, and the conditions for such application, shall be determined by the Awards Committee. Any holder of the Century Club Award submitting forged or altered confirmations must forfeit his right to be considered for further endorsements.

12) OPERATING ETHICS: Fair play and good sportsmanship in operating are required of all amateurs working toward the DX Century Club Award. In the event of specific objections relative to continued poor operating ethics an individual may be disqualified from the DXCC by action of the ARRL Awards Committee.

13) Sufficient postage for the return of confirmations must be forwarded with the application. In order to insure the safe return of large batches of confirmations, it is suggested that enough postage be sent to make possible their return by first-class mail, registered.

14) Decisions of the ARRL Awards Committee regarding interpretation of the rules as here printed or later amended shall be final.

15) Address all applications and confirmations to the Communications Department, ARRL, 38 La Salle Road, West Hartford 7, Conn.

WAC Award

The International Amateur Radio Union issues WAC (Worked All Continents) certificates to all members of member-societies who submit proof of two-way communication with at least one station on each continent. Foreign amateurs submit their proof direct to member-societies of the IARU. Others may make application to ARRL, headquarters society of the Union. A c.w. and a telephony certificate are available. Also, special endorsements will be placed on certificates upon receipt of request accompanied by proof of having worked all continents on the 3.5- or 50-Mc. bands.

Code Proficiency Award

Many hams can follow the general idea of a contact "by ear" but when pressed to "write it down" they "muff" the copy. The Code Proficiency Award invites every amateur to prove himself as a proficient operator, and sets up a system of awards for step-by-step gains in copying proficiency. It enables every amateur to check his code proficiency, to better that proficiency, and to receive a certification of his receiving speed.

This program is a whale of a lot of fun. The League will give a certificate to any licensed radio amateur who demonstrates that he can copy perfectly, for at least one minute, plain-language Continental code at 10, 15, 20, 25, 30 or

35 words per minute, as transmitted during special monthly transmissions from W1AW and W6OWP.

As part of the ARRL Code Proficiency program W1AW transmits plain-language practice material each evening at speeds from 5 to 35 w.p.m. All amateurs are invited to use these transmissions to increase their code-copying ability. Non-amateurs are invited to utilize the lower speeds, 5, 7½ and 10 w.p.m., which are transmitted for the benefit of persons studying the code in preparation for the amateur license examination. Refer to any issue of *QST* for details of the practice schedule.

Rag Chewers Club

The Rag Chewers Club is designed to encourage friendly contacts and discourage the "hello-good-by" type of QSO. Its purpose is to bond together operators interested in honest-to-goodness rag-chewing over the air. Membership certificates are available.

How To Get in: (1) Chew the rag with a member of the club for at least a solid half hour. This does not mean a half hour spent in trying to get a message over through bad QRM or QRN, but a solid half hour of conversation or message handling. (2) Report the conversation by card to The Rag Chewers Club, ARRL, Communications Department, West Hartford, Conn., and ask the member station you talk with to do the same. When both reports are received you will be sent a membership certificate entitling you to all the privileges of a Rag Chewer.

How To Stay in: (1) Be a conversationalist on the air instead of one of those tongue-tied infants who don't know any words except "cuagn" or "cul," or "QRU" or "nil." Talk to the fellows you work with and get to know them. (2) Operate your station in accordance with the radio laws and ARRL practice. (3) Observe rules of courtesy on the air (4) Sign "RCC" after each call so that others may know you can talk as well as call.

A-1 Operator Club

The A-1 Operator Club should include in its ranks every good operator. To become a member, one must be nominated by at least two operators who already belong. General keying or voice technique, procedure, copying ability, judgment and courtesy all count in rating candidates under the club rules detailed at length in *Operating an Amateur Radio Station*. Aim to make yourself a fine operator, and one of these days you may be pleasantly surprised by an invitation to belong to the A-1 Operator Club, which carries a worth-while certificate in its own right.

Brass Pounders League

Every individual reporting more than a specified minimum in official monthly traffic totals is given an honor place in the *QST* listing known as the Brass Pounders League and a certificate to recognize his performance is furnished by the SCM.

The value to amateurs in operator training, and the utility of amateur message handling to the members of the fraternity itself as well as to the general public, make message-handling work of prime importance to the fraternity. Fun, enjoyment, and the feeling of having done something really worth while for one's fellows is accentuated by pride in message files, records, and letters from those served.

Old Timers Club

The Old Timers Club is open to anyone who holds an amateur call at the present time, and who held an amateur license (operator or station) 20-or-more years ago. Lapses in activity during the intervening years are permitted.

If you can qualify as an "Old Timer," send us a brief chronology of your ham career, being sure to indicate the date of your first amateur license, and your present call. If the evidence submitted proves you eligible for the OTC, you will be added to the roster and will receive a membership certificate.

● INVITATION

Amateur radio is capable of giving enjoyment, self-training, social and organization benefits in proportion to what the individual amateur puts into his hobby. All amateurs are invited to become ARRL members, to work toward awards, and to accept the challenge and invitation offered in field-organization appointments. Drop a line to ARRL Headquarters for the booklet *Operating an Amateur Radio Station*, which has detailed information on the field-organization appointments and awards. Accept today the invitation to take full part in all League activities and organization work.

SEE NEXT PAGE ➔

▶ *Operating an Amateur Radio Station* covers the details of practical amateur operating. In it you will find information on Operating Practices, Emergency Communication, ARRL Operating Activities and Awards, the ARRL Field Organization, Handling Messages, Network Organization, "Q" Signals and Abbreviations used in amateur operating, important extracts from the FCC Regulations, and other helpful material. It's a handy reference that will serve to answer many of the questions concerning operating that arise during your activities on the air.

▶ *Emergency Communications* is the "bible" of the Amateur Radio Emergency Corps. Within its eight pages are contained the fundamentals of emergency communication which every amateur interested in public service work should know, including a complete diagrammatical plan adaptable for use in any community, explanation of the role of the American Red Cross and FCC's regulations concerning amateur operation in emergencies. The Radio Amateur Civil Emergency Service (RACES) comes in for special consideration, including a complete table of RACES frequencies on the front cover. If you don't already have an up-to-date copy of this manual, we suggest you take steps to obtain one immediately.

The two publications described above may be obtained without charge by any *Handbook* reader. Either or both will be sent upon request.

AMERICAN RADIO RELAY LEAGUE
38 La Salle Road
West Hartford 7, Connecticut, U. S. A.

Please send me, without charge, the following:

☐ **OPERATING AN AMATEUR RADIO STATION**
☐ **EMERGENCY COMMUNICATIONS**

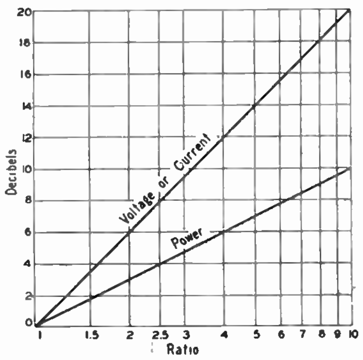
***Name*...**
(Please Print)

***Address*..**

....

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 a 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.** A change of one decibel (abbreviated **db.**) in the power level is just detectable as a change in loudness under ideal conditions. The power ratio and decibels are related by the following formula:

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

Common logarithms (base 10) are used.

Note that the decibel is based on *power* ratios. Voltage or current ratios can be used, but *only when the impedance is the same for both values of voltage, or current.* The gain of an amplifier cannot be expressed correctly in db. if it is based on the ratio of the output voltage to the input voltage unless both voltages are measured across the same value of impedance. When the impedance at both points of measurement is the same, the following formula may be used for voltage or current ratios:

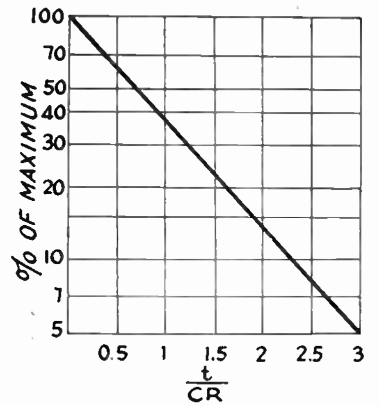
$$Db. = 20 \log \frac{V_2}{V_1}$$

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

The two formulas are shown graphically in the accompanying chart for ratios from 1 to 10. Gains (increases) expressed in decibels may be added arithmetically; losses (decreases) may be subtracted. A power decrease is indicated by prefixing the decibel figure with a minus sign. Thus +6 db. means that the power has been multiplied by 4, while −6 db. means that the power has been divided by 4. The chart may be used for other ratios by adding (or subtracting, if a loss) 10 db. each time the ratio scale is multiplied by 10, for power ratios; or by adding (or subtracting) 20 db. each time the scale is multiplied by 10 for voltage or current ratios.

VOLTAGE DECAY IN *RC* CIRCUITS

The accompanying chart enables calculation of the instantaneous voltage across the terminals of a condenser discharging through a resistance. The voltage is given in terms of percentage of the voltage to which the condenser is initially charged. To obtain the voltage-decay time in seconds, multiply the factor (t/CR) by the time constant of the resistor-condenser circuit.



Example: A 0.01-μfd. condenser is charged to 150 volts and then allowed to discharge through a 0.1-megohm resistor. How long will it take the voltage to fall to 10 volts? In percentage, 10/150 = 6.7%. From the chart, the factor corresponding to 6.7% is 2.7. The time constant of the circuit is equal to $CR = 0.01 \times 0.1 = 0.001$. The time is therefore $2.7 \times 0.001 = 0.0027$ second, or 2.7 milliseconds.

Example: An *RC* circuit is desired in which the voltage will fall to 50% of the initial value in 0.1 second. From the chart, $t/CR = 0.7$ at the 50%-voltage point. Therefore $CR = t/0.7 = 0.1/0.7 = 1.43$. Any combination of resistance and capacitance whose product (*R* in megohms and *C* in microfarads) is equal to 1.43 can be used; for example, *C* could be 1 μfd. and *R* 1.43 megohms.

STANDARD METAL GAUGES

Gauge No.	American or B. & S.[1]	U. S. Standard[2]	Birmingham or Stubs[3]
1	.2893	.28125	.300
2	.2576	.265625	.284
3	.2294	.25	.259
4	.2043	.234375	.238
5	.1819	.21875	.220
6	.1620	.203125	.203
7	.1443	.1875	.180
8	.1285	.171875	.165
9	.1144	.15625	.148
10	.1019	.140625	.134
11	.09074	.125	.120
12	.08081	.109375	.109
13	.07196	.09375	.095
14	.06408	.078125	.083
15	.05707	.0703125	.072
16	.05082	.0625	.065
17	.04526	.05625	.058
18	.04030	.05	.049
19	.03589	.04375	.042
20	.03196	.0375	.035
21	.02846	.034375	.032
22	.02535	.03125	.028
23	.02257	.028125	.025
24	.02010	.025	.022
25	.01790	.021875	.020
26	.01594	.01875	.018
27	.01420	.0171875	.016
28	.01264	.015625	.014
29	.01126	.0140625	.013
30	.01003	.0125	.012
31	.008928	.0109375	.010
32	.007950	.01015625	.009
33	.007080	.009375	.008
34	.006350	.00859375	.007
35	.005615	.0078125	.005
36	.005000	.00703125	.004
37	.004453	.006640626	....
38	.003965	.00625	....
39	.003531	......	....
40	.003145	......	....

[1] Used for aluminum, copper, brass and nonferrous alloy sheets, wire and rods.
[2] Used for iron, steel, nickel and ferrous alloy sheets, wire and rods.
[3] Used for seamless tubes; also by some manufacturers for copper and brass.

MUSICAL SCALE

Approximate frequencies of notes of the musical scale, based on A-440.

(Bottom Octave)

	Note	Frequency		Note	Frequency
	A-1	28	Middle C — First octave above Middle C	C3	262
	A#-1	29		C#3	277
	B-1	31		D3	294
Third octave below Middle C	Co	33		D#3	311
	C#o	35		E3	330
	Do	37		F3	349
	D#o	39		F#3	370
	Eo	41		G3	392
	Fo	44		G#3	415
	F#o	46		A3	440
	Go	49		A#3	466
	G#o	52		B3	494
	Ao	55	Second octave above Middle C	C4	523
	A#o	58		C#4	554
	Bo	62		D4	587
Second octave below Middle C	C1	65		D#4	622
	C#1	69		E4	659
	D1	73		F4	698
	D#1	78		F#4	740
	E1	82		G4	784
	F1	87		G#4	831
	F#1	93		A4	880
	G1	98		A#4	932
	G#1	104		B4	988
	A1	110	Third octave above Middle C	C5	1047
	A#1	117		C#5	1109
	B1	123		D5	1175
First octave below Middle C	C2	131		D#5	1245
	C#2	139		E5	1319
	D2	147		F5	1397
	D#2	156		F#5	1480
	E2	165		G5	1568
	F2	175		G#5	1661
	F#2	185		A5	1760
	G2	196		A#5	1865
	G#2	208		B5	1976
	A2	220	Fourth octave above Middle C	C6	2093
	A#2	233		C#6	2217
	B2	247		D6	2349
				D#6	2489
				E6	2637
				F6	2794
				F#6	2960
				G6	3136
				G#6	3322
				A6	3520
				A#6	3729
				B6	3951
				C7	4186

GREEK ALPHABET

Greek Letter	Greek Name	English Equivalent	Greek Letter	Greek Name	English Equivalent
Α α	Alpha	a	Ν ν	Nu	n
Β β	Beta	b	Ξ ξ	Xi	x
Γ γ	Gamma	g	Ο ο	Omicron	ŏ
Δ δ	Delta	d	Π π	Pi	p
Ε ε	Epsilon	e	Ρ ρ	Rho	r
Ζ ζ	Zeta	z	Σ σ	Sigma	s
Η η	Eta	ē	Τ τ	Tau	t
Θ θ	Theta	th	Υ υ	Upsilon	u
Ι ι	Iota	i	Φ φ	Phi	ph
Κ κ	Kappa	k	Χ χ	Chi	ch
Λ λ	Lambda	l	Ψ ψ	Psi	ps
Μ μ	Mu	m	Ω ω	Omega	ō

COPPER-WIRE TABLE

Gauge No. B. & S.	Diam. in Mils[1]	Circular Mil Area	Turns per Linear Inch[2]				Turns per Square Inch[2]			Feet per Lb.		Ohms per 1000 ft. 25° C.	Current Carrying Capacity[3] at 700 C.M. per Amp.	Diam. in mm.	Nearest British S.W.G. No.
			Enamel	S.S.C.	D.S.C. or S.C.C.	D.C.C.	S.C.C.	Enamel S.C.C.	D.C.C.	Bare	D.C.C.				
1	289.3	83690	—	—	—	—	—	—	—	3.947	—	.1264	119.6	7.348	1
2	257.6	66370	—	—	—	—	—	—	—	4.977	—	.1593	94.8	6.544	3
3	229.4	52640	—	—	—	—	—	—	—	6.276	—	.2009	75.2	5.827	4
4	204.3	41740	—	—	—	—	—	—	—	7.914	—	.2533	59.6	5.189	5
5	181.9	33100	—	—	—	—	—	—	—	9.980	—	.3195	47.3	4.621	7
6	162.0	26250	—	—	—	—	—	—	—	12.58	—	.4028	37.5	4.115	8
7	144.3	20820	—	—	—	—	—	—	—	15.87	—	.5080	29.7	3.665	9
8	128.5	16510	7.6	—	7.4	7.1	—	—	—	20.01	19.6	.6405	23.6	3.264	10
9	114.4	13090	8.6	—	8.2	7.8	—	—	—	25.23	24.6	.8077	18.7	2.906	11
10	101.9	10380	9.6	—	9.3	8.9	87.5	84.8	80.0	31.82	30.9	1.018	14.8	2.588	12
11	90.74	8234	10.7	—	10.3	9.8	110	105	97.5	40.12	38.8	1.284	11.8	2.305	13
12	80.81	6530	12.0	—	11.5	10.9	136	131	121	50.59	48.9	1.619	9.33	2.053	14
13	71.96	5178	13.5	—	12.8	12.0	170	162	150	63.80	61.5	2.042	7.40	1.828	15
14	64.08	4107	15.0	—	14.2	13.8	211	198	183	80.14	77.3	2.575	5.87	1.628	16
15	57.07	3257	16.8	—	15.8	14.7	262	250	223	101.4	97.3	3.247	4.65	1.450	17
16	50.82	2583	18.9	18.9	17.9	16.4	321	306	271	127.9	119	4.094	3.69	1.291	18
17	45.26	2048	21.2	21.2	19.9	18.1	397	372	329	161.3	150	5.163	2.93	1.150	18
18	40.30	1624	23.6	23.6	22.0	19.8	493	454	399	203.4	188	6.510	2.32	1.024	19
19	35.89	1288	26.4	26.4	24.4	21.8	592	553	479	256.5	237	8.210	1.84	.9116	20
20	31.96	1022	29.4	29.4	27.0	23.8	775	725	625	323.4	298	10.35	1.46	.8118	21
21	28.46	810.1	33.1	32.7	29.8	26.0	940	895	754	407.8	370	13.05	1.16	.7230	22
22	25.35	642.4	37.0	36.5	34.1	30.0	1150	1070	910	514.2	461	16.46	.918	.6438	23
23	22.57	509.5	41.3	40.6	37.6	31.6	1400	1300	1080	648.4	584	20.76	.728	.5733	24
24	20.10	404.0	46.3	45.3	41.5	35.6	1700	1570	1260	817.7	745	26.17	.577	.5106	25
25	17.90	320.4	51.7	50.4	45.6	38.6	2060	1910	1510	1031	903	33.00	.458	.4547	26
26	15.94	254.1	58.0	55.6	50.2	41.8	2500	2300	1750	1300	1118	41.62	.363	.4049	27
27	14.20	201.5	64.9	61.5	55.0	45.0	3030	2780	2020	1639	1422	52.48	.288	.3606	29
28	12.64	159.8	72.7	68.6	60.2	48.5	3670	3350	2310	2067	1759	66.17	.228	.3211	30
29	11.26	126.7	81.6	74.8	65.4	51.8	4300	3900	2700	2607	2207	83.44	.181	.2859	31
30	10.03	100.5	90.5	83.3	71.5	55.5	5040	4660	3020	3287	2534	105.2	.144	.2546	33
31	8.928	79.70	101	92.0	77.5	59.2	5920	5280	—	4145	2768	132.7	.114	.2268	34
32	7.950	63.21	113	101	83.6	62.6	7060	6250	—	5227	3137	167.3	.090	.2019	36
33	7.080	50.13	127	110	90.3	66.3	8120	7360	—	6591	4697	211.0	.072	.1798	37
34	6.305	39.75	143	120	97.0	70.0	9600	8310	—	8310	6168	266.0	.057	.1601	38
35	5.615	31.52	158	132	104	73.5	10900	8700	—	10480	6737	335.0	.045	.1426	38–39
36	5.000	25.00	175	143	111	77.0	12200	10700	—	13210	7877	423.0	.036	.1270	39–40
37	4.453	19.83	198	154	118	80.3	—	—	—	16660	9309	533.4	.028	.1131	41
38	3.965	15.72	224	166	126	83.6	—	—	—	21010	10666	672.6	.022	.1007	42
39	3.531	12.47	248	181	133	86.6	—	—	—	26500	11907	848.1	.018	.0897	43
40	3.145	9.88	282	194	140	89.7	—	—	—	33410	14222	1069	.014	.0799	44

[1] A mil is 1/1000 (one-thousandth) of an inch.

[2] The figures given are approximate only, since the thickness of the insulation varies with different manufacturers.

[3] 700 circular mils per ampere is a satisfactory design figure for small transformers, but values from 500 to 1000 C.M. are commonly used. For 1000 C.M./amp. divide the circular mil area (third column) by 1000; for 500 C.M./amp. divide circular mil area by 500.

FILTERS

The filter sections shown on the facing page can be used alone or, if greater attenuation and sharper cut-off are required, several sections can be connected in series. In the low- and high-pass filters, f_c represents the cut-off frequency, the highest (for the low-pass) or the lowest (for the high-pass) frequency transmitted without attenuation. In the bandpass-filter designs, f_1 is the low-frequency cut-off and f_2 the high-frequency cut-off. The units for L, C, R and f are henrys, farads, ohms and cycles, respectively.

All of the types shown are for use in an unbalanced line (one side grounded), and thus they are suitable for use in coaxial line or any other unbalanced circuit. To transform them for use in balanced lines (e.g., 300-ohm transmission line, or push-pull audio circuits), the series reactances should be equally divided between the two legs. Thus the balanced constant-k π-section low-pass filter would use two inductances of a value equal to $L_k/2$, while the balanced constant-k π-section high-pass filter would use two condensers of a value equal to $2C_k$.

If several low- (or high-) pass sections are to be used, it is advisable to use m-derived end sections on either side of a constant-k section, although an m-derived center section can be used. The factor m relates the ratio of the cut-off frequency and f_∞, a frequency of high attenuation. Where only one m-derived section is used, a value of 0.6 is generally used for m, although a deviation of 10 or 15 per cent from this value is not too serious in amateur work. For a value of $m = 0.6$, f will be $1.25f_c$ for the low-pass filter and $0.8f_c$ for the high-pass filter. Other values can be found from

$$m = \sqrt{1 - \left(\frac{f_c}{f_\infty}\right)^2} \text{ for the low-pass filter and}$$

$$m = \sqrt{1 - \left(\frac{f_\infty}{f_c}\right)^2} \text{ for the high-pass filter}$$

The filters shown should be terminated in a resistance $= R$, and there should be little or no reactive component in the termination.

Simple audio filters can be made with powdered-iron-core chokes and paper condensers. Sharper cut-off characteristics will be obtained with more sections. The values of the components can vary by ±5% with little or no reduction in performance. The more sections there are to a filter the greater is the need for accuracy in the values of the components. High-performance audio filters can be built with only two sections by winding the inductances on toroidial powdered-iron forms — it generally takes three sections to obtain the same results when using other inductances.

Sideband filters are usually designed to operate in the range 10 to 20 kc. Their attenuation requirements are such that usually at least a five-section filter is required. The coils should be as high-Q as possible, and mica condensers are the most suitable capacitors.

Low-pass and high-pass filters for harmonic suppression and receiver-overload prevention in the television frequencies range are usually made with self-supporting coils and mica or ceramic condensers, depending upon the power requirements.

In any filter, there should be no magnetic or capacity coupling between sections of the filter unless the design specifically calls for it. This requirement makes it necessary to shield the coils from each other in some applications, or to mount them at right angles to each other.

Further information on filter design can be found in the following articles:

Bennett, "Audio Filters for Eliminating QRM," *QST*, July, 1949.

Berry, "Filter Design for the Single-Sideband Transmitter," *QST*, June, 1949.

Buchheim, "Low-Pass Audio Filters," *QST*, July, 1948.

Grammer, "Pointers on Harmonic Reduction," *QST*, April, 1949; "High-Pass Filters for TVI Reduction," *QST*, May, 1949.

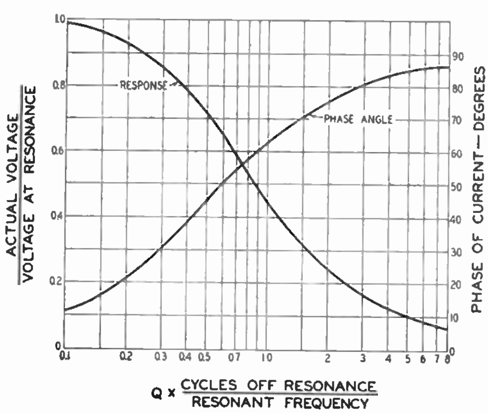
Mann, "An Inexpensive Sideband Filter," *QST*, March, 1949.

Rand, "The Little Slugger," *QST*, February, 1949.

Smith, "Premodulation Speech Clipping and Filtering," *QST*, February, 1946; "More on Speech Clipping," *QST*, March, 1947.

TUNED-CIRCUIT RESPONSE

The graph below gives the response and phase angle of a high-Q parallel-tuned circuit.

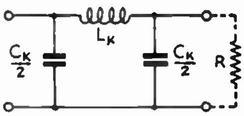


Circuit Q is equal to

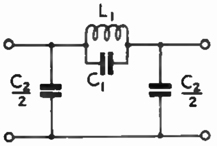
$$2\pi fRC \quad \text{or} \quad \frac{R}{2\pi fL}$$

where L and C are the inductance and capacitance at the resonant frequency, f, and R is the parallel resistance across the circuit. The curves above become more accurate as the circuit Q is higher, but the error is not especially great for values as low as $Q = 10$.

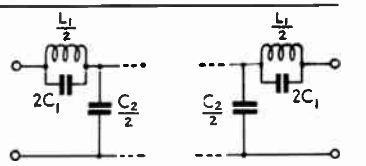
LOW-PASS FILTERS



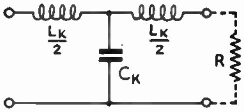
Constant-k π section



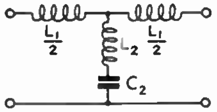
m-derived π section



m-derived end sections for use with intermediate π section



Constant-k T section



m-derived T section

m-derived end sections for use with intermediate T section

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

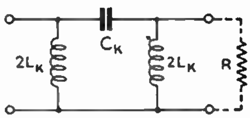
$L_1 = m L_K$ $C_1 = \frac{1-m^2}{4m} C_K$

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

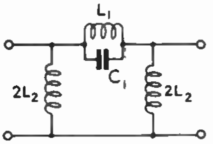
$L_2 = \frac{1-m^2}{4m} L_K$ $C_2 = m C_K$

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

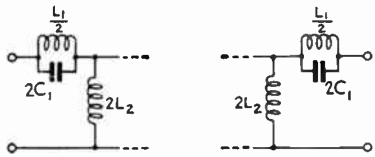
HIGH-PASS FILTERS



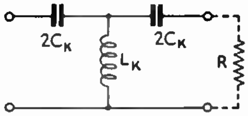
Constant-k π section



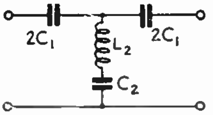
m-derived π section



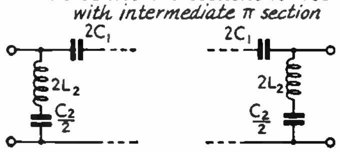
m-derived end sections for use with intermediate π section



Constant-k T section



m-derived T section



m-derived end section for use with intermediate T section

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

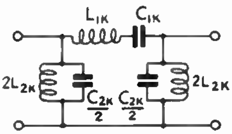
$L_1 = \frac{4m}{1-m^2} L_K$ $C_1 = \frac{C_K}{m}$

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

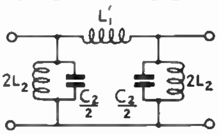
$L_2 = \frac{L_K}{m}$ $C_2 = \frac{4m}{1-m^2} C_K$

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

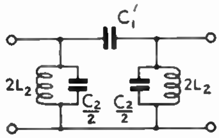
BANDPASS FILTERS



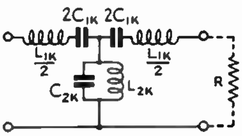
Constant-k π section



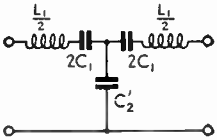
Three-element π section



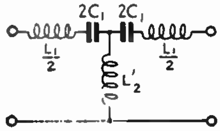
Three-element π section



Constant-k T section



Three-element T section



Three-element T section

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

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

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

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

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

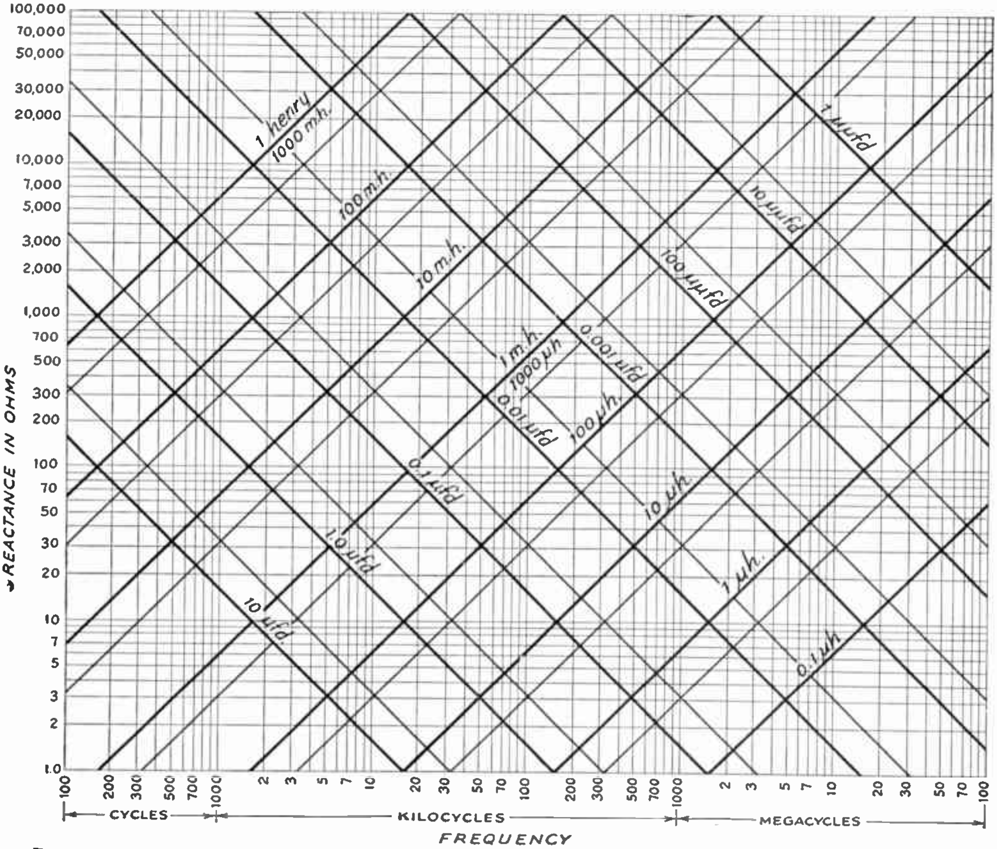
$C_1' = \frac{f_1+f_2}{4\pi f_1 f_2 R}$ $L_2 = L_{2K}$

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

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

In the above formulas *R* is in ohms, *C* in farads, *L* in henrys, and *f* in cycles per second.

INDUCTIVE AND CAPACITIVE REACTANCE *VS.* FREQUENCY CHART



By use of the chart above, the approximate reactance of any capacitance from 1.0 μμfd. to 10 μfd. at any frequency from 100 cycles to 100 megacycles, or the reactance of any inductance from 0.1 μh. to 1.0 henry, can be read directly. Intermediate values can be estimated by interpolation. In making interpolations, remember that the rate of change between lines is logarithmic. Use the frequency or reactance scales as a guide in estimating intermediate values on the capacitance or inductance scales.

This chart also can be used to find the approximate resonance frequencies of *LC* combinations, or the frequency to which a given coil-and-condenser combination will tune. First locate the respective slanting lines for the capacitance and inductance. The point where they intersect, i.e., where the reactances are equal, is the resonant frequency (projected downward and read on the frequency scale).

ELECTRICAL CONDUCTIVITY OF METALS

	Relative Conductivity [1]	*Temp. Coef.* [2] *of Resistance*
Aluminum (2S; pure)	59	0.0049
Aluminum (alloys):		
Soft-annealed	45–50	
Heat-treated	30–45	
Brass	28	0.002–0.007
Cadmium	19	
Chromium	55	
Climax	1.83	
Cobalt	16.3	
Constantin	3.24	0.00002
Copper (hard drawn)	89.5	0.004
Copper (annealed)	100	
Everdur	6	
German Silver (18%)	5.3	0.00019
Gold	65	
Iron (pure)	17.7	0.006
Iron (cast)	2–12	
Iron (wrought)	11.4	

[1] At 20° C., based on copper as 100. [2] Per °C. at 20° C.

	Relative Conductivity	*Temp. Coef.* [2] *of Resistance*
Lead	7	0.0041
Manganin	3.7	0.00002
Mercury	1.66	0.00089
Molybdenum	33.2	0.0033
Monel	4	0.0019
Nichrome	1.45	0.00017
Nickel	12–16	0.005
Phosphor Bronze	36	0.004
Platinum	15	
Silver	106	0.004
Steel	3–15	
Tin	13	0.0042
Tungsten	28.9	0.0045
Zinc	28.2	0.0035

Approximate relations

An increase of 1 in A. W. G. or B. & S. wire size increases resistance 25%.
An increase of 2 increases resistance 60%.
An increase of 3 increases resistance 100%.
An increase of 10 increases resistance 10 times.

PILOT-LAMP DATA

Lamp No.	Bead Color	Base (Miniature)	Bulb Type	RATING Volts	RATING Amp.
40	Brown	Screw	T-3¼	6–8	0.15
40A[1]	Brown	Bayonet	T-3¼	6–8	0.15
41	White	Screw	T-3¼	2.5	0.5
42	Green	Screw	T-3¼	3.2	**
43	White	Bayonet	T-3¼	2.5	0.5
44	Blue	Bayonet	T-3¼	6–8	0.25
45	*	Bayonet	T-3¼	3.2	**
46[2]	Blue	Screw	T-3¼	6–8	0.25
47[1]	Brown	Bayonet	T-3¼	6–9	0.15
48	Pink	Screw	T-3¼	2.0	0.06
49[3]	Pink	Bayonet	T-3¼	2.0	0.06
[4]	White	Screw	T-3¼	2.1	0.12
49A[3]	White	Bayonet	T-3¼	2.1	0.12
50	White	Screw	G-3½	6–8	0.2
51[2]	White	Bayonet	G-3½	6–8	0.2
—	White	Screw	G-4½	6–8	0.4
55	White	Bayonet	G-4½	6–8	0.4
292[5]	White	Screw	T-3¼	2.9	0.17
292A[5]	White	Bayonet	T-3¼	2.9	0.17
1455	Brown	Screw	G-5	18.0	0.25
1455A	Brown	Bayonet	G-5	18.0	0.25

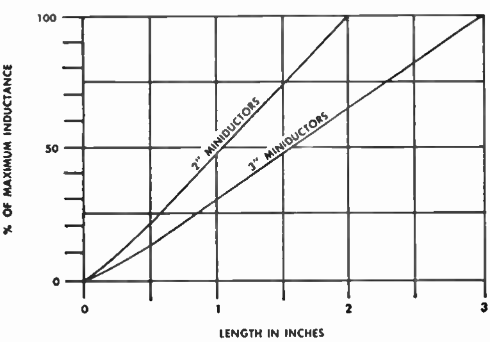
* White in G.E. and Sylvania; green in National Union, Raytheon and Tung-Sol.

** 0.35 in G.E. and Sylvania; 0.5 in National Union, Raytheon and Tung-Sol.

[1] **40A** and **47** are interchangeable.
[2] Have frosted bulb.
[3] **49** and **49A** are interchangeable.
[4] Replace with No. **48.**
[5] Use in 2.5-volt sets where regular bulb burns out too frequently.

MINIDUCTOR DATA

The chart and table below, furnished through courtesy of Barker & Williamson, can be used to determine the approximate inductance of coils made of Miniductor material. The curves show the percentage of the total inductance (given in the right-hand column of the table) of the coil as supplied, when cut to various lengths.

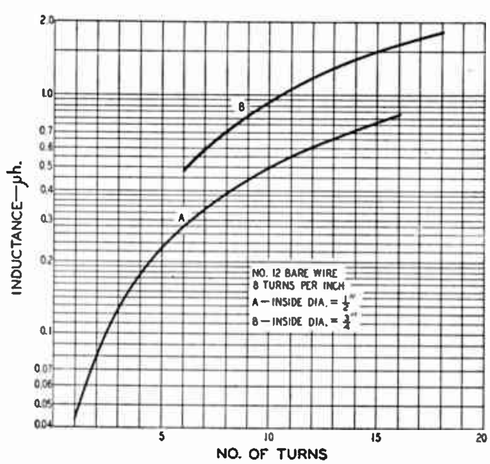


MINIDUCTOR SPECIFICATIONS

Catalog Number	Diam. (Inches)	Turns Per Inch	Approx. Length (Inches)	Approx. Inductance (μh.)
3001	½	4	2	0.4
3002	½	8	2	0.96
3003	½	16	2	3.2
3004	½	32	2	13.7
3005	⅝	4	2	0.56
3006	⅝	8	2	1.4
3007	⅝	16	2	4.9
3008	⅝	32	2	19.2
3009	¾	4	3	0.94
3010	¾	8	3	2.9
3011	¾	16	3	10.9
3012	¾	32	3	42.5
3013	1	4	3	1.9
3014	1	8	3	4.8
3015	1	16	3	19.9
3016	1	32	3	73.0

INDUCTANCE OF SMALL COILS

Most inductance formulas lose accuracy when applied to the small coils used in v.h.f. work and in low-pass filters built for reducing harmonic interference to television, because the conductor thickness is no longer negligible in comparison with the size of the coil. The accompanying chart shows the measured inductance of typical coils used for these purposes, and may be used as a basis for circuit design. Two curves are given: curve *A* is for coils wound to an inside diameter of ½ inch; curve *B* is for coils of ¾-inch inside diameter. In both curves the wire size is No. 12, winding pitch 8 turns to the inch (⅛ inch center-to-center turn spacing). The inductance values given include leads ½ inch long.



Measured inductance of coils wound with No. 12 bare wire, 8 turns to the inch. The values include half-inch leads. Where smaller inductance values are required, they should be obtained experimentally by adjusting to the proper resonance frequency with the specified capacitance. Coils of larger inductance can be wound from the common formulas.

Vacuum Tubes and Semiconductors

For the convenience of the designer, the receiving-type tubes listed in this chapter are grouped by filament voltages and construction types (glass, metal, miniature, etc.). For example, all 6.3-volt metal tubes are listed in Table I, all lock-in base tubes are in Table III, all miniatures are in Table XI, and so on.

Transmitting tubes are divided into triodes and tetrodes-pentodes, then listed according to rated plate dissipation. This permits direct comparison of ratings of tubes in the same power classification.

For quick reference, all tubes are listed in numerical-alphabetical order in the index beginning on the following page.

Tube Ratings

Vacuum tubes are designed to be operated within definite maximum (and minimum) ratings. These ratings are the maximum safe operating voltages and currents for the electrodes, based on inherent limiting factors such as permissible cathode temperature, emission, and power dissipation in electrodes.

In the transmitting-tube tables, maximum ratings for electrode voltage, current and dissipation are given separately from the typical operating conditions for the recommended classes of operation. In the receiving-tube tables, because of space limitations, ratings and operating data are combined. Where only one set of operating conditions appears, the positive electrode voltages shown (plate, screen, etc.) are, in general, also the maximum rated voltages for those electrodes.

For certain air-cooled transmitting tubes, there are two sets of maximum values, one designated as CCS (Continuous Commercial Service) ratings, the other ICAS (Intermittent Commercial and Amateur Service) ratings. Continuous Commercial Service is defined as that type of service in which long tube life and reliability of performance under continuous operating conditions are the prime consideration. Intermittent Commercial and Amateur Service is defined to include the many applications where the transmitter design factors of minimum size, light weight, and maximum power output are more important than long tube life. ICAS ratings are considerably higher than CCS ratings. They permit the handling of greater power, and although such use involves some sacrifice in tube life, the period over which tubes will continue to give satisfactory performance in intermittent service can be extremely long.

Typical Operating Conditions

The typical operating conditions given for transmitting tubes represent, in general, maximum ICAS ratings where such ratings have been given by the manufacturer. They do not represent the *only* possible method of operation of a particular tube type. Other values of plate voltage, plate current, grid bias, etc., may be used so long as the maximum ratings for a particular voltage or current are not exceeded.

INDEX TO TUBE TABLES

BASE TYPE DESIGNATIONS

A = Acorn
B = Glass-button miniature
B_8 = Glass-button subminiature
J = Jumbo
L = Lock-in
M = Medium
X = None or special type
O = Octal
S = Small
W = Wafer

INDEX TO VACUUM-TUBE TYPES

For convenience in locating data on specific tube types the index below lists all tubes in numerical-alphabetical order, showing the page number where individual tubes may be found in the classified-data section (pages V14–V63) and the identifying base-diagram number in the base-diagram section (pages V5–V13).

Type	Page	Base
6SV7	V15	7AZ
6SZ7	V15	8Q
6T4	V29	7DK
6T5	V19	6R
6T6GM	V17	6Z
6T7	V15	7V
6T8	V29	9E
6U3	V42	9BM
6U4GT	V42	4CG
6U5	V19	6R
6U6GT	V17	7AC
6U7G	V17	7R
6U8	V29	9AE
6V3	V42	9BD
6V4	V42	9M
6V5GT	V17	6AO
6V6	V15	7AC
6V6GT	V56	7AC
6V7G	V17	7V
6V8	V29	9AH
6W4GT	V42	4CG
6W5G	V42	6S
6W6GT	V17	7AC
6W7G	V17	7R
6X4	V42	7CF
6X5	V42	6S
6X6G	V17	7AL
6X8	V29	9AK
6Y3G	V42	4AC
6Y5	V42	6J
6Y6G	V17	7AC
6Y7G	V17	8B
6Z3	V42	4G
6Z4	V43	5D
6Z5	V42	6K
6Z7G	V17	8B
6ZY5G	V42	6S
7A4	V17	5AC
7A5	V17	6AA
7A6	V17	7AJ
7A7	V17	8V
7A8	V17	8U
7AB7	V26	8BO
7AD7	V18	8V
7AF7	V18	8AC
7AG7	V18	8V
7AH7	V18	8V
7AJ7	V18	8V
7AK7	V18	8V
7AP4	V38	5AJ
7B4	V18	5AC
7B5	V18	6AE
7B6	V18	8W
7B7	V18	8V
7B8	V18	8X
7BP1	V38	5AN
7C4	V26	4AH
7C5	V18	6AA
7C6	V18	8W
7C7	V18	8V
7CP1	V38	6AZ
7D7	V18	8AR
7DP4	V38	12C
7E5	V26	8BN
7E6	V18	8W
7E7	V18	8AE
7EP4	V38	11N
7F7	V18	8AC
7F8	V18	8BW
7G7	V18	8V
7G8	V18	8BV
7GP4	V38	Fig. 47
7H7	V18	8V
7J7	V18	8AR
7JP1	V38	14G
7JP4	V38	14G
7K7	V18	8BF
7L7	V18	8V
7MP7	V38	12D
7N7	V18	8AC
7NP4	V38	14N
7Q7	V18	8AL
7QP4	V38	12D
7R7	V18	8AE
7RP4	V38	12D
7S7	V18	8BL
7T7	V18	8V
7TP4	V38	12C
7V7	V18	8V
7VP1	V38	14A
7W7	V18	8BJ
7WP4	V38	14N
7X6	V42	7AJ
7X7	V18	8BZ
7Y4	V42	5AB
7Z4	V42	5AB
8AP4	V39	12H
8BP4	V39	14G
9AP4	V39	6AL
9BM5	V29	7DQ
9BW6	V29	9AM
9CP4	V39	4AF
9JP1	V39	8BR
10	V25	4D
10	V46	4D
10BP4	V39	12D
10EP4	V39	12D
10FP4	V39	12D
10HP4	V39	14G
10KP7	V39	12D
10SP4	V39	12C
10Y	V46	4D
11/12	V25	4F
12A4	V30	9AG
12A5	V23	7F
12A6	V23	7AC
12A7	V23	7K
12A7	V42	7K
12A8GT	V23	8A
12AH7GT	V23	8BE
12AH8	V30	9BP
12AL5	V30	6BT
12AP4	V39	6AL
12AQ5	V30	7BZ
12AT6	V30	7BT
12AT7	V30	9A
12AU6	V30	7CC
12AU7	V30	9A
12AU7	V45	9A
12AV6	V30	7BT
12AV7	V30	9A
12AW6	V30	7CM
12AW7	V30	7CM
12AX4GT	V43	4CG
12AX7	V30	9A
12AY7	V30	9A
12AZ7	V30	9A
12B4	V30	9AG
12B6M	V23	6Y
12B7	V23	8V
12B7ML	V23	8V
12B8GT	V23	8T
12BA6	V30	7CC
12BA7	V30	8CT
12BD6	V30	7CC
12BE6	V30	7CH
12BF6	V30	7BT
12BH7	V30	9A
12BK6	V30	7BT
12BN6	V30	7DF
12BT6	V30	7BT
12BU6	V30	7BT
12BY7	V30	9BF
12BZ7	V30	9A
12C8	V23	8E
12CP4	V39	4AF
12DP4	V39	5AN
12E5GT	V23	6Q
12F5GT	V23	5M
12G4	V30	6BG
12G7G	V23	7V
12H6	V23	7Q
12J5GT	V23	6Q
12J7GT	V23	7R
12K7GT	V23	7R
12K8	V23	8K
12KP4A	V39	Fig. 35
12L8GT	V23	8BU
12LP4	V39	12D
12Q7GT	V23	7V
12QP4	V39	Fig. 35
12RP4	V39	12D
12S8GT	V23	8CB
12SA7	V23	8R
12SC7	V23	8S
12SF5	V23	6AB
12SF7	V23	7AZ
12SG7	V23	8BK
12SH7	V23	8BK
12SJ7	V23	8N
12SK7	V23	8N
12SL7GT	V23	8BD
12SN7GT	V23	8BD
12SP7	V39	12D
12SQ7	V23	8Q
12SR7	V23	8Q
12SW7	V23	8Q
12SX7	V23	8BD
12SY7	V23	8R
12TP4	V39	12D
12UP4	V39	12D
12V6GT	V23	7AC
12X4	V43	5BS
12Z3	V43	4G
12Z5	V43	7L
14A4	V23	5AC
14A5	V23	6AA
14A7	V23	8V
14AF7	V23	8AC
14AF7	V25	8AC
14B6	V23	8W
14B8	V23	8X
14BP4	V39	Fig. 35
14C5	V23	6AA
14C7	V23	8V
14CP4	V39	12D
14DP4	V39	12D
14E6	V24	8W
14E7	V24	8AE
14EP4	V39	12D
14F7	V24	8AC
14F8	V24	8BW
14GP4	V39	Fig. 42
14H7	V24	8V
14HP4	V39	Fig. 43
14J7	V24	8BL
14N7	V24	8AC
14Q7	V24	8AL
14R7	V24	8AE
14S7	V24	8BL
14V7	V24	8V
14W7	V24	8BJ
14X7	V24	8BZ
14Y4	V43	5AB
14Z3	V43	4G
15	V21	5F
15A6	V30	9AR
15AP4	V39	12D
15CP4	V39	Fig. 35
15DP4	V39	12D
15E	V46	T-4AF
16A5	V30	9BL
16ADP4	V39	Fig. 69
16AP4	V39	Fig. 35
16CP4	V39	Fig. 35
16EP4A	V39	12D
16FP4	V39	Fig. 35
16GP4	V39	12D
16GP4B	V39	12D
16GP4C	V39	12D
16HP4	V39	Fig. 35
16JP4	V39	12D
16KP4	V39	12D
16LP4	V40	Fig. 35
16MP4	V40	12D
16RP4	V40	12D
16SP4A	V40	12D
16TP4	V40	Fig. 35
16UP4	V40	12D
16VP4	V40	12D
16WP4A	V40	12D
16ZP4	V40	12D
17	V36	3G
17AP4	V40	12D
17BP4A	V40	Fig. 45
17BP4B	V40	12D
17CP4	V40	12D
17FP4	V40	Fig. 42
17GP4	V40	Fig. 43
17HP4	V40	Fig. 42
17JP4	V40	Fig. 45
17KP4	V40	Fig. 45
17LP4	V40	Fig. 42
17QP4	V40	12D
17RP4	V40	Fig. 66
17SP4	V40	12D
17YP4	V40	Fig. 45
17Z3	V43	9CB
18	V24	6B
19	V21	6C
19AP4	V40	Fig. 35
19AP4A	V40	12D
19AQ5	V30	7BL
19BG6G	V24	5BT
19C8	V30	9E
19DP4A	V40	12D
19EP4	V40	Fig. 35
19FP4	V40	Fig. 35
19GP4	V40	12D
19J6	V31	7BF
19JP4	V40	12D
19T8	V31	9E
19V8	V31	9AH
19X3	V43	9BM
19X8	V31	9AK
19Y3	V43	9BM
20	V25	4D
20BP4	V40	12D
20CP4	V40	Fig. 44
20CP4A	V40	Fig. 44
20DP4	V40	Fig. 44
20FP4	V40	Fig. 66
20GP4	V40	Fig. 42
20HP4	V40	Fig. 66
20JP4	V40	Fig. 45
20J8GM	V24	8H
20LP4	V40	Fig. 43
20MP4	V40	Fig. 42
21A6	V31	9AS
21A7	V24	8AR
21AP4	V40	Fig. 44
21EP4A	V40	Fig. 44
21FP4A	V40	Fig. 43
21KP4A	V40	Fig. 45
21MP4	V40	Fig. 43
22	V25	4K
22AP4	V40	Fig. 35
24-A	V20	5E
24AP4A	V41	12D
24BP4	V41	Fig. 43
24-G	V47	2D
24XH	V41	Fig. 1
25A6	V24	7S
25A7GT	V24	8F
25A7GT	V43	8F
25AC5GT	V24	6Q
25AV5GT	V24	6CK
25B5	V24	6D
25B6G	V24	7S
25B8GT	V24	8T
25BK5	V31	9BQ
25BQ6GT	V24	6AM
25C6G	V24	7AC
25CD6G	V24	5BT
25D8GT	V24	8AF
25L6	V24	7AC
25N6G	V24	7W
25S	V21	6M
25T	V47	3G
25W4GT	V43	4CG
25X6GT	V43	7Q
25Y4GT	V43	5AA
25Y5	V43	6E
25Z3	V43	4G
25Z4	V43	5AA
25Z5	V43	6E
25Z6	V43	7Q
26	V25	4D
26A6	V31	7BK
26A7GT	V24	8BU
26BK6	V31	7BT
26C6	V31	7BT
26CG6	V31	7BK
26D6	V31	7CH
26Z5W	V43	9BS
27	V20	5A
27AP4	V41	Fig. 43
27EP4	V41	12D
28D7	V25	8BS
28Z5	V43	5AB
30	V21	4D
30BP4	V41	12D
31	V21	4D
32	V21	4K
32L7GT	V24	8Z
32L7GT	V43	8Z
33	V21	5K
34	V21	4M
35/51	V20	5E
35A5	V24	6AA
35B5	V31	7BZ
35C5	V31	7CV
35L6GT	V24	7AC
35T	V48	3G
35TG	V48	2D
35W4	V43	5BQ
35Y4	V43	5AL
35Z3	V43	4Z
35Z4GT	V43	5AA
35Z5G	V43	6AD
35Z6G	V43	7Q
36	V19	5E
37	V19	5A
38	V19	5F
39/44	V19	5F
40	V26	4D
40Z5GT	V43	6AD
41	V19	6B
42	V19	6B
43	V24	6B
45	V20	4D
45Z3	V43	5AM
45Z5GT	V43	6AD
46	V20	5C
47	V20	5B
48	V24	6A
49	V21	5C
50	V26	4D
50A5	V24	6AA
50AX6G	V43	7Q
50B5	V31	7BZ
50C5	V31	7CV
50C6GT	V24	7AC
50L6GT	V24	7AC
50T	V50	2D
50X6	V43	7AJ
50Y6GT	V43	7Q
50Y7GT	V43	8AN
50Z6G	V43	7Q
50Z7G	V43	8AN
51	V20	5E
52	V19	5C
53	V20	7B
53A	V47	T-4B
55	V20	6G
56	V20	5A
56AS	V19	5A
57	V20	6F
57AS	V19	6F
58	V20	6F
58AS	V19	6F
59	V20	7A
70A7GT	V24	8AB
70A7GT	V43	8AB
70L7GT	V24	8AA
70L7GT	V43	8AA
71-A	V26	4D
72	V43	4P
73	V43	4Y
75	V19	6G
75TH	V50	2D
75TL	V50	2D
76	V19	5A
77	V19	6F
78	V19	6F
79	V19	6H
80	V43	4C
81	V43	4B
82	V43	4C
83	V43	4C
83-V	V43	4AD
84/6Z4	V43	5D
85	V19	6G
85AS	V19	6G
89	V19	6F
99	V26	4D
100TH	V51	2D
100TL	V51	2D
111H	V50	2D
112-A	V26	4D
117L7GT	V24	8AO
117L7GT	V43	8AO
117M7GT	V24	8AO
117M7GT	V43	8AO
117N7GT	V24	8AV
117N7GT	V43	8AV
117P7GT	V25	8AV
117P7GT	V43	8AV
117Z3	V43	4BR
117Z4GT	V43	5AA
117Z6GT	V43	7Q
128AS	V36	5A
150T	V53	2N
152TH	V53	4BC
152TL	V53	4BC
182-B	V26	4D
183	V26	4D
203-A	V51	4E
203-H	V51	3N
204-A	V55	T-1A
205D	V46	4D
211	V51	4E
212-E	V55	T-2A
217A	V43	4AT
217C	V43	4AT
227A	V51	T-4B
241B	V55	T-2AA
242A	V50	4E
242B	V51	4E
242C	V52	4E
249B	V43	Fig. 53
250TH	V54	2N
250TL	V54	2N
254A	V57	T-4C
254B	V58	T-4C
261A	V52	4E
270A	V55	T-1A
276A	V52	4E
282A	V59	4-TC
284B	V52	3N
284D	V50	4E
295A	V52	4E
300T	V55	2N
303A	V51	4E
304A	V55	T-1A
304B	V48	2D
304TH	V55	4BC
304TL	V55	4BC
305A	V59	T-4CE
306A	V57	T-5CB
307A	V57	T-5C
308B	V55	T-2A
310	V46	4D
311	V51	4E
311CH	V52	Fig. 57
312A	V59	T-6C
312E	V55	T-2AA
316A	V47	—
327A	V51	T-4AD
327B	V50	T-4AD
342B	V51	4E
356A	V49	T-4BD
361A	V52	4E
376A	V52	4E
410R	V61	Fig. 58
417A	V31	9V
482B	V26	4D
483	V26	4D
485	V26	5A
527	V55	T-4B
559	V26	Fig. 18
575A	V44	4AT
592	V54	Fig. 52
703A	V46	—
705A	V43	T-3AA
707B	V62	Fig. 61
715B	V59	—
717A	V17	8BK
723AB	V61	Fig. 60
756	V48	4D
800	V47	2D
801	V46	4D
801A	V46	4D
802	V56	6BM
803	V60	5J
804	V59	T-5C
805	V52	3N
805	V53	3N
806	V54	2N
807	V58	5AW
807W	V58	5AW
808	V49	2D
809	V47	3G
810	V53	2N
811	V49	3G
811A	V49	3G
812	V49	3G
812A	V49	3G
812H	V50	3G
813	V60	5BA
814	V59	T-5D
815	V58	8BY
816	V43	4P
822	V54	3N
822S	V54	2N
826	V49	7BO
828	V60	5J
829	V58	7BP
829A	V58	7BP
829B	V58	7BP
830	V48	4D
830B	V49	3G
831	V55	T-1AA
832	V57	7BP
832A	V57	7BP
833A	V55	T-1AB
834	V49	2D
835	V51	4E
836	V43	4P
837	V56	6BM
838	V52	6E
840	V21	5J
841	V46	4D
841A	V48	3G
841SW	V49	3G
843	V46	5A
844	V57	5AW
849	V55	T-1A
850	V60	T-3B
852	V52	2D
860	V60	T-4CB
861	V61	T-1B
864	V26	4D
865	V57	T-4C
866	V43	4P
866A	V43	4P
866B	V43	4P
866jr	V43	4B
871	V44	4P
872	V44	4AT
872A	V44	4AT
874	V36	4S
876	V36	—
878	V44	4P
879	V44	4P
884	V36	6Q
885	V36	5A
886	V36	—
902	V41	Fig. 1
903	V41	6AL
904	V41	Fig. 3
905	V41	Fig. 6
906P1	V37	7AN
907	V41	Fig. 6
908	V41	7AN
908A	V41	7CE
909	V41	Fig. 6

Type	Page	Base
910	V41	7AN
911	V41	7AN
912	V41	Fig. 8
913	V41	Fig. 1
914	V41	Fig. 12
930B	V49	3G
938	V52	4E
950	V21	5K
951	V21	4M
954	V26	5BB
955	V26	5BC
955	V45	5BC
956	V26	5BB
957	V26	5BD
958	V26	5BD
958A	V26	5BD
958A	V45	5BD
959	V26	5BE
967	V36	3G
975A	V44	4AT
991	V36	—
1003	V44	4R
1005	V44	5AQ
1006	V44	4C
1201	V26	8BN
1203	V26	4AH
1204	V26	8BO
1206	V18	8BV
1221	V20	6F
1223	V17	7R
1229	V21	4K
1230	V21	4D
1231	V18	8V
1232	V18	8V
1247	V33	—
1265	V36	4AJ
1266	V36	4AJ
1267	V36	4V
1273	V18	8V
1274	V44	6S
1275	V44	4C
1276	V24	4D
1280	V25	8V
1284	V25	8V
1291	V22	7BE
1293	V22	4AA
1294	V22	4AH
1299	V22	6BB
1602	V46	4D
1603	V20	6F
1608	V46	4D
1609	V26	5B
1610	V56	T-5CA
1611	V15	7S
1612	V15	7T
16[illegible]	V56	7S
16[illegible]	V57	7AC
16[illegible]	V44	4P
1619	V57	T-9H
1620	V15	7R
1621	V15	7S
1622	V15	7AC
1623	V46	3G
1624	V58	T-5DC
1625	V58	5AZ
1626	V45	6Q
1627	V53	2N
1628	V47	T-4BB
1629	V25	6RA
1631	V25	7AC
1632	V25	7AC
1633	V25	8BD
1634	V25	8S
1635	V17	8B
1641	V44	T-4AG
1642	V19	7BH
1644	V25	Fig. 7
1654	V44	2Z
1800	V41	6AL
1801	V41	Fig. 13
1802P1	V37	11A
1803P4	V39	6AL
1804P4	V39	6AL
1805P1	V37	11A
1806P1	V37	11A
1809P1	V39	8BR
1811P1	V38	6AZ
1816P4A	V41	Fig. 65
1851	V15	7R
1852	V14	8N
1853	V14	8N
2001	V41	4AA
2002	V41	Fig. 1
2005	V41	Fig. 1
2050	V36	8BA
2051	V36	8BA
2523N/128AS	V36	5A
5514	V49	4BO
5516	V57	7CL
5517	V44	5BU
5556	V45	4D
5562	V59	Fig. 54
5590	V31	7BD
5591	V31	7BD
5608	V31	7BD
5608A	V20	7B
5610	V31	6CG
5618	V56	7CU
5633	V34	—
5634	V34	—
5635	V35	8DB
5636	V35	8DC
5637	V34	—
5638	V34	—
5639	V35	8DL
5640	V34	—
5641	V34	—
5641	V35	6CJ
5642	V34	—
5643	V35	8DD
5644	V35	4CN
5645	V34	—
5646	V35	—
5647	V35	B1
5648	V52	—
5651	V36	5BO
5654	V31	7BD
5656	V31	9F
5663	V36	7CE
5670	V31	8CJ
5672	V35	—
5675	V45	Fig. 36
5676	V35	—
5677	V35	—
5678	V35	—
5679	V18	7CX
5686	V56	Fig. 29
5686	V31	Fig. 29
5687	V31	9H
5690	V44	Fig. 74
5691	V17	8BD
5692	V17	8BD
5693	V15	8N
5694	V17	8CS
5697	V35	—
5702	V35	—
5703	V35	—
5704	V35	—
5718	V35	8DK
5719	V35	8DK
5722	V31	5CB
5725	V31	7CM
5726	V31	6BT
5727	V36	7BN
5731	V26	5BC
5744	V35	—
5749	V31	7BK
5750	V31	7CH
5751	V31	9A
5755	V31	9J
5763	V56	9K
5764	V45	Fig. 36
5765	V45	Fig. 36
5766	V45	Fig. 36
5767	V45	Fig. 36
5768	V26	Fig. 36
5784	V35	—
5785	V35	—
5794	V45	Fig. 36
5812	V56	7CQ
5812	V31	7CQ
5814	V31	9A
5823	V36	4CK
5824	V25	7S
5825	V44	4P
5836	V62	—
5837	V62	—
5840	V35	8DL
5842	V31	9V
5844	V31	7BF
5845	V31	5CA
5847	V31	9X
5851	V35	6CL
5866	V53	Fig. 5
5867	V55	—
5871	V20	7AC
5876	V45	Fig. 36
5879	V31	9AD
5881	V57	7AC
5881	V17	7AC
5890	V36	12J
5893	V45	Fig. 36
5894A	V58	Fig. 10
5896	V35	8DJ
5897	V35	8DK
5898	V35	8DK
5899	V35	8DL
5900	V35	8DL
5901	V35	8DL
5902	V35	8DL
5903	V35	8DJ
5904	V35	8DK
5905	V35	8DL
5906	V35	8DL
5907	V35	8DL
5908	V35	8DC
5910	V31	6AR
5915	V31	7CH
5916	V35	8DC
5920	V32	Fig. 70
5933	V58	5AW
5961	V15	8R
5962	V36	2AG
5963	V32	9A
5964	V32	7BF
5965	V31	9A
5977	V35	8DK
5987	V35	8DM
5993	V44	Fig. 71
6005	V32	7B7
6026	V45	—
6057	V32	9A
6058	V32	6BT
6059	V32	9BC
6060	V32	9A
6061	V32	9AM
6062	V32	9K
6062	V56	9K
6063	V44	5BS
6064	V32	7BD
6066	V32	7BT
6067	V32	9A
6072	V32	9A
6073	V36	5BO
6074	V36	5BO
6080	V17	8BD
6082	V25	8BD
6084	V32	9BJ
6085	V32	9A
6086	V32	9BK
6111	V35	8DG
6132	V32	9BA
6135	V32	6BG
6136	V32	7BK
6137	V15	8N
6140	V36	9BY
6141	V36	9BZ
6146	V58	7CK
6155	V60	5BK
6156	V61	5BK
6157	V44	Fig. 72
6158	V32	9A
6159	V58	7CK
6173	V26	Fig. 67
6201	V32	9A
6216	V32	Fig. 73
6216	V56	Fig. 73
7000	V17	7R
7193	V45	4AM
7700	V20	6F
8000	V54	2N
8001	V59	7BM
8003	V52	3N
8005	V51	3G
8008	V44	Fig. 11
8010-R	V48	—
8012	V47	T-4BB
8013-A	V44	4P
8016	V42	4AC
8020	V44	4P
8025	V47	4AQ
9001	V32	7PM
9002	V32	7TM
9002	V45	7TM
9003	V32	7PM
9004	V26	4BJ
9005	V26	5BG
9006	V32	6BH
AT-340	V61	5BK
AX9900	V53	Fig. 5
AX9901	V55	—
AX9903	V58	Fig. 10
AX9905	V57	Fig. 34
BA	V42	4J
BH	V42	4J
BR	V42	4H
CE220	V42	4P
CK501	V33	—
CK502	V33	—
CK503	V33	—
CK504	V33	—
CK505	V33	—
CK506	V33	—
CK507	V33	—
CK509	V33	—
CK510	V33	—
CK512	V33	—
CK515BX	V33	—
CK520AX	V33	—
CK521AX	V33	—
CK522AX	V34	—
CK523AX	V34	—
CK524AX	V34	—
CK525AX	V34	—
CK526AX	V34	—
CK527AX	V34	—
CK529AX	V34	—
CK551AXA	V34	—
CK553AXA	V34	—
CK556AX	V34	—
CK568AX	V34	—
CK569AX	V34	—
CK605CX	V34	—
CK606BX	V34	—
CK608CX	V34	—
CK619CX	V34	—
CK624CX	V34	—
CK650AX	V34	—
CK705	V65	—
CK706	V65	—
CK707	V65	—
CK708	V65	—
CK710	V65	—
CK1005	V44	5AQ
CK1006	V44	4C
CK1007	V44	T-9G
CK1009	V44	—
CK5672	V34	—
DR3B27	V42	4P
DR123C	V52	Fig. 26
DR200	V53	2N
EF50	V26	9C
F123A	V52	Fig. 26
F127A	V54	Fig. 26
G84	V42	4B
GL2C44	V26	Fig. 17
GL2C44	V45	Fig. 17
GL5C24	V54	Fig. 26
GL5D24	V61	5BK
GL146	V53	T-4BG
GL152	V53	T-4BG
GL159	V55	T-4BG
GL169	V55	T-4BG
GL446A	V26	Fig. 19
GL446A	V45	Fig. 19
GL446B	V26	Fig. 19
GL446B	V45	Fig. 19
GL464A	V26	Fig. 17
GL464A	V45	Fig. 17
GL559	V26	Fig. 18
GL592	V54	Fig. 52
GL8012A	V47	T-4BB
HD203A	V53	3N
HF60	V50	2D
HF75	V50	2D
HF100	V50	2D
HF120	V51	4F
HF125	V51	—
HF130	V52	—
HF140	V51	4F
HF150	V52	—
HF175	V52	T-3AC
HF200	V53	2N
HF250	V53	2N
HF300	V54	2N
HK24	V47	3G
HK54	V48	2D
HK57	V59	Fig. 64
HK154	V48	2D
HK158	V48	2D
HK252L	V53	4BC
HK253	V43	4AT
HK254	V51	2N
HK257	V60	7BM
HK257B	V60	7BM
HK304L	V55	4BC
HK354	V53	2N
HK354C	V53	2N
HK354D	V53	2N
HK354E	V53	2N
HK354F	V53	2N
HK454H	V55	2N
HK454L	V55	2N
HK654	V55	2N
HV12	V54	3N
HV18	V53	2N
HV27	V54	3N
HY6J5GTX	V45	6Q
HY6L6GT	V57	7AC
HY6V6GT	V57	7AC
HY24	V45	4D
HY25	V47	3G
HY30Z	V47	4BO
HY31Z	V47	T-4D
HY40	V48	3G
HY40Z	V48	3G
HY51A	V49	3G
HY51B	V49	3G
HY51Z	V49	4BO
HY57	V48	3G
HY60	V57	5AW
HY61	V58	5AW
HY63	V56	T-8DB
HY65	V57	T-8DB
HY67	V59	T-5DB
HY69	V58	T-5D
HY75	V46	2T
HY75A	V46	2T
HY113	V34	5K
HY114B	V45	2T
HY115	V34	5K
HY123	V34	5K
HY125	V34	5K
HY145	V34	5K
HY155	V34	5K
HY615	V45	T-8AG
HY801A	V46	4D
HY866jr.	V44	4P
HY1231Z	V47	T-4D
HY1269	V59	T-5DB
HYE1148	V45	T-8AG
KY21	V36	—
KY866	V37	Fig. 8
M54	V34	—
M64	V34	—
M74	V34	—
NU-2C35	V26	Fig. 38
PE340	V60	5BK
QK140	V62	—
QK159	V62	Fig. 63
QK174C	V63	—
QK226	V62	—
QK227	V62	—
QK289	V62	—
QK290	V62	—
QK291	V62	—
QK292	V62	—
QK293	V62	—
QK294	V62	—
QK295	V62	—
QK306	V63	—
QK312	V63	—
RK2J22	V63	—
RK2J23	V63	—
RK2J24	V63	—
RK2J25	V63	—
RK2J26	V63	—
RK2J27	V63	—
RK2J28	V63	—
RK2J29	V63	—
RK2J30	V63	—
RK2J31	V63	—
RK2J32	V63	—
RK2J33	V63	—
RK2J34	V63	—
RK2J36	V63	—
RK2J38	V63	—
RK2J39	V63	—
RK2J48	V63	—
RK2J49	V63	—
RK2J50	V63	—
RK2J51	V63	—
RK2J54	V63	—
RK2J55	V63	—
RK2J56	V63	—
RK2J58	V63	—
RK2J61A	V63	—
RK2J62A	V63	—
RK2J66	V63	—
RK2J67	V63	—
RK2J68	V63	—
RK2J69	V63	—
RK2J70	V63	—
RK2J71	V63	—
RK4J31	V63	—
RK4J32	V63	—
RK4J33	V63	—
RK4J34	V63	—
RK4J35	V63	—
RK4J36	V63	—
RK4J37	V63	—
RK4J38	V63	—
RK4J39	V63	—
RK4J40	V63	—
RK4J41	V63	—
RK4J43	V63	—
RK4J44	V63	—
RK4J53	V63	—
RK4J54	V63	—
RK4J55	V63	—
RK4J56	V63	—
RK4J57	V63	—
RK4J58	V63	—
RK4J59	V63	—
RK5J26	V63	—
RK725A	V63	—
RK10	V46	4D
RK11	V47	3G
RK12	V47	3G
RK15	V20	4D
RK16	V20	5A
RK17	V20	5F
RK18	V47	3G
RK19	V44	4AT
RK20	V58	T-5C
RK20A	V58	T-5C
RK21	V44	4P
RK22	V44	T-4AG
RK23	V56	6BM
RK24	V21	4D
RK24	V45	4D
RK25	V56	6BM
RK25B	V56	6BM
RK28	V60	5J
RK28A	V60	5J
RK30	V47	2D
RK31	V48	3G
RK32	V48	2D
RK33	V45	T-7DA
RK34	V46	T-7DC
RK35	V48	2D
RK36	V51	2D
RK37	V48	2D
RK38	V51	2D
RK39	V57	5AW
RK41	V57	5AW
RK42	V22	4D
RK43	V22	6C
RK44	V56	6BM
RK46	V58	T-5C
RK47	V59	T-5D
RK48	V60	T-5D
RK48A	V60	T-5D
RK49	V57	6A
RK51	V49	3G
RK52	V49	3G
RK56	V56	5AW
RK57	V52	3N
RK58	V51	3N
RK59	V46	T-4D
RK60	V44	T-4AG
RK61	V34	—
RK61	V36	—
RK62	V36	4D
RK63	V54	2N
RK63A	V54	2N
RK64	V56	5AW
RK65	V61	T-3BC
RK66	V58	T-5C
RK73DA	V63	—
RK75	V57	T-5C
RK106	V46	T-6B
RK705A	V43	T-3AA
RK726C	V62	—
RK866	V44	4P
RK5586	V63	—
RK5609	V63	—
RK5657	V63	—
RK5721	V62	—
RK5976	V62	—
RK5981	V62	—
RK5982	V63	—
RK6013	V62	—
RK6115	V62	—
RM208	V36	—
RM209	V37	—
RX21A	V44	—
SD917A	V34	—
SD828A	V34	—
SD828E	V34	—
SD1103	V62	—
SD1104	V62	—
SN944	V34	—
SN946	V34	—
SN947D	V34	—
SN948C	V34	—
SN953D	V34	—
SN954	V34	—
SN955B	V34	—
SN956B	V34	—
SN957A	V34	—
SN1006	V34	—
SN1007B	V34	—
T20	V46	3G
T21	V57	6A
T40	V48	3G
T55	V49	3G
T60	V49	2D
T100	V50	2D
T125	V52	2N
T200	V54	2N
T300	V54	—
T814	V54	3N
T822	V54	3N
TB35	V58	Fig. 54
TUF20	V46	2T
TW75	V50	2D
TW150	V53	2N
TZ20	V46	3G

SEMICONDUCTORS

VACUUM-TUBE BASE DIAGRAMS

The diagrams on the following pages show standard socket connections corresponding to the base designations given in the column headed "Socket Connections" in the classified tube-data tables. Bottom views are shown throughout. Terminal designations are as follows:

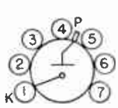
A = Anode	CL = Collector	H = Heater	P = Plate (Anode)	Ref = Reflector
B = Beam	D = Deflecting Plate	IC = Internal Con.	P_1 = Starter-Anode	S = Shell
BP = Bayonet Pin	F = Filament	IS = Internal Shield	P_{BF} = Beam Plates	TA = Target
BS = Base sleeve	FE = Focus Elect.	K = Cathode	RC = Ray-Control Electrode	U = Unit
C = Ext. Coating	G = Grid	NC = No Connection		● = Gas-Type Tube

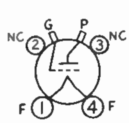
Alphabetical subscripts D, P, T and HX indicate, respectively, diode unit, pentode unit, triode unit or hexode unit in multiunit types. Subscript M, T or CT indicates filament or heater tap.

Generally when the No. 1 pin of a metal-type tube in Table I, with the exception of all triodes, is shown connected to the shell, the No. 1 pin in the glass (G or GT) equivalent is connected to an internal shield.

R.E.T.M.A. TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are shown above.


2AG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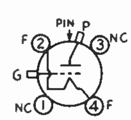
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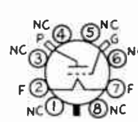
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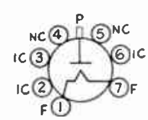
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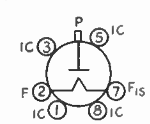
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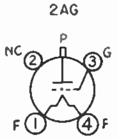
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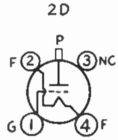
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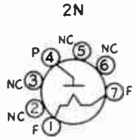
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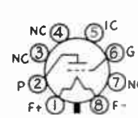
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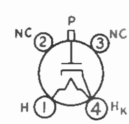
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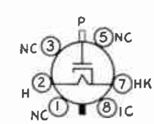
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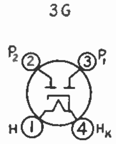
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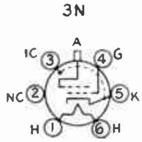
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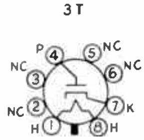
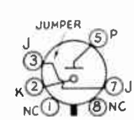
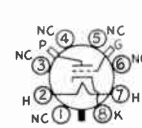
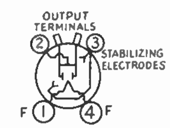

4AH

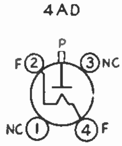


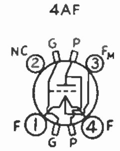
4AJ

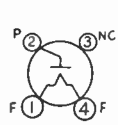

4AM

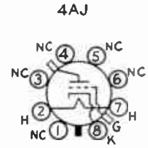


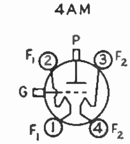
4AP

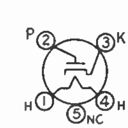

4AT

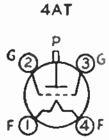

4AQ

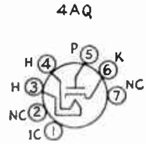

4B

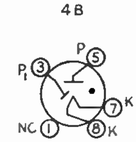

4BB

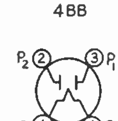

4BC

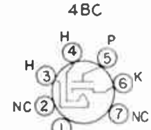

4BJ

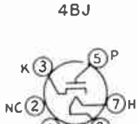

4BO


4BR


4BU

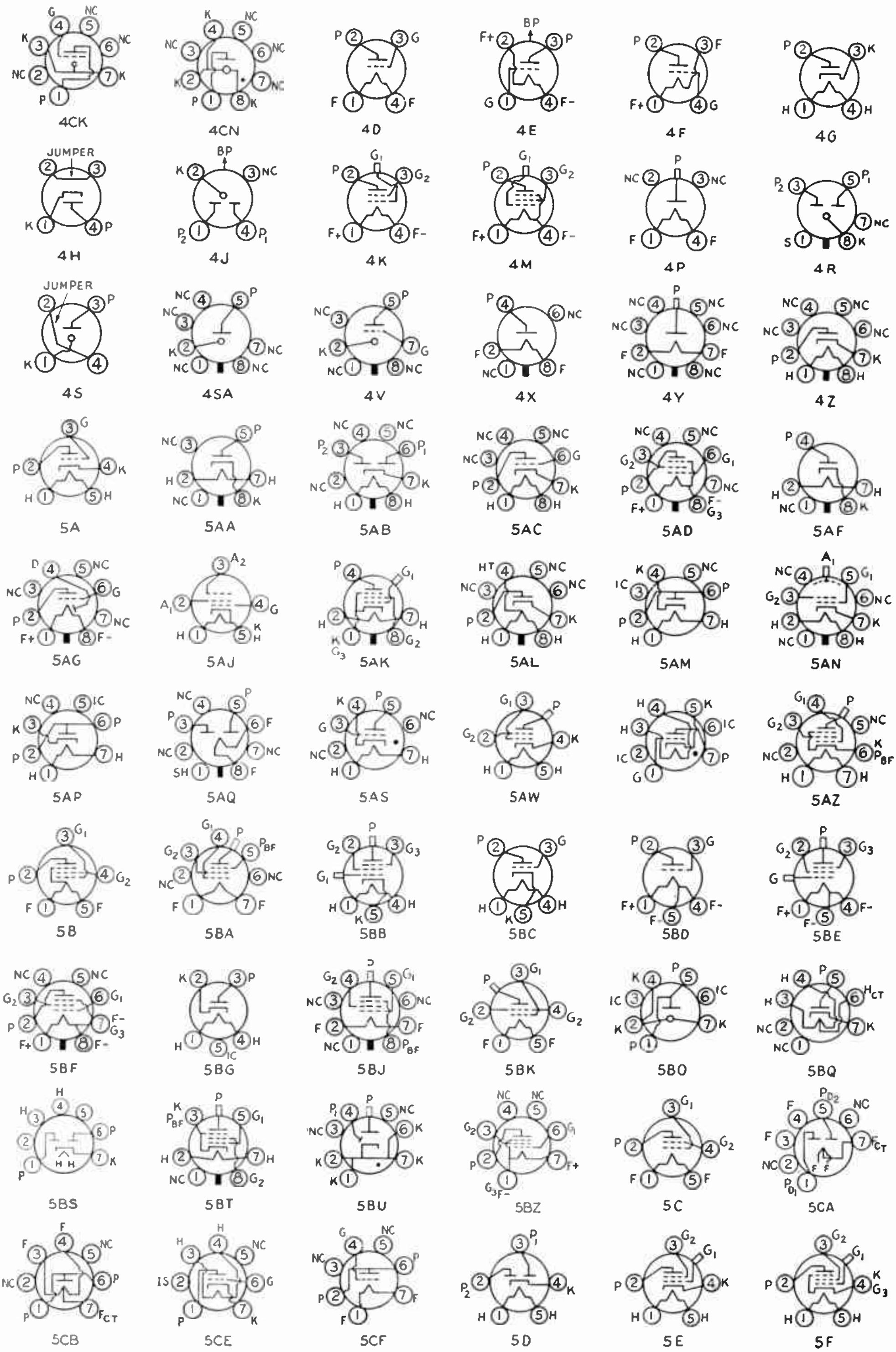

4C


4CB


4CG

TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.



4CK 4CN 4D 4E 4F 4G

4H 4J 4K 4M 4P 4R

4S 4SA 4V 4X 4Y 4Z

5A 5AA 5AB 5AC 5AD 5AF

5AG 5AJ 5AK 5AL 5AM 5AN

5AP 5AQ 5AS 5AW 5AX 5AZ

5B 5BA 5BB 5BC 5BD 5BE

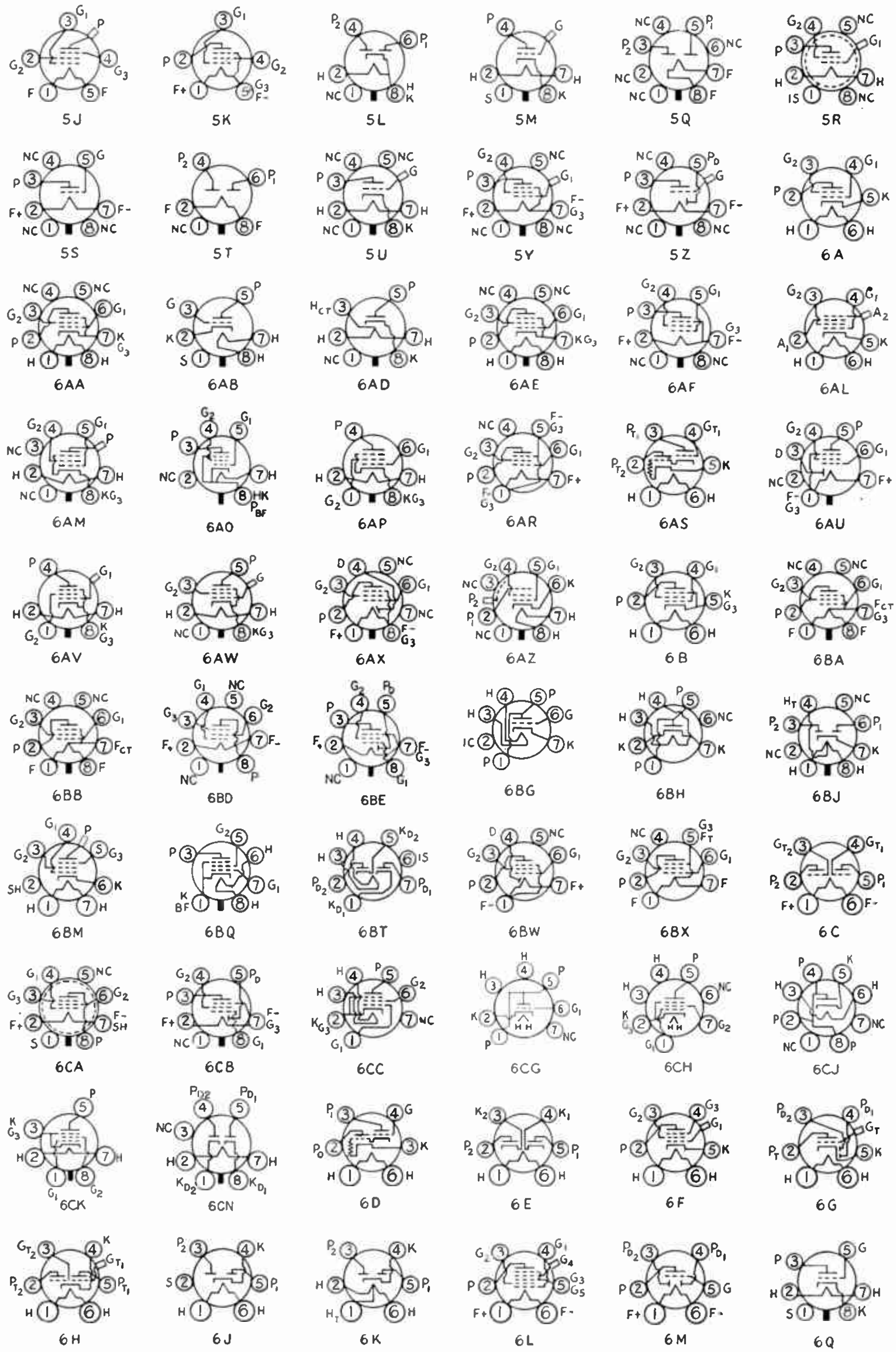
5BF 5BG 5BJ 5BK 5BO 5BQ

5BS 5BT 5BU 5BZ 5C 5CA

5CB 5CE 5CF 5D 5E 5F

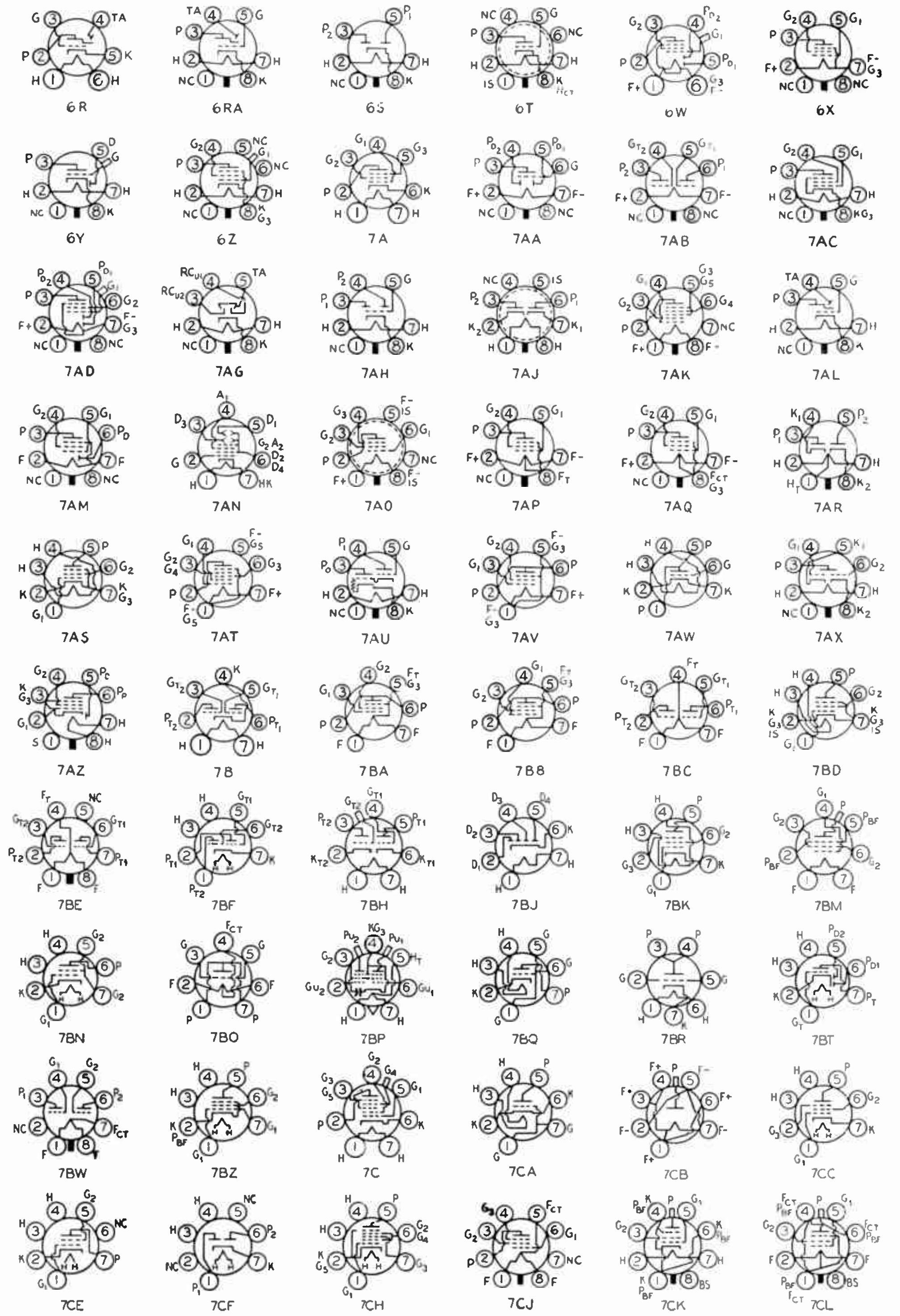
TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.



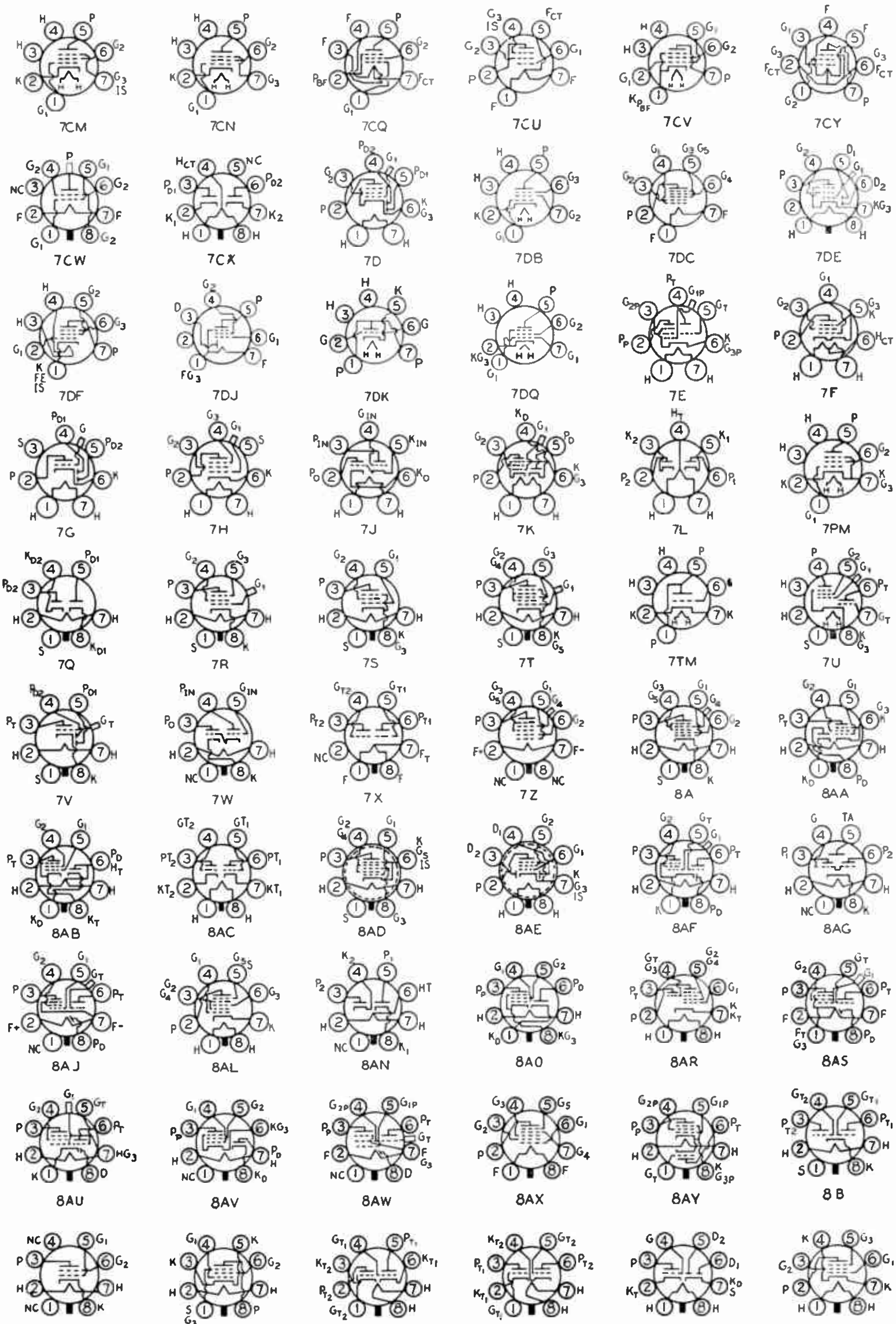
TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.



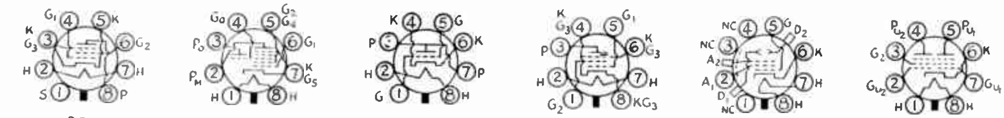
TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.

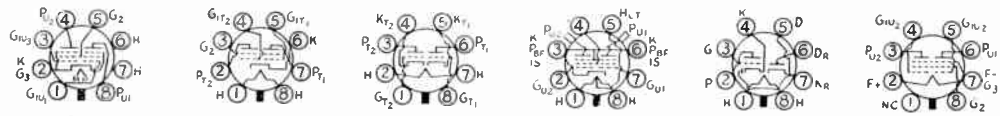


TUBE BASE DIAGRAMS

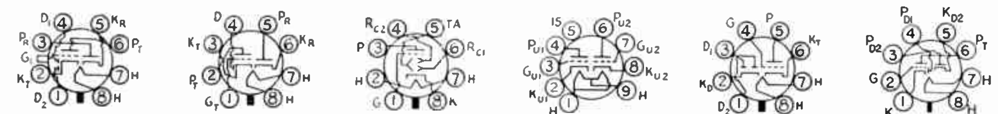
Bottom views are shown. Terminal designations on sockets are given on page V5.



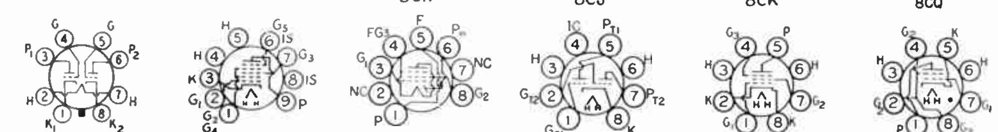
8BK 8BL 8BN 8BO 8BR 8BS



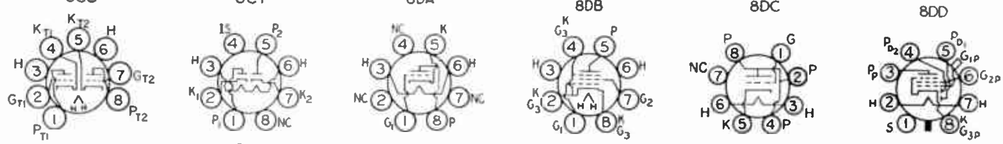
8BU 8BV 8BW 8BY 8BZ 8C



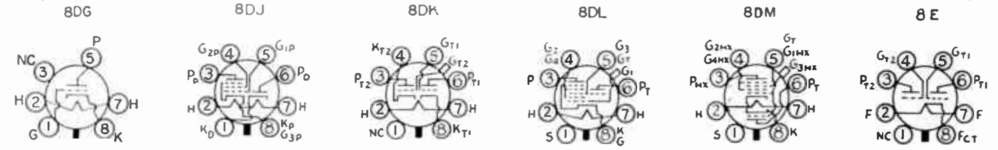
8CB 8CG 8CH 8CJ 8CK 8CQ



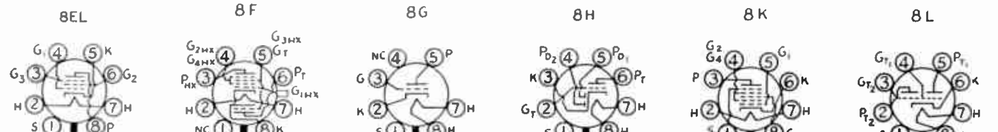
8CS 8CT 8DA 8DB 8DC 8DD



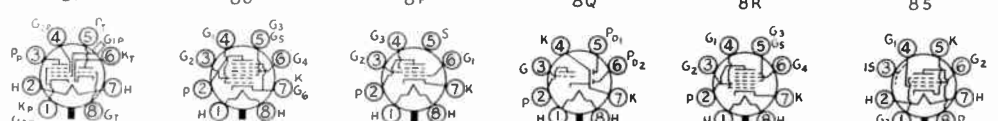
8DG 8DJ 8DK 8DL 8DM 8E



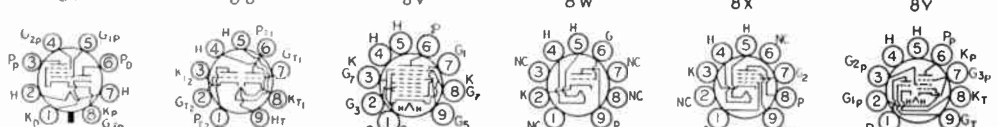
8EL 8F 8G 8H 8K 8L



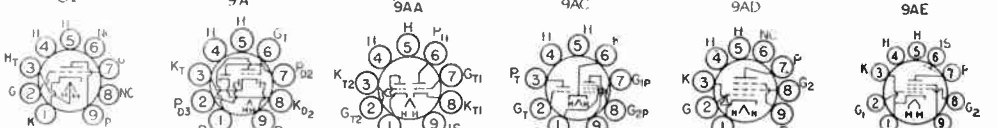
8N 8O 8P 8Q 8R 8S



8T 8U 8V 8W 8X 8Y



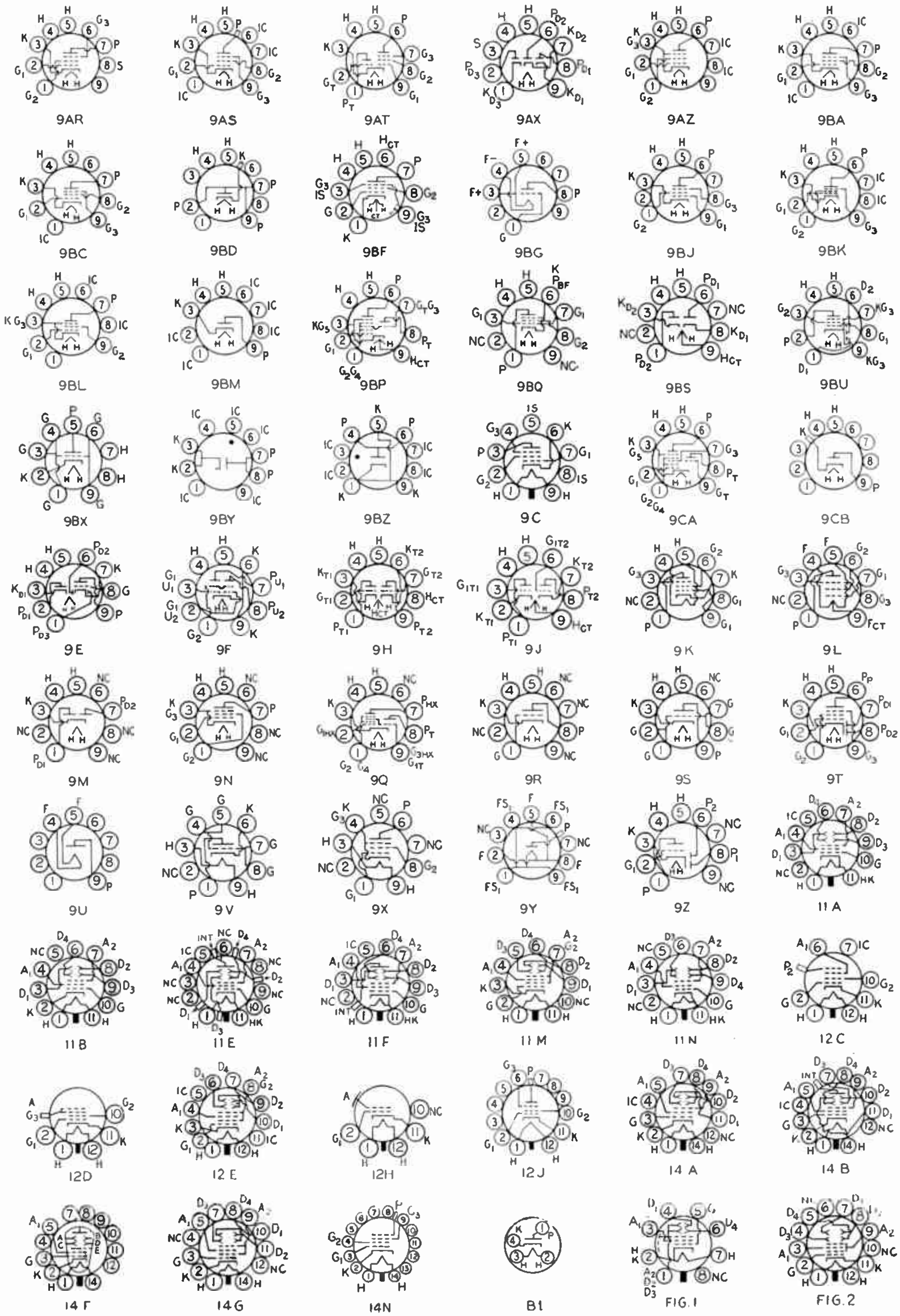
8Z 9A 9AA 9AC 9AD 9AE



9AG 9AH 9AJ 9AK 9AM 9AQ

TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.



TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.

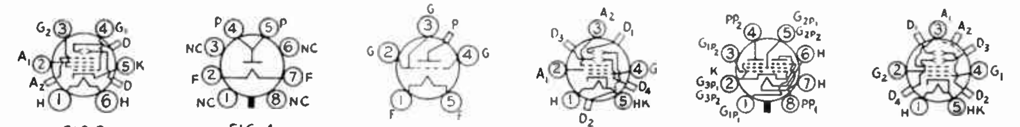


FIG 3 · FIG. 4 · FIG.5 · FIG 6 · FIG. 7 · FIG. 8

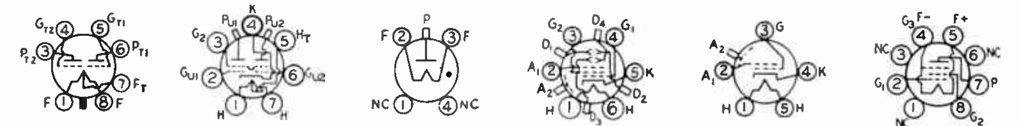


FIG. 9 · FIG. 10 · FIG. 11 · FIG. 12 · FIG 13 · FIG. 14

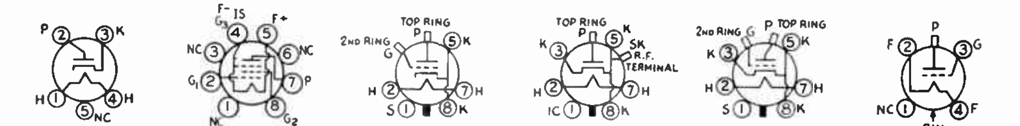


FIG 15 · FIG. 16 · FIG 17 · FIG 18 · FIG 19 · FIG. 26

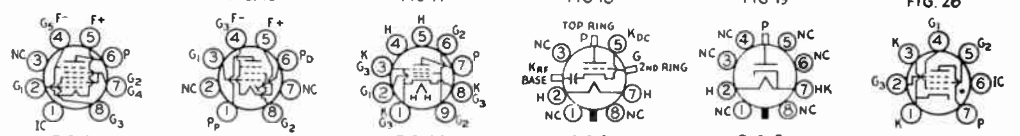


FIG 27 · FIG 28 · FIG. 29 · FIG. 30 · FIG. 31 · FIG. 33

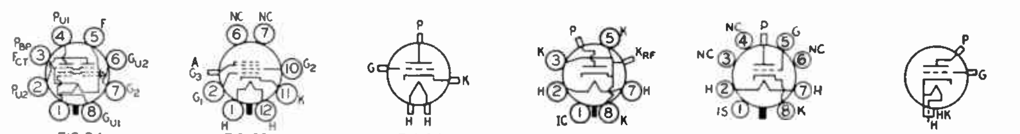


FIG 34 · FIG 35 · FIG. 36 · FIG. 37 · FIG 38 · FIG. 39

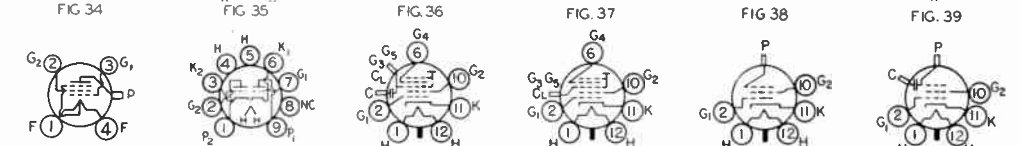


FIG 40 · FIG. 41 · Fig. 42 · Fig. 43 · Fig. 44 · Fig. 45

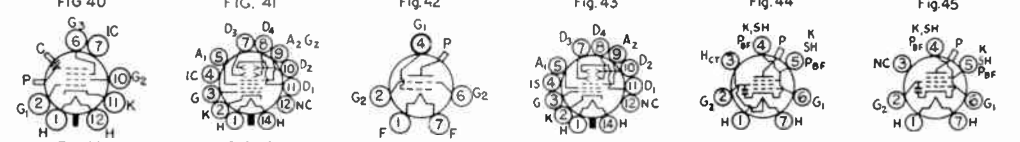


Fig 46 · FIG 47 · Fig. 48 · FIG 49 · FIG. 50 · FIG. 51

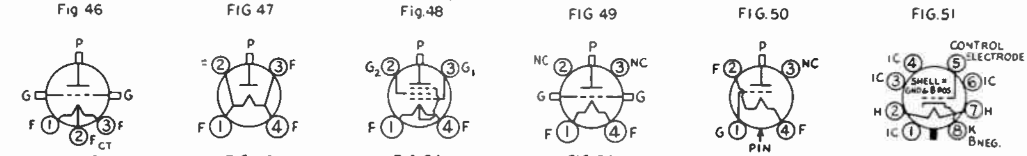


FIG. 52 · FIG. 53 · FIG. 54 · FIG. 56 · FIG. 57 · FIG. 58

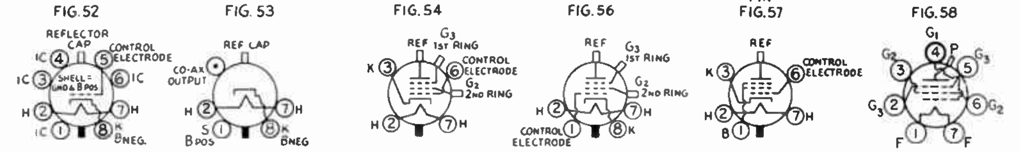


FIG 59 · FIG. 60 · FIG. 61 · FIG. 62 · FIG. 63 · Fig 64

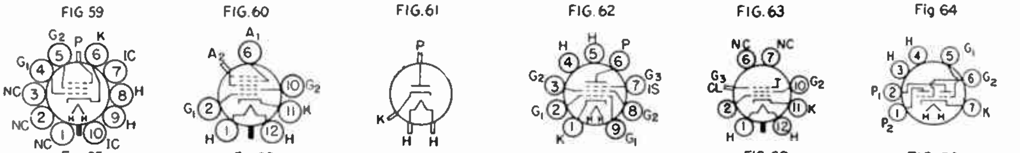


Fig 65 · Fig 66 · FIG. 67 · FIG. 68 · FIG. 69 · FIG. 70

TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page V5.

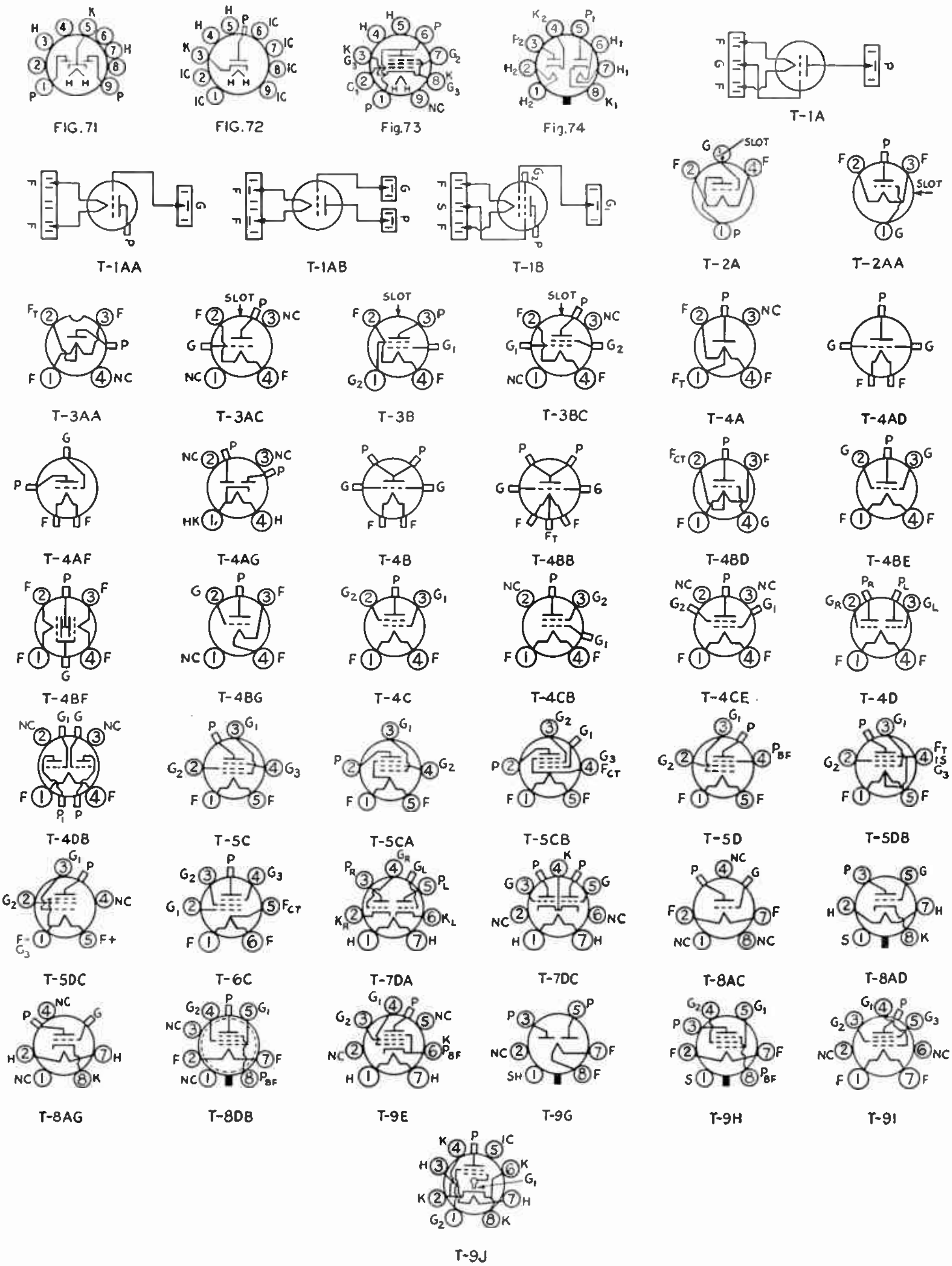


TABLE I—METAL RECEIVING TUBES

Characteristics given in this table apply to all tubes having type numbers shown, including metal tubes, glass tubes with "G" suffix, and bantam tubes with "GT" suffix. For "G" and "GT" tubes not listed (not having metal counterparts), see Tables II, VII, VIII and IX.

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6A8	Pentagrid Converter	8A	6.3	0.3	Osc. Grid leak = 50000Ω			Converter	250	−3.0	100	2.7	3.5	Anode-grid (No. 2) 250 volts max. thru 20,000 ohms					6A8
6AB7 1853	Remote Cut-off Pentode	8N	6.3	0.45	8	5	0.015	Class-A Amp.	300	−3.0	200	3.2	12.5	700000	5000	3500	—	—	6AB7 1853
6AC7 1852	Sharp Cut-off Pentode	8N	6.3	0.45	11	5	0.015	Class-A Amp.	300	160*	150	2.5	10	1000000	9000	6750	—	—	6AC7 1852
6AG7	Power Pentode	8Y	6.3	0.65	13	7.5	0.06	Class-A_1 Amp.	300	−3.0	150	7/9	30/30.5	130000	11000	—	10000	3.0	6AG7
6AJ7	Sharp Cut-off Pentode	8N	6.3	0.45	—	—	—	Class-A Amp.	300	160*	300	2.5	10	1000000	9000	—	—	—	6AJ7
6AK7	Pentode Power Amp.	8Y	6.3	0.65	13	7.5	0.06	Class-A Amp.	300	−3	150	7	30	130000	11000	—	10000	3.0	6AK7
6B8	Duplex-Diode Pentode	8E	6.3	0.3	6	9	0.005	Class-A Amp.	250	−3.0	125	2.3	9.0	650000	1125	730	—	—	6B8
6C5	Triode	6Q	6.3	0.3	3	11	2	Class-A Amp.	250	−8.0	—	—	8.0	10000	2000	20	—	—	6C5
								Bias Detector	250	−17.0	—	—	Plate current adjusted to 0.2 ma. with no signal						
6F5	High-μ Triode	5M	6.3	0.3	5.5	4	2.3	Class-A Amp.	250	−1.3	—	—	0.2	66000	1500	100	—	—	6F5
6F6	Pentode Power Amplifier	7S	6.3	0.7	6.5	13	0.2	Class-A_1 Pent.[3]	250 315	−16.5 −22.0	250 315	6.5 8.0	36[7] 42	80000 75000	2500 2650	200 200	7000 7000	3.2 5.0	6F6
								Class-A_1 Triode[1]	250	−20.0	—	—	34[7]	2600	2600	6.8	4000	0.85	
								Class-AB_1 Amp.[6] Class-AB_2 Amp.[6]	375 375	340* −26.0	250 250	8/18 5/19.5	54/77 34/82	Power output for 2 tubes at stated load, plate-to-plate			10000[8] 10000[8]	19.0 18.5	
								Class-AB_2 Amp.[1] [6]	350 350	730* −38	—	—	50/61 48/92	—	—	—	10000[8] 6000[8]	9 13	
6H6	Twin Diode	7Q	6.3	0.3	—	—	—	Rectifier	Max. a.c. voltage per plate = 150 r.m.s. Max. output current 8.0 ma. d.c.										6H6
6J5	Triode	6Q	6.3	0.3	3.4	3.6	3.4	Class-A Amp.	250	−8.0	—	—	9	7700	2600	20	—	—	6J5
6J7	Sharp Cut-off Pentode	7R	6.3	0.3	7	12	0.005	R.F. Amp.	250	−3.0	100	0.5	2.0	1.5 meg.	1225	1500	—	—	6J7
								Bias Detector	250	−4.3	100	Cathode current 0.43 ma.			—	—	0.5 meg.	—	
6K7	Variable-μ Pentode	7R	6.3	0.3	7	12	0.005	R.F. Amp.	250	−3.0	125	2.6	10.5	600000	1650	990	—	—	6K7
								Mixer	250	−10.0	100	—	—	—	Oscillator peak volts = 7.0				
6K8	Triode-Hexode	8K	6.3	0.3	—	—	—	Converter	250	−3.0	100	6	2.5	Triode Plate (No. 2) 100 volts, 3.8 ma.					6K8
6L6	Beam Power Amplifier	7AC	6.3	0.9	10	12	0.4	Single Tube Class A_1	250 300	170* 220*	250 200	5.4/7.2 3.0/4.6	75/78 51/54.5	—	—	—	2500 4500	6.5 6.5	6L6
								Single Tube Class A_1	250 350	−14.0 −18.0	250 250	5.0/7.3 2.5/7.0	72/79 54/66	22500 33000	6000 5200	—	2500 4200	6.5 10.8	
								P.P. Class A_1[6]	270	125*	270	11/17	134/145	—	—	—	5000[8]	18.5	
								P.P. Class A_1[6]	250 270	−16.0 −17.5	250 270	10/16 11/17	120/140 134/155	24500 23500	5500 5700	—	5000[8] 5000[8]	14.5 17.5	
								P.P. Class AB_1[5]	360	250*	270	5/17	88/100	Power output for 2 tubes. Load plate-to-plate			9000[8]	24.5	
								P.P. Class AB_1[6]	360	−22.5	270	5/15	88/132				6600[8]	26.5	
								P.P. Class AB_2[6]	360 360	−18.0 −22.5	225 270	3.5/11 5/16	78/142 88/205				6000[8] 3800[8]	31.0 47.0	
6L7	Pentagrid Mixer Amplifier	7T	6.3	0.3	—	—	—	R.F. Amp.	250	−3.0	100	5.5	5.3	800000	1100	—	—	—	6L7
								Mixer	250	−6.0	150	8.3	3.3	Over 1 meg.	Oscillator-grid (No. 3) voltage = −15				
6N7	Twin Triode	8B	6.3	0.8	—	—	—	Class-B Amp.	300	0	—	—	35/70	—	—	—	8000	10.0	6N7
6Q7	Duplex-Diode Triode	7V	6.3	0.3	5	3.8	1.4	Triode Amp.	250	−3.0	—	—	1.1	58000	1200	70	—	—	6Q7
6R7	Duplex-Diode Triode	7V	6.3	0.3	4.8	3.8	2.4	Triode Amp.	250	−9.0	—	—	9.5	8500	1900	16	10000	0.28	6R7
6S7	Remote Cut-off Pentode	7R	6.3	0.15	6.5	10.5	0.005	Class-A Amp.	250	−3.0	100	2.0	8.5	1000000	1750	—	—	—	6S7
6SA7	Pentagrid Converter	8R[2]	6.3	0.3	—	—	—	Converter	250	0[3]	100	8.0	3.4	800000	Grid No. 1 resistor 20000 ohms				6SA7
6SB7Y	Pentagrid Converter	8R	6.3	0.3	9.6	9.2	—	Converter	100	−1	100	10.2	3.6	500000	900	—	—	—	6SB7Y
								Converter	250	−1	100	10	3.8	1000000	950	—	—	—	
					Osc. Section in 88–108 Mc. Serv.				250	22000[9]	12000[9]	12.6/12.5	6.8/6.5	—	—	—	—	—	
6SC7	Twin-Triode	8S	6.3	0.3	—	—	—	Class-A Amp.	250	−2.0	—	—	2.0	53000	1325	70	—	—	6SC7

TABLE I—METAL RECEIVING TUBES—*Continued*

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6SF5	High-μ Triode	6AB	6.3	0.3	4	3.6	2.4	Class-A Amp.	250	− 2.0	—	—	0.9	66000	1500	100	—	—	6SF5
6SF7	Diode Variable-μ Pentode	7AZ	6.3	0.3	5.5	6	0.004	Class-A Amp.	250	− 1.0	100	3.3	12.4	700000	2050	—	—	—	6SF7
6SG7	Semivariable-μ Pentode	8BK	6.3	0.3	8.5	7	0.003	H.F. Amp.	250	− 2.5	150	3.4	9.2	Over 1 meg.	4000	—	—	—	6SG7
6SH7	Sharp Cut-off Pentode	8BK	6.3	0.3	8.5	7	0.003	Class-A Amp.	250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	6SH7
6SJ7[4]	Sharp Cut-off Pentode	8N	6.3	0.3	6	7	0.005	Class-A Amp.	250	− 3.0	100	0.8	3	1500000	1650	2500	—	—	6SJ7
6SK7	Variable-μ Pentode	8N	6.3	0.3	6	7	0.003	Class-A Amp.	250	− 3.0	100	2.4	9.2	800000	2000	1600	—	—	6SK7
6SQ7	Duplex-Diode Triode	8Q	6.3	0.3	3.2	3.0	1.6	Class-A Amp.	250	− 2.0	—	—	0.8	91000	1100	100	—	—	6SQ7
6SR7	Duplex-Diode Triode	8Q	6.3	0.3	3.6	2.8	2.40	Class-A Amp.	250	− 9.0	—	—	9.5	8500	1900	16	—	—	6SR7
6SS7	Variable-μ Pentode	8N	6.3	0.15	5.5	7.0	0.004	Class-A Amp.	250	− 3.0	100	2.0	9.0	1000000	1850	—	—	—	6SS7
6ST7	Duplex-Diode Triode	8Q	6.3	0.15	2.8	3	1.50	Class-A Amp.	250	− 9.0	—	—	9.5	8500	1900	16	—	—	6ST7
6SV7	Diode R.F. Pentode	7AZ	6.3	0.3	6.5	6	0.004	Class-A Amp.	250	− 1	150	2.8	7.5	800000	3400	—	—	—	6SV7
6SZ7	Duplex-Diode Triode	8Q	6.3	0.15	2.6	2.8	1.10	Class-A Amp.	250	− 3	—	—	1.0	58000	1200	70	—	—	6SZ7
6T7	Duplex-Diode Triode	7V	6.3	0.15	1.8	3.1	1.70	Class-A Amp.	250	− 3.0	—	—	1.2	62000	1050	65	—	—	6T7
6V6	Beam Power Amplifier	7AC	6.3	0.45	2.0	7.5	0.7	Class-A1 Amp.[5]	250	−12.5	250	4.5/7.0	45/47	52000	4100	218	5000	4.5	6V6
								Class-AB1 Amp.[6]	250	−15.0	250	5/13	70/79	60000	3750	—	10000[8]	10.0	
									285	−19.0	285	4/13.5	70/92	65000	3600	—	8000[8]	14.0	
1611	Pentode Power Amplifier	7S	6.3	0.7	—	—	—	Audio Amp.		Characteristics same as 6F6									1611
1612	Pentagrid Amplifier	7T	6.3	0.3	7.5	11	0.001	Class-A Amp.	250	− 3.0	100	6.5	5.3	600000	1100	880	—	—	1612
1620	Sharp Cut-off Pentode	7R	6.3	0.3	—	—	—	Class-A Amp.		Characteristics same as 6J7									1620
1621	Power Amplifier Pentode	7S	6.3	0.7	—	—	—	Class-AB2 Amp.[6]	300	−30.0	300	6.5/13	38/69	—	—	—	4000[8]	5.0	1621
								Class-A1 Amp.[6, 1]	330	500*	—	—	55/59	—	—	—	5000[8]	2.0	
1622	Beam Power Amplifier	7AC	6.3	0.9	—	—	—	Class-A1 Amp.	300	−20.0	250	4/10.5	86/125	—	—	—	4000	10.0	1622
1851	Television Amp. Pentode	7R	6.3	0.45	11.5	5.2	0.02	Class-A Amp.	300	− 2.0	150	2.5	10	750000	9000	6750	—	—	1851
5693	Sharp Cut-off Pentode	8N	6.3	0.3	5.3	6.2	0.005	Class-A Amp.	250	− 3	100	0.85	3.0	1000000	1650	—	—	—	5693
5961	Pentagrid Converter	8R	6.3	0.3	Osc. Grid 20K Ω			Converter	250	− 2	100	8.5	3.5	1000000	Conversion Gm = 450				5961
6137	Remote Cut-off Pentode	8N	6.3	0.3	5.0	6.5	0.003	Class-A1 Amp.	250	− 3	100	2.6	9.2	800000	2000	—	—	—	6137

* Cathode resistor—ohms.
[1] Screen tied to plate.
[2] For 6SA7GT use base diagram 8AD.
[3] Grid bias—2 volts if separate oscillator excitation is used.
[4] Also Type "6SJ7Y."
[5] Values are for single tube.
[6] Values are for two tubes in push-pull.
[7] Max.-signal value.
[8] Plate-to-plate value.
[9] Osc. grid leak—Scrn res.

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES

(For "G" and "GT"-Type Tubes Not Listed Here, See Equivalent Type in Table I; Characteristics and Connections Will Be Identical)

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
2B22	Diode	Fig. 37	6.3	0.75	2.2	—	—	U.h.f. Detector		Average cathode Ma. = 5; Output volts = 50 d.c.; Load resistance = 10000.									2B22
2C22	Triode	4AM	6.3	0.3	2.2	0.7	3.60	Class-A Amp.	300	−10.5	—	—	11	6600	3000	20	—	—	2C22
6A5GT	Triode Power Amplifier	6T	6.3	1.0	—	—	—	Class-A Amp.[4]	250	−45.0	—	—	60	800		4.2	2500	3.75	6A5G
								P.P. Class AB[5]	325	−68.0	—	—	80	—	5250	—	3000[6]	15.0	
								P.P. Class AB[5]	325	850*			80	—		—	5000[6]	10.0	
6AB6G	Direct-Coupled Amplifier	7AU	6.3	0.5	—	—	—	Class-A Amp.	250	0	Input		5.0	40000	1800	72	8000	3.5	6AB6G
									250	0	Output		34						
6AC5GT	High-μ Power-Amplifier Triode	6Q	6.3	0.4	—	—	—	P.P. Class B[5]	250	0	—	—	5.0	36700	3400	125	10000[6]	8.0	6AC5GT
								Dyn.-Coupled	250	—	—	—	32				7000	3.7	
6AC6G	Direct-Coupled Amplifier	7AU	6.3	1.1	—	—	—	Class-A Amp.	180	0	Input		7.0	—	3000	54	4000	3.8	6AC6G
									180	0	Output		45						
6AD5G	High-μ Triode	6Q	6.3	0.3	4.1	3.9	3.3	Class-A Amp.	250	− 2.0	—	—	0.9	—	1500	100	—	—	6AD5G
6AD6G[10]	Electron-Ray Tube	7AG	6.3	0.15	—	—	—	Indicator	100		0 for 90°; —23 for 135°; 45 for 0°. Target current 1.5 ma. for 0°.								6AD6G

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6AD7G	Triode-Pentode	8AY	6.3	0.85	—	—	—	Triode Amp.	250	−25.0	—	—	4.0	19000	325	6.0	—	—	6AD7G
								Pentode Amp.	250	−16.5	250	6.5	34	80000	2500	—	7000	3.2	
6AE5G[10]	Triode Amplifier	6Q	6.3	0.3	—	—	—	Class-A Amp.	95	−15.0	—	—	7.0	3500	1200	4.2	—	—	6AE5G
6AE6G	Single-Grid Twin-Plate Triode	7AH	6.3	0.15	—	—	—	Plate No. 1	250	1.5	—	—	6.5	25000	1000	25	—	—	6AE6G
								Plate No. 2	250	1.5	—	—	4.5	35000	950	33	—	—	
6AE7GT[10]	Twin-Input Triode	7AX	6.3	0.5	—	—	—	Driver Amplifier	250	−13.5	—	—	5.0	9300	1500	14	—	—	6AE7GT
6AF5G	Triode	6Q	6.3	0.3	—	—	—	Class-A Amplifier	180	−18.0	—	—	7.0	—	1500	7.4	—	—	6AF5G
6AF7G	Twin Electron Ray	8AG	6.3	0.3	—	—	—	Indicator Tube											6AF7G
6AG6G[10]	Power-Amplifier Pentode	7S	6.3	1.25	—	—	—	Class-A Amplifier	250	− 6.0	250	6.0	32	—	10000	—	8500	3.75	6AG6G
6AH4GT	Triode	8EL	6.3	0.75	7.5	3.2	4.2	Class-A Amplifier	250	−23	—	—	30	1780	4500	8	—	—	6AH4GT
6AH5G	Beam Power Amplifier	6AP	6.3	0.9	—	—	—	Class-A Amplifier	350	−18	250	—	—	33000	5200	—	4200	10.8	6AH5G
6AH7GT	Twin Triode	8BE	6.3	0.3	—	—	—	Converter & Amp.	250	− 9.0	—	—	12[1]	6600	2400	16	—	—	6AH7GT
6AL6G	Beam Power Amplifier	6AM	6.3	0.9	—	—	—	Class-A Amplifier	250	−14.0	250	5.0	72	22500	6000	—	2500	6.5	6AL6G
6AL7GT	Electron-Ray Tube	8CH	6.3	0.15	—	—	—	Indicator	Outer edge of any of the three illuminated areas displaced 1/16 in. min. outward with +5 volts to its electrode. Similar inward disp. with —5 volts. No pattern with —6 volts grid.										6AL7GT
6AQ7GT	Duplex Diode Triode	8CK	6.3	0.3	2.3	1.5	2.8	Class-A Amplifier	250	− 2.0	—	—	2.3	44000	1600	70	—	—	6AQ7GT
6AR6	Beam Power Amp.	6BQ	6.3	1.2	11	7	0.55	Class-A Amplifier	250	−22.5	250	5	77	21000	5400	95	—	—	6AR6
6AR7GT	Duo-Diode Remote Pentode	7DE	6.3	0.3	5.5	7.5	.003	Class-A Amplifier	250	− 2	100	1.8	7.0	1200000	2500	—	—	—	6AR7GT
6AS7G	Low-Mu Twin Triode	8BD	6.3	2.5	—	—	—	D.C. Amplifier	135	250*	—	—	125	280	7500	2.1	—	—	6AS7G
								Class-A_1 Amp. P.P.	250	2500*	—	—	100/106	280	225[9]	—	6000[8]	13	
6AU5GT	Beam Pentode	6CK	6.3	1.25	11.3	7	0.5	Horz. Def. Amp.	450[11]	−50[11]	—	—	100[11]	Peak pos. plate pulse = 5000 volts.					6AU5GT
6AV5GT	Beam Pentode	6CK	6.3	1.2	—	—	—	Horz. Def. Amp.	500[11]	−50[11]	175[11]	—	100[11]	Peak pos. plate pulse = 4500 volts.					6AV5GT
6AW7GT	Twin Triode	8CQ	6.3	0.3	—	—	—	Class-A Amplifier	100	0	—	—	1.4	—	1200	80	—	—	6AW7GT
6B4G	Triode Power Amplifier	5S	6.3	1.0	—	—	—	Power Amplifier	Characteristics same as Type 6A3—Table IV							—	—	—	6B4G
6B6G	Duplex-Diode High-μ Triode	7V	6.3	0.3	1.7	3.8	1.7	Detector-Amplifier	Characteristics same as Type 75—Table IV							—	—	—	6B6G
6BD5GT	Beam Pentode	6CK	6.3	0.9	—	—	—	Horz. Def. Amp.	325[11]	—	325[11]	—	100[11]	Peak pos. plate pulse = 4000 volts.					6BD5GT
6BL7GT	Double Triode	8BD	6.3	1.5	4.4	1.1	4	Class-A Amp.	250	− 9	—	—	40[1]	2000	7000	14	—	—	6BL7GT
6BQ6GT	Beam Pentode	6AM	6.3	1.2	—	—	—	Deflection Amp.	550[11]	—	150	—	100[11]	Peak pos. plate pulse = 4000 volts.					6BQ6GT
6BG6G	Beam Power Amplifier	5BT	6.3	0.9	11	6.5	0.5	Deflection Amp.	700[11]	−50[11]	350	—	100[11]	Peak pos. plate pulse = 6000 volts.					6BG6G
6BX7GT	Twin Triode	8BD	6.3	1.5	4.4	1.1	4.2	Class-A Amplifier	250	390*	—	—	42	1300	7600	10	—	—	6BX7GT
6C8G	Twin Triode	8G	6.3	0.3	—	—	—	Amp. 1 Section	250	− 4.5	—	—	3.1	26000	1450	38	—	—	6C8G
6CD6G	Beam Pentode	5BT	6.3	2.5	26	10	1.0	Horz. Def. Amp.	700[11]	−50[11]	175[11]	—	170[11]	Peak pos. plate pulse = 6000 volts.					6CD6G
6D8G	Pentagrid Converter	8A	6.3	0.15	—	—	—	Converter	250	− 3.0	100	Cathode current 13.0 Ma. Anode grid (No. 2) Volts = 250[3]							6D8G
6E8G[10]	Triode-Hexode Converter	8O	6.3	0.3	—	—	—	Converter	250	− 2.0	Triode Plate 150 volts								6E8G
6F8G	Twin Triode	8G	6.3	0.6	—	—	—	Amplifier	250	− 8.0	—	—	9[1]	7700	2600	20	—	—	6F8G
6G6G	Pentode Power Amplifier	7S	6.3	0.15	—	—	—	Class-A Amplifier	180	− 9.0	180	2.5	15	175000	2300	400	10000	1.1	6G6G
								Class-A Amplifier[2]	180	−12.0	—	—	—	4750	2000	9.5	12000	0.25	
6H4GT	Diode Rectifier	5AF	6.3	0.15	—	—	—	Detector	100	—	—	—	4.0	—	—	—	—	—	6H4GT
6H8G	Duo-Diode High-μ Pentode	8E	6.3	0.3	—	—	—	Class-A Amplifier	250	− 2.0	100	—	8.5	650000	2400	—	—	—	6H8G
6J8G[10]	Triode Heptode	8H	6.3	0.3	—	—	—	Converter	250	− 3.0	100	2.8	1.2	Anode-grid (No. 2) 250 volts max.[3] 5 ma.					6J8G
6K5GT[10]	High-μ Triode	5U	6.3	0.3	2.4	3.6	2.0	Class-A Amplifier	250	− 3.0	—	—	1.1	50000	1400	70	—	—	6K5GT
6K6GT	Pentode Power Amplifier	7S	6.3	0.4	—	—	—	Class-A Amplifier	Characteristics same as Type 41—Table IV										6K6GT
6L5G	Triode Amplifier	6Q	6.3	0.15	2.8	5.0	2.8	Class-A Amplifier	250	− 9.0	—	—	8.0	—	1900	17	—	—	6L5G
6M6G[10]	Power Amplifier Pentode	7S	6.3	1.2	—	—	—	Class-A Amplifier	250	− 6.0	250	4.0	36	—	9500	—	7000	4.4	6M6G
6M7G	Pentode Amplifier	7R	6.3	0.3	—	—	—	R.F. Amplifier	250	− 2.5	125	2.8	10.5	900000	3400	—	—	—	6M7G
6M8GT	Diode Triode Pentode	8AU	6.3	0.6	—	—	—	Triode Amplifier	100	—	—	—	0.5	91000	1100	—	—	—	6M8GT
								Pentode Amplifier	100	− 3.0	100	—	8.5	200000	1900	—	—	—	
6N6G[10]	Direct-Coupled Amplifier	7AU	6.3	0.8	—	—	—	Power Amplifier	Characteristics same as Type 6B5—Table IV							—	—	—	6N6G
6P5GT[10]	Triode Amplifier	6Q	6.3	0.3	3.4	5.5	2.6	Class-A Amplifier	250	−13.5	—	—	5.0	9500	1450	13.8	—	—	6P5GT

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

Type	Name	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6P7G[10]	Triode-Pentode	7U	6.3	0.3	—	—	—	Class-A Amplifier	Characteristics same as 6F7—Table IV										6P7G
6P8G	Triode-Hexode Converter	8K	6.3	0.8	—	—	—	Converter	250	−2.0	75	1.4	1.5	Triode Plate 100 v. 2.2 ma.					6P8G
6Q6G	Diode-Triode	6Y	6.3	0.15	—	—	—	Class-A Amplifier	250	−3.0	—	—	1.2	—	1050	65	—	—	6Q6G
6R6G	Pentode Amplifier	6AW	6.3	0.3	4.5	11	0.007	Class-A Amplifier	250	−3.0	100	1.7	7.0	—	1450	1160	—	—	6R6G
6S6GT	Remote Cut-off Pentode	5AK	6.3	0.45	—	—	—	R.F. Amplifier	250	−2.0	100	3.0	13	350000	4000	—	—	—	6S6GT
6S8GT	Triple Diode Triode	8CB	6.3	0.3	1.2	5	2	Class-A Amplifier	250	−2.0	—	—	0.9	91000	1100	100	—	—	6S8GT
6SD7GT	Medium Cut-off Pentode	8M	6.3	0.3	9	7.5	.0035	R.F. Amplifier	250	−2.0	100	1.9	6.0	1000000	3600	—	—	—	6SD7GT
6SE7GT	Sharp Cut-off Pentode	8N	6.3	0.3	8	7.5	.005	R.F. Amplifier	250	−1.5	100	1.5	4.5	1100000	3400	3750	—	—	6SE7GT
6SH7L	Pentode R.F. Amp.	8BK	6.3	0.3	—	—	—	Class-A Amplifier	250	−1.0	150	4.1	10.8	900000	4900	—	—	—	6SH7L
6SL7GT	Twin Triode	8BD	6.3	0.3	—	—	—	Class-A Amplifier	250	−2.0	—	—	2.3[1]	44000	1600	70	—	—	6SL7GT
6SN7GT 6SN7GTA	Twin Triode	8BD	6.3	0.6	—	—	—	Class-A Amplifier	250	−8.0	—	—	9.0[1]	7700	2600	20	—	—	6SN7GT 6SN7GTA
6SU7GTY	Twin Triode	8BD	6.3	0.3	—	—	—	Class-A Amplifier	250	−2.0	—	—	2.3	44000	1600	70	—	—	6SU7GTY
6T6GM[10]	Amplifier	6Z	6.3	0.45	—	—	—	Class-A Amplifier	250	−1.0	100	2.0	10	1000000	5500	—	—	—	6T6GM
6U6GT	Beam Power Amplifier	7AC	6.3	0.75	—	—	—	Class-A Amplifier	200	−14.0	135	3.0	56	20000	6200	—	3000	5.5	6U6GT
6U7G	Variable-μ Pentode	7R	6.3	0.3	5	9	.007	Class-A Amplifier	Characteristics same as Type 6D6—Table III										6U7G
6V5GT	Beam Power Amplifier	6AO	6.3	0.45	9.0	10	0.6	Class-A Amplifier	315	−13	225	6.0	35	77000	3750	—	8500	5.5	6V5GT
6V7G[10]	Duplex Diode-Triode	7V	6.3	0.3	2	3.5	1.7	Detector-Amplifier	Characteristics same as Type 85—Table III										6V7G
6W6GT	Beam Power Amplifier	7AC	6.3	1.25	—	—	—	Class-A Amplifier	135	−9.5	135	12.0	61.0	—	9000	215	2000	3.3	6W6GT
6W7G	Pentode Det. Amplifier	7R	6.3	0.15	5	8.5	.007	Class-A Amplifier	250	−3.0	100	2.0	0.5	1500000	1225	1850	—	—	6W7G
6X6G	Electron-Ray Tube	7AL	6.3	0.3	—	—	—	Indicator Tube	250	0 v. for 300°, 2 ma. —8 v. for 0°, 0 ma. Vane grid 125 v.									6X6G
6Y6G	Beam Power Amplifier	7AC	6.3	1.25	15	8	0.7	Class-A Amplifier	135	−13.5	135	3.0	60.0	9300	7000	—	2000	3.6	6Y6G
6Y7G[10]	Twin Triode Amplifier	8B	6.3	0.3	—	—	—	Class-B Amplifier	Characteristics same as Type 79—Table IV										6Y7G
6Z7G	Twin Triode Amplifier	8B	6.3	0.3	—	—	—	Class-B Amplifier	180	0	—	—	8.4	—	—	—	12000	4.2	6Z7G
									135	0	—	—	6.0	—	—	—	9000	2.5	
717A	Sharp Cut-off Pentode	8BK	6.3	0.175	—	—	—	Class-A Amplifier	120	−2.0	120	2.5	7.5	390000	4000	—	—	—	717A
1223	Sharp Cut-off Pentode	7R	6.3	0.3	—	—	—	Class-A Amplifier	Characteristics same as 6C6—Table IV										1223
1635	Twin Triode Amplifier	8B	6.3	0.6	—	—	—	Class-B Amplifier	300	0	—	—	6.6/54	—	—	—	12000	10.4	1635
5691	Hi-Mu Twin Triode	8BD	6.3	0.6	2.4[7] 2.7[8]	2.3[7] 2.7[8]	3.6[7] 3.6[8]	Class-A Amp.	250	−2	—	—	2.3[1]	44000	1600	70	—	—	5691
5692	Medium-Mu Twin Triode	8BD	6.3	0.6	2.3[7] 2.6[8]	2.5[7] 2.7[8]	3.5[7] 3.3[8]	Class-A Amp.	250	−9	—	—	6.5[1]	9100	2200	18	—	—	5692
5694	Twin Triode	8CS	6.3	0.8	—	—	—	Class-A Amplifier	300	−6	—	—	7.0	11000	3200	35	—	—	5694
5881	Beam Power Amp.	7AC	6.3	0.9	—	—	—	Audio Amplifier	Characteristics same as 6L6, Table I										5881
6080	Low-Mu Twin Triode	8BD	6.3	2.5	6.4	2.2	8.4	D.C. Amplifier	135	250*	—	—	125[1]	280	7000	2	—	—	6080
7000	Low-Noise Amplifier	7R	6.3	0.3	—	—	—	Class-A Amplifier	Characteristics same as Type 6J7—Table I										7000

* Cathode resistor-ohms.
[1] Per plate.
[2] Screen tied to plate.
[3] Through 20,000-ohm dropping resistor.
[4] Values are for single tube.
[5] Values are for two tubes in push-pull.
[6] Plate-to-plate value.
[7] No. 1 triode.
[8] No. 2 triode.
[9] Peak a.f. volts G-G.
[10] Discontinued.
[11] Max. value.

TABLE III—7-VOLT LOCK-IN-BASE TUBES—For other lock-in-base types see Tables VIII, IX, and X

Type	Name	Socket Connec-tions	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
7A4	Triode Amplifier	5AC	7.0	0.32	3.4	3	4	Class-A Amplifier	250	−8.0	—	—	9.0	7700	2600	20	—	—	7A4
7A5	Beam Power Amplifier	6AA	7.0	0.75	13	7.2	0.44	Class-A_1 Amplifier	125	−9.0	125	3.2/8	37.5/40	17000	6100	—	2700	1.9	7A5
7A6	Twin Diode	7AJ	7.0	0.16	—	—	—	Rectifier	Max. A.C. volts per plate—150. Max. Output current—10 ma.										7A6
7A7	Remote Cut-off Pentode	8V	7.0	0.32	6	7	.005	Class-A Amplifier	250	−3.0	100	2.0	8.6	800000	2000	1600	—	—	7A7
7A8	Multigrid Converter	8U	7.0	0.16	7.5	9.0	0.15	Converter	250	−3.0	100	3.1	3.0	50000	Anode-grid 250 volts max.[1]				7A8

TABLE III—7-VOLT LOCK-IN-BASE TUBES—*Continued*

Type	Name	Socket Connections	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
7AD7	Pentode	8V	6.3	0.6	11.5	7.5	0.03	Class-A$_1$ Amp.	300	68*	150	7.0	28.0	300000	9500	—	—	—	7AD7
7AF7	Twin Triode	8AC	6.3	0.3	2.2	1.6	2.3	Class-A Amp.	250	−10	—	—	9.0	7600	2100	16	—	—	7AF7
7AG7	Sharp Cut-off Pentode	8V	7.0	0.16	7.0	6.0	0.005	Class-A$_1$ Amp.	250	250*	250	2.0	6.0	750000	4200	—	—	—	7AG7
7AH7	Pentode Amplifier	8V	6.3	0.15	7.0	6.5	0.005	Class-A$_1$ Amplifier	250	250*	250	1.9	6.8	1000000	3300	—	—	—	7AH7
7AJ7	Sharp Cut-off Pentode	8V	6.3	0.3	6.0	6.5	0.007	Class-A$_1$ Amp.	250	− 3	100	0.7	2.2	1 Meg.	1575	—	—	—	7AJ7
									100	− 1	100	1.8	5.5	400000	2275	—	—	—	
7AK7	Sharp Cut-off Pentode	8V	6.3	0.8	12	9.5	4	Class-A$_1$ Amp.	150	0	90	21	40	11500	5500	—	—	—	7AK7
7B4	High-μ Triode	5AC	7.0	0.32	3.6	3.4	1.6	Class-A Amplifier	250	− 2.0	—	—	0.9	66000	1500	100	—	—	7B4
7B5	Pentode Power Amplifier	6AE	7.0	0.43	3.2	3.2	1.6	Class-A$_1$ Amplifier	250	−18.0	250	5.5/10	32/33	68000	2300	—	7600	3.4	7B5
7B6	Duo-Diode Triode	8W	7.0	0.32	3.0	2.4	1.6	Class-A Amplifier	250	− 2.0	—	—	1.0	91000	1100	100	—	—	7B6
7B7	Remote Cut-off Pentode	8V	7.0	0.16	5	7	.005	Class-A Amplifier	250	− 3.0	100	2.0	8.5	700000	1700	1200	—	—	7B7
7B8	Pentagrid Converter	8X	7.0	0.32	10.0	9.0	0.2	Converter	250	− 3.0	100	2.7	3.5	360000	Anode-grid 250 volts max.[1]				7B8
7C5	Tetrode Power Amplifier	6AA	7.0	0.48	9.5	9.0	0.4	Class-A$_1$ Amplifier	250	−12.5	250	4.5/7	45/47	52000	4100	—	5000	4.5	7C5
7C6	Duo-Diode Triode	8W	7.0	0.16	2.4	3	1.4	Class-A Amplifier	250	− 1.0	—	—	1.3	100000	1000	100	—	—	7C6
7C7	Pentode Amplifier	8V	7.0	0.16	5.5	6.5	.007	Class-A Amplifier	250	− 3.0	100	0.5	2.0	2 meg.	1300	—	—	—	7C7
7D7	Triode-Hexode Converter	8AR	7.0	0.48	—	—	—	Converter	250	− 3.0	Triode Plate (No. 3) 150 v. 3.5 ma.								7D7
7E6	Duo-Diode Triode	8W	7.0	0.32	—	—	—	Class-A Amplifier	250	− 9.0	—	—	9.5	8500	1900	16	—	—	7E6
7E7	Duo-Diode Pentode	8AE	7.0	0.32	4.6	4.6	.005	Class-A Amplifier	250	− 3.0	100	1.6	7.5	700000	1300	—	—	—	7E7
7F7	Twin Triode	8AC	7.0	0.32	—	—	—	Class-A Amplifier[2]	250	− 2.0	—	—	2.3	44000	1600	70	—	—	7F7
7F8	Twin Triode	8BW	6.3	0.30	2.8	1.4	1.2	R.F. Amplifier	250	− 2.5	—	—	10.0	10400	5000	—	—	—	7F8
									180	− 1.0			12.0	8500	7000				
7G7/1232	Sharp Cut-off Pentode	8V	7.0	0.48	9	7	.007	Class-A Amplifier	250	− 2.0	100	2.0	6.0	800000	4500	—	—	—	7G7/1232
7G8/1206	Dual Tetrode	8BV	6.3	0.30	3.4	2.6	0.15	R.F. Amplifier[2]	250	− 2.5	100	0.8	4.5	225000	2100	—	—	—	7G8/1206
7H7	Semi-Variable-μ Pentode	8V	7.0	0.32	8	7	.007	R.F. Amplifier	250	− 2.5	150	2.5	9.0	1000000	3500	—	—	—	7H7
7J7	Triode-Heptode Converter	8AR	7.0	0.32	—	—	—	Converter	250	− 3.0	100	2.9	1.3	Triode Plate 250 v. Max.[1]					7J7
7K7	Duo-Diode High-μ Triode	8BF	7.0	0.32	—	—	—	Class-A Amplifier	250	− 2.0	—	—	2.3	44000	1600	70	—	—	7K7
7L7	Sharp Cut-off Pentode	8V	7.0	0.32	8	6.5	.01	Class-A Amplifier	250	− 1.5	100	1.5	4.5	100000	3100	Cathode Resistor 250 ohms			7L7
7N7	Twin Triode	8AC	7.0	0.6	3.4[3] 2.9[4]	2.0[3] 2.4[4]	3.0[3] 3.0[4]	Class-A Amplifier[2]	250	− 8.0	—	—	9.0	7700	2600	20	—	—	7N7
7Q7	Pentagrid Converter	8AL	7.0	0.32	—	—	—	Converter	250	0	100	8.0	3.4	800000	Grid No. 1 resistor 20000 ohms				7Q7
7R7	Duo-Diode Pentode	8AE	7.0	0.32	5.6	5.3	.004	Class-A Amplifier	250	− 1.0	100	1.7	5.7	1000000	3200	—	—	—	7R7
7S7	Triode Hexode Converter	8BL	7.0	0.32	—	—	—	Converter	250	− 2.0	100	2.2	1.7	2000000	Triode Plate 250 v. Max.[1]				7S7
7T7	Pentode Amplifier	8V	7.0	0.32	8	7	.005	Class-A Amplifier	250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	7T7
7V7	Sharp Cut-off Pentode	8V	7.0	0.48	9.5	6.5	.004	Class-A Amplifier	300	160*	150	3.9	10	300000	5800	—	—	—	7V7
7W7	Sharp Cut-off Pentode	8BJ	7.0	0.48	9.5	7.0	.0025	Class-A Amplifier	300	− 2.2	150	3.9	10	300000	5800	—	—	—	7W7
7X7	Duo-Diode Triode	8BZ	6.3	0.3	—	—	—	Class-A Amplifier	250	− 1.0	—	—	1.9	67000	1500	100	—	—	7X7
1231	Pentode Amplifier	8V	6.3	0.45	8.5	6.5	.015	Class-A Amplifier	300	200*	150	2.5	10	700000	5500	3850	—	—	1231
1273	Nonmicrophonic Pentode	8V	7.0	0.32	6.0	6.5	.007	Class-A$_1$ Amplifier	250	− 3.0	100	0.7	2.2	1000000	1575	—	—	—	1273
									100	− 1.0	100	1.8	5.7	400000	2275	—	—	—	
5679	Twin Diode	7CX	6.3	0.15	—	—	—	V.T.V.M. Rectifier	Same as 7A6										5679
XXL	Triode Oscillator	5AC	7.0	0.32	—	—	—	Oscillator	250	− 8.0	—	—	8.0	—	2300	20	—	—	XXL

* Cathode resistor—ohms. [1] Applied through 20000-ohm dropping resistor. [2] Each section. [3] Triode No. 1. [4] Triode No. 2.

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
2C21/1642	Twin-Triode Amplifier	M.	7BH	6.3	0.6	—	—	—	Class-A Amp.	250	−16.5	—	—	8.3	7600	1375	10.4	—	—	2C21/1642
6A3	Triode Power Amplifier	M.	4D	6.3	1.0	7.0	5.0	16.0	Class-A Amp.	250	−45	—	—	60	800	5250	4.2	2500	3.5	6A3
									Class AB$_1$ Amp.[10]	300 300	−62 850[*]	Fixed Bias Self Bias		80 80	—	—	—	3000[11] 5000[11]	15 10	
6A4	Pentode Power Amplifier	M.	5B	6.3	0.3	—	—	—	Class-A Amp.	180	−12.0	180	3.9	22	60000	2500	150	8000	1.5	6A4
6A6	Twin Triode Amplifier	M.	7B	6.3	0.8	—	—	—	Class-B Amp. P.P	250 300	0 0	—	—	Power output is for one tube at stated load, plate-to-plate				8000 10000	8.0 10.0	6A6
6A7	Pentagrid Converter	S.	7C	6.3	0.3	8.5	9.0	0.3	Converter	250	− 3.0	100	2.2	3.5	360000	Anode grid (No. 2) 200 volts max.				6A7
6AB5/6N5	Electron-Ray Tube	S.	6R	6.3	0.15	—	—	—	Indicator Tube	180	Cut-off Grid Bias = −12 v.			0.5	Target Current 2 ma.				—	6AB5/6N5
6AF6G	Electron-Ray Tube Twin Indicator Type	S.	7AG	6.3	0.15	—	—	—	Indicator Tube	135 100	Ray Control Voltage = 81 for 0° Shadow Angle. Target current 1.5 ma. Ray Control Voltage = 60 for 0° Shadow Angle. Target current 0.9 ma.									6AF6G
6B5	Direct-Coupled Power Amplifier	M.	6AS	6.3	0.8	—	—	—	Class-A Amp.[9] Push-Pull Amp.[10]	300 400	0 −13.0	— —	6[1] 4.5[1]	45 40	241000 —	2400 —	58 —	7000 10000[11]	4.0 20	6B5
6B7	Duplex-Diode Pentode	S.	7D	6.3	0.3	3.5	9.5	.007	Pentode R.F. Amp.	250	− 3.0	125	2.3	9.0	650000	1125	730	—	—	6B7
6C6	Sharp Cut-off Pentode	S.	6F	6.3	0.3	5	6.5	.007	R.F. Amplifier	250	− 3.0	100	0.5	2.0	1500000	1225	1500	—	—	6C6
6C7	Duplex Diode Triode	S.	7G	6.3	0.3	—	—	—	Class-A Amp.	250	− 9.0	—	—	4.5	—	20	1250	—	—	6C7
6D6	Variable-μ Pentode	S.	6F	6.3	0.3	4.7	6.5	.007	R.F. Amplifier	250	− 3.0	100	2.0	8.2	800000	1600	1280	—	—	6D6
6D7	Sharp Cut-off Pentode	S.	7H	6.3	0.3	5.2	6.8	.01	Class-A Amp.	250	− 3.0	100	0.5	2.0	—	1600	1280	—	—	6D7
6E5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	250	0	—	—	0.25	Target Current 4 ma.				—	6E5
6E6	Twin Triode Amplifier	M.	7B	6.3	0.6	—	—	—	Class-A Amp.	250	−27.5	Per plate—18.0			3500	1700	6.0	14000	1.6	6E6
6E7	Variable-μ Pentode	S.	7H	6.3	0.3	—	—	—	R.F. Amplifier	Characteristics same as 6U7G—Table II										6E7
6F7	Triode Pentode	S.	7E	6.3	0.3	—	—	—	Triode Unit Amp.	100	− 3.0	—	—	3.5	16000	500	8	—	—	6F7
									Pentode Unit Amplifier	250	− 3.0	100	1.5	6.5	850000	1100	900	—	—	
6U5/6G5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	250 100	Cut-off Grid Bias = −22 v. Cut-off Grid Bias = −8 v.			0.24 0.19	Target Current 4 ma. Target Current 1 ma.				—	6U5/6G5
6H5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	Same characteristics as Type 6G5—Circular Pattern										6H5
6T5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	250	Cut-off Grid Bias = −12 v.			0.24	Target Current 4 ma.				—	6T5
36	Tetrode R.F. Amplifier	S.	5E	6.3	0.3	3.8	9	.007	R.F. Amplifier	250	− 3.0	90	1.7	3.2	550000	1080	595	—	—	36
37	Triode Detector Amplifier	S.	5A	6.3	0.3	3.5	2.9	2	Class-A Amp.	250	−18.0	—	—	7.5	8400	1100	9.2	—	—	37
38	Pentode Power Amplifier	S.	5F	6.3	0.3	3.5	7.5	0.3	Class-A Amp.	250	−25.0	250	3.8	22.0	100000	1200	120	10000	2.5	38
39/44	Remote Cut-off Pentode	S.	5F	6.3	0.3	3.8	10	.007	R.F. Amplifier	250	− 3.0	90	1.4	5.8	1000000	1050	1050	—	—	39/44
41	Pentode Power Amplifier	S.	6B	6.3	0.4	—	—	—	Class-A Amp.	250	−18.0	250	5.5	32.0	68000	2200	150	7600	3.4	41
42	Pentode Power Amplifier	M.	6B	6.3	0.7	—	—	—	Class-A Amp.	250	−16.5	250	6.5	34.0	100000	2200	220	7000	3.0	42
52	Dual Grid Triode	M.	5C	6.3	0.3	—	—	—	Class-A Amp.[4]	110	0	—	—	43.0	1750	3000	5.2	2000	1.5	52
									Class-B, 2 tubes[5]	180	0	—	—	3.0[12]				10000	5.0	
56AS	Triode Amplifier	S.	5A	6.3	0.4	—	—	—	Class-A Amp.	Characteristics same as 56										56AS
57AS	Sharp Cut-off Pentode	S.	6F	6.3	0.4	—	—	—	R.F. Amplifier	Characteristics same as 57										57AS
58AS	Remote Cut-off Pentode	S.	6F	6.3	0.4	—	—	—	R.F. Amplifier	Characteristics same as 58										58AS
75	Duplex-Diode Triode	S.	6G	6.3	0.3	1.7	3.8	1.7	Triode Amplifier	250	− 1.35	—	—	0.4	91000	1100	100	—	—	75
76	Triode Detector Amplifier	S.	5A	6.3	0.3	3.5	2.5	2.8	Class-A Amp.	250	−13.5	—	—	5.0	9500	1450	13.8	—	—	76
77	Sharp Cut-off Pentode	S.	6F	6.3	0.3	4.7	11	.007	R.F. Amplifier	250	− 3.0	100	0.5	2.3	1500000	1250	1500	—	—	77
78	Variable-μ Pentode	S.	6F	6.3	0.3	4.5	11	.007	R.F. Amplifier	250	− 3.0	100	1.7	7.0	800000	1450	1160	—	—	78
79	Twin Triode Amplifier	S.	6H	6.3	0.6	—	—	—	Class-B Amp.	250	0	—	—	10.6[12]	Power output is for one tube			14000	8.0	79
85	Duplex-Diode Triode	S.	6G	6.3	0.3	1.5	4.3	1.5	Class-A Amp.	250	−20.0	—	—	8.0	7500	1100	8.3	20000	0.35	85
85AS	Duplex-Diode Triode	S.	6G	6.3	0.3	—	—	—	Class-A Amp.	250	− 9.0	—	—	5.5	—	1250	20	—	—	85AS
89	Power Amplifier Pentode	S.	6F	6.3	0.4	—	—	—	Triode Amp.[2]	250	−31.0	—	—	32.0	2600	1800	4.7	5500	0.9	89
									Pentode Amp.[8]	250	−25.0	250	5.5	32.0	70000	1800	125	6750	3.4	

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1221	Pentode R.F. Amplifier	S.	6F	6.3	0.3	—	—	—	Class-A Amp.	Special non-microphonic. Characteristics same as 6C6										1221
1603[3]	Sharp Cut-off Pentode	M.	6F	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6C6										1603
5871	Beam Power Amplifier	O.	7AC	6.3	0.45	9.5	7.5	0.7	Class-A Amp.	315	−13	225	2.2	34	77000	3750	—	8500	5.5	5871
7700[3]	Sharp Cut-off Pentode	S.	6F	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6C6										7700

* Cathode bias resistor—ohms.
∮ Discontinued.
[1] Current to input plate (P_1).
[2] Grids Nos. 2 and 3 connected to plate.
[3] Low noise, nonmicrophonic tubes.
[4] G_2 tied to plate.
[5] G_1 tied to G_2.
[6] Osc. grid leak ohms.
[7] Screen dropping resistor ohms.
[8] Grid No. 2, screen; grid No. 3, suppressor.
[9] Values for single tube.
[10] Values for two tubes in push-pull.
[11] Plate-to-plate value.
[12] No signal value.

TABLE V—2.5-VOLT RECEIVING TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
2S/4S	Duodiode	M.	5D	2.5	1.35	—	—	—	Detector	At 50 d.c. Volts per plate, cathode ma. = 80										2S/4S
2A3	Triode Power Amplifier	M.	4D	2.5	2.5	7.5	5.5	16.5	Class-A Amp.	Characteristics same as Type 6A3, Table IV										2A3
2A5	Pentode Power Amplifier	M.	6B	2.5	1.75	—	—	—	Class-A Amp.	Characteristics same as Type 42, Table IV										2A5
2A6	Duplex-Diode Triode	S.	6G	2.5	0.8	1.7	3.8	1.7	Class-A Amp.	Characteristics same as Type 75, Table IV										2A6
2A7	Pentagrid Converter	S.	7C	2.5	0.8	—	—	—	Converter	Characteristics same as Type 6A7, Table IV										2A7
2B6	Direct-Coupled Amplifier	M.	7J	2.5	2.25	—	—	—	Amplifier	250	−24.0	—	—	40.0	5150	3500	18.0	5000	4.0	2B6
2B7	Duplex-Diode Pentode	S.	7D	2.5	0.8	3.5	9.5	.007	Pentode Amp.	Characteristics same as Type 6B7—Table IV										2B7
2E5	Electron-Ray Tube	S.	6R	2.5	0.8	—	—	—	Indicator Tube	Characteristics same as Type 6E5—Table IV										2E5
2G5	Electron-Ray Tube	S.	6R	2.5	0.8	—	—	—	Indicator Tube	Characteristics same as 6U5/6G5—Table IV										2G5
24-A	Tetrode R.F. Amplifier	M.	5E	2.5	1.75	5.3	10.5	.007	Screen-Grid R.F. Amplifier	250	− 3.0	90	1.7	4.0	600000	1050	630	—	—	24-A
									Bias Detector	250	− 5.0	20/45	Plate current adjusted to 0.1 ma. with no signal							
27	Triode Detector-Amplifier	M.	5A	2.5	1.75	3.1	2.3	3.3	Class-A Amp.	250	−21.0	—	—	5.2	9250	975	9.0	—	—	27
									Bias Detector	250	−30.0	—	Plate current adjusted to 0.2 ma. with no signal							
35/51	Remote Cut-off Pentode	M.	5E	2.5	1.75	5.3	10.5	.007	Screen-Grid R.F. Amplifier	250	− 3.0	90	2.5	6.5	400000	1050	420	—	—	35/51
45	Triode Power Amplifier	M.	4D	2.5	1.5	4	3	7	Class-A Amp.	275	−56.0	—	—	36.0	1700	2050	3.5	4600	2.00	45
46	Dual-Grid Power Amp.	M.	5C	2.5	1.75	—	—	—	Class-A Amp.[2]	250	−33.0	—	—	22.0	2380	2350	5.6	6400	1.25	46
									Class-B Amp.[3]	400	0	—	—	12	Power output for 2 tubes			5800	20.0	
47	Pentode Power Amplifier	M.	5B	2.5	1.75	8.6	13	1.2	Class-A Amp.	250	−16.5	250	6.0	31.0	60000	2500	150	7000	2.7	47
53	Twin Triode Amplifier	M.	7B	2.5	2.0	—	—	—	Class-B Amp.	Characteristics same as Type 6A6, Table IV										53
55	Duplex-Diode Triode	S.	6G	2.5	1.0	1.5	4.3	1.5	Class-A Amp.	Characteristics same as Type 85, Table IV										55
56	Triode Amplifier, Detector	S.	5A	2.5	1.0	3.2	2.4	3.2	Class-A Amp.	Characteristics same as Type 76, Table IV										56
57	Sharp Cut-off Pentode	S.	6F	2.5	1.0	—	—	—	R.F. Amplifier	250	− 3.0	100	0.5	2.0	1500000	1225	1500	—	—	57
58	Remote Cut-off Pentode	S.	6F	2.5	1.0	4.7	6.3	.007	Screen-Grid R.F. Amplifier	250	− 3.0	100	2.0	8.2	800000	1600	1280	—	—	58
59	Pentode Power Amplifier	M.	7A	2.5	2.0	—	—	—	Class-A Triode[4]	250	−28.0	—	—	26.0	2300	2600	6.0	5000	1.25	59
									Class-A Pentode[5]	250	−18.0	250	9.0	35.0	40000	2500	100	6000	3.0	
5608A	Twin Triode	M.	7B	2.5	2.0	—	—	—	Class-A Amp.[6]	300	− 6	—	—	6.0	13000	2450	32	—	—	5608A
RK15	Triode Power Amplifier	M.	4D[1]	2.5	1.75	—	—	—	Characteristics same as Type 46 with Class-B connections											RK15
RK16	Triode Power Amplifier	M.	5A	2.5	2.0	—	—	—	Characteristics same as Type 59 with Class-A triode connections											RK16
RK17	Pentode Power Amplifier	M.	5F	2.5	2.0	—	—	—	Characteristics same as Type 2A5											RK17

[1] Grid connection to cap; no connection to No. 3 pin.
[2] Grid No. 2 tied to plate.
[3] Grids Nos. 1 and 2 tied together.
[4] Grids Nos. 2 and 3 connected to plate.
[5] Grid No. 2, screen; grid No. 3, suppressor.
[6] Values each section.

TABLE VI—2.0-VOLT BATTERY RECEIVING TUBES

Type	Name	Base	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1A4P	Variable-μ Pentode	S.	4M	2.0	0.06	5	11	.007	R.F. Amplifier	180	− 3.	67.5	0.8	2.3	1000000	750	750	—	—	1A4P
1A4T	Variable-μ Tetrode	S.	4K	2.0	0.06	5	11	.007	R.F. Amplifier	180	− 3.0	67.5	0.7	2:3	960000	750	720	—	—	1A4T
1A6	Pentagrid Converter	S.	6L	2.0	0.06	—	—	—	Converter	180	− 3.	67.5	2.4	1.3	500000	Anode grid (No. 2) 180 max. volts				1A6
1B4/951	Pentode R.F. Amplifier	S.	4M	2.0	0.06	5	11	.007	R.F. Amplifier	180 90	− 3.0 − 3.0	67.5 67.5	0.6 0.7	1.7 1.6	1500000 1000000	650 600	1000 550	—	—	1B4/951
1B5/25S	Duplex-Diode Triode	S.	6M	2.0	0.06	1.6	1.9	3.6	Triode Class-A	135	− 3.0	—	—	0.8	35000	575	20	—	—	1B5/25S
1C6	Pentagrid Converter	S.	6L	2.0	0.12	10	10	—	Converter	180	− 3.0	67.5	2.0	1.5	750000	Anode grid (No. 2) 135 max. volts				1C6
1F4	Pentode Power Amplifier	M.	5K	2.0	0.12	—	—	—	Class-A Amp.	135	− 4.5	135	2.6	8.0	200000	1700	340	16000	0.34	1F4
1F6	Duplex-Diode Pentode	S.	6W	2.0	0.06	4	9	.007	R.F. Amplifier	180	− 1.5	67.5	0.6	2.0	1000000	650	650	—	—	1F6
									A.F. Amplifier	135	− 1.0	135	Plate, 0.25 megohm; screen, 1.0 megohm					Amp. = 48		
15 #	Sharp Cut-off Pentode	S.	5F	2.0	0.22	2.3	7.8	0.01	R.F. Amplifier	135	− 1.5	67.5	0.3	1.85	800000	750	600	—	—	15
19	Twin-Triode Amplifier	S.	6C	2.0	0.26	—	—	—	Class-B Amp.	135	0	—	—	—	Load plate-to-plate			10000	2.1	19
30	Triode Detector Amplifier	S.	4D	2.0	0.06	—	—	—	Class-A Amp.	180	−13.5	—	—	3.1	10300	900	9.3	—	—	30
31	Triode Power Amplifier	S.	4D	2.0	0.13	3.5	2.7	5.7	Class-A Amp.	180	−30.0	—	—	12.3	3600	1050	3.8	5700	0.375	31
32	Sharp Cut-off Pentode	M.	4K	2.0	0.06	5.3	10.5	.015	R.F. Amplifier	180	− 3.0	67.5	0.4	1.7	1200000	650	780	—	—	32
33	Pentode Power Amplifier	M.	5K	2.0	0.26	8	12	1	Class-A Amp.	180	−18.0	180	5.0	22.0	55000	1700	90	6000	1.4	33
34	Variable-μ Pentode	M.	4M	2.0	0.06	6	11	.015	R.F. Amplifier	180	− 3.0	67.5	1.0	2.8	1000000	620	620	—	—	34
49	Dual-Grid Power Amp.	M.	5C	2.0	0.12	—	—	—	Class-A Amp.[1]	135	−20.0	—	—	6.0	4175	1125	4.7	11000	0.17	49
									Class-B Amp.[2]	180	0	—	—	Power output for 2 tubes				12000	3.5	
840	Pentode	S.	5J	2.0	0.13	—	—	—	Class-A Amp.	180	− 3.0	67.5	0.7	1.0	1000000	400	400	—	—	840
950	Pentode Power Amplifier	M.	5K	2.0	0.12	—	—	—	Class-A Amp.	135	−16.5	135	2.0	7.0	100000	1000	125	13500	0.575	950
RK24	Triode	M.	4D	2.0	0.12	—	—	—	Class-A Amp.	180	−13.5	—	—	8.0	5000	1600	8.0	12000	0.25	RK24
1229	Tetrode	M.	4K	2.0	0.06	—	—	—	Special Type 32 for low grid-current applications											1229
1230	Triode	M.	4D	2.0	0.06	3.0	2.1	6.0	Special Type 30 for low grid-current applications											1230

Discontinued. [1] Grid No. 2 tied to plate. [2] Grids Nos. 1 and 2 tied together.

TABLE VII—2.0-VOLT BATTERY TUBES WITH OCTAL BASES

Type	Name	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1C7G	Heptode	7Z	2.0	0.06	10	14	0.26	Converter	Characteristics same as Type 1C6—Table VI										1C7G
1D5GP	Variable-μ Pentode	5Y	2.0	0.06	5	11	.007	R.F. Amplifier	Characteristics same as Type 1A4P—Table VI										1D5GP
1D5GT #	Variable-μ Tetrode	5R	2.0	0.06	—	—	—	R.F. Amplifier	180	− 3.0	67.5	0.7	2.2	600000	650				1D5GT
1D7G	Pentagrid Converter	7Z	2.0	0.06	10.5	9.0	0.25	Converter	Characteristics same as Type 1A6—Table VI										1D7G
1E5GP	Pentode Amplifier	5Y	2.0	0.06	5	11	.007	R.F. Amplifier	Characteristics same as Type 1B4—Table VI										1E5GP
1E7G	Double Pentode Power Amp.	8C	2.0	0.24	—	—	—	Class-A Amplifier	135	− 7.5	135	2.0[1]	6.5[1]	220000	1600	350	24000	0.65	1E7G
1F5G	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	Characteristics same as Type 1F4—Table VI										1F5G
1F7G[2]	Duplex-Diode Pentode	7AD	2.0	0.06	3.8	9.5	0.01	Detector-Amplifier	Characteristics same as Type 1F6—Table VI										1F7G
1G5G	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	135	−13.5	135	2.5	8.7	160000	1550	250	9000	0.55	1G5G
1H4G	Triode Amplifier	5S	2.0	0.06	—	—	—	Detector-Amplifier	Characteristics same as Type 30—Table VI										1H4G
1H6G	Duplex-Diode Triode	7AA	2.0	0.06	1.6	1.9	3.6	Detector-Amplifier	Characteristics same as Type 1B5—Table VI										1H6G
1J5G #	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	135	−16.5	135	2.0	7.0	—	950	100	13500	0.45	1J5G
1J6GT	Twin Triode	7AB	2.0	0.24	—	—	—	Class-B Amplifier	Characteristics same as Type 19—Table VI										1J6GT
4A6G	Twin Triode	8L	2.0	0.12	—	—	—	Class-A, 1 section	90	− 1.5	—	—	1.1	26600	750	20			4A6G
			4.0	0.06				Class-B, 2 sections	90	− 1.5	—	—	10.8[3]				8000	1.0	

Discontinued. [1] Total current for both sections; no signal. [2] Type GV has 7AF base. [3] Max. signal.

TABLE VIII—1.5-VOLT FILAMENT BATTERY TUBES

See also Table X for Special 1.4-volt Tubes

Type	Name	Base	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output M-watts	Type
1A5GT	Pentode Power Amplifier	O.	6X	1.4	0.05	—	—	—	Class-A_1 Amp.	90	−4.5	90	0.8	4.0	300000	850	240	25000	115	1A5GT
1A7GT	Pentagrid Converter	O.	7Z	1.4	0.05	Osc. Grid leak 200000Ω			Converter	90	0	45	0.7	0.6	600000	250	Anode-grid volts 90		—	1A7GT
1AB5	Pentode R.F. Amplifier	L.	5BF	1.2	0.05	2.8	4.2	0.25	R.F. Amplifier	90 150	0 −1.5	90 150	0.8 2.0	3.5 6.8	275000 125000	1100 1350	—	—	—	1AB5
1B7GT ‡	Heptode	O.	7Z	1.4	0.1	—	—	—	Converter	90	0	45	1.3	1.5	350000	Grid No. 1 resistor 200,000 ohms				1B7GT
1B8GT	Diode Triode Pentode	O.	8AW	1.4	0.1	—	—	—	Triode Amplifier Pentode Amp.	90 90	0 −6.0	— 90	— 1.4	0.15 6.3	240000 —	275 1150	— —	— 14000	— 210	1B8GT
1C5GT	Pentode Power Amplifier	O.	6X	1.4	0.1	—	—	—	Class-A_1 Amp.	90	−7.5	90	1.6	7.5	115000	1550	165	8000	240	1C5GT
1D8GT	Diode Triode Pentode	O.	8AJ	1.4	0.1	—	—	—	Triode Amp. Pentode Amp.	90 90	0 −9.0	— 90	— 1.0	1.1 5.0	43500 200000	575 925	25 —	— —	— —	1D8GT
1E4G	Triode Amplifier	O.	5S	1.4	0.05	2.4	6	2.40	Class-A Amp.	90 90	0 −3.0	—	—	4.5 1.5	11000 17000	1325 825	14.5 14	—	—	1E4G
1G4GT	Triode Amplifier	O.	5S	1.4	0.05	2.2	3.4	2.80	Class-A Amp.	90	−6.0	—	—	2.3	10700	825	8.8	—	—	1G4GT
1G6GT	Twin Triode	O.	7AB	1.4	0.1	—	—	—	Class-A Amp. Class-B Amp.	90 90	0 0	— —	— —	1.0 1/7	45000 34 volts input per grid	675	30	— 12000	— 675	1G6GT
1H5GT	Diode High-μ Triode	O.	5Z	1.4	0.05	1.1	6	1.00	Class-A Amp.	90	0	—	—	0.14	240000	275	65	—	—	1H5GT
1LA4	Pentode Power Amplifier	L.	5AD	1.4	0.05	—	—	—	Class-A Amp.	90	Characteristics same as 1A5GT									1LA4
1LA6	Pentagrid Converter	L.	7AK	1.4	0.05	Osc. Grid leak 200000Ω			Converter	90	0	45	0.6	0.55	750000	250	Anode Grid Volts 90			1LA6
1LB4	Pentode Power Amplifier	L.	5AD	1.4	0.05	—	—	—	Class-A Amp.	90	−9	90	1.0	5.0	200000	925	—	12000	200	1LB4
1LB6	Heptode Converter	L.	8AX	1.4	0.05	—	—	—	Converter	90	0	67.5	2.2	0.4	Grid No. 4—67.5 v., No. 5—0 v.					1LB6
1LC5	Remote Cut-off Pentode	L.	7AO	1.4	0.05	3.2	7	.007	R.F. Amplifier	90	0	45	0.2	1.15	1500000	775	—	—	—	1LC5
1LC6	Pentagrid Converter	L.	7AK	1.4	0.05	Osc. Grid leak 200000Ω			Converter	90	0	35[1]	0.7	0.75	650000	275	Anode Grid Volts 45			1LC6
1LD5	Diode Pentode	L.	6AX	1.4	0.05	3.2	6	0.18	Class-A Amp.	90	0	45	0.1	0.6	950000	600	—	—	—	1LD5
1LE3	Triode Amplifier	L.	4AA	1.4	0.05	1.7	3	1.70	Class-A Amp.	90 90	0 −3	—	—	4.5 1.3	11200 19000	1300 760	14.5	—	—	1LE3
1LF3	Triode	L.	4AA	1.4	0.05	1.7	3	1.7	Class-A Amp.	90	−3	—	—	1.4	—	760	14.5	—	—	1LF3
1LG5	Pentode R.F. Amp.	L.	7AO	1.4	0.05	—	—	—	Class-A Amp.	90	0	45	0.4	1.7	1000000	800	—	—	—	1LG5
1LH4	Diode High-μ Triode	L.	5AG	1.4	0.05	1.1	6	1.00	Class-A Amp.	90	0	—	—	0.15	240000	275	65	—	—	1LH4
1LN5	Remote Cut-off Pentode	L.	7AO	1.4	0.05	3.4	8	.007	Class-A Amp.	90	0	90	0.3	1.2	1500000	750	—	—	—	1LN5
1N5GT	Remote Cut-off Pentode	O.	5Y	1.4	0.05	3	10	.007	Class-A Amp.	90	0	90	0.3	1.2	1500000	750	1160	—	—	1N5GT
1N6G ‡	Diode-Power-Pentode	O.	7AM	1.4	0.05	—	—	—	Class-A Amp.	90	−4.5	90	0.6	3.1	300000	800	—	25000	100	1N6G
1P5GT	Pentode	O.	5Y	1.4	0.05	3	10	.007	R.F. Amplifier	90	0	90	0.7	2.3	800000	800	640	—	—	1P5GT
1Q5GT	Tetrode Power Amplifier	O.	6AF	1.4	0.1	—	—	—	Class-A Amp.	85 90	−5.0 −4.5	85 90	1.2 1.6	7.2 9.5	70000 75000	1950 2100	—	9000 8000	250 270	1Q5GT
1R4/1294	U.h.f. Diode	L.	4AH	1.4	0.15	—	—	—	Rectifier	Max. r.m.s. voltage per plate—30					Max. d.c. output current—340 μa.					1R4/1294
1SA6GT	Medium Cut-off Pentode	O.	6CA	1.4	0.05	5.2	8.6	0.01	R.F. Amplifier	90	0	67.5	0.68	2.45	800000	970	—	—	—	1SA6GT
1SB6GT	Diode Pentode	O.	6CB	1.4	0.05	3.2	3	0.25	Class-A Amp. R.C. Amplifier	90 90	0 0	67.5 90	0.38 Screen resistor 5 meg., grid 10 meg.	1.45	700000	665	—	— 1 meg.	— 110[2]	1SB6GT
1T5GT	Beam Power Amplifier	O.	6AF	1.4	0.05	4.8	8	0.50	Class-A Amp.	90	−6.0	90	1.4	6.5	—	1150	—	14000	170	1T5GT
3B7/1291	U.h.f. Twin Triode	L.	7BE	2.8[3]	0.11	1.4	2.6	2.6	Class-A Amp.	90	0	—	—	5.2	11350	1850	21	—	—	3B7/1291
1293	U.h.f. Triode	L.	4AA	1.4	0.11	1.7	3.0	1.7	Class-A Amp.	90	0	—	—	4.7	10750	1300	14	—	—	1293
3D6/1299	U.h.f. Tetrode	L.	6BB	2.8[3]	0.11	7.5	6.5	0.30	Class-A Amp.	135	−6	90	0.7	5.7	—	2200	—	13000	500	3D6/1299
3E6	R.F. Pentode	L.	7CJ	1.4 2.8	0.10 0.05	5.5	7.5	0.007	Class-A Amp.	90	0	90	1.3	3.8	300000	2100	—	—	—	3E6
RK42	Triode Amplifier	S.	4D	1.5	0.6	—	—	—	Class-A Amp.	Characteristics same as Type 30—Table VI										RK42
RK43	Twin Triode Amplifier	S.	6C	1.5	0.12	—	—	—	Class-A Amp.	135	−3	—	—	4.5	14500	900	13	—	—	RK43

‡ Discontinued. [1] Through series resistor. Screen voltage must be at least 10 volts lower than oscillator anode. [2] Voltage gain. [3] Center-tap filament permits 1.4-volt operation.

TABLE IX—HIGH-VOLTAGE HEATER TUBES

Type	Name	Base	Socket Connec-tions	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
2C52	High-μ Twin Triode	O.	8BD	12.6	0.3	2.3	0.75	2.7	Class-A Amp.	250	−2	—	—	1.3	—	1900	100	—	—	2C52
12A5[8]	Pentode Power Amplifier	M.	7F	12.6 6.3	0.3 0.6	9.0	9.0	0.3	Class-A_1 Amp.[6]	100 180	−15 −25	100 180	3/6.5 8/14	17/19 45/48	50000 35000	1700 2400	— —	4500 3300	0.8 3.4	12A5
12A6	Beam Power Amplifier	O.	7AC	12.6	0.15	—	—	—	Class-A Amp.	250	−12.5	250	3.5	30	70000	3000	—	7500	3.4	12A6
12A7	Rectifier-Amplifier	M.	7K	12.6	0.3	—	—	—	Class-A Amp.	135	−13.5	135	2.5	9.0	102000	975	100	13500	0.55	12A7
12A8GT	Heptode	O.	8A	12.6	0.15	9.5	12	0.26	Converter	Characteristics same as 6A8—Table I										12A8GT
12AH7GT	Twin Triode	O.	8BE	12.6	0.15	Each Triode Sect.			Class-A Amp.	180	− 6.5	—	—	7.6	8400	1900	16	—	—	12AH7GT
12B6M	Diode Triode	O.	6Y	12.6	0.15	—	—	—	Class-A Amp.	250	− 2.0	—	—	0.9	91000	1100	100	—	—	12B6M
12B7ML	Pentode Amplifier	O.	8V	12.6	0.15	—	—	—	Class-A Amp.	250	− 3.0	100	2.6	9.2	800000	2000	—	—	—	12B7ML
12B8GT[8]	Triode-Pentode	O.	8T	12.6	0.3	Triode Section Pentode Section			Class-A Amp. Class-A Amp.	100 100	− 1 − 3	— 100	— 2	0.6 8	73000 170000	1500 2100	110 360	— —	— —	12B8GT
12C8	Duplex-Diode Pentode	O.	8E	12.6	0.15	6	9	.005	Class-A Amp.	Characteristics same as 6B8—Table I										12C8
12E5GT	Triode Amplifier	O.	6Q	12.6	0.15	3.4	5.5	2.60	Class-A Amp.	250	−13.5	—	—	50	—	1450	13.8	—	—	12E5GT
12F5GT	Triode Amplifier	O.	5M	12.6	0.15	1.9	3.4	2.40	Class-A Amp.	Characteristics same as 6SF5—Table I										12F5GT
12G7G	Duplex-Diode Triode	O.	7V	12.6	0.15	—	—	—	Class-A Amp.	250	− 3.0	—	—	—	58000	1200	70	—	—	12G7G
12H6	Twin Diode	O.	7Q	12.6	0.15	—	—	—	Rectifier	Characteristics same as 6H6—Table I										12H6
12J5GT	Triode Amplifier	O.	6Q	12.6	0.15	3.4	3.6	3.40	Class-A Amp.	Characteristics same as 6J5—Table I										12J5GT
12J7GT	Sharp Cut-off Pentode	O.	7R	12.6	0.15	4.2	5.0	3.8	Class-A Amp.	Characteristics same as 6J7—Table I										12J7GT
12K7GT	Remote Cut-off Pentode	O.	7R	12.6	0.15	4.6	12	005	R.F. Amplifier	Characteristics same as 6K7—Table I										12K7GT
12K8	Triode Hexode Converter	O.	8K	12.6	0.15	—	—	—	Converter	Characteristics same as 6K8—Table I										12K8
12L8GT	Twin Pentode	O.	8BU	12.6	0.15	5	6	0.70	Class-A_1 Amp.	180	− 9.0	180	2.8	13.0	160000	2150	—	10000	1.0	12L8GT
12Q7GT	Duplex-Diode Triode	O.	7V	12.6	0.15	2.2	5	1.60	Class-A Amp.	Characteristics same as 6Q7—Table I										12Q7GT
12S8GT	Triple-Diode Triode	O.	8CB	12.6	0.15	2.0	3.8	1.2	Class-A Amp.	250	− 2.0	—	—	0.9	91000	1100	100	—	—	12S8GT
12SA7	Heptode	O.	8R	12.6	0.15	9.5	12	0.13	Converter	Characteristics same as 6SA7—Table I										12SA7
12SC7	Twin Triode	O.	8S	12.6	0.15	2.2	3.0	2.0	Class-A Amp.	Characteristics same as 6SC7—Table I										12SC7
12SF5	High-μ Triode	O.	6AB	12.6	0.15	4	3.6	2.40	Class-A Amp.	Characteristics same as 6SF5—Table I										12SF5
12SF7	Diode Variable-μ Pentode	O.	7AZ	12.6	0.15	5.5	6.0	.004	Class-A Amp.	Characteristics same as 6SF7—Table I										12SF7
12SG7	Medium Cut-off Pentode	O.	8BK	12.6	0.15	8.5	7.0	.003	Class-A Amp.	Characteristics same as 6SG7—Table I										12SG7
12SH7	Sharp Cut-off Pentode	O.	8BK	12.6	0.15	8.5	7.0	.003	H-F Amplifier	Characteristics same as 6SH7—Table I										12SH7
12SJ7	Sharp Cut-off Pentode	O.	8N	12.6	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SJ7—Table I										12SJ7
12SK7	Remote Cut-off Pentode	O.	8N	12.6	0.15	6.0	7.0	.003	R.F. Amplifier	Characteristics same as 6SK7—Table I										12SK7
12SL7GT	Twin Triode	O.	8BD	12.6	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SL7GT—Table II										12SL7GT
12SN7GT	Twin Triode	O.	8BD	12.6	0.3	—	—	—	Class-A Amp.	Characteristics same as 6SN7GT—Table II										12SN7GT
12SQ7	Duplex-Diode Triode	O.	8Q	12.6	0.15	3.2	3.0	1.60	Class-A Amp.	Characteristics same as 6SQ7—Table I										12SQ7
12SR7	Duplex-Diode Triode	O.	8Q	12.6	0.15	3.6	2.8	2.40	Class-A Amp.	Characteristics same as 6R7—Table I										12SR7
12SW7	Duplex-Diode Triode	O.	8Q	12.6	0.15	3.0	2.8	2.4	Class-A_1 Amp.	250	− 9	—	—	9.5	8500	1900	16	—	—	12SW7
12SX7	Twin Triode	O.	8BD	12.6	0.3	3.0	0.8	3.6	Class-A_1 Amp.[5]	250	− 8	—	—	9	7700	2600	20	—	—	12SX7
12SY7	Heptode Converter	O.	8R	12.6	0.15	Osc.-Grid leak 20000 ohms			Converter	250	− 2	100	8.5	3.5	1000000	450	—	—	—	12SY7
12V6GT	Beam Pentode	O.	7AC	12.6	0.225	9	7.5	0.7	Class-A Amp.	315	−13	225	2.2/6	34/35	80000	3750	—	8500	5.5	12V6GT
14A4	Triode Amplifier	L.	5AC	14	0.16	3.4	3.0	4.00	Class-A Amp.	Characteristics same as 7A4—Table III										14A4
14A5	Beam Power Amplifier	L.	6AA	14	0.16	—	—	—	Class-A_1 Amp.	250	−12.5	250	3.5/5.5	30/32	70000	3000	—	7500	2.8	14A5
14A7/ 12B7	Remote Cut-off Pentode	L.	8V	14	0.16	6.0	7.0	.005	Class-A Amp.	250	− 3.0	100	2.6	9.2	800000	2000	—	—	—	14A7/ 12B7
14AF7	Twin Triode	L.	8AC	14	0.16	2.2	1.6	2.30	Class-A Amp.	250	−10	—	—	9	7600	2100	16	—	—	14AF7
14B6	Duplex-Diode Triode	L.	8W	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7B6—Table III										14B6
14B8	Pentagrid Converter	L.	8X	14	0.16	Ic2 —4 Ma.			Converter	Characteristics same as 7B8—Table III										14B8
14C5	Beam Power Amplifier	L.	6AA	14	0.24	—	—	—	Class-A Amp.	Characteristics same as 6V6—Table I										14C5
14C7	R.F. Pentode	L.	8V	14	0.16	6.0	6.5	.007	Class-A Amp.	250	− 3.0	100	0.7	2.2	1000000	1575	—	—	—	14C7

TABLE IX—HIGH-VOLTAGE HEATER TUBES—*Continued*

Type	Name	Base	Socket Connec-tions	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
14E6	Duplex-Diode Triode	L.	8W	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7E6—Table III										14E6
14E7	Duplex-Diode Pentode	L.	8AE	14	0.16	4.6	5.3	.005	Class-A Amp.	Characteristics same as 7E7—Table III										14E7
14F7	Twin Triode	L.	8AC	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7F7—Table III										14F7
14F8	Twin Triode	L.	8BW	12.6	0.15	2.8	1.4	1.2	Class-A$_1$ Amp.	Characteristics same as 7F8										14F8
14H7	Semi-Variable-μ Pentode	L.	8V	14	0.16	8.0	7.0	.007	Class-A Amp.	250	−2.5	150	3.5	9.5	800000	3800	—	—	—	14H7
14J7	Triode-Hexode Converter	L.	8BL	14	0.16	Ipt=5 Ma.			Converter	Characteristics same as 7J7—Table III										14J7
14N7	Twin Triode	L.	8AC	14	0.32	—	—	—	Class-A Amp.	Characteristics same as 7N7—Table III										14N7
14Q7	Heptode Pentagrid Converter	L.	8AL	14	0.16	—	—	—	Converter	Characteristics same as 7Q7—Table III										14Q7
14R7	Duplex-Diode Pentode	L.	8AE	14	0.16	5.6	5.3	.004	Class-A Amp.	Characteristics same as 7R7—Table III										14R7
14S7	Triode Heptode	L.	8BL	14	0.16	Ipt=5 Ma.			Converter	250	−2.0	100	3	1.8	1250000	525	—	—	—	14S7
14V7	H.f. Pentode	L.	8V	14	0.24	—	—	—	Class-A Amp.	300	−2.0	150	3.9	9.6	300000	5800	—	—	—	14V7
14W7	Pentode	L.	8BJ	14	0.24	Rk=160 ohms			Class-A Amp.	300	−2.2	150	3.9	10	300000	5800	—	—	—	14W7
14X7	Twin Diode Triode	L.	8BZ	12.6	0.15	—	—	—	Class-A Amp.	250	−1	—	—	1.9		1500	100	—	—	14X7
18	Pentode	M.	6B	14	0.30	—	—	—	Class-A Amp.	Characteristics same as 6F6G										18
19BG6G	Beam Power Amp.	O.	5BT	18.9	0.3	11	6.5	0.65	Deflection Amp.	400	Peak surge E_P=4000 V. Peak surge E_{G1}=−100 V. I_{G2}=6 ma. I_P=70 ma.									19BG6G
20J8GM	Triode Heptode Converter	O.	8H	20	0.15	—	—	—	Converter	250	−3.0	100	3.4	1.5	Triode Plate (No. 6) 100 v. 1.5 ma.					20J8GM
21A7	Triode Hexode Converter	L.	8AR	21	0.16	—	—	—	Converter	250	−3.0	100	2.8	1.3	—	275	—	—	—	21A7
										150	−3.0	Triode		3.5	—	1900	32	—	—	
25A6	Pentode Power Amplifier	O.	7S	25	0.3	8.5	12.5	0.20	Class-A Amp.	135	−20.0	135	8	37	35000	2450	85	4000	2.0	25A6
25A7GT[8]	Rectifier Power Pentode	O.	8F	25	0.3	—	—	—	Class-A Amp.	100	−15.0	100	4	20.5	50000	1800	90	4500	0.77	25A7GT
25AC5GT	Triode Power Amplifier	O.	6Q	25	0.3	—	—	—	Class-A Amp.	110	+15.0	—	—	45	—	3800	58	2000	2.0	25AC5GT
										165	Used in dynamic-coupled circuit with 6AF5G driver							3500	3.3	
25AV5GT	Beam Pentode	O.	6CK	25	0.3	—	—	—	Horz. Def. Amp.	250[9]	−50[9]	175[9]	—	100[9]	Peak pos. plate pulse=4500 volts.					25AV5GT
25B5[8]	Direct-Coupled Triodes	S.	6D	25	0.3	—	—	—	Class-A Amp.	110	0	110	7	45	11400	2200	25	2000	2.0	25B5
25B6G[8]	Pentode Power Amplifier	O.	7S	25	0.3	—	—	—	Class-A Amp.	95	−15.0	95	4	45	—	4000	—	2000	1.75	25B6G
25B8GT[8]	Triode Pentode	O.	8T	25	0.15	—	—	—	Class-A Amp.		Characteristics same as 12B8GT									25B8GT
25BQ6GT	Beam Pentode	O.	6AM	25	0.3	—	—	—	Deflection Amp.	250	47*	150	2.1	45	—	5500	—	—	—	25BQ6GT
25C6G[8]	Beam Power Amplifier	O.	7AC	25	0.3	—	—	—	Class-A$_1$ Amp.	135	−13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	25C6G
25CD6G	Beam Pentode	O.	5BT	25	0.6	26	10	1.0	Deflection Amp.	500	Peak Pos. Pulse=6000 volts. E_{G2}=170. I_P=92 ma. I_{G2}=15.5 ma.									25CD6G
25D8GT	Diode Triode Pentode	O.	8AF	25	0.15	—	—	—	Triode Amp.	100	−1.0	—	—	0.5	91000	1100	100	—	—	25D8GT
									Pentode Amp.	100	−3.0	100	2.7	8.5	200000	1900	—	—	—	
25L6	Beam Power Amplifier	O.	7AC	25	0.3	16	13.5	0.30	Class-A$_1$ Amp.	110	−8.0	110	3.5/10.5	45/48	10000	8000	80	2000	2.2	25L6
25N6G[8]	Direct-Coupled Triodes	O.	7W	25	0.3	—	—	—	Class-A Amp.	110	0	110	7	45	11400	2200	25	2000	2.0	25N6G
26A7GT	Twin Beam-Power Audio Amplifier	O.	8BU	26.5	0.6	Each Unit Push-Pull			Class-A Amp.	26.5	−4.5	26.5	2/5.5	20/20.5	2500	5500	—	1500	0.2	26A7GT
									Class-AB Amp.[3]	26.5	−7.0	26.5	2/8.5	19/30	—	—	—	2500[4]	0.5	
32L7GT	Diode-Beam Tetrode	O.	8Z	32.5	0.3	—	—	—	Class-A Amp.	110	−7.5	110	3	40	15000	6000	—	2500	1.5	32L7GT
35A5	Beam Power Amplifier	L.	6AA	35	0.15	—	—	—	Class-A$_1$ Amp.	110	−7.5	110	3/7	40/41	14000	5800	—	2500	1.5	35A5
35L6GT	Beam Power Amplifier	O.	7AC	35	0.15	13	9.5	0.80	Class-A$_1$ Amp.	110	−7.5	110	3/7	40/41	13800	5800	—	2500	1.5	35L6GT
43	Pentode Power Amplifier	M.	6B	25	0.3	8.5	12.5	0.20	Class-A Amp.	95	−15.0	95	4.0	20.0	45000	2000	90	4500	0.90	43
48[8]	Tetrode Power Amplifier	M.	6A	30	0.4	—	—	—	Class-A Amp.	96	−19.0	96	9.0	52.0	—	3800	—	1500	2.0	48
50A5	Beam Power Amplifier	L.	6AA	50	0.15	—	—	—	Class-A$_1$ Amp.	110	−7.5	110	4/11	49/50	10000	8200	—	2000	2.2	50A5
50C6GT	Beam Power Amplifier	O.	7AC	50	0.15	—	—	—	Class-A$_1$ Amp.	135	−13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	50C6GT
50L6GT	Beam Power Amplifier	O.	7AC	50	0.15	—	—	—	Class-A Amp.	110	−7.5	110	4/11	49/50	—	8200	82	2000	2.2	50L6GT
70A7GT	Diode-Beam Tetrode	O.	8AB[1]	70	0.15	—	—	—	Class-A Amp.	110	−7.5	110	3.0	40	—	5800	80	2500	1.5	70A7GT
70L7GT	Diode-Beam Tetrode	O.	8AA	70	0.15	—	—	—	Class-A$_1$ Amp.	110	−7.5	110	3/6	40/43	15000	7500	—	2000	1.8	70L7GT
117L7GT/117M7GT	Rectifier-Amplifier	O.	8AO	117	0.09	—	—	—	Class-A Amp.	105	−5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117L7GT/117M7GT
117N7GT	Rectifier-Amplifier	O.	8AV	117	0.09	—	—	—	Class-A Amp.	100	−6.0	100	5.0	51	16000	7000	—	3000	1.2	117N7GT

TABLE IX—HIGH-VOLTAGE HEATER TUBES—*Continued*

Type	Name	Base	Socket Connections	Heater Volts	Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
117P7GT	Rectifier-Amplifier	O.	8AV	117	0.09	—	—	—	Class-A Amp.	105	−5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117P7GT
1280	Pentode	L.	8V	12.6	0.15	6.0	6.5	0.007	Class-A1 Amp.	Same as 14C7 (Special Non-microphonic)										1280
1284	U.h.f. Pentode	L.	8V	12.6	0.15	5.0	6.0	0.01	Class-A Amp.	250	−3.0	100	2.5	9.0	800000	2000	—	—	—	1284
1629	Electron-Ray Tube	O.	6RA	12.6	0.15	—	—	—	Indicator Tube	Characteristics same as 6E5—Table IV										1629
1631	Beam Power Amplifier	O.	7AC	12.6	0.45	—	—	—	Class-A Amp.	Characteristics same as 6L6—Table I										1631
1632	Beam Power Amplifier	O.	7AC	12.6	0.6	—	—	—	Class-A Amp.	Characteristics same as 25L6										1632
1633	Twin Triode	O.	8BD	25	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SN7GT—Table II										1633
1634	Twin Triode	O.	8S	12.6	0.15	—	—	—	Class-A Amp.	Characteristics same as 6SC7—Table I										1634
1644	Twin Pentode	O.	Fig. 7	12.6	0.15	—	—	—	Class-A Amp.	180	−9.0	180	2.8/4.6	13	160000	2150	—	10000	1.0	1644
XXD/ 14AF7	Twin Triode	L.	8AC	12.6	0.15	—	—	—	Class-A Amp.	250	−10	—	—	9.0	—	2100	16	—	—	XXD/ 14AF7
28D7	Double Beam Power Amplifier	L.	8BS	28.0	0.4	—	—	—	Class-A Amp.	28	390* 180*	28[2] 28[3]	0.7[2] 1.2[3]	9.0[2] 18.5[3]	— —	— —	— —	4000[4] 6000[4]	0.08[2] 0.175[3]	28D7
5824	Pentode	O.	7S	25	0.3	—	—	—	Class-A1 Amp.	135	−22	135	2.5/14.5	61/69	15000	5000	—	1700	4.3	5824
6082	Low-Mu Twin Triode	O.	8BD	26.5	0.6	6.4	2.2	8.4	D.C. Amp.[5]	135	250*	—	—	125	280	7000	2	—	—	6082

* Cathode resistor—ohms.
[1] 6.3-volt pilot lamp must be connected between Pins 6 and 7.
[2] Per section—resistance-coupled.
[3] P.p. operation—values for both sections.
[4] Plate to plate.
[5] Values are for each unit.
[6] Values are for single tube.
[7] Grids 2 and 3 connected to plate.
[8] Discontinued.
[9] Max. value.

TABLE X—SPECIAL RECEIVING TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
00-A [7]	Triode Detector	M.	4D	5.0	0.25	3.2	2.0	8.50	Grid-Leak Det.	45	—	—	—	1.5	30000	666	20	—	—	00-A
01-A [7]	Triode Detector Amplifier	M.	4D	5.0	0.25	—	—	—	Class-A Amp.	135	−9.0	—	—	3.0	10000	800	8.0	—	—	01-A
3A8GT	Diode Triode Pentode	O.	8AS	1.4 2.8	0.1 0.05	2.6 3.0	4.2 10.0	2.0 0.012	Class-A Triode Class-A Pentode	90 90	0 0	— 90	— 0.3	0.15 1.2	240000 600000	275 750	65 —	— —	— —	3A8GT
3B5GT	Beam Power Amplifier	O.	7AP	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	67.5	−7.0	67.5	0.6 0.5	8.0 6.7	100000	1650 1500	—	5000	0.2 0.18	3B5GT
3C5GT	Power Output Pentode	O.	7AQ	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	−9.0	90	1.4	6.0	—	1550 1450	—	8000 10000	0.24 0.26	3C5GT
3C6	Twin Triode	L.	7BW	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	0	—	—	4.5	11200	1300	14.5	—	—	3C6
3LE4	Power Amplifier Pentode	L.	6BA	2.8	0.05	—	—	—	Class-A Amp.	90	−9.0	90	1.8	9.0	110000	1600	—	6000	0.30	3LE4
3LF4	Beam Pentode	L.	6BB	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	−4.5	90	1.3 1.0	9.5 8.0	75000 80000	2200 2000	—	8000 7000	0.27 0.23	3LF4
3Q5GT	Beam Power Amplifier	O.	7AQ	1.4 2.8	0.1 0.05	Parallel Filaments Series Filaments			Class-A Amp.	90	−4.5	90	1.3 1.0	9.5 7.5	—	2100 1800	—	8000	0.27 0.25	3Q5GT
4A6G	Twin Triode Amplifier	O.	8L	4.0 2.0	0.06 0.12	Triodes Parallel Both Sections			Class-A Amp. Class-B Amp.	90 90	−1.5 0	— —	— —	2.2 4.6	13300 —	1500 —	20 —	— 8000	— 1.0	4A6G
6F4	Acorn Triode	A.	7BR	6.3	0.225	2.0	0.6	1.90	Class-A Amp.	80	150*	—	—	13.0	2900	5800	17	—	—	6F4
6L4	U.H.F. Triode	A.	7BR	6.3	0.225	1.8	0.5	1.6	Class-A1 Amp.	80	150*	—	—	9.5	4400	6400	28	—	—	6L4
10	Triode Power Amplifier	M.	4D	7.5	1.25	4.0	3.0	7.00	Class-A Amp.	425	−39.0	—	—	18.0	5000	1600	8.0	10200	1.6	10
11/12 [7]	Triode Detector Amplifier	M.	4F/4D	1.1	0.25	—	—	—	Class-A Amp.	135	−10.5	—	—	3.0	15000	440	6.6	—	—	11/12
20 [7]	Triode Power Amplifier	S.	4D	3.3	0.132	2.0	2.3	4.10	Class-A Amp.	135	−22.5	—	—	6.5	6300	525	3.3	6500	0.11	20
22 [7]	Tetrode R.F. Amplifier	M.	4K	3.3	0.132	3.5	10	0.02	Class-A Amp.	135	−1.5	67.5	1.3	3.7	325000	500	160	—	—	22
26	Triode Amplifier	M.	4D	1.5	1.05	2.8	2.5	8.10	Class-A Amp.	180	−14.5	—	—	6.2	7300	1150	8.3	—	—	26

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
40 [7]	Triode Voltage Amplifier	M.	4D	5.0	0.25	2.8	2.2	2.00	Class-A Amp.	180	− 3.0	—	—	0.2	150000	200	30	—	—	40
50	Triode Power Amplifier	M.	4D	7.5	1.25	4.2	3.4	7.10	Class-A Amp.	450	−84.0	—	—	55.0	1800	2100	3.8	4350	4.6	50
71-A	Triode Power Amplifier	M.	4D	5.0	0.25	3.2	2.9	7.50	Class-A Amp.	180	−43.0	—	—	20.0	1750	1700	3.0	4800	0.79	71-A
99 [6] [7]	Triode Detector Amplifier	S.	4D	3.3	0.063	2.5	2.5	3.30	Class-A Amp.	90	− 4.5	—	—	2.5	15500	425	6.6	—	—	99
112A [7]	Triode Detector Amplifier	M.	4D	5.0	0.25	—	—	—	Class-A Amp.	180	−13.5	—	—	7.7	4700	1800	8.5	—	—	112A
182B/ 482B	Triode Amplifier	M.	4D	5.0	1.25	—	—	—	Class-A Amp.	250	−35.0	—	—	18.0	—	1500	5.0	—	—	182B/ 482B
183/483 [7]	Power Triode	M.	4D	5.0	1.25	—	—	—	Class-A Amp.	250	−60.0	—	—	25.0	18000	1800	3.2	4500	2.0	183/483
485 [7]	Triode	S.	5A	3.0	1.3	—	—	—	Class-A Amp.	180	− 9.0	—	—	6.0	9300	1350	12.5	—	—	485
864	Triode Amplifier	S.	4D	1.1	0.25	—	—	—	Class-A Amp.	90	− 4.5	—	—	2.9	13500	610	8.2	—	—	864
954	Pentode Detector, Amplifier	A.	5BB	6.3	0.15	3.4	3.0	0.007	Class-A Amp. Bias Detector	250 250	− 3.0 − 6.0	100 100	0.7 —	2.0 Plate current to be adjusted to 0.1 ma. with no signal	1.5 meg.	1400	2000	—	— —	954
955	Triode Detector, Amplifier, Oscillator	A.	5BC	6.3	0.15	1.0	0.6	1.40	Class-A Amp.	250 90	− 7.0 − 2.5	— —	— —	6.3 2.5	11400 14700	2200 1700	25 25	— —	— —	955
956	Variable-μ Pentode R.F. Amplifier	A.	5BB	6.3	0.15	3.4	3.0	0.007	Class-A Amp. Mixer	250 250	− 3.0 −10.0	100 100	2.7 —	6.7 —	700000 Oscillator peak volts—7 min.	1800	1440	—	—	956
957	Triode Detector, Amplifier, Oscillator	A.	5BD	1.25	0.05	0.3	0.7	1.20	Class-A Amp.	135	− 5.0	—	—	2.0	20800	650	13.5	—	—	957
958 958-A	Triode A.F. Amplifier, Oscillator	A.	5BD	1.25	0.1	0.6	0.8	2.60	Class-A Amp.	135	− 7.5	—	—	3.0	10000	1200	12	—	—	958 958-A
959	Pentode Detector, Amplifier	A.	5BE	1.25	0.05	1.8	2.5	0.015	Class-A Amp.	145	− 3.0	67.5	0.4	1.7	800000	600	480	—	—	959
7E5/1201	U.h.f. Triode	L.	8BN	6.3	0.15	3.6	2.8	1.50	Class-A Amp.	180	− 3	—	—	5.5	12000	—	36	—	—	7E5/1201
7C4/1203	U.h.f. Diode	L.	4AH	6.3	0.15	—	—	—	Rectifier		Max. r.m.s. voltage—150				Max. d.c. output current—8 ma.					7C4/1203
7AB7/ 1204	Sharp Cut-off Pentode	L.	8BO	6.3	0.15	3.5	4.0	0.06	Class-A Amp.	250	− 2	100	0.6	1.75	800000	1200	—	—	—	7AB7/ 1204
1276	Triode Power Amplifier	M.	4D	4.5	1.14	—	—	—	Class-A Amp.				Characteristics similar to 6A3							1276
1609	Pentode Amplifier	S.	5B	1.1	0.25	—	—	—	Class-A Amp.	135	− 1.5	67.5	0.65	2.5	400000	725	300	—	—	1609
5731	Acorn Triode	A.	5BC	6.3	0.15	1.0	0.4	1.3	Class-A Amp.	250	− 7	—	—	6.3	11400	2200	25	—	—	5731
5768	U.h.f. "Rocket" Triode	N.	Fig. 36	6.3	0.4	1.2	0.01	1.3	1000-3000-Mc. Amplifier	250	− 1	—	—	9.3	—	4500	85	—	—	5768
6173	U.h.f. "Pencil" Diode	N.	Fig. 67	6.3	0.135	Plate to K—1.1			Rectifier		Peak inverse—375 Volts. Peak Ip—50 Ma. Max. d.c. output—5.5 Ma.									6173
9004	U.h.f. Diode	A.	4BJ	6.3	0.15	—	—	—	Detector			Max. a.c. voltage—117. Max. d.c. output current—5 ma.								9004
9005	U.h.f. Diode	A.	5BG	3.6	0.165	—	—	—	Detector			Max. a.c. voltage—117. Max. d.c. output current—1 ma.								9005
EF-50	Sharp Cut-off Pentode	L.	9C	6.3	0.3	8	5	0.007	I.F.-R.F. Amp.	250	150*	250	3.1	10	600000	6300	—	—	—	EF-50
GL-2C44 GL-464A	U.h.f. Triode	O.	Fig. 17	6.3	0.75	—	—	—	Class-A Amp. and Modulator	250	100*	—	—	25.0	—	7000	—	—	—	GL-2C44 GL-464A
GL-446A GL-446B	U.h.f. Triode	O.	Fig. 19	6.3	0.75	—	—	—	Oscillator, Amp. or Converter	250	200*	—	—	15.0	—	4500	45	—	—	GL-446A GL-446B
559 GL-559	U.h.f. Diode	O.	Fig. 18	6.3	0.75	—	—	—	Detector or trans. line switch	5.0	—	—	—	24.0	—	—	—	—	—	559 GL-559
NU-2C35	Special Hi-Mu Triode	O.	Fig. 38	6.3	0.3	5.2	2.3	0.62	Shunt Voltage Regulator	8000	−200	—	—	5.0	525000	950	500	—	—	NU-2C35
VT52	Triode	M.	4D	7.0	1.18	5.0	3.0	7.7	Class-A_1 Amp.	220	−43.5	—	—	29.0	1650	2300	3.8	3800	1.0	VT52
X6030	Diode	L.	Fig. 4	3.0	0.6	—	—	—	Noise Diode	90	—	—	—	4.0	—	—	—	—	—	X6030
XXB	Twin-Triode Frequency Converter	L.	Fig. 9	2.8/ 1.4 3.2 [3]/ 1.6	0.05/ 0.10 — —	—	—	—	Converter [2]	90 [1]	0 − 3	— —	— —	4.5 [4] 4.5 [5] 1.4 [4] 1.4 [5]	11200 [4] 11200 [5] 1900 [4] 1900 [5]	1300 [4] 1300 [5] 760 [4] 760 [5]	14.5 [1] 14.5 [1]	— —	— —	XXB
XXFM	Twin-Diode Triode	L.	8BZ	6.3	0.3	—	—	—	Class-A Amp.	250 100	− 1 0	— —	— —	1.9 1.2	6700 85000	1500 1000	100 85	— —	— —	XXFM

* Cathode resistor—ohms.
[1] Both sections.
[2] Section No. 2 recommended for h.f.a.
[3] Dry battery operation.
[4] Section No. 1.
[5] Section No. 2.
[6] Same as X99. Type V99 is same, but socket connections are 4E.
[7] Discontinued.

TABLE XI—MINIATURE RECEIVING TUBES—Other miniature types in Tables XIII and XV

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor [4]	Load Resistance Ohms	Power Output Watts	Prototype
1A3	H. F. Diode	B.	5AP	1.4	0.15	—	—	—	Detector F.M. Discrim.	Max. a.c. voltage per plate—117. Max. output current—0.5 ma.										—
1AB6	Pentagrid Converter	B.	7DH	1.4	0.025	7.6	8.4	0.36	Converter	—	—	—	—	—	—	—	—	—	—	—
1AC6	Pentagrid Converter	B.	7DH	1.4	0.05	7.5	8.4	0.36	Converter	—	—	—	—	—	—	—	—	—	—	—
1AE4	Sharp Cut-off Pentode	B.	6AR	1.25	0.1	3.6	4.4	0.008	Class-A_1 Amp.	90	0	90	1.2	3.5	500000	1550	—	—	—	—
1AF4	Pentode	B.	6AR	1.4	0.025	3.8	7.6	.008	Class-A_1 Amp.	90	0	90	0.5	1.65	1800000	950	—	—	—	—
1AF5	Diode Pentode	B.	6AU	1.4	0.025	—	—	—	Class-A_1 Amp.	90	0	90	0.4	1.1	2000000	600	—	—	—	—
1AH5	Diode A.F. Pentode	B.	7DJ	1.4	0.025	2.1	2.9	0.3	Class-A Amp.	85	—	—	0.015	0.05	1000000	—	62	—	—	—
1C3	Triode	B.	5CF	1.4	0.05	0.9	4.2	1.8	Class-A_1 Amp.	90	-3	—	—	1.4	19000	760	14.5	—	—	1LE3
1E3	U.h.f. Triode	B.	9BG	1.25	0.22	1.25	0.75	1.5	Class-A Amp.	150	-3.5	—	—	20	—	3500	14	—	—	—
1L4	Sharp Cut-off Pentode	B.	6AR	1.4	0.05	3.6	7.5	.008	Class-A Amp.	90	0	90	2.0	4.5	350000	1025	—	—	—	1N5GT
1L6	Pentagrid Converter	B.	7DC	1.4	0.05	7.5	12	0.3	Converter	90	0	45	0.6	0.5	650000	300	—	—	—	1LA6
1R5	Pentagrid Converter	B.	7AT	1.4	0.05	—	—	—	Converter	90	0	67.5	3.0	1.7	500000	300	Grid No. 1 100000 ohms			1A7GT
1S4	Pentagrid Power Amp.	B.	7AV	1.4	0.1	—	—	—	Class-A Amp.	90	-7.0	67.5	1.4	7.4	100000	1575	—	8000	0.270	1Q5GT
1S5	Diode Pentode	B.	6AU	1.4	0.05	—	—	—	Class-A Amp.	67.5	0	67.5	0.4	1.6	600000	625	—	—	—	—
									R-Coupled Amp.	90	0	90	Screen resistor 3 meg., grid 10 meg.					1 meg.	0.050	
1T4	Variable-μ Pentode	B.	6AR	1.4	0.05	3.6	7.5	0.01	Class-A Amp.	90	0	67.5	1.4	3.5	500000	900	—	—	—	1P5GT
1U4	Sharp Cut-off Pentode	B.	6AR	1.4	0.05	3.6	7.5	0.01	Class-A Amp.	90	0	90	0.5	1.6	1500000	900	—	—	—	1N5GT
1U5	Diode Pentode	B.	6BW	1.4	0.05	—	—	—	Class-A Amp.	67.5	0	67.5	0.4	1.6	600000	625	—	—	—	—
1U6	Pentagrid Converter	B.	7DC	1.4	0.025	8	12	0.4	Converter	90	0	45	0.55	0.55	600000	275	—	—	—	—
1W4	Power Amplifier Pentode	B.	5BZ	1.4	0.05	3.6	7	0.1	Class-A_1 Amp.	90	-9	90	1	5	300000	925	—	12000	0.2	1LB4
2C51	Twin Triode	B.	8CJ	6.3	0.3	2.2	1.0	1.3	Class-A_1 Amp.	150	-2	—	—	8.2 [1]	—	5500	35	—	—	7F8
2E30	Beam Power Pentode	B.	7CQ	6.0	0.7	10	4.5	0.5	Class-A_1 Single	250	450*	250	7.4 [2]	44 [2]	63000	3700	40 [5]	4500	4.5	—
									Class-A_1 Amp.[3]	250	225*	250	14.8 [2]	88 [2]	—	—	80 [5]	9000 [6]	9	
									Class-AB_1 Amp.[3]	250	-25	250	13.5 [2]	80 [2]	—	—	48 [5]	8000 [6]	12.5	
									Class-AB_2 Amp.[3]	250	-30	250	20 [2]	120 [2]	—	—	40 [5]	3800 [6]	17	
3A4	Power Amplifier Pentode	B.	7BB	1.4 2.8	0.2 0.1	4.8	4.2	0.34	Class-A_1 Amp.	135 150	-7.5 -8.4	90 90	2.6 2.2	14.9 [2] 14.1 [2]	90000 100000	1900	—	8000	0.6 0.7	—
3A5	H.F. Twin Triode	B.	7BC	1.4 2.8	0.22 0.11	0.9	1.0	3.20	Class-A Amp.	90	-2.5	—	—	3.7	8300	1800	15	—	—	—
3C4	Power Pentode	B.	6BX	1.4	0.05	4.9	4.4	0.3	Class-A Amp.	85	-5.2	85	5.0	1.1	125000	1350	—	13000	0.2	—
3E5	Power Amplifier Pentode	B.	6BX	1.4 2.8	0.05 .025	—	—	—	Class-A_1 Amp.	90	-8	90	1.5	5.5	120000	1100	—	8000	.175	—
3Q4	Power Amplifier Pentode	B.	7BA	1.4	0.1	Parallel Filaments			Class-A Amp.	90	-4.5	90	2.1	9.5	100000	2150	—	10000	0.27	3Q5GT
				2.8	0.05	Series Filaments							1.7	7.7	120000	2000			0.24	
3S4	Power Amplifier Pentode	B.	7BA	1.4	0.1	Parallel Filaments			Class-A Amp.	90	-7.0	67.5	1.4	7.4	100000	1575	—	8000	0.27	3Q5GT
				2.8	0.05	Series Filaments							1.1	6.1		1425			0.235	
3V4	Power Amplifier Pentode	B.	6BX	1.4	0.1	Parallel Filaments			Class-A Amp.	90	-4.5	90	2.1	9.5	100000	2150	—	10000	0.27	3Q5GT
				2.8	0.05	Series Filaments			Class-A Amp.	90	-4.5	90	1.7	7.7	120000	2000	—	10000	0.24	
6AB4	U.h.f. Triode	B.	5CE	6.3	0.15	2.2	0.5	1.5	Class-A Amp.	250	200*	—	—	10	10900	5500	60	—	—	Single unit 12AT7
6AB8	Triode-Pentode	B.	9AT	6.3	0.3	4.6	4.7	0.2	Triode	100	-2	—	—	4	—	1350	18	—	—	—
						—	—	—	Pentode	200	-7.7	200	3.3	17.5	150000	3400	—	11000	1.4	—
6AD8	Dual Diode Pentode	B.	9T	6.3	0.3	4.0	4.6	0.002	Class-A Amp.	250	-2	85	2.3	6.7	1000000	1100	—	—	—	—
6AE8	Triode Hexode	B.	9Q	6.3	0.3	—	—	—	Freq. Converter	—	—	—	—	—	—	—	—	—	—	6K8
6AF4	U.h.f. Triode	B.	7DK	6.3	0.225	2.2	0.45	1.9	Class-A_1 Amp.	80	150*	—	—	16	2270	6600	15	—	—	—
									Osc. at 950 Mc.	100	10000Ω	—	0.4 [10]	22	—	—	—	—	—	
6AG5	Sharp Cut-off Pentode	B.	7BD	6.3	0.3	—	—	—	Class-A Amp.	250 100	200* 100*	150 100	2.0 1.6	7.0 5.5	800000 300000	5000 4750	— —	— —	—	6SH7GT
6AH6	Sharp Cut-off Pentode	B.	7CC	6.3	0.45	10	2	0.03	Pentode Amp.	300	160*	150	2.5	10	500000	9000	—	—	—	6AC7
									Triode Amp.[7]	150	160*	—	—	12.5	3600	11000	40	—	—	

Type	Name	Base	Socket Connections	Fil. or Heater		Capacitance $\mu\mu$fd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor *	Load Resistance Ohms	Power Output Watts	Prototype
				Volts	Amp.	In	Out	Plate-Grid												
6AJ4	U.h.f Triode	B.	9BX	6.3	0.225	4.4	0.18	2.4	Class-A$_1$ Amp.	125	68*	—	—	16	4200	10000	42	—	—	—
6AJ5	Sharp Cut-off Pentode	B.	7PM	6.3	0.175	—	—	—	R.F. Amplifier	28	200*	28	1.2	3.0	90000	2750	250	—	—	—
									Class-AB Amp.[3]	180	− 7.5	75	—	—	—	—	—	28000[6]	1.0	
6AJ8	Triode-Heptode	B.	9CA	6.3	0.3	4.8	7.9	0.01	Heptode Mixer	250	− 2	22K Ω	6.7	3.25	1000000	775	—	—	—	—
									Triode Amp.	100	0	—	—	13.5	—	3700	22	—	—	
6AK5	Sharp Cut-off Pentode	B.	7BD	6.3	0.175	4.3	2.1	0.03		180	200*	120	2.4	7.7	690000	5100	3500	—	—	—
									R.F. Amplifier	150	330*	140	2.2	7.0	420000	4300	1800	—	—	
										120	200*	120	2.5	7.5	340000	5000	1700	—	—	
6AK6	Power Amplifier Pentode	B.	7BK	6.3	0.15	3.6	4.2	0.12	Class-A Amp.	180	− 9.0	180	2.5	15.0	200000	2300	—	10000	1.1	—
6AK8	Triple Diode Triode	B.	9E	6.3	0.45	1.9	1.6	2.2	Class-A Amp.	250	− 3	—	—	1	58000	1200	70	—	—	—
6AL5	U.h.f. Twin Diode	B.	6BT	6.3	0.3	—	—	—	Detector	Max. r.m.s. voltage—150. Max. d.c. output current—10 ma.[1]										6H6GT
6AM4	U.h.f. Triode	B.	9BX	6.3	0.225	4.4	0.16	2.4	Class-A Amp.	150	100*	—	—	7.5	10000	9000	90	—	—	—
6AM5	Power Amplifier Pentode	B.	6CH	6.3	0.2	—	—	—	Class-A$_1$ Amp.	250	−13.5	250	2.4	16	130000	2600	—	16000	1.4	—
6AM6	Pentode	B.	7DB	6.3	0.3	7.5	3.25	0.01	Class-A$_1$ Amp.	250	− 2	250	2.5	10	1000000	7500	—	—	—	—
6AN4	U.h.f. Triode	B.	7DK	6.3	0.225	2.2	0.17	1.7	Class-A Amp.	200	100*	—	—	13	—	9000	70	—	—	—
6AN5	Power Amp. Pentode	B.	7BD	6.3	0.5	9.0	4.8	0.05	Class-A$_1$ Amp.	120	− 6	120	12	35	12500	8000	—	—	—	6AG7
6AN6	Twin Diode	B.	7BJ	6.3	0.2	—	—	—	Detector	R.m.s. voltage per plate = 75 volts; d.c. output = 3.5 ma. with 25000 ohms and 8 $\mu\mu$fd. load; peak current per plate = 10 ma.; peak inverse voltage = 210.										—
6AN7	Triode Hexode	B.	9Q	6.3	0.23	3.8	9.2	0.1	Converter	250	− 2	85	3	3	—	750	—	—	—	—
6AQ5	Beam Power Tetrode	B.	7BZ	6.3	0.45	7.6	6.0	0.35	Class-A$_1$ Amp.	180	− 8.5	180	4.0[2]	30[2]	58000	3700	29[5]	5500	2.0	6V6GT
										250	12.5	250	7.0[2]	47[2]	52000	4100	45[5]	5000	4.5	
6AQ6	Duodiode Hi-mu Triode	B.	7BT	6.3	0.15	1.7	1.5	1.80	Class-A Triode	250	− 3.0	—	—	1.0	58000	1200	70	—	—	6T7G
										100	− 1.0	—	—	0.8	61000	1150	70	—	—	
6AR5	Pentode Power Amp.	B.	6CC	6.3	0.4	—	—	—	Class-A$_1$ Amp.	250	−18	250	5.5[2]	33[2]	68000	2300	—	7600	3.4	6K6GT
										250	−16.5	250	5.5[2]	35[2]	65000	2400	—	7000	3.2	
6AS5	Beam Pentode	B.	7CV	6.3	0.8	12	6.2	0.6	Class-A$_1$ Amp.	150	− 8.5	110	2/6.5	35/36	—	5600		4500	2.2	—
6AS6	Sharp Cut-off Pentode	B.	7CM	6.3	0.175	4.0	3.0	0.02	Class-A Amp.	120	− 2	120	3.5	5.2	—	3200	—	—	—	—
6AT6	Duplex Diode Triode	B.	7BT	6.3	0.3	2.3	1.1	2.10	Class-A Amp.	250	− 3	—	—	1.0	58000	1200	70	—	—	6Q7GT
6AU6	Sharp Cut-off Pentode	B.	7BK	6.3	0.3	5.5	5.0	.0035	Class-A Amp.	250	− 1	150	4.3	10.8	2000000	5200	—	—	—	6SH7GT
6AV6	Duodiode Hi-mu Triode	B.	7BT	6.3	0.3	—	—	—	Class-A$_1$ Amp.	250	− 2	—	—	1.2	62500	1600	100	—	—	6SQ7GT
6BA6	Remote Cut-off Pentode	B.	7CC	6.3	0.3	5.5	5.0	.0035	Class-A Amp.	250	68*	100	4.2	11	1500000	4400	—	—	—	6SG7GT
6BA7	Pentagrid Converter	B.	8CT	6.3	0.3	9.5	8.3	—	Converter	250	− 1	100	10	3.8	1000000	950	—	—	—	6SB7Y
6BC5	Pentode	B.	7BD	6.3	0.3	6.6	3.1	.02	Class-A$_1$ Amp.	250	180*	150	1.4	4.7	600000	4900	—	—	—	—
6BC7	Triple Diode	B.	9AX	6.3	0.45	—	—	—	FM/AM Det.	Max. diode current per plate = 12 Ma. Max. htr.-cath volts = 200										—
6BD6	Remote Cut-off Pentode	B.	7CC	6.3	0.3	—	—	—	Class-A Amp.	100	− 1	100	5	13	120000	2350	—	—	—	6SK7GT
										250	− 3	100	3.5	9	700000	2000	—	—	—	
6BD7	Duodiode Hi-mu Triode	B.	9Z	6.3	0.23	2.4	1.3	1.3	Class-A$_1$ Amp.	250	− 3	—	—	1.0	58000	1200	70	—	—	—
6BE6	Pentagrid Converter	B.	7CH	6.3	0.3	Osc. Grid 50000 Ω			Converter	250	− 1.5	100	7.8	3.0	1000000	475	—	—	—	6SA7GT
6BE7	Heptode Limiter-Disc.	B.	9AA	6.3	0.2	—	—	—	FM Limiter-Discriminator	250	− 4.4	20	1.5	0.28	5000000	—	—	—	—	—
6BF5	Beam Power Pentode	B.	7BZ	6.3	1.2	—	—	—	Class-A$_1$ Amp.	110	− 7.5	110	4.0/8.5	49/50	10000	7500	—	2500	1.9	—
6BF6	Duplex-Diode Triode	B.	7BT	6.3	0.3	1.8	1.1	2.0	Class-A$_1$ Amp.	250	− 9	—	—	9.5	8500	1900	16	10000	—	6SR7GT
6BH5	Remote Cut-off Pentode	B.	9AZ	6.3	0.2	4.9	5.5	0.002	Class-A Amp.	250	− 2.5	90K Ω	1.7	6	1100000	2200	—	—	—	—
6BH6	Sharp Cut-off Pentode	B.	7CM	6.3	0.15	5.4	4.4	0.0035	Class-A$_1$ Amp.	250	− 1	150	2.9	7.4	1400000	4600	—	—	—	—
6BJ5	Pentode	B.	6CH	6.3	0.64	—	—	—	Power Amp.	250	− 5	250	5.5	35	40000	10500	420	7000	4.0	—
6BJ6	Remote Cut-off Pentode	B.	7CM	6.3	0.15	4.5	5.0	.0035	Class-A$_1$ Amp.	250	− 1	100	3.3	9.2	1300000	3800	—	—	—	6SS7GT
6BK5	Beam Power Pentode	B.	9BQ	6.3	1.2	13	5.0	0.6	Class-A$_1$ Amp.	250	− 5	250	3.5/10	35/37	100000	8500	—	6500	3.5	—
6BK6	Duodiode Triode	B.	7BT	6.3	0.3	—	—	—	Class-A$_1$ Amp.[11]	250	− 2	—	—	1.2	80000	1250	100	—	—	—
6BK7	U.h.f. Twin Triode	B.	9AJ	6.3	0.45	3.0	1.1	1.9	Class-A$_1$ Amp.	150	56*	—	—	18	4700	8500	40	—	—	—
6BK7A	U.h.f. Twin Triode	B.	9AJ	6.3	0.45	3.0	1.9	1.8	Cascode Amp.	150	56*	—	—	18	4600	9300	43	—	—	6BK7

TABLE XI — MINIATURE RECEIVING TUBES — *Continued*

Type	Name	Base	Socket Connections	Fil. or Heater		Capacitance μμfd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor [4]	Load Resistance Ohms	Power Output Watts	Prototype
				Volts	Amp.	In	Out	Plate-Grid												
6BM5	Power Pentode	B.	7DO	6.3	0.45	8	5.5	0.5	Class-A Amp.	250	−6	250	3	30	60000	7000	—	7000	3.5	—
6BN6	Gated-beam Disc.	B.	7DF	6.3	0.3	4.2	3.3	.004	FM Disc.	80	−1.3	60	5	0.23	—	—	—	68000	—	—
6BN7	Dual Triode	B.	Fig. 41	6.3	0.75	5.5[7]	1.6[7]	3[7]	Class-A_1 Amp.[7]	250	−15	—	—	24	2200	5500	12	—	—	—
						1.4[8]	0.3[8]	0.7[8]	Class-A_1 Amp.[8]	120	−1	—	—	5	14000	2000	28	—	—	
6BQ7	Double Triode	B.	9AJ	6.3	0.4	2.55	1.3	1.15	Class-A_1 Amp.[11]	150	220*	—	—	9.0	5800	6000	35	—	—	—
6BQ7A	Dual Triode	B.	9AJ	6.3	0.4	—	—	1.15	Class-A Amp.[11]	150	220*	—	—	9	6100	6400	39	—	—	6BQ7
6BR7	Sharp Cut-off Pentode	B.	9BC	6.3	0.15	4.25	4.0	0.01	Class-A Amp.	250	−3	100	0.6	2.1	2500000	1250	—	—	—	6J7
6BT6	Duodiode Triode	B.	7BT	6.3	0.03	—	—	—	Class-A_1 Amp.	250	−3	—	—	1	58000	1200	70	—	—	—
6BU6	Duodiode Triode	B.	7BT	6.3	0.3	—	—	—	Class-A_1 Amp.	250	−9	—	—	9.5	8500	1900	16	10000	0.3	—
6BV7	Duodiode Pentode	B.	9BU	6.3	0.8	11.5	9.5	0.5	Class-A Amp.	250	−5	250	6	38	100000	1000	—	8000	4.0	—
6BW6	Beam Pentode	B.	9AM	6.3	0.45	—	—	—	Class-A_1 Amp.	315	−13	225	6	35	77000	3750	—	8500	5.5	—
										250	−12.5	250	7	47	52000	4100	—	5000	4.5	
6BX6	R.F. Pentode	B.	9AQ	6.3	0.3	7.2	3.4	0.007	Class-A Amp.	170	−2	170	2.5	10	400000	7200	—	—	—	—
6BY7	R.F. Pentode	B.	9AQ	6.3	0.3	7.2	3.7	0.007	Class-A Amp.	250	−2	100	2.5	10	500000	6000	—	—	—	—
6BZ7	U.h.f. Twin Triode	B.	9AJ	6.3	0.4	2.85	2.27	1.15	Class-A Amp.[11]	150	220*	—	—	10	5600	6800	38	—	—	6BQ7
6C4	Triode Amplifier	B.	6BG	6.3	0.15	1.8	1.3	1.60	Class-A_1 Amp.	250	−8.5	—	—	10.5	7700	2200	17	—	—	6J5GT
6CB6	Sharp Cut-off Pentode	B.	7CM	6.3	0.3	6.3	1.9	0.02	Class-A_1 Amp.	200	180*	150	2.8	9.5	600000	6200	—	—	—	—
6CF6	Sharp Cut-off Pentode	B.	7CM	6.3	0.3	6.3	1.9	0.02	Class-A_1 Amp.	200	180*	150	2.8	9.5	600000	6200	—	—	—	—
6CG6	Remote Cut-off Pentode	B.	7BK	6.3	0.3	5	5	0.008	Class-A_1 Amp.	250	−8	150	2.3	9.0	720000	2000	—	—	—	—
6CH6	R.F. Pentode	B.	9BA	6.3	0.75	14	5	0.25	Class-A Amp.	250	−4.5	250	6	40	50000	11000	—	—	—	—
6CJ6	Audio Pentode	B.	9AS	6.3	1.05	14.7	6.0	0.8	Class-A Amp.	250	−38.5	250	2.4	32	15000	4600	—	—	—	—
6CK6	R.F. Pentode	B.	9AR	6.3	0.71	11.2	6.6	0.1	Class-A Amp.	250	−5.5	250	5	36	130000	10000	—	—	—	—
6CL6	Power Pentode	B.	Fig. 68	6.3	0.65	11	5.5	0.12	Class-A_1 Amp.	250	−3	150	7/7.2	30/31	15000	11000	—	7500	2.8	6AG7
6CS6	Heptode	B.	7CH	6.3	0.3	5.5	7.5	0.05	Sync. Separator	100	−1	30	1.1	0.75	1000000	950	—	—	—	—
										Grid #2 current = 1.1 Ma.										
6J4	U.h.f. Grounded-Grid R.F. Amplifier	B.	7BQ	6.3	0.4	5.5	0.24	4.0	Grounded-Grid	150	200*	—	—	15.0	4500	12000	55	—	—	—
									Class-A_1 Amp.	100	100*	—	—	10.0	5000	11000	55	—	—	
6J6	Twin Triode	B.	7BF	6.3	0.45	2.2	0.4	1.6	Class-A_1 Amp. Mixer, Oscillator	100	50*	—	—	8.5	7100	5300	38	—	—	—
6M5	Power Amplifier Pentode	B.	9N	6.3	0.71	10	6.2	1	Class-A_1 Amp.	250	170*	250	5.2	36	40000	10000	—	7000	3.9	—
6N4	U.h.f. Triode Amplifier	B.	7CA	6.3	0.2	3.0	1.6	1.10	Class-A Amp.	180	−3.5	—	—	12	—	6000	32	—	—	—
6N8	Duodiode Pentode	B.	9T	6.3	0.3	4	4.6	.002	Class-A_1 Amp.	250	−2	85	—	1	1600000	2200	—	—	—	—
6Q4	Grnd.-Grid Triode	B.	9S	6.3	0.48	5.4	.06	3.4	Class-A_1 Amp.	250	−1.5	—	—	15	—	12000	80	—	—	—
6R4	U.h.f. Triode	B.	9R	6.3	0.2	1.7	0.5	1.5	Class-A_1 Amp.	150	−2	—	—	30	—	5500	16	—	—	—
6R8	Triple Diode Triode	B.	9E	6.3	0.45	1.5	1.1	2.4	Class-A_1 Amp.	250	−9	—	—	9.5	8500	1900	16	10000	0.3	—
6S4	Triode	B.	9AC	6.3	0.6	—	—	—	Class-A_1 Amp.	250	−8	—	—	26	3600	4500	16	—	—	—
6T4	Triode	B.	7DK	6.3	0.225	2.4	0.45	1.8	Class-A Amp.	80	150*	—	—	18	—	7000	13	—	—	—
6T8	Triple-Diode Triode	B.	9E	6.3	0.45	1.5	1.1	2.4	Class-A_1 Amp.	250	−3	—	—	1.0	5800	1200	70	—	—	—
										100	−1	—	—	0.8	5400	1300	70	—	—	
6U8	Triode	B.	9AE	6.3	0.45	2.5	1.0	1.8	Class-A_1 Amp.	150	56*	—	—	18	5000	8500	40	—	—	—
	Pentode					5.0	2.6	0.01	Class-A_1 Amp.	250	68*	110	3.5	10	400000	5200	—	—	—	
6V8	Triple-Diode Triode	B.	9AH	6.3	0.45	—	—	—	Class-A_1 Amp.	100	−1	—	—	0.8	54000	1300	70	—	—	—
										250	−3	—	—	1.0	58000	1200	70	—	—	
									Diode	Max. diode #2 and #3 Ma. = 10 each. Max. diode #1 Ma. = 1.0										
6X8	Medium Mu Triode	B.	9AK	6.3	0.45	2.6	1.0	1.4	Triode Osc.	150	2700Ω	—	—	13	—	—	—	—	—	—
	Sharp Cut-off Pentode					4.5	1.2	0.008	Pentode Mix.	150	−3.5	150	1.1	4.6	—	1600	—	—	—	
9BM5	Power Pentode	B.	7DQ	9.5	0.3	8	5.5	0.5	Class-A Amp.	250	−6	250	3	30	60000	7000	420	7000	3.5	—
9BW6	Beam Pentode	B.	9AM	9.45	0.3	—	—	—	Class-A Amp.	315	−13	225	2.2	34	77000	3750	—	8500	5.5	6BW6

TABLE XI—MINIATURE RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor [4]	Load Resistance Ohms	Power Output Watts	Prototype
12A4	Triode	B.	9AG	6.3	0.6	—	—	—	Class-A$_1$ Amp.	150	−17	—	—	30	1200	5200	6.5	—	—	—
				12.6	0.3															
12AH8	Triode Heptode	B.	9BP	12.6	0.15	—	—	—	Heptode	250	—	100	4.4	2.6	1500000	550	—	—	—	—
				6.3	0.3				Triode	100	−3	—	—	I_K = 12.5 Ma.		3500	17	—	—	
12AL5	Twin Diode	B.	6BT	12.6	0.15	2.5	—	—	Detector	R.m.s. voltage per plate = 117; d.c. output = 9 ma. per plate; peak ma. per plate = 54; peak inverse voltage = 330.										12H6GT
12AQ5	Beam Pentode	B.	7BZ	12.6	0.225	8.3	8.2	0.35	Class-A$_1$ Amp.	250	−12.5	250	4.5/7	45/47	52000	4100	—	5000	4.5	6AQ5
12AT6	Duplex Diode Triode	B.	7BT	12.6	0.15	2.3	1.1	2.10	Class-A Amp.	250	−3.0	—	—	1.0	58000	1200	70	—	—	12Q7GT
12AT7	Double Triode	B.	9A	6.3	0.3	2.5[7]	0.45[7]	1.45[7]	Class-A$_1$ Amp.	250	−2	—	—	10	10000	5500	55	—	—	—
				12.6	0.15	2.5[8]	0.35[8]	1.45[8]	Each Unit	180	−1	—	—	11	9400	6600	62	—	—	
12AU6	Sharp Cut-off Pentode	B.	7CC	12.6	0.15	5.5	5.0	.0035	Class-A$_1$ Amp.	250	−1.0	150	4.3	10.8	1 meg.	5200	—	—	—	12SH7GT
12AU7	Twin-Triode Amplifier	B.	9A	6.3	0.3	1.6[7]	0.5[7]	1.5[7]	Class-A$_1$ Amp.	250	−8.5	—	—	10.5	7700	2200	17	—	—	12SN7GT
				12.6	0.15	1.6[8]	0.35[8]	1.5[8]												
12AV6	Duodiode Hi-mu Triode	B.	7BT	12.6	0.15	—	—	—	Class-A$_1$ Amp.	250	−2	—	—	1.2	62500	1600	100	—	—	—
12AV7	Double Triode	B.	9A	12.6	0.225	3.1	0.5[7]	1.9	Class-A$_1$ Amp.[11]	100	120*	—	—	9.0	6100	6100	37	—	—	—
				6.3	0.45		0.4[8]			150	56*	—	—	18	4800	8500	41	—	—	
12AW6	Sharp Cut-off Pentode	B.	7CM	12.6	0.15	6.5	1.5	0.025	Pentode Amp.	250	200*	150	2.0	7.0	800000	5000	—	—	—	—
									Triode Amp.[9]	250	825*	—	—	5.5	11000	3800	42	—	—	
12AW7	Sharp Cut-off Pentode	B.	7CM	12.6	0.15	6.5	1.5	0.025	Class-A$_1$ Amp.	250	200*	150	2.0	7.0	0.8 meg.	5000	—	—	—	—
12AX7	Double Triode	B.	9A	12.6	0.15	1.6[7]	0.46[7]	1.7[7]	Class-A Amp.	250	−2	—	—	1.2[1]	62500	1600	100	—	—	—
				6.3	0.3	1.6[8]	0.34[8]	1.7[8]	Class-B Amp.	300	0	—	—	14/40	—	—	—	16000	7.5	
12AY7	Dual Triode	B.	9A	12.6	0.15	1.3	0.6	1.3	Class-A Amp.	250	−4	—	—	3	—	1750	40	—	—	—
				6.3	0.3				Lo-Level Amp.	150	2700*	Plate resistor = 20000 Ω. Grid resistor = 0.1 Meg. V.G. = 12.5								
12AZ7	Double Triode	B.	9A	12.6	0.225	3.1[7]	0.5[7]	1.9[7]	Class-A$_1$ Amp.	100	270*	—	—	3.7	15000	4000	60	—	—	—
				6.3	0.45	3.1[8]	0.4[8]	1.9[8]		250	200*	—	—	10.0	10900	5500	60	—	—	
12B4	Triode	B.	9AG	12.6	0.3	6.4	7	4.3	Class-A Amp.	150	−17.5	—	—	35	—	6500	6.5	—	—	—
				6.3	0.6															
12BA6	Remote Cut-off Pentode	B.	7CC	12.6	0.15	5.5	5.0	.0035	Class-A Amp.	250	68*	100	4.2	11.0	1500000	4400	—	—	—	12SG7G
12BA7	Pentagrid Converter	B.	8CT	12.6	0.15	9.5	8.3	—	Converter	250	−1	100	10	3.8	1000000	3.5	—	—	—	—
12BD6	Remote Cut-off Pentode	B.	7CC	12.6	0.15	4.3	5.0	.004	Class-A Amp.	250	−3	100	3.5	9.0	700000	2000	—	—	—	12SK7GT
12BE6	Pentagrid Converter	B.	7CH	12.6	0.15	Osc. Grid 50000 Ω			Converter	250	−1.5	100	7.8	3.0	1000000	475	—	—	—	12SA7GT
12BF6	Duodiode Triode	B.	7BT	12.6	0.15	1.8	1.1	2.00	Class-A Amp.	250	−9	—	—	9.5	8500	1900	16	—	—	12SR7GT
12BH7	Dual Triode	B.	9A	6.3	0.6	3	2.6	2.4	Class-A$_1$ Amp.	250	−9.5	—	—	11.5	—	3250	18	—	—	6SN7GT
				12.6	0.3															
12BK6	Duodiode Triode	B.	7BT	12.6	0.15	—	—	—	Class-A$_1$ Amp.	250	−2	—	—	1.2	63000	1600	100	—	—	—
12BN6	Gated-beam Disc.	B.	7DF	12.6	0.15	4.2	3.3	.004	FM Disc.	Same as 6BN6										—
12BT6	Duodiode Triode	B.	7BT	12.6	0.15	—	—	—	Class-A$_1$ Amp.	Same as 6BT6										—
12BU6	Duodiode Triode	B.	7BT	12.6	0.15	—	—	—	Class-A$_1$ Amp.	Same as 6BU6										—
12BY7	Sharp Cut-off Pentode	B.	9BF	12.6	0.3	10.7	4.0	0.063	Class-A Amp.	250	100*	180	5.0	24	110000	12000	30[12]	—	—	—
				6.3	0.6															
12BZ7	Dual Triode	B.	9A	12.6	0.15	6.5	0.7	0.45	Class-A$_1$ Amp.[11]	250	−2	—	—	2.5	31800	3200	100	—	—	—
				6.3	0.3															
12G4	Triode	B.	6BG	12.6	0.15	2.4	0.9	3.4	Class-A Amp.	250	−8	—	—	9	—	2600	—	—	—	12J5GT
15A6	R.F. Pentode	B.	9AR	15.0	0.3	10	7	0.1	Video Amp.	180	−2.9	180	4.6	36	100000	10000	—	—	—	—
16A5	A.F. Pentode	B.	9BL	16.5	0.3	11	5.9	1.0	Class-A Amp.	170	−10.4	170	8.5	45	24000	7600	—	3000	4.2	—
19AQ5	Beam Pentode	B.	7BZ	18.9	0.15	—	—	—	Class-A Amp.	Same as 6AQ5										—
19C8	Triple-Diode Triode	B.	9E	18.9	0.15	—	—	—	Class-A$_1$ Amp.	100	−1	—	—	0.5	80000	1250	100	—	—	—
									Diode	Max. diode current = 6 Ma. each.										

TABLE XI—MINIATURE RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor [4]	Load Resistance Ohms	Power Output Watts	Prototype
19J6	Twin Triode	B.	7BF	18.9	0.15	2.0	0.4	1.5	Class-A$_1$ Amp.	100	50*	—	—	8.5 [1]	7100	5300	38	—	—	—
19T8	Triple-Diode Triode	B.	9E	18.9	0.15	1.5	1.1	2.4	Class-A$_1$ Amp.	250	−3	—	—	1.0	5800	1200	70	—	—	—
19V8	Triple-Diode Triode	B.	9AH	18.9	0.15	—	—	—	Characteristics same as 6V8											—
19X8	Triode Pentode	B.	9AK	18.9	0.15	4.3	0.7	0.09	Triode Osc.	150	2700 Ω gridleak			13	3.6 Ma. grid current					—
									Pentode Mixer	150	−3.5	150	1.8	6.2	—	2100	—	—	—	
21A6	R.F. Pentode	B.	9AS	21.5	0.3	14.3	6.5	0.4	Horizontal Time Base	180	−23	180	3	45	—	6500	—	—	—	—
										180	0	180	29	430	—	—	—	—	—	
25BK5	Beam Power Amp.	B.	9BQ	25	0.3	13	5.0	0.6	Class-A$_1$ Amp.	250	−5.0	250	3.5/10	35/37	100000	8500	—	6500	3.5	—
26A6	Remote Cut-off Pentode	B.	7BK	26.5	0.07	6.0	5.0	.0035	Class-A$_1$ Amp.	250	125*	100	4	10.5	1000000	4000	—	—	—	—
26BK6	Duodiode Triode	B.	7BT	26.5	0.07	—	—	—	Class-A$_1$ Amp.	Same as 6BK6										—
26C6	Duplex-Diode Triode	B.	7BT	26.5	0.07	1.8	1.4	2	Class-A$_1$ Amp.	250	−9	—	—	9.5	8500	1900	16	—	—	—
26CG6	Semi-Remote Cut-off Pentode	B.	7BK	26.5	0.07	5.0	5.0	0.008	Class-A$_1$ Amp.	250	−8	150	2.3	9.0	720000	2000	—	—	—	—
26D6	Pentagrid Converter	B.	7CH	26.5	0.07	Osc. Grid 20000 Ω			Converter	250	−1.5	100	7.8	3.0	1000000	475	—	—	—	—
35B5	Beam Power Amplifier	B.	7BZ	35	0.15	11	6.5	0.4	Class-A$_1$ Amp.	110	−7.5	110	7 [4]	41 [2]	—	5800	40 [5]	2500	1.5	35L6GT
35C5	Beam Power Amplifier	B.	7CV	35	0.15	12	6.2	0.57	Class-A$_1$ Amp.	110	−7.5	110	3/7	40/41	—	5800		2500	1.5	—
50B5	Beam Power Amplifier	B.	7BZ	50	0.15	13	6.5	0.50	Class-A Amp.	110	−7.5	110	4.0	49.0	14000	7500	—	3000	1.9	50L6GT
50C5	Beam Power Amplifier	B.	7CV	50	0.15	—	—	—	Class-A$_1$ Amp.	110	−7.5	110	4/8.5	49/50	10000	7500	—	2500	1.9	—
5590	Pentode	B.	7BD	6.3	0.15	3.4	2.9	0.01	Class-A$_1$ Amp.	90	820*	90	1.4	3.9	300000	2000	—	—	—	—
5591	R.F. Pentode	B.	7BD	6.3	0.15	3.9	2.85	0.01	Class-A$_1$ Amp.	180	200*	120	2.4	1.7	690000	5100	3500	—	—	—
5608	Sharp Cut-off Pentode	B.	7BD	6.3	1.75	4	2.9	0.02	Class-A Amp.	120	−12	120	2.5	7.5	340000	5000	—	—	—	—
5610	Triode	B.	6CG	6.3	0.15	—	—	—	Class-A Amp.	90	−1.5	—	—	17	3500	4000	14	—	—	—
5654	Sharp Cut-off Pentode	B.	7BD	6.3	0.175	4	2.9	0.02	Class-A$_1$ Amp.	120	200*	120	2.5	7.5	340000	5000	—	—	—	—
5656	Double Tetrode	B.	9F	6.3	0.4	3.6	1.5	0.06	Class-A$_1$ Amp.[11]	150	−2	120	2.7	15	60000	5800	—	—	—	—
5670	Dual Triode	B.	8CJ	6.3	0.35	2.2	1.0	1.3	Class-A$_1$ Amp.	150	240*	—	—	8.2	—	5500	35	—	—	7F8
5686	Power Pentode	B.	Fig. 29	6.3	0.35	6.4	4.0	0.11	Class-A$_1$ Amp.	250	−12.5	250	5	27	—	3100	—	9000	2.7	—
5687	Dual Triode	B.	9H	12.6	0.45	4	0.45	3.1	Class-A Amp.	250	−12.5	—	—	16	4000	4100	16.5	—	—	—
				6.3	0.9					120	−2	—	—	34	2000	10000	20	—	—	
5722	Noise Generating Diode	B.	5CB	2/5.5	1.6	—	1.5	—	Noise Generator	200	—	—	—	35	—	—	—	—	—	—
5725	Semi-Remote Cut-off Pentode	B.	7CM	6.3	.175	—	—	—	Class-A$_1$ Amp.	120	−2	120	3.5	5.2	—	3200	—	—	—	—
5726	Twin Diode	B.	6BT	6.3	0.3	—	3.2	—	Rectifier	Maximum a.c. voltage per plate = 117; Maximum d.c. Ma. per plate = 9.										—
5749	Remote Cut-off Pentode	B.	7BK	6.3	0.30	5.5	5.0	.0035	Class-A$_1$ Amp.	250	68*	100	4.2	11	1 Meg.	4400	—	—	—	—
5750	Pentagrid Converter	B.	7CH	6.3	0.30	Osc. Grid 20000 Ω			Converter	250	−1.5	100	7.5	2.6	1 Meg.	475	0.5 [10]	—	—	—
5751	Dual Triode	B.	9A	12.6	.175	—	—	—	Class-A$_1$ Amp.	250	−3	—	—	1.1	58000	1200	70	—	—	12SL7GT
5755	Double Triode	B.	9J	12.6	0.18	—	—	—	D.C. Amp.	310	150K*	—	—	0.15	140000	500	70	900000	—	—
				6.3	0.6															
5812	Beam Pentode	B.	7CQ	6.3	0.65	9	7.4	0.2	Class-A$_1$ Amp.	250	−23	250	1.8	40	55000	4100	—	—	—	—
5814	Dual Triode	B.	9A	6.3	0.35	1.6	0.5	1.5	Class-A$_1$ Amp.	250	−8.5	—	—	10.5	6250	2200	19.5	—	—	12SN7GT
				12.6	.175															
5842 417A	Triode	B.	9V	6.3	0.3	9.0	0.48	1.8	Class-A$_1$ Amp.	150	62*	—	—	26	1800	24000	43	—	—	—
5844	Twin Triode	B.	7BF	6.3	0.3	2.4	0.5	2.7	Class-A$_1$ Amp.	100	470*	—	—	4.8	7950	3400	27	—	—	6J6
5845	Double Triode	B.	5CA	4.3	0.435	—	—	—	Noise Generator	300	(Plates tied together)							600000	—	—
5847	Sharp Cut-off Pentode	B.	9X	6.3	0.3	7.1	2.9	0.04	Class-A$_1$ Amp.	160	−8.5	160	4.5	—	—	12500	—	—	—	—
5879	Sharp Cut-off Pentode	B.	9AD	6.3	0.15	2.7	2.4	0.11	Class-A$_1$ Amp.	250	−3	100	0.4	1.8	2 Meg.	1000	—	—	—	—
5910	Sharp Cut-off Pentode	B.	6AR	1.4	0.05	3.6	7.5	0.008	Class-A$_1$ Amp.	90	0	90	0.45	—	1.5 Meg.	900	—	—	—	—
5915	Dual Control Sharp Cut-off Heptode	B.	7CH	6.3	0.3	7.2	8.6	0.3	Switch	30	−5.5	75	8.25	6	—	—	—	—	—	—

TABLE XI—MINIATURE RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Filament Volts	Filament Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output M-Watts	Type
5920	Twin Triode	B.	Fig. 70	6.3	0.4	3.1	0.3	2.6	Class-A Amp.[11]	100	−1.8	—	—	8.5	—	5500	25	—	—	—
5963	Dual Triode	B.	9A	12.6 6.3	0.15 0.3	1.9	—	1.5	Class-A_1 Amp.	67.5	0	—	—	7[1]	7850	2800	22	—	—	—
5964	Dual Triode	B.	7BF	6.3	0.45	2.1	—	1.3	Class-A_1 Amp.	100	50*	—	—	9.5[1]	6500	6000	39	—	—	—
5965	Twin Triode	B.	9A	12.6 6.3	0.225 0.45	3.8	0.5	3.0	Class-A Amp.[11]	150	220*	—	—	8.2	7250	6500	47	—	—	—
6005	Beam Power Amplifier	B.	7BZ	6.3	0.45	—	—	—	Class-A_1 Amp. Class-AB_2 [3]	250 250	−12.5 −15	250 250	4.5/7 5/13	45/47 70/79	52000 60000	4100 3750	— —	5000 10000	4.5 10	—
6057	Twin Triode	B.	9A	12.6 6.3	0.15 0.3	1.6	0.46 0.34	1.7	Class-A Amp.[11]	250	−2	—	—	1.2	62500	1600	100	—	—	12AX7
6058	Twin Triode	B.	6BT	6.3	0.3	—	3.2	—	Detector	Max. r.m.s. volts = 150; Max. d.c. Ma. = 9[1]										6AL5
6059	H.F. Pentode	B.	9BC	6.3	0.15	4.25	4.0	0.01	Class-A_1 Amp.	250	−3	100	0.6	2.1	2500000	1250	—	—	—	6J7
6060	Dual Triode	B.	9A	12.6 6.3	0.15 0.3	2.25	0.4	1.6	Class-A Amp.[11]	250	−2	—	—	10	10000	5500	55	—	—	12AT7
6061	Beam Pentode	B.	9AM	6.3	0.45	—	—	—	Class-A Amp.	315	13	225	2.2	34	77000	3750	—	8500	5.5	6V6GT
6062	R.F. Beam Pentode	B.	9K	6.0	0.75	9.5	4.5	0.3	Class-C Amp.	300	−60	250	5	50	—	—	—	—	8	5763
6064	R.F. Pentode	B.	7BD	6.3	0.3	7.8	3.9	0.01	Class-A Amp.	250	−2	250	—	10	1000000	7500	—	—	—	6AM6
6066	Dual Triode	B.	7BT	6.3	0.3	2.3	1.1	2.1	Class-A Amp.[11]	250	−3	—	—	1.0	58000	1200	70	—	—	6AT6
6067	Twin Triode	B.	9A	12.6 6.3	0.15 0.3	1.6	0.5 0.35	1.5	Class-A Amp.[11]	250	−8.5	—	—	10.5	7700	2200	17	—	—	12AU7
6072	Lo-Noise Twin Triode	B.	9A	6.3	0.35	1.4	0.5	1.4	Class-A_1 Amp.	250	−4.0	—	—	3.0	25000	1750	44	—	—	12AY7
6084	Pentode	B.	9BJ	6.3	0.3	5.1	7.1	0.025	Class-A Amp.	250	−2	100	0.55	3	1800000	1850	—	—	—	—
6085	Dual Triode	B.	9A	12.6 6.3	0.3 0.6	2.8 2.7	1.2 1.3	2.6 2.75	Class-A Amp.[11]	250	−5.5	—	—	6	—	2700	30	—	—	—
6086	Pentode	B.	9BK	18	0.1	8.8	3.6	0.015	Class-A Amp.	210	180*	120	1.7	8.3	440000	8200	—	20000	0.66	—
6132	R.F. Pentode	B.	9BA	6.3	0.75	14	5	0.25	Video Amp.	250	−4.5	250	6	40	50000	11000	—	—	—	6CH6
6135	Med.-Mu Triode	B.	6BG	6.3	0.175	1.5	0.7	1.4	Class-A_1 Amp	250	−8.5	—	—	10.5	7700	2200	17	—	—	6C4
6136	Sharp Cut-off Pentode	B.	7BK	6.3	0.3	6.0	5.0	0.0035	Class-A_1 Amp.	250	68*	150	4.3	10.6	1000000	5200	—	—	—	6AU6
6158	Twin Triode	B.	9A	12.6 6.3	0.3 0.6	2.3	0.95	2.1	Class-A Amp.[11]	250	−4.6	—	—	6	14000	2300	32	—	—	—
6201	U.h.f. Triode	B.	9A	6.3	0.3	2.3	0.4	1.6	Class-A_1 Amp.	250	200*	—	—	10	10900	5500	60	—	—	12AT7
6216	Filter Reactor—Beam Power Tube	B.	Fig. 73	6.3	1.2	12.3	6.7	0.37	Filter Reactor	400	−1	100	D.C. output = 335 volts at 100 Ma. d.c. load.							—
									Class-A Amp.	200	−6	100	2/4	47/51	38800	8800	—	4500	3.8	
9001	Sharp Cut-off Pentode	B.	7PM	6.3	0.15	3.6	3.0	0.01	Class-A Amp.	250	−3.0	100	0.7	2.0	1 meg.+	1400	—	—	—	—
									Mixer	250	−5.0	100	Osc. peak voltage 4 volts			550	—	—	—	
9002	Triode Detector, Amplifier, Oscillator	B.	7TM	6.3	0.15	1.2	1.1	1.40	Class-A Amp.	250 90	−7.0 −2.5	— —	— —	6.3 2.5	11400 14700	2200 1700	25 25	— —	— —	—
9003	Remote Cut-off Pentode	B.	7PM	6.3	0.15	3.6	3.0	0.01	Class-A Amp.	250	−3.0	100	2.7	6.7	700000	1800	—	—	—	—
									Mixer	250	−10.0	100	Osc. peak voltage 9 volts			600	—	—	—	
9006	U.h.f. Diode	B.	6BH	6.3	0.15	—	—	—	Detector	Max. a.c. voltage—270. Max. d.c. output current—5 ma.										—

Ω Oscillator gridleak ohms.
* Cathode resistor—ohms.
[1] Per Plate.
[2] Maximum-signal current for full-power output.
[3] Values are for two tubes in push-pull
[4] Unless otherwise noted.
[5] No signal plate ma.
[6] Effective plate-to-plate.
[7] Triode No. 1.
[8] Triode No. 2.
[9] Grid No. 2 tied to plate and No. 3 to cathode.
[10] Oscillator grid current Ma.
[11] Values for each section.
[12] Between G_1 and G_2.

TABLE XII—SUB-MINIATURE TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
1AC5	Power Pentode	Bs.	Fig. 14	1.25	0.04	—	—	—	Class-A_1 Amp.	67.5	−4.5	67.5	0.4	2.0	150000	750	—	25000	0.05	1AC5
1AD4	Pentode	1	2	1.25	0.1	4.5	4.5	0.01	Class-A_1 Amp.	45	0	45	0.8	3.0	500000	2000	—	—	—	1AD4
1AD5	Sharp Cut-off Pentode	Bs.	Fig. 16	1.25	0.04	1.8	2.8	0.01	Class-A_1 Amp.	67.5	0	67.5	0.75	1.85	700000	735	—	—	—	1AD5
1AE5	Heptode	1	2	1.25	0.06	4.9	2.1	4.0	Mixer	45	0	45	2.0	0.9	200000	200	—	—	—	1AE5
1AH4	R.F. Pentode	1	2	1.25	0.04	3.5	4.5	0.01	Class-A_1 Amp.	67.5	0	67.5	0.2	0.75	2000000	750	—	—	—	1AH4
1AJ5	Diode Pentode	1	2	1.25	0.04	1.7	2.4	0.1	Class-A_1 Amp.	45	0	45	0.3	1.0	300000	425	—	—	—	1AJ5
1C8	Heptode	—	—	1.25	0.04	6.5	4.0	0.25	Converter	30	0	30	0.75	0.32	300000	100	—	—	—	1C8
1D3	Triode	1	2	1.25	0.3	1.0	1.0	2.6	Class-A Amp.	90	−5	—	—	12.5	—	3400	8.7	—	—	1D3
1E8	Pentagrid Converter	Bs.	Fig. 27	1.25	0.04	6	—	—	Converter	67.5	0	67.5	1.5	1.0	—	150	—	—	—	1E8
1Q6	Diode Pentode	Bs.	8CO	1.25	0.04	1.8	4.2	0.085	Class-A Amp.	67.5	0	67.5	0.4	1.6	400000	600	—	—	—	1Q6
1S6	Diode Pentode	Bs.	8DA	1.25	0.04	—	—	—	Detector Amp.	67.5	0	67.5	0.4	1.6	400000	600	—	—	—	1S6
1T6	Diode-Pentode	Bs.	Fig. 28	1.25	0.04	—	—	—	Class-A_1 Amp.	67.5	0	67.5	0.4	1.6	400000	600	—	—	—	1T6
1V5	Audio Pentode	1	2	1.25	0.04	—	—	—	Class-A_1 Amp.	67.5	−4.5	67.5	0.4	2.0	150000	750	—	25000	0.05	1V5
1V6	Triode Pentode	1	2	1.25	0.04	—	—	—	Pentode	45	0	45	0.15	0.4	1000000	200	—	—	—	1V6
									Triode Osc.	45	Osc. grid current = 12 μ amp. through 1 Meg. gridleak									
1W5	Sharp Cut-off Pentode	1	2	1.25	0.04	2.3	3.5	0.01	Class-A_1 Amp.	67.5	0	67.5	0.75	1.85	700000	735	—	—	—	1W5
2B5	Twin Triode	1	2	1.2	0.26	0.8	0.8	1.2	Class-A Amp.	90	−1	—	—	2.6	18700	1150	21.5	—	—	2B5
				2.4	0.13															
2E31	R.F. Pentode	1	2	1.25	0.05				Class-A_1 Amp.	22.5	0	22.5	0.3	0.4	—	500	—	—	—	2E31
2E32	R.F. Pentode	1	2	1.25	0.05				Class-A Amp.	22.5	0	22.5	0.3	0.4	350000	500	—	—	—	2E32
2E35	Audio Pentode	1	2	1.25	0.03				Class-A_1 Amp.	22.5	0	22.5	0.07	0.27	—	385	—	—	0.0012	2E35
2E36	Audio Pentode	1	2	1.25	0.03				Class-A_1 Amp.	22.5	0	22.5	0.07	0.27	220000	385	—	150000	0.0012	2E36
										45	−1.25	45	0.11	0.45	250000	500	—	100000	0.00	
2E41	Diode Pentode	1	2	1.25	0.03	—	—	—	Detector Amp.	22.5	0	22.5	0.12	0.35	—	—	—	—	—	2E41
2E42	Diode Pentode	1	2	1.25	0.03				Detector Amp.	22.5	0	22.5	0.12	0.35	250000	375	—	1 meg.	—	2E42
2G21	Triode Heptode	1	2	1.25	0.05	—	—	—	Converter	22.5	—	22.5	0.2	0.3	—	75	—	—		2G21
2G22	Converter	1	2	1.25	0.05				Converter	22.5	0	22.5	0.3	0.2	500000	60	—	—		2G22
6AD4	Triode	Bs.	2	6.3	0.15	2.8	3.2	1.31	Class-A_1 Amp.	100	820*	—	—	1.4	26000	2700	70	—	—	6AD4
6AZ5	Dual Diode	1	2	6.3	0.15	—	—	—	Rectifier	Max. a.c. volts—150. Peak inverse volts—420. Peak Ma.—24. Av. Ma.—4.0										6AZ5
6BA5	Pentode	1	2	6.3	0.15	4.0	6.5	0.19	Class-A_1 Amp.	100	270*	100	1.25	4.8	150000	3300	—	—	—	6BA5
6BF7	Dual Triode	Bs.	8DG	6.3	0.3	2.0	1.6	1.5	R.F. Amp.	100	100*	—	—	8.0	7000	4800	35	—	—	6BF7
6BG7	Dual Triode	Bs.	8DG	6.3	0.3	2.0	1.6	1.5	R.F. Amp.	100	100*	—	—	8.0	7000	4800	35	—	—	6BG7
6K4	Triode	1	2	6.3	0.15	2.4	0.8	2.4	Class A_1 Amp.	200	680*	—	—	11.5	4650	3450	16	—	—	6K4
1247	Diode	1	2	0.7	0.065	—	—	—	R.F. Probe	Max. a.c. volts—300 r.m.s. D.C. plate current—0.4 Ma.										1247
CK501	Pentode Voltage Amplifier	—1	2	1.25	0.033	—	—	—	Class-A Amp.	30	0	30	0.06	0.3	1000000	325	—	—	—	CK501
										45	−1.25	45	0.055	0.28	1500000	300				
CK502	Pentode Output Amplifier	—1	2	1.25	0.033	—	—	—	Class-A Amp.	30	0	30	0.13	0.55	500000	400	—	60000	0.003	CK502
CK503	Pentode Output Amplifier	—1	2	1.25	0.033	—	—	—	Class-A Amp.	30	0	30	0.33	1.5	150000	600	—	20000	0.006	CK503
CK504	Pentode Output Amplifier	—1	2	1.25	0.033	—	—	—	Class-A Amp.	30	−1.25	30	0.09	0.4	500000	350	—	60000	0.003	CK504
CK505	Pentode Voltage Amplifier	—1	2	0.625	0.03	—	—	—	Class-A Amp.	30	0	30	0.07	0.17	1100000	140	—	—	—	CK505
										45	−1.25	45	0.08	0.2	2000000	150				
CK506	Pentode Output Amplifier	—1	2	1.25	0.05	—	—	—	Class-A_1 Amp.	45	−4.5	45	0.4	1.25	120000	500	—	30000	0.025	CK506
CK507	Pentode Output Amplifier	—1	2	1.25	0.05	—	—	—	Class-A_1 Amp.	45	−2.5	45	0.21	0.6	360000	500	—	50000	0.010	CK507
CK509	Triode Voltage Amplifier	—1	2	0.625	0.03	—	—	—	Class-A Amp.	45	0	—	—	0.15	150000	160	16	1000000	—	CK509
CK510	Dual Space-Charge Tetrode	—1	2	0.625	0.05	—	—	—	Class-A Amp.	45	0	0.2	200 μa	60 μa	500000	65	32.5	—	—	CK510
CK512	Low Microphonic Pentode	1	2	0.625	0.02	—	—	—	Voltage Amp.	22.5	0	22.5	0.04	0.125	—	160	—	—	—	CK412
CK515BX	Triode Voltage Amplifier	—1	2	0.625	0.03	—	—	—	Class-A Amp.	45	0	—	—	0.15	—	160	24	1000000	—	CK515BX
CK520AX	Audio Pentode	1	2	0.625	0.05	—	—	—	Class-A_1 Amp.	45	−2.5	45	0.07	0.24	—	180	—	—	0.0045	CK520AX
CK521AX	Audio Pentode	1	2	1.25	0.05	—	—	—	Class-A_1 Amp.	22.5	−3	22.5	0.22	0.8	—	400	—	—	0.006	CK521AX

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
CK522AX	Audio Pentode	1	2	1.25	0.02	—	—	—	Class-A$_1$ Amp.	22.5	0	22.5	0.08	0.3	—	450	—	—	0.0012	CK522AX
CK523AX	Pentode Output Amp.	1	—	1.25	0.03	—	—	—	Class-A Amp.	22.5	−1.2	22.5	0.075	0.3	—	360	—	—	0.0025	CK523AX
CK524AX	Pentode Output Amp.	1	—	1.25	0.03	—	—	—	Class-A Amp.	15	−1.75	15	0.125	0.45	—	300	—	—	0.0022	CK524AX
CK525AX	Pentode Output Amp.	1	—	1.25	0.2	—	—	—	Class-A Amp.	22.5	−1.2	22.5	0.06	0.25	—	325	—	—	0.0022	CK525AX
CK526AX	Pentode Output Amp.	1	—	1.25	0.2	—	—	—	Class-A Amp.	22.5	−1.5	22.5	0.12	0.45	—	400	—	—	0.004	CK526AX
CK527AX	Pentode Output Amp.	1	—	1.25	0.015	—	—	—	Class-A Amp.	22.5	0	22.5	0.025	0.1	—	75	—	—	0.0007	CK527AX
CK529AX	Shielded Output Pentode	1	—	1.25	0.02	—	—	—	Class-A Amp.	15	−1.5	15	0.05	0.2	—	275	—	—	0.0012	CK529AX
CK551AXA	Diode Pentode	1	2	1.25	0.03	—	—	—	Detector-Amp.	22.5	0	22.5	0.04	0.17	—	235	—	—	—	CK551AXA
CK553AXA	R.F. Pentode	1	2	1.25	0.05	—	—	—	Class-A$_1$ Amp.	22.5	0	22.5	0.13	0.42	—	550	—	—	—	CK553AXA
CK556AX	U.h.f. Triode	1	2	1.25	0.125	—	—	—	R.F. Oscillator	135	−5	—	—	4.0	—	1600	—	—	—	CK556AX
CK568AX	U.h.f. Triode	1	2	1.25	0.07	—	—	—	R.F. Oscillator	135	−6	—	—	1.9	—	650	—	—	—	CK568AX
CK569AX	R.F. Pentode	1	2	1.25	0.05	—	—	—	Class-A$_1$ Amp.	67.5	0	67.5	0.48	1.8	—	1100	—	—	—	CK569AX
CK605CX	Sharp Cut-off Pentode	1	—	6.3	0.2	—	—	—	Class-A Amp.	120	−2	120	2.5	7.5	—	5000	—	—	—	CK605CX
CK606BX	Single Diode	1	2	6.3	0.15	—	—	—	Detector	150 a.c.	—	—	—	9.0 d.c.	—	—	—	—	—	CK606BX
CK608CX	U.h.f. Triode	1	2	6.3	0.2	—	—	—	500-Mc. Osc.	120	−2	—	—	9.0	—	5000	—	—	0.75	CK608CX
CK619CX	Hi-Mu Triode	1	2	6.3	0.2	—	—	—	Class-A$_1$ Amp.	250	−2	—	—	4.0	—	4000	—	—	—	CK619CX
CK624CX	Sharp Cut-off Pentode	1	—	6.3	0.2	—	—	—	Class-A Amp.	120	−2	120	3.5	5.2	—	3000	—	—	—	CK624CX
CK650AX	Sharp Cut-off Pentode	1	2	6.3	0.2	—	—	—	Class-A$_1$ Amp.	120	−2	120	2.5	7.5	—	5000	—	—	—	CK650AX
CK5672	Pentode Output Amp.	1	—	1.25	0.05	—	—	—	Class-A Amp.	67.5	−6.25	67.5	1.0	2.75	—	625	—	—	0.06	CK5672
HY113 HY123	Triode Amplifier	—[1]	5K	1.4	0.07	—	—	—	Class-A Amp.	45	−4.5	—	—	0.4	25000	250	6.3	40000	0.0065	HY113 HY123
HY115 HY145	Pentode Voltage Amplifier	—[1]	5K	1.4	0.07	—	—	—	Class-A Amp.	45 90	−1.5 −1.5	22.5 45	0.008 0.1	0.03 0.48	5200000 1300000	58 270	300 370	—	—	HY115 HY145
HY125 HY155	Pentode Power Amplifier	—[1]	5K	1.4	0.07	—	—	—	Class-A Amp.	45 90	−3.0 −7.5	45 90	0.2 0.5	0.9 2.6	825000 420000	310 450	255 190	50000 28000	0.0115 0.09	HY125 HY155
M54	Tetrode Power Amplifier	1	2	0.625	0.04	—	—	—	Class-A Amp.	30	0	30	0.06	0.5	130000	200	26	35000	0.005	M54
M64	Tetrode Voltage Amplifier	1	2	0.625	0.02	—	—	—	Class-A Amp.	30	0	—	—	0.03	200000	110	25	—	—	M64
M74	Tetrode Voltage Amplifier	1	2	0.625	0.02	—	—	—	Class-A Amp.	30	0	7.0	0.01	0.02	500000	125	70	—	—	M74
RK61	Gas Triode	1	2	1.4	0.05	—	—	—	Radio Control	45	—	—	—	1.5	—	—	—	—	—	RK61
SD917A 5637	Triode	1	2	6.3	0.15	2.6	0.7	1.4	Class-A$_1$ Amp.	100	820*	—	—	1.4	26000	2700	70	—	—	SD917A 5637
SD828A 5638	Audio Pentode	1	2	6.3	0.15	4.0	3.0	0.22	Class-A$_1$ Amp.	100	270*	100	1.25	4.8	150000	3300	—	—	—	SD828A 5638
SD828E 5634	Sharp Cut-off Pentode	4	—	6.3	0.15	4.4	2.8	0.01	Class-A$_1$ Amp.	100	150*	100	2.5	6.5	240000	3500	—	—	—	SD828E 5634
SN944 5633	Remote Cut-off Pentode	4	—	6.3	0.15	4.0	2.8	0.01	Class-A$_1$ Amp.	100	150*	100	2.8	7.0	200000	3400	—	—	—	SN944 5633
SN946	Diode	1	2	6.3	0.15	1.8	—	—	Rectifier	150	—	—	—	9.0	—	—	—	—	—	SN946
SN947D 5640	Audio Beam Pentode	1	2	6.3	0.45	—	—	—	Class-A$_1$ Amp.	100	−9	100	2.2	31.0	15000	5000	—	3000	1.25	SN947C 5640
SN948C	Voltage Regulator	1	—	—	—	—	—	—	Regulator	Operating voltage = 95; Max. current = 25 Ma.										SN948C
SN953D	Power Pentode	1	—	6.3	0.15	9.5	3.8	0.2	Class-A Amp.	150	100*	100	4/7.5	21/20	50000	9000	—	9000	1.0	SN953D
SN954 5641	Half-Wave Rectifier	1	2	6.3	0.45	—	—	—	Rectifier	300	—	—	—	45.0	—	—	—	—	—	SN954 5641
SN955B	Dual Triode	1	2	6.3	0.45	2.8	1.0	1.3	Class-A$_1$ Amp.[6]	100	100*	—	—	5.5	8000	4250	34	—	—	SN955B
SN956B 5642	H.V. Half-Wave Rectifier		—	1.25	0.14	—	—	—	H.V. Rectifier	Peak inverse V. = 10000 Max. Average Ip = 2 Ma. Peak Ip = 23 Ma.										SN956B 5642
SN957A 5645	Triode	1	2	6.3	0.15	2.0	1.0	1.8	Class-A$_1$ Amp.	100	560*	—	—	5.0	7400	2700	20	—	—	SN957A 5645
SN1006	Triode	1	2	6.3	0.15				Class-A$_1$ Amp.	100	820*	—	—	1.4	29000	2400	70	—	—	SN1006
SN1007B	Mixer	4	—	6.3	0.15	5.0	2.8	0.003	Mixer	100	150*	100	5.0	4.0	230000	900	—	—	—	SN1007B

TABLE XII—SUB-MINIATURE TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amp.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
5635	Dual Triode	Bs.	8DB	6.3	0.45	2.6	1.6	1.2	Class-A Amp.[5]	100	100*	—	—	4.8	10000	3800	38	—	—	5635
5636	Pentode Mixer	Bs.	8DC	6.3	0.15	4.0	1.9	0.034	Class-A Amp.	100	150*	100	4.0	5.6	110000	3200	—	—	—	5636
5639	Video Pentode	[1]	8DL	6.3	0.45	9.5	7.5	0.10	Class-A$_1$ Amp.	150	100*	100	4.0	21	50K	9000	—	9000	1.0	5639
5641	Single Diode	[1]	6CJ	6.3	0.45	—	—	—	H. W. Rectifier	235 volts a.c. max.; 45 Ma. d.c. output.										5641
5643	Tetrode Thyratron	[1]	8DD	6.3	0.15	1.7	1.6	0.1	Relay Tube Grid Contr. Rect.	Peak anode volts = 500; Inv. volts = 500; Peak I_k = 100 Ma.; Avg. = 22 Ma.										5643
5644	Cold Cathode Diode	[1]	4CN	—	—	—	—	—	Voltage Reg.	Starting voltage = 125 max. d.c. Operating voltage = 95. Operating current = 5-25 Ma. Regulation = 4 volts approx.										5644
5646	Triode	[1]	—	6.3	0.15	2.4	3.4	1.2	Class-A Amp.	100	820*	—	—	1.4	29000	2400	70	—	—	5646
5647	Single Diode	[1]	B1	6.3	0.15	2.2	—	—	H. W. Rectifier	150 volts a.c. max; 9 Ma. d.c. output.										5647
5672	Power Pentode	[1]	[2]	1.25	0.05	—	—	—	Class-A Amp.	67.5	−6.5	67.5	1.1	3.25	—	650	—	20000	0.065	5672
5676	Triode	[1]	[2]	1.25	0.12	1.3	4	2	Class-A Amp.	135	−5	—	—	4	—	1600	15	—	—	5676
5677	Triode	[1]	[2]	1.25	0.06	1.3	3.8	2	Class-A Amp.	135	−6	—	—	1.9	—	650	13.5	—	—	5677
5678	Pentode	[1]	[2]	1.25	0.05	3.3	3.8	.01	Class-A Amp.	67.5	0	67.5	0.48	1.8	1000000	1100	—	—	—	5678
5697	Lo-Mu Triode	[1]	—	0.625	0.02	—	—	—	Class-A Amp.	12	−3	—	—	0.22	—	135	2.1	—	—	5697
5702	Remote Cut-off Pentode	[1]	—	6.3	0.2	4.4	3.5	0.03	Class-A Amp.	120	200*	120	2.5	7.5	340000	5000	—	—	—	5702
5703	Medium Mu Triode	[1]	—	6.3	0.2	2.6	0.7	1.2	Class-A Amp.	120	220*	—	—	9	—	5000	25	—	—	5703
5704	Diode	[1]	—	6.3	0.15	—	—	—	Max. RMS plate volts = 150; Max. d.c. Ma. = 9; Peak I_P = 54 Ma.; Inverse volts = 420.											5704
5718	U.h.f. Medium-Mu Triode	[1]	8DK	6.3	0.15	2.2	0.7	1.4	Class-A$_1$ Amp.	150	180*	—	—	13	4150	6500	27	—	—	5718
									U.h.f. Oscillator	150	−12	Fmc. = 500		20		Ig = 3.7 Ma.			0.9	
5719	Hi-Mu Triode	[1]	8DK	6.3	0.15	2.4	0.6	0.7	Class-A$_1$ Amp.	150	680*	—	—	1.7	26000	2700	70	—	—	5719
5744	Hi-Mu Triode	[1]	—	6.3	0.2	—	—	—	Class-A Amp.	250	500*	—	—	4.0	—	4000	70	—	—	5744
5784	Dual-Control Pentode	[1]	—	6.3	0.2	3.9	3.0	0.03	Class-A Amp.	120	−2	120	4.8	3.6	—	1850	—	—	—	5784
5785	High Voltage Diode	[1]	—	—	—	—	—	—	Max. d.c. I_P = 0.1 Ma.; Peak I_P = 0.45 Ma.; Max. peak inverse volts = 3500.											5785
5840	U.h.f. Sharp Cut-off Pent.	[1]	8DL	6.3	0.15	4.2	4.0	0.015	Class-A$_1$ Amp.	100	150*	100	2.4	7.5	230K	5000	—	—	—	5840
5851	Pentode Power Amplifier	[1]	6CL	2.5 1.25	0.055 0.11	2.5	3.0	0.055	Class-A Amp.	125	−7.5	125	0.9	5.5	175000	1600	—	—	—	5851
5896	U.h.f. Dual Diode	[1]	8DJ	6.3	0.3	3.0	—	—	Det.-Rectifier	150 volts a.c. max.; 9 Ma. d.c. output per plate.										5896
5897	U.h.f. Medium-Mu Triode	[1]	8DK	6.3	0.15	2.2	0.7	1.4	Class-A$_1$ Amp.	150	180*	—	—	13	4150	6500	27	—	—	5897
									U.h.f. Oscillator	150	−12	—	—	20	Ig = 3.7 Ma. Fmc. = 500.				0.9	
5898	Hi-Mu Triode	[1]	8DK	6.3	0.15	2.4	0.6	0.7	Class-A$_1$ Amp.	150	680*	—	—	1.7	26000	2700	70	—	—	5898
5899	U.h.f. Semi-Remote Pent.	[1]	8DL	6.3	0.15	4.4	4.0	0.015	Class-A$_1$ Amp.	100	120*	100	2.2	7.2	260K	4500	—	—	—	5899
5900	U.h.f. Semi-Remote Pent.	[1]	8DL	6.3	0.15	4.4	4.0	0.015	Class-A$_1$ Amp.	100	120*	100	2.2	7.2	260K	4500	—	—	—	5900
5901	U.h.f. Sharp Cut-off Pent.	[1]	8DL	6.3	0.15	4.2	4.0	0.015	Class-A$_1$ Amp.	100	150*	100	2.4	7.5	230K	5000	—	—	—	5901
5902	Audio Beam Pentode	[1]	8DL	6.3	0.15	6.5	7.5	0.11	Class-A$_1$ Amp.	110	270*	110	2.2	30	15K	4200	—	3000	1.0	5902
5903	U.h.f. Dual Diode	[1]	8DJ	26.5	0.075	3.0	—	—	Det.-Rectifier	150 volts a.c. max.; 9 Ma. d.c. output per plate.										5903
5904	U.h.f. Medium-Mu Triode	[1]	8DK	26.5	0.045	2.2	0.8	1.8	Class-A$_1$ Amp.	26.5	−3.5	—	—	3	3800	5000	19	—	—	5904
									U.h.f. Oscillator	26.5	0	—	—	20	Ig = 7.5 Ma. Fmc. = 400.				0.06	
5905	U.h.f. Sharp Cut-off Pent.	[1]	8DL	26.5	0.045	4.4	4.2	0.015	Class-A$_1$ Amp.	26.5	2.2[6]	26.5	0.9	2.3	110K	2850	—	—	—	5905
5906	U.h.f. Sharp Cut-off Pent.	[1]	8DL	26.5	0.045	4.2	4.0	0.015	Class-A$_1$ Amp.	100	150*	100	2.4	7.5	230K	5000	—	—	—	5906
5907	U.h.f. Remote Cut-off Pent.	[1]	8DL	26.5	0.045	4.4	4.0	0.015	Class-A$_1$ Amp.	26.5	2.2[6]	26.5	1.1	2.7	125K	3000	—	—	—	5907
5908	U.h.f. Pentode	[1]	8DC	26.5	0.045	4.4	4.6	0.08	Class-A$_1$ Amp.	26.5	2.2[6]	26.5	1.6	2.3	30K	1750	—	—	—	5908
									Mixer	26.5	2.2[6]	26.5	1.6	1.0	100K	800	—	—	—	
5916	U.h.f. Pentode		8DC	26.5	0.045	4.2	4.0	0.015	Class-A$_1$ Amp.	100	150*	100	3.4	4.4	130K	3000	—	—	—	5916
									Mixer	100	150*	100	4.6	2.5	400K	1100	—	—	—	
5977	Triode	Bs.	8DK	6.3	0.15	2.0	0.8	1.3	Class-A Amp.	100	270*	—	—	10	3650	4500	16	—	—	5977
5987	Triode	Bs.	8DM	6.3	0.45	2.8	1.5	3.2	Class-A$_1$ Amp.	150	−24	—	—	22.5/28	2220	1850	4.1	3500	0.75	5987
6111	Twin Triode	Bs.	8DG	6.3	0.3	1.9	0.3	0.009	Class-A Amp.	100	220*	—	—	8.5	4000	5000	20	—	—	6111

* Cathode resistor ohms. [1] No base; tinned wire leads. [2] Leads identified on tube. [3] No screen connection. [4] Double-ended type. [5] Values per triode. [6] Grid leak resistor, megohms.

TABLE XIII—CONTROL AND REGULATOR TUBES

Type	Name	Base	Socket Connec-tions	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Use	Peak Anode Voltage	Max. Anode Ma.	Minimum Supply Voltage	Operating Voltage	Operating Ma.	Grid Resistor	Tube Voltage Drop	Type
0A2 6073	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	—	—	185	150	5–30	—	—	0A2
0A5	Gas Pentode	7-pin B.	Fig. 33	Cold	—	—	Relay or Trigger	Plate—750 V., Screen—90 V., Grid +3 V., Pulse—85 V.							0A5
0B2 6074	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	—	—	133	108	5–30	—	—	0B2
0A4G 1267	Gas Triode Starter-Anode Type	6-pin O.	4V 4V	Cold	—	—	Cold-Cathode Starter-Anode Relay Tube	With 105–120-volt a.c. anode supply, peak starter-anode a.c. voltage is 70, peak r.f. voltage 55. Peak d.c. ma = 100. Average d.c. ma = 25.							0A4G 1267
0G3	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	—	—	125	85	1–6	—	—	0G3
1B47	Voltage Regulator	7-pin B.	—	—	—	—	Voltage Regulator	—	—	225	82	1–2	—	—	1B47
1C21	Gas Triode Glow-Discharge Type	6-pin O.	4V	Cold	—	—	Relay Tube	125–145	25	66 [6]	—	—	—	73	1C21
							Voltage Regulator		0.1 [6]	180 [4]				55	
2A4G	Gas Triode Grid Type	7-pin O.	5S	Fil.	2.5	2.5	Control Tube	200	100	—	—	—	—	15	2A4G
6Q5G	Gas Triode Grid Type	8-pin O.	6Q	Htr.	6.3	0.6	Sweep Circuit Oscillator	300	300	—	—	1.0	0.1–10 [7]	19	6Q5G
2B4		5-pin M.	5A	Htr.	2.5	1.4									2B4
2C4	Gas Triode	7-pin B.	5AS	Fil.	2.5	0.65	Control Tube	Plate volts = 350; Grid volts = −50; Avg. Ma. = 5; Peak Ma. = 20; Voltage drop = 16.							2C4
2D21	Gas Tetrode	7-pin B.	7BN	Htr.	6.3	0.6	Grid-Controlled Rectifier	650	500	—	650	100	0.1–10 [7]	8	2D21
							Relay Tube	400	—	—	—	—	1.0 [7]	—	
3C23	Gas and Mercury Vapor Grid Type	4-pin M.	3G	Fil.	2.5	7.0	Grid-Controlled Rectifier	1000	6000	—	500	1500	−4.5 [8]	15	3C23
										—	100	1500	−2.5 [8]	15	
6D4	Gas Triode	7-pin B.	5AY	Htr.	6.3	0.25	Control Tube	Plate volts = 350; Grid volts = −50; Avg. Ma. = 25; Peak Ma. = 100; Voltage drop = 16.							6D4
17	Mercury Vapor Triode	4-pin M.	3G	Fil.	2.5	5.0	Grid-Controlled Rectifier	7500 [5]	2000	—	—	500	200–3000	—	17
								2500		−5 [3]	1000	250	—	10–24	
874	Voltage Regulator	4-pin M.	4S	—	—	—	Voltage Regulator	—	—	125	90	10–50	—	—	874
876♯	Current Regulator	Mogul	—	—	—	—	Current Regulator	—	—	—	40–60	1.7	—	—	876
884	Gas Triode Grid Type	6-pin O.	6Q	Htr.	6.3	0.6	Sweep Circuit Oscillator	300	300	—	—	2	25000	—	884
							Grid-Controlled Rectifier	350	300	—	—	75	25000	—	
885	Gas Triode Grid Type	5-pin S.	5A	Htr.	2.5	1.4	Same as Type 884	Characteristics same as Type 884							885
886♯	Current Regulator	Mogul	—	—	—	—	Current Regulator	—	—	—	40–60	2.05	—	—	886
967	Mercury Vapor Triode	4-pin M.	3G	Fil.	2.5	5.0	Grid-Controlled Rectifier	2500	500	−5 [3]	—	—	—	10–24	967
991	Voltage Regulator	Bayonet	—	—	—	—	Voltage Regulator	—	—	87	55–60	2.0	—	—	991
1265	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	130	90	5–30	—	—	1265
1266	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	—	70	5–40	—	—	1266
1267	Gas Triode	6-pin O.	4V	Cold	—	—	Relay Tube	Characteristics same as OA4G							1267
2050	Gas Tetrode	8-pin O.	8BA	Htr.	6.3	0.6	Grid-Controlled Rectifier	650	500	—	—	100	0.1–10 [7]	8	2050
2051	Gas Tetrode	8-pin O.	8BA	Htr.	6.3	0.6	Grid-Controlled Rectifier	350	375	—	—	75	0.1–10 [7]	14	2051
2523N1/ 128AS	Gas Triode Grid Type	5-pin M.	5A	Htr.	2.5	1.75	Relay Tube	400	300	—	—	1.0	300 [7]	13	2523N1/ 128AS
5651	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	115	—	115	87	1.5–3.5	—	—	5651
5663	Tetrode Thyratron	7-pin B	7CE	Htr.	6.3	0.15	Control and Relay	Max. peak inv. volts = 500; Peak Ma. = 100; Avg. Ma. = 20.							5663
5727	Gas Tetrode	7-pin B.	7BN	Htr.	6.3	0.6	Grid-Controlled Rectifier	650	500	—	650	100	0.1 [7]	8	5727
5823	Gas Triode	7-pin B.	4CK	Cold	—	—	Relay or Trigger	Max. peak inv. volts = 200; Peak Ma. = 100; Avg. Ma. = 25.							5823
5890	Remote Cut-off Pentode Regulator	—	12J	Htr.	6.3	0.6	Shunt Regulator		E_{G1} = −60 volts; E_{G2} = 200 volts; E_{G3} = 5500 volts. E_P = 30000 volts; I_{G2} = 0 Ma.; I_P Max. = 0.5 Ma.						5890
5962	Voltage Regulator	7-pin B.	2AG	Cold	—	—	Voltage Regulator	—	—	730	700	5/55 [10]	—	—	5962
6140	Voltage Regulator	9-pin B.	9BY	Cold	—	—	Voltage Regulator	—	—	160	100	4–6	—	—	6140
6141	Voltage Regulator	9-pin B.	9BZ	Cold	—	—	Voltage Regulator	—	—	165	100	5–40	—	—	6141
KY21	Gas Triode Grid Type	4-pin M.	—	Fil.	2.5	10.0	Grid-Controlled Rectifier	—	—	—	3000	500	—	—	KY21
RK61	Thyratron	— [9]	—	Fil.	1.4	0.05	Radio-Controlled Relay	45	1.5	30	—	0.5–1.5	3 [7]	30	RK61
RK62	Gas Triode Grid Type	4-pin S.	4D	Fil.	1.4	0.05	Relay Tube	45	1.5	—	30–45	0.1–1.5	—	15	RK62
RM208	Permotron	4-pin M.	—	Fil.	2.5	5.0	Controlled Rectifier [1]	7500 [2]	1000	—	—	—	—	15	RM208

TABLE XIII—CONTROL AND REGULATOR TUBES—*Continued*

Type	Name	Base	Socket Connec-tions	Cathode	Fil. or Heater Volts	Amps.	Use	Peak Anode Voltage	Max. Anode Ma.	Minimum Supply Voltage	Operating Voltage	Operating Ma.	Grid Resistor	Tube Voltage Drop	Type
RM209	Permatron	4-pin M.	—	Fil.	5.0	10.0	Controlled Rectifier[1]	7500[2]	5000	—	—	—	—	15	RM209
OA3/VR75	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	105	75	5–40	—	—	OA3/VR75
OB3/VR90	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	125	90	5–40	—	—	OB3/VR90
OC3/VR105	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	135	105	5–40	—	—	OC3/VR105
OD3/VR150	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	185	150	5–40	—	—	OD3/VR150
KY866	Mercury Vapor Triode	4-pin M.	Fig. 8	Fil.	2.5	5.0	Grid-Controlled Rectifier	10000	1000	0–150	—	—	—	—	KY866

[1] For use as grid-controlled rectifier or with external magnetic control. RM-208 has characteristics of 866, RM-209 of 872.
♯ Discontinued.
[2] When under control peak inverse rating is reduced to 2500.
[3] At 1000 anode volts.
[4] Grid tied to plate.
[5] Peak inverse voltage.
[6] Grid.
[7] Megohms.
[8] Grid voltage.
[9] No base. Tinned wire leads.
[10] Values in μ amperes.

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES

Type	Name	Socket Connec-tions	Heater Volts	Heater Amp.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Ion-Trap Ma.	Max. Input Voltage[1]	Focus Coil Ma.	Deflection Sensitivity[6] $D_1 D_2$	Deflection Sensitivity[6] $D_3 D_4$	Anode No. 3 Voltage	Pattern Color	Type
2AP1[7]-11	Electrostatic Cathode-Ray	11B	6.3	0.6	Oscillograph Television	2″	1000	250	−60	—	—	660	—	0.11	0.13	—	Green	2AP1-11
							500	125	−30	—	—			0.22	0.26			
2BP1-11	Electrostatic Cathode-Ray	12E	6.3	0.6	Oscillograph	2″	2000	300/560	−135	—	—	500	—	270[3]	174[3]	—	Green	2BP1-11
							1000	150/280	−67.5	—	—	500	—	135[3]	87[3]	—		
3AP1/ 906-P1-4-5-11[7]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″	1500	430	−50	—	—	550	—	0.22	0.23	—	Green Blue White	3AP1/ 906-P1-4-5-11
							1000	285	−33	—	—			0.33	0.35			
							600	170	−20	—	—			0.55	0.58			
3BP1-4-11	Electrostatic Cathode-Ray	14A	6.3	0.6	Oscillograph	3″	2000	575	−60	—	—	550	—	0.13	0.17	—	Green	3BP1-4-11
							1500	430	−45	—	—			0.17	0.23			
3DP1	Electrostatic Cathode-Ray	Fig. 49	6.3	0.6	Oscillograph	3″	2000	575	−60	—	—	550	—	200[3]	148[3]	—	Green	3DP1
							1500	430	−40	—	—			150[3]	111[3]	—		
3EP1/ 1806-P1	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph Television	3″	2000	575	−60	—	—	550	—	0.115	0.154	—	Green	3EP1/ 1806-P1
							1500	430	−45	—	—			0.153	0.205			
3FP7-A	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph	3″	4000	400/690	−90	2000	—	—	—	212[3]	153[3]	—	—	3FP7-A
3GP1-4-5-11	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph	3″	1500	350	−50	—	—	550	—	0.21	0.24	—	White Green Blue	3GP1-4-5-11
							1000	234	−33	—	—		—	0.32	0.36			
3JP1-2-4-7-11	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph	3″	2000	575	−60	—	—	550	—	0.13	0.17	4000	Green Blue White	3JP1-2-4-7-11
							1500	430	−45	—	—		—	0.17	0.23	3000		
3KP1-11	Electrostatic Cathode-Ray	11M	6.3	0.6	Oscillograph	3″	1000	300	−45	1000	—	500	—	68[3]	136[5]	—	Green	3KP1-11
							2000	600	−90	2000	—			52[3]	104[3]			
3MP1	Electrostatic Cathode-Ray	Fig. 2	6.3	0.6	Oscillograph	3″	1000	200/350	−68	—	—	—	—	190[3]	180[3]	—	Green	3MP1
3RP1	Electrostatic Cathode-Ray	12E	6.3	0.6	Oscillograph	3″	1000	165/310	−67.5	—	—	—	—	73/99[3]	52/70[3]	—	Green	3RP1
							2000	330/620	−135	—	—	—	—	146/198[3]	104/140[3]	—		
3SP1-7	Electrostatic Cathode-Ray	12E	6.3	0.6	Oscillograph	3″	2000	330/620	−58/−135	—	—	—	—			—	Green	3SP1
5ABP1-7-11	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph	5″	2000	400/690	−52/−87	—	—	—	—	26/36[3]	18/24[3]	4000	—	5ABP1-7-11
5AP1/ 1805-P1 5AP4/ 1805-P4[7]	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph Television	5″	2000	575	−35	—	—	500	—	0.17	0.21	—	Green White	5AP1/ 1805-P1 5AP4/ 1805-P4
							1500	430	−27					0.23	0.28	—		
5BP1/ 1802-P1-2-4-5-11	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph	5″	2000	450	−40	—	—	500	—	0.3	0.33	—	Green White Blue	5BP1/ 1802-P1-2-4-5-11
							1500	337	−30	—	—			0.4	0.45			

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater Volts	Heater Amps.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Ion-Trap Ma.	Max. Input Voltage[1]	Focus Coil Ma.	Deflection Sensitivity[6] D1 D2	Deflection Sensitivity[6] D3 D4	Anode No. 3 Voltage	Pattern Color	Type
5CP1-2-4-5-7-11	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph Television	5″	2000	575	−60	—	—	550	—	0.28	0.32	4000	White Green Blue	5CP1-2-4-5-7-11
							1500	430	−45	—	—		—	0.37	0.43	3000		
							2000	575	−60	—	—		—	0.36	0.41	2000		
5FP1-2-4-11-14	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Oscillograph Television	5″	7000	250	−45	—	—	—	—	—	—	—	Green White Blue	5FP1-2-4-11-14
							4000	250	−45	—	—		—	—	—			
5HP1 5HP4[7]	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph	5″	2000	425	−40	—	—	500	—	0.3	0.33	—	Green	5HP1
							1500	310	−30					0.4	0.44	—	White	5HP4
5JP1-2-4-5-11	Electrostatic Cathode-Ray	11E	6.3	0.6	Oscillograph	5″	2000	520	−75	—	—	500	—	0.25	0.28	4000	White Green Blue	5JP1-2-4-5-11
							1500	390	−56	—	—		—	0.33	0.37	3000		
5LP1-2-4-5-11	Electrostatic Cathode-Ray	11F	6.3	0.6	Oscillograph Television	5″	2000	500	−60	—	—	500	—	0.25	0.28	4000	White Green Blue	5LP1-2-4-5-11
							1500	375	−45	—	—		—	0.33	0.37	3000		
							1000	250	−30	—	—		—	0.49	0.56	2000		
5MP1-4-5-11	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	5″	1500	375	−50	—	—	660	—	0.39	0.42	—	White Green Blue	5MP1-4-5-11
							1000	250	−33	—	—		—	0.58	0.64			
5RP1-2-4-7-11	Electrostatic Cathode-Ray	14F	6.3	0.6	Oscillograph	5″	3000	—	−90	—	—	1200	—	0.12	0.12	15000	Green White Blue	5RP1-2-4-7-11
							2000	575	−60	—	—		—	0.18	0.18	10000		
5TP4	Projection Kinescope	12C	6.3	0.6	Television	5″	27000	4900	−70	200	—	—	—	—	—	—	White	5TP4
5UP1-7-11	Electrostatic Cathode-Ray	12E	6.3	0.6	Oscillograph	5″	2500	640	−90	—	—	500	—	38.5[3]	77[3]	—	Green Yel-low Blue	5UP1-7-11
							2500	340	−90	—	—	500	—	28[3]	56[3]	—		
							1000	320	−45	—	—	500	—	31[3]	62[3]	—		
							1000	170	−45	—	—	500	—	23[3]	46[3]	—		
5WP11	Transcriber Kinescope	12C	6.3	0.6	Television	5″	27000	5400	−42/−98	200	—	—	—	—	—	—	Blue	5WP11
5WP15	Flying-Spot Cathode-Ray	12C	6.3	0.6	Vid. Sig. Gen.	5″	20000	3000/3800	−42/−98	200	—	—	—	—	—	—	Blue Green	5WP15
5ZP16	Flying-Spot Cathode-Ray	Fig. 46	6.3	0.6	Vid. Sig. Gen.	5″	20000	4700	−70	200	—	—	—	—	—	—	—	5ZP16
7AP4	Electromagnetic Picture Tube	5AJ	2.5	2.1	Television	7″	3500	1000	−67.5	—	—	—	—	—	—	—	White	7AP4
7BP1-[7] 2-4-7-11	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Oscillograph Television	7″	7000	250	−45	—	—	—	—	—	—	—	White Green Blue	7BP1-2-4-7-11
							4000	250	−45	—	—							
7CP1/[5] 1811-P1	Electromagnetic Cathode-Ray	6AZ	6.3	0.6	Oscillograph	7″	7000	1470	−45	250	—	—	—	—	—	—	Green	7CP1/1811-P1
							4000	840	−45	250	—			—	—			
7DP4	Kinescope	12C	6.3	0.6	Television	7″	6000	1430	−45	250	—	—	—	—	—	—	White	7DP4
7EP4	Electrostatic Cathode-Ray	11N	6.3	0.6	Television	7″	2500	650	−60	—	—	—		110[3]	95[3]	—	White	7EP4
7GP4[5]	Electrostatic Kinescope	Fig. 47	6.3	0.6	Television	7″	3000	1200	−84	3000	—	—	—	123[3]	102[3]	—	White	7GP4
7JP1	Electrostatic Cathode-Ray	14G	6.3	0.6	Oscillograph	7″	2000	800	−56	—	—	—	—	62/82[3]	50/68[3]	—	Green	7JP1
							4000	1600	−112	—	—	—	—	124/164[3]	100/136[3]	—		
7JP4	Electrostatic Kinescope	14G	6.3	0.6	Television	7″	6000	2400	−168					246[3]	204[3]	—	White	7JP4
7MP7	Electromagnetic Cathode-Ray	12D	6.3	0.6	Oscillograph Radar	7″	—	7000	−27/−63	250	—	—	85	—	—	—	Gr'nish-Yellow	7MP7
							—	4000	−27/−63	250	—	—	62	—	—	—		
7NP4	Projection Kinescope	14N	6.6	0.62	Television	7″	75000	16000/18000	−155	400/600	—	—	—	—	—	—	White	7NP4
7QP4	Electromagnetic Kinescope	12D	6.3	0.6	Monitor	7″	—	912/1368	−67.5	250	—	—	—	—	—	6000	White	7QP4
7RP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	7″	—	9000	−27/−63	250	—	—	120	—	—	—	White	7RP4
7TP4	Monitor Kinescope	12C	6.3	0.6	Television	7″	10000	1040/1400	−22/−52	200	0/8[8]	—	—	—	—	—	White	7TP4
7VP1	Electrostatic Cathode-Ray	14A	6.3	0.6	Oscillograph	7″	3000	800/1200	−84	—	—	—	—	93/123[3]	75/102[3]	—	Green	7VP1
7WP4	Projection Kinescope	14N	6.6	0.62	Television	7″	75000	8000	−155	400/600	—	—	—	—	—	—	White	7WP4

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater Volts	Heater Amps.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Ion-Trap Ma.	Max. Input Voltage[1]	Focus Coil Ma.	Deflection Sensitivity[6] D1 D2	Deflection Sensitivity[6] D3 D4	Anode No. 3 Voltage	Pattern Color	Type
8AP4	Electromagnetic Picture Tube	12H	6.3	0.6	Television	8″	—	7000	−27/−63	—	45[8]	—	115	—	—	—	White	8AP4
8BP4	Electrostatic Picture Tube	14G	6.3	0.6	Television	8″	— —	2400 1620	−72/−168 −72/−168	6000 6000	— —	— —	— —	146/198[3]	124/168[3]	— —	White	8BP4
9AP4/ 1804-P4	Electromagnetic Kinescope	6AL	2.5	2.1	Television	9″	7000 6000	1425 1225	− 40 − 38	250	—	—	—	—	—	—	White	9AP4/ 1804-P4
9CP4	Electromagnetic Kinescope	4AF	2.5	2.1	Television	9″	7000	—	−110	—	—	—	—	—	—	—	White	9CP4
9JP1/ 1809-P1	Electrostatic-Magnetic Cathode-Ray	8BR	2.5	2.1	Oscillograph	9″	5000 2500	1570 785	− 90 − 45	—	—	3000	—	0.136 0.272	— —	—	Green	9JP1/ 1809-P1
10BP4	Magnetic Kinescope	12D	6.3	0.6	Television	10″	—	9000	− 45	250	—	—	—	—	—	—	White	10BP4
10EP4	Magnetic-Focus Cathode-Ray	12D	6.3	0.6	Television	10½″	—	8000	− 45	250	—	—	—	—	—	—	White	10EP4
10FP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	10″	—	9000	−27/−63	250	—	—	—	—	—	—	White	10FP4
10HP4	Electrostatic Cathode-Ray	14G	6.3	0.6	Television	10″	—	5000	−60/−140	1800	—	—	—	130[3]	100[3]	—	White	10HP4
10KP7	Magnetic Cathode-Ray	12D	6.3	0.6	Oscillograph	10″	—	9000	−27/−63	250	—	—	—	—	—	—	—	10KP7
10SP4	Monitor Kinescope	12C	6.3	0.6	Television	10″	14000	1640/ 2225	−18/−48	200	—	—	—	—	—	—	White	10SP4
12AP4/ 1803-P4	Electromagnetic Picture Tube	6AL	2.5	2.1	Television	12″	7000 6000	1460 1240	− 75	250	25	—	10	—	—	—	White	12AP4/ 1803-P4
12CP4[7]	Electromagnetic Picture Tube	4AF	2.5	2.1	Television	12″	7000	—	−110	—	25	—	10	—	—	—	White	12CP4
12DP4-7	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Television	12″	7000 4000	250 250	− 45 − 45	— —	— —	— —	— —	— —	— —	— —	White	12DP4
12KP4-A	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	12″	—	11000	−27/−63	250	—	—	—	—	—	—	White	12KP4-A
12LP4[7]	Electromagnetic Kinescope	12D	6.3	0.6	Television	12″	—	11000	−27/−63	250	—	—	—	—	—	—	White	12LP4
12QP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	12″	—	10000	−27/−63	250	80	—	135	—	—	—	White	12QP4
12RP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	12″	—	10000	−27/−63	250	52[8]		135	—	—	—	White	12RP4
12SP7	Electromagnetic Cathode-Ray	12D	6.3	0.6	Oscillograph	12″	—	10000	−27/−63	250	—	—	107	—	—	—	Gr'nish-Yellow	12SP7
12TP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	12″	—	11000	−27/−63	250	120	—	110	—	—	—	White	12TP4
12UP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	12″	—	11000	−27/−63	250	—	—	110	—	—	—	White	12UP4
14BP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	14″	—	11000	−27/−63	250	120	—	110	—	—	—	White	14BP4
14CP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	14″	—	12000	−33/−77	250	32[8]	—	105	—	—	—	White	14CP4
14DP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	14″	—	11000	−27/−63	250	120	—	100	—	—	—	White	14DP4
14EP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	14″	—	12000	−33/−77	—	110	—	110	—	—	—	White	14EP4
14GP4	Electrostatic-Magnetic Kinescope	Fig. 42	6.3	0.6	Television	14″	—	12000	−33/−77	300						2940[2]	White	14GP4
14HP4	Electrostatic-Magnetic Kinescope	Fig. 43	6.3	0.6	Television	14″	12000	−48/264	−33/−77	300	70	—	—	—	—	—	White	14HP4
15AP4	Electromagnetic Cathode-Ray	12D	6.3	0.6	Television	15″	—	8000	− 45	250	—	—	—	—	—	—	White	15AP4
15CP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	15″	—	9000	− 45	250	109	—	115	—	—	—	White	15CP4
15DP4[7]	Electromagnetic Picture Tube	12D	6.3	0.6	Television	15″	—	13000	−27/−63	250	105	—	146	—	—	—	White	15DP4
16ADP4	Electromagnetic Cathode-Ray	Fig. 69	6.3	0.6	Oscillograph	16″	—	12000	−27/−63	250	—	—	—	—	—	—	Gr'nish-Yellow	16ADP4
16AP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	12000	−33/−77	300	—	—	—	—	—	—	White	16AP4
16CP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	12000	−27/−63	250	120	—	110	—	—	—	White	16CP4
16EP4A	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	−33/−77	300	—	—	105	—	—	—	White	16EP4A
16FP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	13000	−27/−63	250	105	—	146	—	—	—	White	16FP4
16GP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	−33/−77	300	23[8]	—	100	—	—	—	White	16GP4
16GP4B	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	−33/−77	300	35[8]	—	100	—	—	—	White	16GP4B
16GP4C	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	−33/−77	300	45[8]	—	100	—	—	—	White	16GP4C
16HP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	12000	−33/−77	300	120	—	110	—	—	—	White	16HP4
16JP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	11000	−27/−63	250	120	—	115	—	—	—	White	16JP4
16KP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	14000	−33/−77	300	30[8]	—	90	—	—	—	White	16KP4

Type	Name	Socket Connec-tions	Heater Volts	Heater Amps.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Ion-Trap Ma.	Max. Input Voltage[1]	Focus Coil Ma.	Deflection Sensitivity[6] $D_1 D_2$	Deflection Sensitivity[6] $D_3 D_4$	Anode No. 3 Voltage	Pattern Color	Type
16LP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	12000	-33/-77	300	120	—	110	—	—	—	White	16LP4
16MP4	Electromagnetic Kinescope	12D	6.3	0.6	Television	16″	—	12000	-33/-77	300	120	—	110	—	—	—	White	16MP4
16RP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-33/-77	300	120	—	100	—	—	—	White	16RP4
16SP4A	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-33/-77	300	120	—	110	—	—	—	White	16SP4A
16TP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	12000	-33/-77	300	45[8]	—	115	—	—	—	White	16TP4
16UP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-27/-63	300	23[8]	—	100	—	—	—	White	16UP4
16VP4	Electromagnetic Kinescope	12D	6.3	0.6	Television	16″	—	12000	-27/-63	250	120	—	110	—	—	—	White	16VP4
16WP4A	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-27/-63	250	120	—	110	—	—	—	White	16WP4A
16ZP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-33/-77	300	120	—	110	—	—	—	White	16ZP4
17AP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	17″	—	12000	-33/-77	300	75	—	100	—	—	—	White	17AP4
17BP4A	Electromagnetic Kinescope	Fig. 45	6.3	0.6	Television	17″	—	14000	-33/-77	300	50[8]	—	99	—	—	—	White	17BP4A
17BP4B	Electromagnetic Picture Tube	12D	6.3	0.6	Television	17″	—	12000	-33/-77	300	35[8]	—	100	—	—	—	White	17BP4B
17CP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	17″	—	14000	-33/-77	300	50[8]	—	104	—	—	—	White	17CP4
17FP4	Electrostatic-Magnetic Kinescope	Fig. 42	6.3	0.6	Television	17″	16000	3100/ 4100	-33/-77	300	40[8]	—	—	—	—	—	White	17FP4
17GP4	Electrostatic-Magnetic Kinescope	Fig. 43	6.3	0.6	Television	17″	—	14000	-33/-77	300	40[8]	—	—	—	—	3620[2]	White	17GP4
17HP4	Electrostatic-Magnetic Kinescope	Fig. 42	6.3	0.6	Television	17″	14000	0–350	-33/-77	300	85	—	—	—	—	—	White	17HP4
17JP4	Electromagnetic Kinescope	Fig. 45	6.3	0.6	Television	17″	—	16000	-33/-77	300	45[8]	—	—	—	—	—	White	17JP4
17KP4	Electrostatic-Magnetic Kinescope	Fig. 45	6.3	0.6	Television	17″	—	12000	-33/-77	300	0/8[8]	—	—	—	—	—	White	17KP4
17LP4	Electrostatic-Magnetic Kinescope	Fig. 42	6.3	0.6	Television	17″	—	16000	-33/-77	300	50[8]	—	—	—	—	—	White	17LP4
17QP4	Electromagnetic Kinescope	12D	6.3	0.6	Television	17″	—	12000	-33/-77	300	35[8]	—	100	—	—	—	White	17QP4
17RP4	Electrostatic-Magnetic Kinescope	Fig. 66	6.3	0.6	Television	17″	14000	0	-33/-77	300	35[8]	—	—	—	—	—	White	17RP4
17SP4	Electrostatic-Magnetic	12D	6.3	0.6	Television	17″	—	13000	-33/-66	250	120	—	—	—	—	—	White	17SP4
17YP4	Electromagnetic Kinescope	Fig. 45	6.3	0.6	Television	17″	—	12000	-33/-77	300	35[8]	—	92	—	—	—	White	17YP4
19AP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	13000	-27/-63	250	105	—	146	—	—	—	White	19AP4
19AP4A	Electromagnetic Picture Tube	12D	6.3	0.6	Television	16″	—	12000	-33/-77	300	75	—	140	—	—	—	White	19AP4A
19DP4A	Electromagnetic Picture Tube	12D	6.3	0.6	Television	19″	—	13000	-26/-63	250	105	—	146	—	—	—	White	19DP4A
19EP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	19″	—	13000	-26/-63	250	105	—	146	—	—	—	White	19EP4
19FP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	16″	—	13000	-27/-68	250	100	—	100/130	—	—	—	White	19FP4
19GP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	19″	—	13000	-27/-63	250	105	—	110/130	—	—	—	White	19GP4
19JP4	Electromagnetic Kinescope	12D	6.3	0.6	Television	19″	—	12000	-33/-77	300	75	—	95	—	—	—	White	19JP4
20BP4	Electromagnetic Cathode-Ray	12D	6.3	0.6	Television	20″	—	15000	-45	250	—	—	—	—	—	—	White	20BP4
20CP4	Electromagnetic Picture Tube	Fig. 44	6.3	0.6	Television	20″	—	12000	-33/-77	300	75	—	95	—	—	—	White	20CP4
20CP4A	Electromagnetic Kinescope	Fig. 44	6.3	0.6	Television	20″	—	12000	-33/-77	300	75	—	95	—	—	—	White	20CP4A
20DP4	Electromagnetic Kinescope	Fig. 44	6.3	0.6	Television	20″	—	12000	-33/-77	300	75	—	95	—	—	—	White	20DP4
20FP4	Electrostatic-Magnetic Kinescope	Fig. 66	6.3	0.6	Television	20″	12000	2300/ 3100	-33/-77	300	75	—	—	—	—	—	White	20FP4
20GP4	Electrostatic-Magnetic Kinescope	Fig. 42	6.3	0.6	Television	20″	—	16000	-33/-77	300	40[8]					4270[2]	White	20GP4
20HP4	Electrostatic-Magnetic Kinescope	Fig. 66	6.3	0.6	Television	20″	14000	-56/310	-33/-77	300	85	—	—	—	—	—	White	20HP4
20JP4	Electrostatic-Magnetic Kinescope	Fig. 45	6.3	0.6	Television	20″	—	12000	-33/-77	300	0/8[8]	—	—	—	—	—	White	20JP4
20LP4	Electrostatic-Magnetic Kinescope	Fig. 43	6.3	0.6	Television	20″	14000	0	-33/-77	300	35[8]	—	—	—	—	—	White	20LP4
20MP4	Electrostatic-Magnetic Kinescope	Fig. 42	6.3	0.6	Television	20″	—	16000	-33/-77	300	50[8]	—	—	—	—	—	White	20MP4
21AP4	Electromagnetic Kinescope	Fig. 44	6.3	0.6	Television	21″	—	16000	-33/-77	300	50[8]	—	110	—	—	—	White	21AP4
21EP4A	Electromagnetic Kinescope	Fig. 44	6.3	0.6	Television	21″	—	12000	-33/-77	300	70	—	95	—	—	—	White	21EP4A
21FP4A	Electrostatic-Magnetic Kinescope	Fig. 43	6.3	0.6	Television	21″	14000	±200	-33/-77	300	40[8]	—	—	—	—	—	White	21FP4A
21KP4A	Electrostatic-Magnetic Kinescope	Fig. 45	6.3	0.6	Television	21″	—	12000	-33/-77	300	50	—	—	—	—	—	White	21KP4A
21MP4	Electrostatic-Magnetic Kinescope	Fig. 43	6.3	0.6	Television	21″	—	16000	-33/-77	300	50[8]	—	—	—	—	—	White	21MP4
22AP4	Electromagnetic Picture Tube	Fig. 35	6.3	0.6	Television	22″	—	14000	-33/-77	300	35[8]	—	117	—	—	—	White	22AP4

TABLE XIV—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater		Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Ion-Trap Ma.	Max. Input Voltage[1]	Focus Coil Ma.	Deflection Sensitivity[6]		Anode No. 3 Voltage	Pattern Color	Type
			Volts	Amp.										$D_1 D_2$	$D_3 D_4$			
24AP4A	Electromagnetic Picture Tube	12D	6.3	0.6	Television	24"	—	12000	-33/-77	300	32[8]	—	97	—	—	—	White	24AP4A
24BP4	Electrostatic-Magnetic Kinescope	Fig. 43	6.3	0.6	Television	24"	14000	-56/310	-33/-77	300	85	—	—	—	—	—	White	24BP4
27AP4	Electrostatic-Magnetic Kinescope	Fig. 43	6.3	0.6	Television	27"	15000	-60/300	-33/-77	300	85	—	—	—	—	—	White	27AP4
27EP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	27"	——	14000	-33/-77	300	85	—	110	—	—	—	White	27EP4
30BP4	Electromagnetic Picture Tube	12D	6.3	0.6	Television	30"	——	12000	-33/-77	300	75	—	95	—	—	—	White	30BP4
902[7]	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	2"	600	150	– 60	—	—	350	—	0.19	0.22	—	Green	902
903[5]	Electromagnetic Cathode-Ray	6AL	2.5	2.1	Oscillograph	9"	7000	1360	–120	250	—	—	—	—	—	—	Green	903
904	Electrostatic-Magnetic Cathode-Ray	Fig. 3	2.5	2.1	Oscillograph	5"	4600	970	– 75	250	—	4000	—	0.09	—	—	Green	904
905[7]	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5"	2000	450	– 35	—	—	1000	—	0.19	0.23	—	Green	905
907	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5"	Characteristics same as Type 905							—	—	—	Blue	907
908[7]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3"	Characteristics same as Type 3AP1/906P1							—	—	—	Blue	908
908-A	Electrostatic Cathode-Ray	7CE	2.5	2.1	Oscillograph	3"	1500	430	– 50	——		500	—	0.223	0.233	—	Blue	908-A
							1000	287	– 33	—		500	—	0.334	0.348	—		
909[5]	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5"	Characteristics same as Type 905							—	—	—	Blue	909
910[5]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3"	Characteristics same as Type 3AP1/906P1							—	—	—	Blue	910
911[5]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3"	Characteristics same as Type 3AP1/906P1							—	—	—	Green	911
912	Electrostatic Cathode-Ray	Fig. 8	2.5	2.1	Oscillograph	5"	10000	2000	– 66	250	—	7000	—	0.041	0.051	—	Green	912
913	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	1"	500	100	– 65	—	—	250	—	0.07	0.10	—	Green	913
914[7]	Electrostatic Cathode-Ray	Fig. 12	2.5	2.1	Oscillograph	9"	7000	1450	– 50	250	—	3000	—	0.073	0.093	—	Green	914
1800[5]	Electromagnetic Kinescope	6AL	2.5	2.1	Television	9"	6000	1250	– 75	250	—	—	—	—	—	—	Yellow	1800
1801[5]	Electromagnetic Kinescope	Fig. 13	2.5	2.1	Television	5"	3000	450	– 35	—	—	—	—	—	—	—	Yellow	1801
1816P4-A	Electromagnetic Kinescope	Fig. 65	6.3	0.6	Monitor	10"		9000	– 63	250	—	—	—	—	—	—	White	1816P4-A
2001	Electrostatic Cathode-Ray	4AA	6.3	0.6	Oscillograph	1"	Characteristics essentially same as 913											2001
2002	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	2"	600	120	—	—	—	—	—	0.16	0.17	—	Green	2002
2005	Electrostatic Cathode-Ray	Fig. 1[4]	2.5	2.1	Television	5"	2000	1000	– 35	200	—	—	—	0.5	0.56	—	—	2005
24-XH	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscilloscope	2"	600	120	– 60	—	—	—	—	0.14	0.16	—	Blue	24-XH

[1] Between Anode No. 2 and any deflecting plate.
[2] Grid No. 4 voltage.
[3] D.c. Volts/in.
[4] Cathode connected to Pin 7.
[5] Discontinued.
[6] In mm./volt d.c.
[7] Superseded by same type with suffix "A."
[8] Ion-trap gausses.

TABLE XV—RECTIFIERS—RECEIVING AND TRANSMITTING

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
BA	Full-Wave Rectifier	4-pin M.	4J	Cold	—	—	350	350	Tube drop 80 v.		G
BH	Full-Wave Rectifier	4-pin M.	4J	Cold	—	—	350	125	Tube drop 90 v.		G
BR	Half-Wave Rectifier	4-pin M.	4H	Cold	—	—	300	50	Tube drop 60 v.		G
CE-220	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	3.0	—	20	20000	100	HV
OY4	Half-Wave Rectifier	5-pin O.	4BU	Cold	Connect Pins 7 and 8		95	75	300	500	G
OZ4	Full-Wave Rectifier	5-pin O.	4R	Cold	—	—	350	30–75	1250	200	G
1	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	1000	400	MV
1AX2	Half-Wave Rectifier	9-pin B.	9Y	Fil.	1.4	0.65	20000	1.0	25000	11	HV
1-V	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	—	—	HV
1V2	Half-Wave Rectifier	9-pin B.	9U	Fil.	.625	0.3	—	0.5	7500	10	HV
1B3GT/8016	Half-Wave Rectifier	6-pin O.	3C	Fil.	1.25	0.2	—	2.0	4000	17	HV
1B48	Half-Wave Rectifier	7-pin B.	—	Cold	—	—	800	6	2700	50	G
1X2	Half-Wave Rectifier	9-pin B.	9Y	Fil.	1.25	0.2	—	1	15000	10	HV
1X2A	Half-Wave Rectifier	9-pin B.	9Y	Fil.	1.25	0.2	—	1.1	20000	11	HV
1X2B	Fly-Back Rectifier	9-pin B.	9Y	Fil.	1.25	0.2	—	0.5	22000	45	HV
1Y2	Half-Wave Rectifier	4-pin M.	4P	Fil.	1.5	0.29	—	2	50000	10	HV
1Z2	Half-Wave Rectifier	7-pin B.	7CB	Fil.	1.5	0.3	7800	2	20000	10	HV
2B25	Half-Wave Rectifier	7-pin B.	3T	Fil.	1.4	0.11	1000	1.5	—	9	HV
2V3G	Half-Wave Rectifier	6-pin O.	4Y	Fil.	2.5	5.0	—	2.0	16500	12	HV
2W3	Half-Wave Rectifier	5-pin O.	4X	Fil.	2.5	1.5	350	55	—	—	HV
2X2/879[10]	Half-Wave Rectifier	4-pin S.	4AB	Htr.	2.5	1.75	4500	7.5	—	—	HV
2X2-A	Half-Wave Rectifier	4-pin S.	4AB	Same as 2X2/879 but will withstand severe shock & vibration							HV
2Y2	Half-Wave Rectifier	4-pin M.	4AB	Fil.	2.5	1.75	4400	5.0	—	—	HV
2Z2/G84	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	1.5	350	50	—	—	HV
3B24	Half-Wave Rectifier	4-pin M.	T-4A	Fil.	5.0 2.5[9]	3.0 3.0	— —	60 30	20000 20000	300 150	HV
3B25	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	500	4500	2000	G
3B26	Half-Wave Rectifier	8-pin O.	Fig. 31	Htr.	2.5	4.75	—	20	15000	8000	HV
DR-3B27	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	3000	250	8500	1000	HV
3B28	Half-Wave Rectifier	4-pin-M	4P	Fil.	2.5	5.0	1700 3500	500 250	5000 10000	2000 1000	G
5AX4GT	Full-Wave Rectifier	5-pin O.	5T	Fil.	5	2.5	350[4] 500[7]	175	1400	525	HV
5AZ4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0		Same as Type 80			HV
5R4GY	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	900[4] 950[7]	150[4] 175[7]	2800	650	HV
5T4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	3.0	450	250	1250	800	HV
5U4G	Full-Wave Rectifier	8-pin O.	5T	Fil.	5.0	3.0		Same as Type 5Z3			HV
5V4G	Full-Wave Rectifier	8-pin O.	5L	Htr.	5.0	2.0		Same as Type 83V			HV
5W4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	1.5	350	110	1000	—	HV
5X3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	2.0	1275	30	—	—	HV
5X4G	Full-Wave Rectifier	8-pin O.	5Q	Fil.	5.0	3.0		Same as 5Z3			HV
5Y3G	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0		Same as Type 80			HV
5Y3WGT	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	375	120	1550	375	HV
5Y4G	Full-Wave Rectifier	8-pin O.	5Q	Fil.	5.0	2.0		Same as Type 80			HV
5Z3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	3.0	500	250	1400	—	HV
5Z4	Full-Wave Rectifier	5-pin O.	5L	Htr.	5.0	2.0	400	125	1100	—	HV
6AU4GT	Damper Diode	6-pin O.	4CG	Htr.	6.3	1.8	—	175	4500	1050	HV
6AX4GT	Damper Diode	6-pin O.	4CG	Htr.	6.3	1.2	—	125	4000	600	HV
6AX5GT	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	1.2	450	125	1250	375	HV
6AX6G	Full-Wave Rectifier	7-pin O.	7Q	Htr.	6.3	2.5	350	250	1250	600	HV
6BY5G	Full-Wave Rectifier	7-pin O.	6CN	Htr.	6.3	1.6	375[4]	175	1400	525	HV
6U3	Damper Diode	9-pin B.	9BM	Htr.	6.3	0.9	—	180	4000	400	HV
6U4GT	Half-Wave Rectifier	5-pin O.	4CG	Htr.	6.3	1.2	—	138	1375	660	HV
6V3	Half-Wave Rectifier	9-pin B.	9BD	Htr.	6.3	1.75	350	125	6000	600	HV
6V4	Full-Wave Rectifier	9-pin B.	9M	Htr.	6.3	0.6	350	90	—	—	HV
6W4GT	Damper Service Half-Wave Rectifier	6-pin O.	4CG	Htr.	6.3	1.2	— 350	125 125	2000 1250	600 600	HV
6W5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.9	350	100	1250	350	HV
6X4 6X5	Full-Wave Rectifier	7-pin B. 6-pin O.	7CF 6S	Htr.	6.3	0.6	325[4] 450[7]	70	1250	210	HV
6Y3G	Half-Wave Rectifier	5-pin O.	4AC	Htr.	6.3	0.7	5000	7.5	—	—	HV
6Y5[10]	Full-Wave Rectifier	6-pin S.	6J	Htr.	6.3	0.8	350	50	—	—	HV
6Z3	Half-Wave Rectifier	4-pin M.	4G	Fil.	6.3	0.3	350	50	—	—	HV
6Z5[10]	Full-Wave Rectifier	6-pin S.	6K	Htr.	6.3	0.6	230	60	—	—	HV
6ZY5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.3	350	35	1000	150	HV
7X6	Full-Wave Rectifier	8-pin O.	7AJ	Htr.	6.3	1.2	235	150	700	450	HV
7Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	6.3	0.5	350	60	—	—	HV
7Z4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	6.3	0.9	450[1] 325[4]	100	1250	300	HV
12A7	Rectifier-Pentode	7-pin S.	7K	Htr.	12.6	0.3	125	30	—	—	HV

TABLE XV—RECTIFIERS—RECEIVING AND TRANSMITTING—*Continued*

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
12AX4GT	Damper Diode	6-pin O.	4CG	Htr.	12.6	0.6	—	125	4000	600	HV
12X4	Full-Wave Rectifier	7-pin B.	5BS	Htr.	12.6	0.3	650[1] 900[7]	70 70	1250 1250	210 210	HV
12Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	12.6	0.3	250	60	—	—	HV
12Z5	Voltage Doubler	7-pin M.	7L	Htr.	12.6	0.3	225	60	—	—	HV
14Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	12.6	0.3	450[1] 325[4]	70	1250	210	HV
14Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	12.6	0.3	250	60	—	—	HV
17Z3	Damper Diode	9-pin B.	9CB	Htr.	17	0.3	—	150	4500	450	HV
19X3	Damper Diode	9-pin B.	9BM	Htr.	19	0.3	—	180	4500	400	HV
19Y3	Half-Wave Rectifier	9-pin B.	9BM	Htr.	19	0.3	—	180	700	—	HV
25A7G[10]	Rectifier-Pentode	8-pin O.	8F	Htr.	25	0.3	125	75	—	—	HV
25W4GT	Half-Wave Rectifier	6-pin O.	4CG	Htr.	25	0.3	350	125	1250	600	HV
25X6GT	Voltage Doubler	7-pin O.	7Q	Htr.	25	0.15	125	60	—	—	HV
25Y4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	25	0.15	125	75	—	—	HV
25Y5[10]	Voltage Doubler	6-pin S.	6E	Htr.	25	0.3	250	85	—	—	HV
25Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	25	0.3	250	50	—	—	HV
25Z4	Half-Wave Rectifier	6-pin O.	5AA	Htr.	25	0.3	125	125	—	—	HV
25Z5	Rectifier-Doubler	6-pin S.	6E	Htr.	25	0.3	125	100	—	500	HV
26Z5W	Full-Wave Rectifier	9-pin B.	9BS	Htr.	26.5	0.2	325[4] 450[7]	100 100	1250	300	HV
25Z6	Rectifier-Doubler	7-pin O.	7Q	Htr.	25	0.3	125	100	—	500	HV
28Z5	Full-Wave Rectifier	8-pin L.	5AB	Htr.	28	0.24	450[7] 325[4]	100	—	300	HV
32L7GT	Rectifier-Tetrode	8-pin O.	8Z	Htr.	32.5	0.3	125	60	—	—	HV
35W4	Half-Wave Rectifier	7-pin B.	5BQ	Htr.	35[2]	0.15	125	100[8]	330	600	HV
35Y4	Half-Wave Rectifier	8-pin O.	5AL	Htr.	35[2]	0.15	235	60 100[5]	700	600	HV
35Z3	Half-Wave Rectifier	8-pin L.	4Z	Htr.	35	0.15	250[5]	100	700	600	HV
35Z4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	35	0.15	250	100	700	600	HV
35Z5G	Half-Wave Rectifier	6-pin O.	6AD	Htr.	35[2]	0.15	125	60 100[8]	—	—	HV
35Z6G	Voltage Doubler	6-pin O.	7Q	Htr.	35	0.3	125	110	—	500	HV
40Z5GT	Half-Wave Rectifier	6-pin O.	6AD	Htr.	40[2]	0.15	125	60 100[8]	—	—	HV
45Z3	Half-Wave Rectifier	7-pin B.	5AM	Htr.	45	0.075	117	65	350	390	HV
45Z5GT	Half-Wave Rectifier	6-pin O.	6AD	Htr.	45[2]	0.15	125	60 100[8]	—	—	HV
50AX6G	Full-Wave Rectifier	7-pin O.	7Q	Htr.	50	0.3	350	250	1250	600	HV
50X6	Voltage Doubler	8-pin L.	7AJ	Htr.	50	0.15	117	75	700	450	HV
50Y6GT	Full-Wave Rectifier	7-pin O.	7Q	Htr.	50	0.15	125	85	—	—	HV
50Y7GT	Voltage Doubler	8-pin L.	8AN	Htr.	50[2]	0.15	117	65	700		HV
50Z6G	Voltage Doubler	7-pin O.	7Q	Htr.	50	0.3	125	150	—	—	HV
50Z7G[10]	Voltage Doubler	8-pin O.	8AN	Htr.	50	0.15	117	65	—	—	HV
70A7GT	Rectifier-Tetrode	8-pin O.	8AB	Htr.	70	0.15	125[5]	60	—	—	HV
70L7GT	Rectifier-Tetrode	8-pin O.	8AA	Htr.	70	0.15	117	70	—	350	HV
72	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	3.0	—	30	20000	150	HV
73	Half-Wave Rectifier	8-pin O.	4Y	Fil.	2.5	4.5	—	20	13000	3000	HV
80	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	2.0	350[4] 500[7]	125 125	1400	375	HV
81	Half-Wave Rectifier	4-pin M.	4B	Fil.	7.5	1.25	700	85	—	—	HV
82	Full-Wave Rectifier	4-pin M.	4C	Fil.	2.5	3.0	500	125	1400	400	MV
83	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	3.0	500	250	1400	800	MV
83-V	Full-Wave Rectifier	4-pin M.	4AD	Htr.	5.0	2.0	400	200	1100	—	HV
84/6Z4	Full-Wave Rectifier	5-pin S.	5D	Htr.	6.3	0.5	350	60	1000	—	HV
117L7GT/ 117M7GT	Rectifier-Tetrode	8-pin O.	8AO	Htr.	117	0.09	117	75	—	—	HV
117N7GT	Rectifier-Tetrode	8-pin O.	8AV	Htr.	117	0.09	117	75	350	450	HV
117P7GT	Rectifier-Tetrode	8-pin O.	8AV	Htr.	117	0.09	117	75	350	450	HV
117Z3	Half-Wave Rectifier	7-pin B.	4BR	Htr.	117	0.04	117	90	330	—	HV
117Z4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	117	0.04	117	90	350	—	HV
117Z6GT	Voltage Doubler	7-pin O.	7Q	Htr.	117	0.075	235	60	700	360	HV
217-A[10]	Half-Wave Rectifier	4-pin J.	4AT	Fil.	10	3.25	—	—	3500	600	HV
217-C	Half-Wave Rectifier	4-pin J.	4AT	Fil.	10	3.25	—	—	7500	600	HV
2225	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	250	10000	1000	MV
249-B	Half-Wave Rectifier	4-pin M.	Fig. 53	Fil.	2.5	7.5	3180	375	10000	1500	MV
HK253	Half-Wave Rectifier	4-pin J.	4AT	Fil.	5.0	10	—	350	10000	1500	HV
705A RK-705A	Half-Wave Rectifier	4-pin W.	T-3AA	Fil.	2.5[9] 5.0	5.0 5.0	— —	50 100	35000 35000	375 750	HV
816	Half-Wave Rectifier	4-pin S.	4P	Fil.	2.5	2.0	2200	125	7500	500	MV
836	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	5.0	—	—	5000	1000	HV
866A/866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	3500	250	10000	1000	MV
866B	Half-Wave Rectifier	4-pin M.	4P	Fil.	5.0	5.0	—	—	8500	1000	MV
866 Jr.	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	2.5	1250	250[3]	—	—	MV

TABLE XV—RECTIFIERS—RECEIVING AND TRANSMITTING—*Continued*

See also Table XIII—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amp.	Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
HY866 Jr.	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	2.5	1750	250[3]	5000	—	MV
RK866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	3500	250	10000	1000	MV
871[10]	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	2.0	1750	250	5000	500	MV
878	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	7100	5	20000	—	HV
879	Half-Wave Rectifier	4-pin S.	4P	Fil.	2.5	1.75	2650	7.5	7500	100	HV
872A/872	Half-Wave Rectifier	4-pin J.	4AT	Fil.	5.0	7.5	—	1250	10000	5000	MV
975A 575A	Half-Wave Rectifier	4-pin J.	4AT	Fil.	5.0	10.0	—	1500	15000	6000	MV
OZ4A/ 1003	Full-Wave Rectifier	5-pin O.	4R	Cold	—	—	—	110	880	—	G
1005/ CK1005	Full-Wave Rectifier	8-pin O.	5AQ	Fil.	6.3	0.1	—	70	450	210	G
1006/ CK1006	Full-Wave Rectifier	4-pin M.	4C	Fil.	1.75	2.25	—	200	1600	—	G
CK1007	Full-Wave Rectifier	8-pin O.	T-9G	Fil.	1.0	1.2	—	110	980	—	G
CK1009/BA	Full-Wave Rectifier	4-pin M.	—	Cold	—	—	—	350	1000	—	G
1274	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.6	Same as 7Y4				HV
1275	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	1.75	Same as 5Z3				HV
1616	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	130	6000	800	HV
1641/ RK60	Full-Wave Rectifier	4-pin M.	T-4AG	Fil.	5.0	3.0	—	50	4500	—	HV
								250	2500	—	
1654	Half-Wave Rectifier	7-pin B.	2Z	Fil.	1.4	0.05	2500	1	7000	6	HV
5517	Half-Wave Rectifier	7-pin B.	5BU	Cold	—	—	1200	6	—	50	G
5690	Half-Wave Rectifier	8-pin O.	Fig. 74	Htr.	6.3	2.4	350	125[4]	1120	750	HV
					12.6	1.2	350	150[7]	1120	750	
5825	Half-Wave Rectifier	4-pin M.	4P	Fil.	1.6	1.25	—	2	60000	40	HV
5993	Full-Wave Rectifier	9-pin B.	Fig. 71	Htr.	6.3	0.8	—	—	1250	230	HV
6063/6X4	Full-Wave Rectifier	7-pin B.	5BS	Htr.	6.3	0.6	325[4]	70	—	—	HV
							450[7]	70	—	—	
6157	Half-Wave Rectifier	9-pin B.	Fig. 72	Htr.	6.3	0.8	500	75	—	—	HV
							350	125	—	—	
8008	Half-Wave Rectifier	4-pin[6]	Fig. 11	Fil.	5.0	7.5	—	1250	10000	5000	MV
8013A	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	20	40000	150	HV
8016	Half-Wave Rectifier	6-pin O.	4AC	Fil.	1.25	0.2	—	2.0	10000	7.5	HV
8020	Half-Wave Rectifier	4-pin M.	4P	Fil.	5.0	5.5	10000	100	40000	750	HV
					5.8	6.5	12500	100	40000	750	
RK19	Full-Wave Rectifier	4-pin M.	4AT	Htr.	7.5	2.5	1250	200[4]	3500	600	HV
RK21	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	4.0	1250	200[4]	3500	600	HV
RK22	Full-Wave Rectifier	4-pin M.	T-4AG	Htr.	2.5	8.0	1250	200[4]	3500	600	HV
RX21A	Half-Wave Rectifier	5-pin M.	—	Fil.	2.5	10.0	—	750	11000	—	MV

[1] With input choke of at least 20 henrys.
[2] Tapped for pilot lamps.
[3] Per pair with choke input.
[4] Condenser input.
[5] With 100 ohms min. resistance in series with plate; without series resistor, maximum r.m.s. plate rating is 117 volts.
[6] Same as 872A/872 except for heavy-duty push-type base. Filament connected to pins 2 and 3, plate to top cap.
[7] Choke input.
[8] Without panel lamp.
[9] Using only one-half of filament.
[10] Discontinued.

TABLE XVI—TRIODE TRANSMITTING TUBES

Type	Max. Plate Dissi-pation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
		Volts	Amp.					Grid to Fil.	Grid to Plate	Plate to Fil.											
958-A	0.6	1.25	0.1	135	7	1.0	12	0.6	2.6	0.8	500	A.	5BD	Class-C Amp.-Oscillator	135	−20	7	1.0	0.035	—	0.6
3B7[2]	—	1.4 2.8	0.22 0.11	180	25	—	20	1.4	2.6	2.6	125	O.	7AP	Class-C Amp. (Telegraphy)	180	0	25	—	—	—	2.8
RK24	1.5	2.0	0.12	180	20	6.0	8.0	3.5	5.5	3.0	125	S.	4D	Class-C Amp.-Oscillator	180	−45	16.5	6.0	0.5	—	2.0
6J6[2]	1.5	6.3	0.45	300	30	16	32	2.2	1.6	0.4	250	B.	7BF	Class-C Amp. (Telegraphy)[2]	150	−10	30	16	0.35	—	3.5
9002	1.6	6.3	0.15	250	8	2.0	25	1.2	1.4	1.1	250	B.	7TM	Class-C Amp.-Oscillator	180	−35	7	1.5	—	—	0.5
955	1.6	6.3	0.15	180	8	2.0	25	1.0	1.4	0.6	250	A.	5BC	Class-C Amp.-Oscillator	180	−35	7	1.5	—	—	0.5
HY114B	1.8	1.4	0.155	180	12	3.0	13	1.0	1.3	1.0	300	O.	2T	Class-C Amp.-Oscillator	180	−30	12	2.0	0.2	—	1.4[3]
														Class-C Amp. (Telephony)	180	−35	12	2.5	0.3	—	1.4[3]
3A5[2]	2.0	1.4 2.8	0.22 0.11	150	30	5.0	15	0.9	3.2	1.0	40	B.	7BC	Class-C Amp.-Oscillator[2]	150	−35	30	5.0	0.2	—	2.2
6F4	2.0	6.3	0.225	150	20	8.0	17	2.0	1.9	0.6	500	A.	7BR	Class-C Amp.-Oscillator	150	−15 550* 2000**	20	7.5	0.2	—	1.8
HY24	2.0	2.0	0.13	180	20	4.5	9.3	2.7	5.4	2.3	60	S.	4D	Class-C Amp. (Telegraphy)	180	−45	20	4.5	0.2	—	2.7
														Class-C Amp. (Telephony)	180	−45	20	4.5	0.3	—	2.5
RK33[1, 2]	2.5	2.0	0.12	250	20	6.0	10.5	3-2	3-2	2.5	60	S.	T-7DA	Class-C Amp.-Oscillator[2]	250	−60	20	6.0	0.54	—	3.5
12AU7[2]	2.75[6]	6.3	0.3	350	12[6]	3.5[6]	18	1.5	1.5	0.5	54	B.	9A	Class-C Amp.-Oscillator[2]	350	−100	24	7	—	—	6.0
6N4	3.0	6.3	0.2	180	12	—	32	3.1	2.35	0.55	500	B.	7CA	Class-C Amp.-Oscillator	180	—	—	—	—	—	—
6026	3.0	6.3	0.2	150	30	10	24	2.2	1.3	0.38	400	N.	—	Class-C Oscillator-400 Mc.	135	1300**	20	9.5	—	—	1.25
HY6J5GTX	3.5	6.3	0.3	330	20	4.0	20	4.2	3.8	5.0	60	O.	6Q	Class-C Amp.-Oscillator	330	−30	20	2.0	0.2	—	3.5
														Class-C Amp. (Telephony)	250	−30	20	2.5	0.3	—	2.5
2C22/7193	3.5	6.3	0.3	500	—	—	20	2.2	3.6	0.7	—	O.	4AM	Class-C Amp. (Telegraphy)	—	—	—	—	—	—	—
HY615 HY-E1148	3.5	6.3	0.175	300	20	4.0	20	1.4	1.6	1.2	300	O.	T-8AG	Class-C Amp.-Oscillator	300	−35	20	2.0	0.4	—	4.0[3]
														Class-C Amp. (Telephony)	300	−35	20	3.0	0.8	—	3.5[3]
GL-446A[1] GL-446B[1]	3.75	6.3	0.75	400	20	—	45	2.2	1.6	0.02	500	O.	Fig. 19	Class-C Amp.-Oscillator	250	—	—	—	—	—	—
GL-2C44[1] GL-464A[1]	5.0	6.3	0.75	500	40	—	—	2.7	2.0	0.1	500	O.	Fig. 17	Class-C Amp.-Oscillator	250	—	—	—	—	—	—
6C4	5.0	6.3	0.15	350	25	8.0	18	1.8	1.6	1.3	54	B.	6BG	Class-C Amp.-Oscillator	300	−27	25	7.0	0.35	—	5.5
1626	5.0	12.6	0.25	250	25	8.0	5.0	3.2	4.4	3.4	30	O.	6Q	Class-C Amp.-Oscillator	250	−70	25	5.0	0.5	—	4.0
2C21/ RK33[2]	5.0	6.3	0.6	250	40	12	—	1.6	1.6	2.0	—	S.	T-7DA	Class-C Amp.-Oscillator[2]	250	−60	40	12	1.0	—	7
2C36	5	6.3	0.4	1500[5]	—	—	25	1.4	2.4	0.36	1200	N.	Fig. 36	Plate-Pulsed 1000-Mc. Osc.	1000[5]	0	900[5]	—	—	—	200[5]
2C37 5766 5767	5	6.3	0.4	350	—	—	25	1.4	1.85	0.02	3300	N.	Fig. 36	1000-Mc. C.W. Oscillator	150	3000**	15	3.6	—	—	0.5
5764	5	6.3	0.4	1500[5]	11.5	—	25	1.4	1.85	0.02	3300	N.	Fig. 36	Plate-Pulsed 3300-Mc. Osc.	1000[5]	0	1300[5]	—	—	—	200[5]
5765	5	6.3	0.4	350	—	—	25	1.3	2.1	0.03	2900	N.	Fig. 36	1900-Mc. C.W. Oscillator	180	10000**	25	—	—	—	0.225
5794	—	6.0	0.16	—	—	—	—	—	—	—	—	N.	Fig. 36	Fixed Tuned Oscillator Approximately 1680 Mc.	85/108	—	—	—	—	—	—
5675	5	6.3	0.135	165	30	8	20	2.3	1.3	0.09	3000	N.	Fig. 36	Grounded-Grid Osc.	120	−8	25	4	—	—	0.05
6N7[2]	5.5[6]	6.3	0.8	350	30[6]	5.0[6]	35	—	—	—	10	O.	8B	Class-C Amp. Oscillator[2, 11]	350	−100	60	10	—	—	14.5
5876	6.25	6.3	0.135	300	25	—	56	2.5	1.4	0.035	1700	N.	Fig. 36	Grounded-Grid Oscillator	250	−2	23	3	—	—	0.75
														Frequency Multiplier	300	−70	17.3	7	—	—	2.0
2C40	6.5	6.3	0.75	500	25	—	36	2.1	1.3	0.05	500	O.	Fig. 19	Class-C Amp.-Oscillator	250	−5	20	0.3	—	—	0.075
5556	7.0	4.5	1.1	350	40	10	8.5	4.0	8.3	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	350	−80	35	2	0.25	—	6
														Class-C Amp. (Telephony	300	−100	30	2	0.3	—	4
5893	8.0	6.0	0.33	400	40	13	27	2.5	1.75	0.07	1000	—	Fig. 36	Class-C Amp. (Telegraphy)	350	−33	35	13	2.4	—	6.5
														Class-C Amp. (Telephony)	300	−45	30	12	2.0	—	6.5

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Interelectrode Capacitances (μμfd.) Grid to Plate	Interelectrode Capacitances (μμfd.) Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
2C43	12	6.3	0.9	500	40	—	48	2.9	1.7	0.05	1250	O.	Fig. 19	Class-C Amp.-Oscillator	470	—	38[7]	—	—	—	9[7]
2C26A	10	6.3	1.10	—	—	—	16.3	2.6	2.8	1.1	250	O.	4BB		—	—	—	—	—	—	—
2C34/RK34[2]	10	6.3	0.8	300	80	20	13	3.4	2.4	0.5	250	M.	T-7DC	Class-C Amp.-Oscillator[2]	300	−36	80	20	1.8	—	16
205D	14	4.5	1.6	400	50	10	7.2	5.2	4.8	3.3	6	M.	4D	Class-C Amp.-Oscillator	400	−112	45	10	1.5	—	10
														Class-C Amp. (Telephony)	350	−144	35	10	1.7	—	7.1
2C25	15	7.0	1.18	450	60	15	8.0	6.0	8.9	3.0	—	M.	4D	Class-C Amp.-Oscillator	450	−100	65	15	3.2	—	19
														Class-C Amp. (Telephony)	350	−100	50	12	2.2	—	12
10Y	15	7.5	1.25	450	65	15	8	4.1	7.0	3.0	8	M.	4D	Class-C Amp.-Oscillator	450	−100	65	15	3.2	—	19
														Class-C Amp. (Telephony)	350	−100	50	12	2.2	—	12
843	15	2.5	2.5	450	40	7.5	7.7	4.0	4.5	4.0	6	M.	5A	Class-C Amp.-Oscillator	450	−140	30	5.0	1.0	—	7.5
														Class-C Amp. (Telephony)	350	−150	30	7.0	1.6	—	5.0
RK59[2]	15	6.3	1.0	500	90	25	25	5.0	9.0	1.0	—	M.	T-4D	Class-C Amp.-Oscillator	500	−60	90	14	1.3	—	32
HY75A	15	6.3	2.6	450	90	25	9.6	1.8	2.6	1.0	175	O.	2T	Class-C Amp. (Telegraphy)	450	−140	90	20	5.2	—	26
														Class-C Amp. (Telephony)	400	−140	90	20	5.2	—	21
HY75	15	6.3	2.5	450	80	20	10	1.8	3.8	1.0	60	O.	2T	Class-C Amp.-Oscillator	450	−50	80	12	—	—	21[3]
														Class-C Amp. (Telephony)	450	−60	80	12	—	—	16[3]
1602[1]	15	7.5	1.25	450	60	15	8.0	4.0	7.0	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	450	−115	55	15	3.3	—	13
														Class-C Amp. (Telephony)	350	−135	45	15	3.5	—	8.0
														Class-B Amp. Audio[7]	425	−50	110[8]	260[9]	2.5[8]	8000	25
841	15	7.5	1.25	450	60	20	30	4.0	7.0	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	450	−34	50	15	1.8	—	15
														Class-C Amp. (Telephony)	350	−47	50	15	2.0	—	11
10[1] RK10[1]	15	7.5	1.25	450	65	15	8.0	3.0	8.0	4.0	— 60	M.	4D	Class-C Amp. (Telegraphy)	450	−100	65	15	3.2	—	19
														Class-C Amp. (Telephony)	350	−100	50	12	2.2	—	12
														Class-B Audio[7]	425	−50	55[8]	130[9]	2.5[8]	8000	25
RK100[1]	15	6.3	0.9	150	250	100	40	23	19	3.0	—	M.	T-6B	Class-C Oscillator	110	—	80	8.0	—	—	3.5
														Class-C Amplifier	110	—	185	40	2.1	—	12
TUF-20	20	6.3	2.75	750	75	20	10	1.8	3.6	0.095	250	O.	2T	Class-C Amp.-Oscillator	750	−150	75	20	1.5/2.5	—	40
1608	20	2.5	2.5	425	95	25	20	8.5	9.0	3.0	45	M.	4D	Class-C Amp. (Telegraphy)	425	−90	95	20	3.0	—	27
														Class-C Amp. (Telephony)	350	−80	85	20	3.0	—	18
														Class-B Amp. Audio[7]	425	−15	190[8]	130[9]	2.2[8]	4800	50
310	20	7.5	1.25	600	70	15	8.0	4.0	7.0	2.2	6	M.	4D	Class-C Amp. (Telegraphy)	600	−150	65	15	4.0	—	25
														Class-C Amp. (Telephony)	500	−190	55	15	4.5	—	18
703-A	20	1.2	4/4.5	350	75	12	8	0.9	1.1	0.6	1400	N.	—	Class-C Amplifier	350	−120	75	12	—	—	2/2.5
801-A/801	20	7.5	1.25	600	70	15	8.0	4.5	6.0	1.5	60	M.	4D	Class-C Amp. (Telegraphy)	600	−150	65	15	4.0	—	25
														Class-C Amp. (Telephony)	500	−190	55	15	4.5	—	18
														Class-B Amp. Audio[7]	600	−75	130	320[9]	3.0[8]	10000	45
HY801-A	20	7.5	1.25	600	70	15	8.0	4.5	6.0	1.5	60	M.	4D	Class-C Amp. (Telegraphy)	600	−200	70	15	4.0	—	30
														Class-C Amp. (Telephony)	500	−200	60	15	4.5	—	22
T20	20	7.5	1.75	750	85	25	20	4.9	5.1	0.7	60	M.	3G	Class-C Amp. (Telegraphy)	750	−85	85	18	3.6	—	44
														Class-C Amp. (Telephony)	750	−140	70	15	3.6	—	38
TZ20	20	7.5	1.75	750	85	30	62	5.3	5.0	0.6	60	M.	3G	Class-C Amp. (Telegraphy)	750	−40	85	28	3.75	—	44
														Class-C Amp. (Telephony)	750	−100	70	23	4.8	—	38
														Class-B Amp. Audio[7]	800	0	40/136	160[9]	1.8[8]	12000	70
15E	20	5.5	4.2	—	—	—	25	1.4	1.15	0.3	600	N.	T-4AF	Class-C Amp. (Telegraphy)	Characteristics similar to 25T						

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
		Volts	Amp.					Grid to Fil.	Grid to Plate	Plate to Fil.											
3-25A3 25T	25	6.3	3.0	2000	75	25	24	2.7	1.5	0.3	60	M.	3G	Class-C Amp.-Oscillator	2000	−130	63	18	4.0	—	100
															1500	− 95	67	13	2.2	—	75
															1000	− 70	72	9	1.3	—	47
														Class-B Amp. Audio [7]	2000	− 80	16/80	270 [9]	0.7 [8]	55500	110
3-25D3 3C24 24G	25	6.3	3.0	2000	75	25	23	2.0 1.7	1.6 1.5	0.2 0.3	150	S.	2D	Class-C Amp.-Oscillator	2000	−170	63	17	4.5	—	100
															1500	−110	67	15	3.1	—	75
															1000	− 80	72	15	2.6	—	47
														Class-B Audio [7]	2000	− 85	16/80	290 [9]	1.1 [8]	55500	110
3C28	25	6.3	3.0	2000	75	25	23	2.1	1.8	0.1	100	S.	Fig. 56	Class-C Amp.-Oscillator	Characteristics same as 3C24						
3C34	25	6.3	3.0	2000	75	25	23	2.5	1.7	0.4	60	S.	3G	Class-C Amp.-Oscillator	Characteristics same as 3C24						
RK11 [1]	25	6.3	3.0	750	105	35	20	7.0	7.0	0.9	60	M.	3G	Class C Amp. (Telegraphy)	750	−120	105	21	3.2	—	55
														Class-C Amp. (Telephony)	600	−120	85	24	3.7	—	38
RK12	25	6.3	3.0	750	105	40	100	7.0	7.0	0.9	60	M.	3G	Class-C Amp. (Telegraphy)	750	−100	105	35	5.2	—	55
														Class-C Amp. (Telephony)	600	−100	85	27	3.8	—	38
HK24	25	6.3	3.0	2000	75	30	25	2.5	1.7	0.4	60	S.	3G	Class-C Amp. (Telegraphy)	2000	−140	56	18	4.0	—	90
														Class-C Amp. (Telephony)	1500	−145	50	25	5.5	—	60
HY25	25	7.5	2.25	800	75	25	55	4.2	4.6	1.0	60	M.	3G	Class-C Amp. (Telegraphy)	750	− 45	75	15	2.0	—	42
														Class-C Amp. (Telephony)	700	− 45	75	17	5.0	—	39
8025	30	6.3	1.92	1000	65	—	18	2.7	2.8	0.35	500	M.	4AQ	Class-C Amp. (Grid. Mod.)	1000	−135	50	4	3.5	—	20
	20				65	20								Class-C Amp. (Telephony)	800	−105	40	10.5	1.4	—	22
	30				80	20								Class-C Amp. (Telegraphy)	1000	− 90	50	14	1.6	—	35
HY30Z [1]	30	6.3	2.25	850	90	25	87	6.0	4.9	1.0	60	M.	4BO	Class-C Amp.-Oscillator	850	− 75	90	25	2.5	—	58
														Class-C Amp. (Telephony)	700	− 75	90	25	3.5	—	47
HY31Z [2] HY1231Z [2]	30	6.3 12.6	3.5 1.7	500	150	30	45	5.0	5.5	1.9	60	M.	T-4D	Class-C Amp. (Telegraphy)	500	− 45	150	25	2.5	—	56
														Class-C Amp. (Telephony)	400	−100	150	30	3.5	—	45
316A VT-191	30	2.0	3.65	450	80	12	6.5	1.2	1.6	0.8	500	N.	—	Class-C Amp. (Telegraphy)	450	—	80	12	—	—	7.5
														Class-C Amp. (Telephony)	400	—	80	12	—	—	6.5
809	30	6.3	2.5	1000	125	—	50	5.7	6.7	0.9	60	M.	3G	Class-C Amp. (Telegraphy)	1000	− 75	100	25	3.8	—	75
														Class-C Amp. (Telephony)	750	− 60	100	32	4.3	—	55
														Class-B Amp. Audio [7]	1000	− 9	40/200	155 [9]	2.7 [8]	11600	145
1623	30	6.3	2.5	1000	100	25	20	5.7	6.7	0.9	60	M.	3G	Class-C Amp.-Oscillator	1000	− 90	100	20	3.1	—	75
														Class-C Amp. (Telephony)	750	−125	100	20	4.0	—	55
														Class-B Amp. Audio	1000	− 40	30/200	230 [9]	4.2 [8]	12000	145
53A	35	5.0	12.5	15000	—	—	35	3.6	1.9	0.4	—	N.	T-4B	Oscillator at 300 Mc.	Approximately 50 watts output						
RK30 [1]	35	7.5	3.25	1250	80	25	15	2.75	2.5	2.75	60	M.	2D	Class-C Amp. (Telegraphy)	1250	−180	90	18	5.2	—	85
														Class-C Amp. (Telephony)	1000	−200	80	15	4.5	—	60
800	35	7.5	3.25	1250	80	25	15	2.75	2.5	2.75	60	M.	2D	Class-C Amp. (Telegraphy)	1250	−175	70	15	4.0	—	65
														Class-C Amp. (Telephony)	1000	−200	70	15	4.0	—	50
														Class-B Amp. Audio [7]	1250	− 70	30/130	300 [9]	3.4 [8]	21000	106
1628 [1]	40	3.5	3.25	1000	60	15	23	2.0	2.0	0.4	500	N.	T-4BB	Class-C Amp.-Oscillator	1000	− 65	50	15	1.7	—	35
														Class-C Amp. (Telephony)	800	−100	40	11	1.6	—	22
														Grid-Modulated Amp.	1000	−120	50	3.5	5.0	—	20
8012 GL-8012-A	40	6.3	2.0	1000	80	20	18	2.7 2.7	2.8 2.5	0.35 0.4	500	N.	T-4BB	Class-C Amp.-Oscillator	1000	− 90	50	14	1.6	—	35
														Class-C Amp. (Telephony)	800	−105	40	10.5	1.4	—	22
														Grid-Modulated Amp.	1000	−135	50	4.0	3.5	—	20
RK18 [1]	40	7.5	3.0	1250	100	40	18	6.0	4.8	1.8	60	M.	3G	Class-C Amp. (Telegraphy)	1250	−160	100	12	2.8	—	95
														Class-C Amp. (Telephony)	1000	−160	80	13	3.1	—	64

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
RK31	40	7.5	3.0	1250	100	35	170	7.0	1.0	2.0	30	M.	3G	Class-C Amp. (Telegraphy)	1250	−80	100	30	3.0	—	90
														Class-C Amp. (Telephony)	1000	−80	100	28	3.5	—	70
HY40[1]	40	7.5	2.25	1000	125	25	25	6.1	5.6	1.0	60	M.	3G	Class-C Amp. (Telegraphy)	1000	−90	125	20	5.0	—	94
														Class-C Amp. (Telephony)	850	−90	125	25	5.0	—	82
														Grid-Modulated Amp.	1000	—	125	—	—	—	20
HY40Z[1]	40	7.5	2.6	1000	125	30	80	6.2	6.3	0.8	60	M.	3G	Class-C Amp. (Telegraphy)	1000	−27	125	25	5.0	—	94
														Class-C Amp. (Telephony)	850	−30	100	30	7.0	—	82
														Grid-Modulated Amp.	1000	—	60	—	—	—	20
T40	40	7.5	2.5	1500	150	40	25	4.5	4.8	0.8	60	M.	3G	Class-C Amp.-Oscillator	1500	−140	150	28	9.0	—	158
														Class-C Amp. (Telephony)	1250	−115	115	20	5.25	—	104
TZ40	40	7.5	2.5	1500	150	45	62	4.8	5.0	0.8	60	M.	3G	Class-C Amp.-Oscillator	1500	−90	150	38	10	—	165
														Class-C Amp. (Telephony)	1250	−100	125	30	7.5	—	116
														Class-B Amp. Audio[7]	1500	−9	250[8]	285[9]	6.0[8]	12000	250
HY57	40	6.3	2.25	850	110	25	50	4.9	5.1	1.7	60	M.	3G	Class-C Amp. (Telegraphy)	850	−48	110	15	2.5	—	70
														Class-C Amp. (Telephony)	700	−45	90	17	5.0	—	47
														Grid-Modulated Amp.	850	—	70	—	—	—	20
756[1]	40	7.5	2.0	850	110	25	8.0	3.0	7.0	2.7	—	M.	4D	Class-C Amplifier	850	—	110	25	—	—	—
830[1]	40	10	2.15	750	110	18	8.0	4.9	9.9	2.2	15	M.	4D	Class-C Amplifier	750	−180	110	18	7.0	—	55
														Grid-Modulated Amp.	1000	−200	50	2.0	3.0	—	15
3-50A4 35T	50	5.0	4.0	2000	150	50	39	4.1	1.8	0.3	100	M.	3G	Class-C Amp. (Telegraphy)	2000	−135	125	45	13	—	200
3-50D4 35TG								2.5	1.8	0.4	100	M.	2D	Class-C Amp. (Telephony)	1500	−150	90	40	11	—	105
														Class-B Amp. Audio[7]	2000	−40	4/167	255[9]	4.0[8]	27500	235
8010-R	50	6.3	2.4	1350	150	20	30	2.3	1.5	0.07	350	N.	—	Class-C Amplifier	—	—	—	—	—	—	—
RK32[1]	50	7.5	3.25	1250	100	25	11	2.5	3.4	0.7	100	M.	2D	Class-C Amp. (Telegraphy)	1250	−225	100	14	4.8	—	90
														Class-C Amp. (Telephony)	1000	−310	100	21	8.7	—	70
RK35[1]	50	7.5	4.0	1500	125	20	9.0	3.5	2.7	0.4	60	M.	2D	Class-C Amp. (Telegraphy)	1500	−250	115	15	5.0	—	120
														Class-C Amp. (Telephony)	1250	−250	100	14	4.6	—	93
														Grid-Modulated Amp.	1500	−180	37	—	2.0	—	25
RK37	50	7.5	4.0	1500	125	35	28	3.5	3.2	0.2	60	M.	2D	Class-C Amp. (Telegraphy)	1500	−130	115	30	7.0	—	122
														Class-C Amp. (Telephony)	1250	−150	100	23	5.6	—	90
														Grid-Modulated Amp.	1500	−50	50	—	2.4	—	26
3-50G2 UH50	50	7.5	3.25	1250	125	25	10.6	2.2	2.6	0.3	60	M.	2D	Class-C Amp. (Telegraphy)	1250	−225	125	20	7.5	—	115
														Class-C Amp. (Telephony)	1250	−325	125	20	10	—	115
														Grid-Modulated Amp.	1250	−200	60	2.0	3.0	—	25
UH51[1]	50	5.0	6.5	2000	175	25	10.6	2.2	2.3	0.3	60	M.	2D	Class-C Amp. (Telegraphy)	2000	−500	150	20	15	—	225
														Class-C Amp. (Telephony)	1500	−400	165	20	15	—	200
														Grid-Modulated Amp.	1500	−400	85	2.0	8.0	—	65
HK54	50	5.0	5.0	3000	150	30	27	1.9	1.9	0.2	100	M.	2D	Class-C Amp. (Telegraphy)	3000	−290	100	25	10	—	250
														Class-C Amp. (Telephony)	2500	−250	100	20	8.0	—	210
														Class-B Amp. Audio[7]	2500	−85	20/150	360[9]	5.0	40000	275
HK154[1]	50	5.0	6.5	1500	175	30	6.7	4.3	5.9	1.1	60	M.	2D	Class-C Amp. (Telegraphy)	1500	−590	167	20	15	—	200
														Class-C Amp. (Telephony)	1250	−460	170	20	12	—	162
														Grid-Modulated Amp.	1500	−450	52	—	5.0	—	28
HK158	50	12.6	2.5	2000	200	40	25	4.7	4.6	1.0	60	M.	2D	Class-C Amp.-Oscillator	2000	−150	125	25	6.0	—	200
														Class-C Amp. (Telephony)	2000	−140	105	25	5.0	—	170
WE304A[1] 304B	50	7.5	3.25	1250	100	25	11	2.0	2.5	0.7	100	M.	2D	Class-C Amp. (Telegraphy)	1250	−200	100	—	—	—	85
														Class-C Amp. (Telephony)	1000	−180	100	—	—	—	65

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
		Volts	Amp.					Grid to Fil.	Grid to Plate	Plate to Fil.											
356A	50	5.0	5.0	1500	120	35	50	2.25	2.75	1.0	60	N.	T-4BD	Class-C Amp. (Telegraphy)	1500	− 60	100	—	—	—	100
														Class-C Amp. (Telephony)	1250	−100	100	35	—	—	85
808	50	7.5	4.0	1500	150	35	47	5.3	2.8	0.15	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	125	30	9.5	—	140
														Class-C Amp. (Telephony)	1250	−225	100	32	10.5	—	105
														Class-B Amp. Audio [7]	1500	− 25	30/190	220 [9]	4.8 [8]	18300	185
834	50	7.5	3.1	1250	100	20	10.5	2.2	2.6	0.6	100	M.	2D	Class-C Amp. (Telegraphy)	1250	−225	90	15	4.5	—	75
														Class-C Amp. (Telephony)	1000	−310	90	17.5	6.5	—	58
841A [1]	50	10	2.0	1250	150	30	14.6	3.5	9.0	2.5	—	M.	3G	Class-C Amplifier	—	—	—	—	—	—	85
841SW	50	10	2.0	1000	150	30	14.6	—	9.0	—	—	M.	3G	Class-C Amplifier	—	—	—	—	—	—	—
T55	55	7.5	3.0	1500	150	40	20	5.0	3.9	1.2	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−170	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1500	−195	125	15	5.0	—	145
811	55	6.3	4.0	1500	150	50	160	5.5	5.5	0.6	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−113	150	35	8.0	—	170
														Class-C Amp. (Telephony)	1250	−125	125	50	11	—	120
														Class-B Amp. Audio [7]	1500	− 9	20/200	150 [9]	3.0 [8]	17600	220
812	55	6.3	4.0	1500	150	35	29	5.3	5.3	0.8	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−175	150	25	6.5	—	170
														Class-C Amp. (Telephony)	1250	−125	125	25	6.0	—	120
														Class-B Amp. Audio [7]	1500	− 45	50/200	232 [9]	4.7 [8]	18000	220
RK51	60	7.5	3.75	1500	150	40	20	6.0	6.0	2.5	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−250	150	31	10	—	170
														Class-C Amp. (Telephony)	1250	−200	105	17	4.5	—	96
														Grid-Modulated Amp.	1500	−130	60	0.4	2.3	—	128
RK52	60	7.5	3.75	1500	130	50	170	6.6	12	2.2	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−120	130	40	7.0	—	135
														Class-C Amp. (Telephony)	1250	−120	115	47	8.5	—	102
														Class-B Amp. Audio [7]	1250	0	40/300	180 [9]	7.5 [8]	10000	250
T-60	60	10	2.5	1600	150	50	20	5.5	5.2	2.5	60	M.	2D	Class-C Amp.-Oscillator	1500	−150	150	50	9.0	—	100
826	55	7.5	4.0	1000	140	40	31	3.0	2.9	1.1	250	N.	7BO	Class-C Amp.-Oscillator	1000	− 70	130	35	5.8	—	90
														Class-C Amp. (Telephony)	1000	−160	95	40	11.5	—	70
														Grid-Modulated Amp.	1000	−125	65	9.5	8.2	—	25
830B 930B	60	10	2.0	1000	150	30	25	5.0	11	1.8	15	M.	3G	Class-C Amp.-Oscillator	1000	−110	140	30	7.0	—	90
														Class-C Amp. (Telephony)	800	−150	95	20	5.0	—	50
														Class-B Amp. Audio [7]	1000	− 35	20/280	270 [9]	6.0 [8]	7600	175
811-A	65	6.3	4.0	1500	175	50	160	5.9	5.6	0.7	60	M.	3G	Class-C Amp. (Telegraphy)	1500	− 70	173	40	7.1	—	200
														Class-C Amp. (Telephony)	1250	−120	140	45	10.0	—	135
														Class-B Amp. Audio [7]	1500	− 4.5	32/313	170 [9]	4.4 [8]	12400	340
812-A	65	6.3	4.0	1500	175	35	29	5.4	5.5	0.77	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−120	173	30	6.5	—	190
														Class-C Amp. (Telephony)	1250	−115	140	35	7.6	—	130
														Class-B Audio [7]	1500	− 48	28/310	270 [9]	5.0	13200	340
HY51A [1] HY51B [1]	65	7.5 10	3.5 2.25	1000	175	25	25	6.5	7.0	1.1	60	M.	3G	Class-C Amp. (Telegraphy)	1000	− 75	175	20	7.5	—	131
														Class-C Amp. (Telephony)	1000	−67.5	130	15	7.5	—	104
														Grid-Modulated Amp.	1000	—	100	—	—	—	33
HY51Z [1]	65	7.5	3.5	1000	175	35	85	7.9	7.2	0.9	60	M.	4BO	Class-C Amp. (Telegraphy)	1000	−22.5	175	35	10	—	131
														Class-C Amp. (Telephony)	1000	− 30	150	35	10	—	104
														Grid-Modulated Amp.	1000	—	100	—	—	—	33
5514	65	7.5	3.0	1500	175	60	145	7.8	7.9	1.0	60	M.	4BO	Class-C Amp. (Telegraphy)	1500	−106	175	60	12	—	200
														Class-C Amp. (Telephony)	1250	− 84	142	60	10	—	135
														Class-B Audio	1500	−4.5	350 [8]	88 [8]	6.5 [8]	10500	400

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
UH35 [1]	70	5.0	4.0	1500	150	35	30	1.4	1.6	0.2	60	M.	3G	Class-C Amp. (Telegraphy)	1500	−170	150	30	7.0	—	170
														Class-C Amp. (Telephony)	1500	−120	100	30	5.0	—	120
V70 V70B	70	10	2.5	1500	140	25	14	5.0	9.0	2.3	—	J. M.	3N 3G	Class-C Amp. (Telegraphy)	1500	−215	130	6.0	3.0	—	140
														Class-C Amp. (Telephony)	1250	−250	130	6.0	3.0	—	120
V70A V70C	70	10	2.5	1500	140	20	25	5.0	9.5	2.0	—	J. M.	3N 3G	Class-C Amp. (Telegraphy)	1000	−110	140	30	7.0	—	90
														Class-C Amp. (Telephony)	800	−150	95	20	5.0	—	50
50T [1]	75	5.0	6.0	3000	100	30	12	2.0	2.0	0.4	—	M.	2D	Class-C Amplifier	3000	−600	100	25	—	—	250
3-75A3 75TH	75	5.0	6.25	3000	225	40	20	2.7	2.3	0.3	40	M.	2D	Class-C Amp. (Telegraphy)	2000	−200	150	32	10	—	225
														Class-B Amp. Audio [7]	2000	− 90	50/225	350 [9]	3 [8]	19300	300
3-75A2 75TL						35	12	2.6	2.4	0.4			2D	Class-C Amp. (Telegraphy)	2000	−300	150	21	8	—	225
														Class-B Amp. Audio [7]	2000	−160	50/250	535 [9]	5 [8]	18000	350
HF-60	75	10	2.5	1600	160	—	28	5.4	5.2	1.5	30	M.	2D	Class-C Amp. (Telegraphy)	1600	−190	158	12	3.5	—	200
														Class-C Amp. (Telephony)	1250	−190	113	8	2.5	—	110
														Class-B Amp. Audio [7]	1600	− 75	50/248	310 [9]	3.0	13800	262
ZB-60	75	10	2.5	1600	160	40	80	6.1	5.8	1.85	30	M.	2D	Class-C Amp. (Telegraphy)	1500	− 95	158	31	6.0	—	190
														Class-B Amp. Audio [7]	1500	− 9	30/305	208 [9]	12.5	11200	320
111H	75	10	2.5	1500	160	30	23	5.0	4.6	2.9	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1250	−250	110	21	8.0	—	105
														Class-B Amp. Audio [7]	1750	− 62	40/270	324 [9]	9.0	16000	350
HF75	75	10	3.25	2000	120	—	12.5	—	2.0	—	75	M.	2D	Class-C Oscillator-Amp.	2000	—	120	—	—	—	150
TW75	75	7.5	4.15	2000	175	60	20	3.35	1.5	0.7	60	M.	2D	Class-C Amp.-Oscillator	2000	−175	150	37	12.7	—	225
														Class-C Amp. (Telephony)	2000	−260	125	32	13.2	—	198
T-100 HF100	75	10	2.5	1500	150	30	23	4.0	4.5	2.6	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1250	−250	110	21	8.0	—	105
														Grid-Modulated Amp.	1500	−280	72	1.5	6.0	—	42
														Class-B Amp. Audio [7]	1750	− 62	40/270	324 [9]	9.0 [8]	16000	350
UE-100	75	10	2.5	1750	150	30	23	3.5	4.5	1.4	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	—	170
														Class-C Amp. (Telephony)	1250	−250	120	21	8.0	—	105
														Class-B Audio [7]	1750	− 62	540 [8]	—	9.0	16000	350
ZB120	75	10	2.0	1250	160	40	90	5.3	5.2	3.2	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−135	160	23	5.5	—	145
														Class-C Amp. (Telephony)	1000	−150	120	21	5.0	—	95
														Grid-Modulated Amp.	1250	—	95	8.0	1.5	—	45
														Class-B Amp. Audio [7]	1500	− 9	60/296	196 [9]	5.0 [8]	11200	300
327B	75	10.5	10.6	—	—	—	30	3.4	2.45	0.3	—	N.	T-4AD	—	—	—	—	—	—	—	—
242A	85	10	3.25	1250	150	50	12.5	6.5	13	4.0	6	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	—	130
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100
284D	85	10	3.25	1250	150	100	4.8	6.0	8.3	5.6	—	J.	4E	Class-C Amp. (Telegraphy)	1250	−500	150	—	—	—	125
														Class-C Amp. (Telephony)	1000	−450	150	50	—	—	100
														Class-B Amp. Audio [7]	1250	−250	30/200	—	—	11200	140
812-H	85	6.3	4.0	1750	200	45	—	5.3	5.3	0.8	30	M.	3G	Class-C Amp. (Telegraphy)	1750	−175	170	26	6.5	—	225
															1250	−125	125	25	5.0	—	116
														Class-C Amp. (Telephony)	1500	−125	165	21	6.0	—	180
															1250	−125	125	25	6.0	—	120
														Class-B Amp. Audio [7]	1500	− 46	42/200	—	—	18000	225

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
8005	85	10	3.25	1500	200	45	20	6.4	5.0	1.0	60	M.	3G	Class-C Amp.-Telegraphy	1500	−130	200	32	7.5	—	220
														Class-C Amp. (Telephony)	1250	−195	190	28	9.0	—	170
														Class-B Amp. Audio[7]	1500	−70	40/310	310[9]	4.0	10000	300
V-70-D	85	7.5	3.25	1750	200	45	—	4.5	4.5	1.7	30	M.	3G	Class-C Amp. (Telegraphy)	1750	−100	170	19	3.9	—	225
															1500	−90	165	19	3.9	—	195
														Class-C Amp. (Telephony)	1500	−90	165	19	3.7	—	185
															1250	−72	127	16	2.6	—	122
RK36[1]	100	5.0	8.0	3000	165	35	14	4.5	5.0	1.0	60	M.	2D	Class-C Amp. (Telegraphy)	2000	−360	150	30	15	—	200
														Class-C Amp. (Telephony)	2000	−360	150	30	15	—	200
														Grid-Modulated Amp.	2000	−270	72	1.0	3.5	—	42
RK38[1]	100	5.0	8.0	3000	165	40	—	4.6	4.3	0.9	60	M.	2D	Class-C Amp. (Telegraphy)	2000	−200	160	30	10	—	225
														Class-C Amp. (Telephony)	2000	−200	160	30	10	—	225
														Grid-Modulated Amp.	2000	−150	80	2.0	5.5	—	60
3-100A4 100TH	100	5.0	6.3	3000	225	60	40	2.9	2.0	0.4	40	M.	2D	Class-C Amp. (Telegraphy)	3000	−200	165	51	18	—	400
														Class-C Amp. (Telephony)							
														Grid-Modulated Amp.	3000	−400	70	3.0	7.0	—	100
														Class-B Amp. (Audio)[7]	3000	−65	40/215	335[9]	5.0[8]	31000	650
3-100A2 100TL	100	5.0	6.3	3000	225	50	14	2.3	2.0	0.4	40	M.	2D	Class-C Amp. (Telegraphy)	3000	−400	165	30	20	—	400
														Class-C Amp. (Telephony)							
														Grid-Modulated Amp.	3000	−560	60	2.0	7.0	—	90
														Class-B Amp. (Audio)[7]	3000	−185	40/215	640[9]	6.0[8]	30000	450
VT127A	100	5.0	10.4	3000	—	—	15.5	2.7	2.3	0.35	150	N.	T-4B	Class-C Amp. (Telegraphy)	2000	−340	210	67	25	—	315
														Class-B Amp. (Audio)[7]	1500	−125	242	44	7.3	3000	200
227A	100	10.5	10.7	—	—	—	31	3.0	2.2	0.30	—	N.	T-4B	Oscillator at 200 Mc.	—	—	—	—	—	—	—
327A	100	10.5	10.7	—	—	—	31	3.4	2.3	0.35	—	N.	T-4AD	Oscillator at 200 Mc.	—	—	—	—	—	—	—
HK254	100	5.0	7.5	4000	200	40	25	3.3	3.4	1.1	50	J.	2N	Class-C Amp. (Telegraphy)	4000	−380	120	35	20	—	475
														Class-C Amp. (Telephony)	3000	−290	135	40	23	—	320
														Grid-Modulated Amp.	3000	—	51	3.0	4.0	—	58
														Class-B Amp. (Audio)[7]	3000	−100	40/240	456[9]	7.0[8]	30000	520
RK58	100	10	3.25	1250	175	70	—	8.5	6.5	10.5	—	J.	3N	Class-C Amp. (Telegraphy)	1250	−90	150	30	6.0	—	130
														Class-C Amp. (Telephony)	1000	−135	150	50	16	—	100
HF120	100	10	3.25	1250	175	50	12	5.5	12.5	3.5	15	J.	4F	Class-C Amp.-Oscillator	1250	300	166	8	3.5	—	148
HF125	100	10	3.25	1500	175	—	25	—	11.5	—	30	J.	—	Class-C Amp.-Oscillator	1500	—	175	—	—	—	200
HF140	100	10	3.25	1250	175	—	12	5.5	13.0	4.5	15	J.	4F	Class-C Amp.-Oscillator	1250	−300	166	8	3.5	—	148
203A 303A	100	10	3.25	1250	175	60	25	6.5	14.5	5.5	15	J.	4E	Class-C Amp. (Telegraphy)	1250	−125	150	25	7.0	—	130
														Class-C Amp. (Telephony)	1000	−135	150	50	14	—	100
														Class-B Amp. (Audio)[7]	1250	−45	26/320	330[9]	11[8]	9000	260
203H	100	10	3.25	1500	175	60	25	6.5	11.5	1.5	15	J.	3N	Class-C Amp. (Telegraphy)	1500	−200	170	12	3.8	—	200
														Class-C Amp. (Telephony)	1250	−160	167	19	5.0	—	160
														Class-B Amp. (Audio)[7]	1500	−52	30/320	304[9]	5.5[8]	11000	340
211 311 835[1]	100	10	3.25	1250	175	50	12	6.0 6.0	14.5 9.25	5.5 5.0	15	J.	4E	Class-C Amp. (Telegraphy)	1250	−225	150	18	7.0	—	130
														Class-C Amp. (Telephony)	1000	−260	150	35	14	—	100
														Class-B Amp. (Audio)[7]	1250	−100	20/320	410[9]	8.0[8]	9000	260
242B 342B	100	10	3.25	1250	150	50	12.5	7.0	13.6	6.0	6	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	—	130
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
		Volts	Amp.					Grid to Fil.	Grid to Plate	Plate to Fil.											
242C	100	10	3.25	1250	150	50	12.5	6.1	13.0	4.7	6	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	150	—	—	—	130
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	− 80	25/150	—	25[8]	7600	200
261A 361A	100	10	3.25	1250	150	50	12	6.5	9.0	4.0	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	125	—	—	—	100
														Class-C Amp. (Telephony)	1000	−160	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	− 90	20/150	—	25[8]	7200	200
276A 376A	100	10	3.0	1250	125	50	12	6.0	9.0	4.0	30	J.	4E	Class-C Amp. (Telegraphy)	1250	−175	125	—	—	—	100
														Class-C Amp. (Telephony)	1000	−160	125	50	—	—	85
														Class-B Amp. (Audio)[7]	1250	− 90	20/125	—	25[8]	9000	175
284B	100	10	3.25	1250	150	100	5.0	4.2	7.4	5.3	—	J.	3N	Class-C Amp. (Telegraphy)	1250	−500	150	—	—	—	125
														Class-C Amp. (Telephony)	1000	−430	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	−245	15/150	—	10[8]	7200	200
295A	100	10	3.25	1250	175	50	25	6.5	14.5	5.5	—	J.	4E	Class-C Amp. (Telegraphy)	1250	−125	150	—	—	—	125
														Class-C Amp. (Telephony)	1000	−125	150	50	—	—	100
														Class-B Amp. (Audio)[7]	1250	− 40	12/160	—	20[8]	9000	250
838 938	100	10	3.25	1250	175	70	—	6.5	8.0	5.0	30	J.	4E	Class-C Amp. (Telegraphy)	1250	− 90	150	30	6.0	—	130
														Class-C Amp. (Telephony)	1000	−135	150	60	16	—	100
														Class-B Amp. (Audio)	1250	0	148/320	200[9]	7.5[8]	9000	260
852	100	10	3.25	3000	150	40	12	1.9	2.6	1.0	30	M.	2D	Class-C Amp. (Telegraphy)	3000	−600	85	15	12	—	165
														Class-C Amp. (Telephony)	2000	−500	67	30	23	—	75
														Class-B Amp. (Audio)[7]	3000	−250	14/160	780[9]	3.5[8]	10250	320
5648[12]	100	6.3	1.1	1000	100	50	100	8.75	1.95	0.035	2500	N.	—	Class-C Amp. (Telegraphy)	1000	− 50	50	18	4	—	30
														Class-C Amp. (Telephony)	600	− 25	55	22	6	—	20
8003	100	10	3.25	1500	250	50	12	5.8	11.7	3.4	30	J.	3N	Class-C Amp.-Oscillator	1350	−180	245	35	11	—	250
														Class-C Amp. (Telephony)	1100	−260	200	40	15	—	167
														Class-B Amp. (Audio)[7]	1350	−100	40/490	480[9]	10.5[8]	6000	460
3X100A11 2C39	100	6.3	1.1	1000	60	40	100	6.5	1.95	0.03	500	N.	—	"Grid Isolation" Circuit	600	− 35	60	40	5.0	—	20
2C39A	100	6.3	1.0	1000	80	50	100	6.5	1.95	.035	500	N.	—	Class-C Amplifier	800	− 20	80	32	6	—	27
														Class-C Amp. (Telephony)	600	− 16	75	40	6	—	18
311-CH	125	10	3.25	1750	200	50	12	5.5	8.0	4.5	30	J.	Fig. 57	Class-C Amp. (Telegraphy)	1750	−200	200	20	4.5	—	260
														Class-C Amp. (Telephony)	1250	−200	166	8	3.5	—	148
														Class-B (Audio)[7]	1500	−110	400[8]	—	—	8200	400
3C22	125	6.3	2.0	1000	150	70	40	4.9	2.4	0.05	500	O.	Fig. 30	Class-C Amp.-Oscillator	1000	−200	150	70	—	—	65
4C36	125	5	7.5	4000	—	—	29	3.2	3.0	0.4	60	J.	Fig. 56	Class-C Amp.-Oscillator	—	—	—	—	18	—	480
F-123-A DR-123C	125	10	4.0	2000	300	75	14.5	6.5	8.5	3.3	—	J.	Fig. 26	Class-C Amp. (Telegraphy)	1500	−250	250	30	11	—	300
														Class-C Amp. (Telephony)	1500	−290	160	25	10	—	200
														Class-B Amp. (Audio)[7]	2000	−130	20/175	217[9]	3.4[8]	13800	522
RK57/805	125	10	3.25	1500	210	70	—	6.5	8.0	5.0	30	J.	3N	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	—	215
														Class-C Amp. (Telephony)	1250	−160	160	60	16	—	140
														Class-B Amp. (Audio)[7]	1500	− 16	84/400	280[9]	7.0[8]	8200	370
T125	125	10	4.5	2500	250	60	25	6.3	6.0	1.3	60	J.	2N	Class-C Amp. (Telegraphy)	2500	−200	240	31	11	—	475
														Class-C Amp. (Telephony)	2000	−215	200	28	10	—	320
HF130	125	10	3.25	1250	210	—	12.5	5.5	9.0	3.5	20	J.	—	Class-C Amp.-Oscillator	1250	−250	200	10	3.5	—	170
HF150	125	10	3.25	1500	210	—	12.5	5.5	7.2	1.9	30	J.	—	Class-C Amp.-Oscillator	1500	−300	200	10	4	—	220
HF175	125	10	4.0	2000	250	—	18	4.8	6.3	2.7	25	J.	T-3AC	Class-C Amp.-Oscillator	2000	−250	200	23	9	—	320

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
GL146	125	10	3.25	1500	200	60	75	7.2	9.2	3.9	15	J.	T-4BG	Class-C Amp.-Oscillator	1250	−150	180	30	—	—	150
														Class-C Amp. (Telephony)	1000	−200	160	40	—	—	100
														Class-B Amp. (Audio) [7]	1250	0	34/320	—	—	8400	250
GL152	125	10	3.25	1500	200	60	25	7.0	8.8	4.0	15	J.	T-4BG	Class-C Amp.-Oscillator	1250	−150	180	30	—	—	150
														Class-C Amp. (Telephony)	1000	−200	160	30	—	—	100
														Class-B Amp. (Audio) [7]	1250	− 40	16/320	—	—	8400	250
805	125	10	3.25	1500	210	70	40/60	8.5	6.5	10.5	30	J.	3N	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	—	215
														Class-C Amp. (Telephony)	1250	−160	160	60	16	—	140
														Class-B Amp. (Audio) [7]	1500	− 16	84/400	280 [9]	7.0 [8]	8200	370
AX9900/ 5866 [12]	135	6.3	5.4	2500	200	40	25	5.8	5.5	0.1	150	N.	Fig. 5	Class-C Amp. (Telegraphy)	2500	−200	200	40	16	—	390
														Class-C Amp. (Telephony)	2000	−225	127	40	16	—	204
														Class-B (Audio) [7]	2500	− 90	80/330	350 [9]	14 [8]	15680	560
3X150A3 3C37	150	6.3	2.5	1000	—	—	23	4.2	3.5	0.6	500	N.	—	—	—	—	—	—	—	—	—
150T [1]	150	5.0	10	3000	200	50	13	3.0	3.5	0.5	—	J.	2N	Class-C Amp. (Telegraphy)	3000	−600	200	35	—	—	450
3-150A3 152TH	150	5/10	12.5 [1]	3000	450	85	20	5.7	4.5	0.8	40	J.	4BC	Class-C Amp. (Telegraphy)	3000	−300	250	70	27	—	600
														Class-B Amp. (Audio) [7]	3000	−150	67/335	430 [9]	3.0 [8]	20300	700
3-150A2 152TL			6.25			75	12	4.5	4.4	0.7			4BC	Class-C Amp. (Telegraphy)	3000	−400	250	40	20	—	600
														Class-B Amp. (Audio) [7]	3000	−260	65/335	675 [9]	3.0 [8]	20400	700
TW150	150	10	4.1	3000	200	60	35	3.9	2.0	0.8	—	J.	2N	Class-C Amp.-Oscillator	3000	−170	200	45	17	—	470
														Class-C Amp. (Telephony)	3000	−260	165	40	17	—	400
HK252-L	150	5/10	13/6.5	3000	500	75	10	7.0	5.0	0.4	125	N.	4BC	Class-C Amp.-Oscillator	3000	−400	250	30	15	—	610
														Class-C Amp. (Telephony)	2500	−350	250	35	16	—	500
DR200 HF200 HV18	150	10–11	3.4	2500	200	50	18	5.2	5.8	1.2	20	J.	2N	Class-C Amp. (Telegraphy)	2500	−300	200	18	8.0	—	380
														Class-C Amp. (Telephony)	2000	−350	160	20	9.0	—	250
														Class-B Amp. (Audio) [7]	2500	−130	60/360	460 [9]	8.0 [8]	16000	600
HD203A	150	10	4.0	2000	250	60	25	—	12	—	15	J.	3N	Class-C Amplifier	—	—	—	—	—	—	375
HF250	150	10.5	4.0	2500	200	—	18	—	5.8	—	20	J.	2N	Class-C Amp.-Oscillator	2500	—	200	—	—	—	375
HK354 HK354C	150	5.0	10	4000	300	50	14	4.5	3.8	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	4000	−690	245	50	48	—	830
														Class-C Amp. (Telephony)	3000	−550	210	50	35	—	525
														Grid-Modulated Amp.	3000	−400	78	3.0	12	—	85
														Class-B Amp. (Audio) [7]	3000	−205	65/313	630 [9]	20 [8]	22000	665
HK354D	150	5.0	10	4000	300	55	22	4.5	3.8	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	3500	−490	240	50	38	—	690
														Class-C Amp. (Telephony)	3500	−425	210	55	36	—	525
HK354E	150	5.0	10	4000	300	60	35	4.5	3.8	1.1	30	J.	2N	Class C Amp. (Telegraphy)	3500	−448	240	60	45	—	690
														Class-C Amp. (Telephony)	3000	−437	210	60	45	—	525
HK354F	150	5.0	10	4000	300	75	50	4.5	3.8	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	3500	−368	250	75	50	—	720
														Class-C Amp. (Telephony)	3000	−312	210	75	45	—	525
UE-468	150	10	4.05	2500	200	60	18	8.8	7.0	1.25	30	J.	Fig. 57	Class-C Amp. (Telegraphy)	2500	−300	200	18	8.0	—	380
														Class-C Amp. (Telephony)	2000	−350	160	20	9.0	—	250
														Class-B (Audio) [7]	2500	−130	320 [8]	410 [9]	2.5	16000	500
810 1627 [1]	175	10 5.0	4.5 9.0	2500	300	75	36	8.7	4.8	12	30	J.	2N	Class-C Amp. (Telegraphy)	2500	−180	300	60	19	—	575
														Class-C Amp. (Telephony)	2000	−350	250	70	35	—	380
														Grid-Modulated Amp.	2250	−140	100	2.0	4.0	—	75
														Class-B Amp. (Audio) [7]	2250	− 60	70/450	380 [9]	13 [8]	11600	725

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
		Volts	Amp.					Grid to Fil.	Grid to Plate	Plate to Fil.											
8000	175	10	4.5	2500	300	45	16.5	5.0	6.4	3.3	30	J.	2N	Class-C Amp.-Oscillator	2500	−240	300	40	18	—	575
														Class-C Amp. (Telephony)	2000	−370	250	37	20	—	380
														Grid-Modulated Amp.	2250	−265	100	0	2.5	—	75
														Class-B Amp. (Audio)[7]	2250	−130	65/450	560[9]	7.9[8]	12000	725
GL-5C24	160	10	5.2	1750	107	—	8	5.6	8.8	3.3	—	N.	Fig. 26	Class-A Amp. (Audio)	1500	−155	107	—	—	8200[5]	55
														Class-AB$_1$ Amp. (Audio)[7]	1750	−200	320[8]	390[9]	—	8000	240
RK63 RK63A	200	5.0 6.3	10 14	3000	250	60	37	2.7	3.3	1.1	—	J.	2N	Class-C Amp. (Telegraphy)	3000	−200	233	45	17	—	525
														Class-C Amp. (Telephony)	2500	−200	205	50	19	—	405
														Grid-Modulated Amp.	3000	−250	100	7.0	12.5	—	100
T200	200	10	5.75	2500	350	80	16	9.5	7.9	1.6	30	J.	2N	Class-C Amp. (Telegraphy)	2500	−280	350	54	25	—	685
														Class-C Amp. (Telephony)	2000	−260	300	54	23	—	460
F-127-A	200	10	4.0	3000	325	70	38	13	4	13	—	J.	Fig. 26	Class-C Amp. (Telegraphy)	3000	−250	250	47	18	—	600
														Class-C Amp. (Telephony)	2500	−300	200	58	25.2	—	420
														Class-B Amp. (Audio)[7]	2800	− 75	20/400	175[9]	6.65[8]	16600	820
822 822S	200	10	4.0	2500	300	60	30	8.5	13.5	2.1	20 30	J.	3N 2N	Class-C Amp. (Telegraphy)	2500	−190	300	51	17*	—	600
														Class-C Amp. (Telephony)	2000	− 75	250	43	13.7	—	405
														Class-B Amp. (Audio)[7]	3000	− 80	450[8]	362[9]	8.0[8]	16000	1000
4C32	200	10	4.5	3000	300	60	30	5.5	5.8	1.1	60	J.	2N	Class-C Amp.-Oscillator	2000	−165	275	20	10	—	400
														Class-C Amp. (Telephony)	2000	−200	250	20	15	—	375
GL-592 3-200A3	200	10	5.0	3500	250	50	25	3.6	3.3	0.29	150	N.	Fig. 52	Class-C Amp. (Telegraphy)	3000	−220	222	25	11	—	466
														Class-C Amp. (Telephony)	2500	−300	200	35	19	—	375
														Class-B (Audio)[7]	2000	− 50	120/500	520[9]	25[3]	8500	600
4C34 HF300	200	11–12	4.0	3000	275	60	23	6.0	6.5	1.4	60 20	J.	2N	Class-C Amp. (Telegraphy)	3000	−400	250	28	16	—	600
														Class-C Amp. (Telephony)	2000	−300	250	36	17	—	385
														Class-B Amp. (Audio)[7]	3000	−115	60/360	450[9]	13[8]	20000	780
T814 HV12	200	10	4.0	2500	200	60	12	8.5	12.8	1.7	30	J	3N	Class-C Amp. (Telegraphy)	2500	−240	300	30	10	—	575
														Class-C Amp. (Telephony)	2000	−370	300	40	20	—	485
														Class-B Amp. (Audio)[7]	2000	−160	50/275	350[9]	7.0[8]	14400	400
T822 HV27	200	10	4.0	2500	300	60	27	8.5	13.5	2.1	30	J.	3N	Class-C Amp. (Telegraphy)	2500	−175	300	50	15	—	585
														Class-C Amp. (Telephony)	2000	−195	250	45	15	—	400
T-300	200	11	6.0	3000	300	—	23	6.0	7.0	1.4	—	—	—	Class-C Amp. (Telegraphy)	3000	−400	250	28	20	—	600
														Class-C Amp. (Telephony)	2000	−300	250	36	17	—	385
														Class-B (Audio)[7]	2500	−100	60/450	—	7.5[8]	—	750
805	225	5.0	10	3300	300	50	12.6	6.1	4.2	1.1	30	J.	2N	Class-C Amp. (Telegraphy)	3300	−600	300	40	34	—	780
														Class-C Amp. (Telephony)	3000	−670	195	27	24	—	460
														Class-B Amp. (Audio)[7]	3300	−240	80/475	930[9]	35[8]	16000	1120
3-250A4 250TH	250	5.0	10.5	4000	350	100	37	5.0	2.9	0.7	40	J.	2N	Class-C Amp. (Telegraphy)	2000	−120	350	100	34	—	500
														Class-C Amp. (Telephony)	3000	−210	330	75	42	—	750
														Grid-Modulated Amp.	3000	−160	125	4.5	20	—	125
														Class-B Amp. (Audio)[7]	3000	− 65	100/560	460[9]	24[8]	12250	1150
3-250A2 250TL	250	5.0	10.5	4000	350	50	14	3.7	3.1	0.7	40	J.	2N	Class-C Amp. (Telegraphy)	3000	−350	335	45	29	—	750
														Class-C Amp. (Telephony)	3000	−350	335	45	29	—	750
														Grid-Modulated Amp.	3000	−450	125	2.0	15	—	125
														Class-B Amp. (Audio)[7]	3000	−175	100/500	840[9]	17[8]	13000	1000

TABLE XVI—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connec-tions	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
GL159	250	10	9.6	2000	400	100	20	11	17.6	5.0	15	J.	T-4BG	Class-C Amp.-Oscillator	2000	−200	400	17	6.0	—	620
														Class-C Amp. (Telephony)	1500	−240	400	23	9.0	—	450
														Class-B Amp. (Audio)[7]	2000	−100	30/660	400[9]	4.0[8]	6880	900
GL169	250	10	9.6	2000	400	100	85	11.5	19	4.7	15	J.	T-4BG	Class-C Amp.-Oscillator	2000	−100	400	42	10	—	620
														Class-C Amp. (Telephony)	1500	−100	400	45	10	—	450
														Class-B Amp. (Audio)[7]	2000	−18	30/660	220[9]	6.0[8]	7000	900
204A 304A	250	11	3.85	2500	275	80	23	12.5	15	2.3	3	N.	T-1A	Class-C Amp. (Telegraphy)	2500	−200	250	30	15	—	450
														Class-C Amp. (Telephony)	2000	−250	250	35	20	—	350
														Class-B Amp. (Audio)[7]	3000	−100	80/372	500[9]	18[8]	20000	700
308B	250	14	4.0	2250	325	75	8.0	13.6	17.4	9.3	1.5	N.	T-2A	Class-C Amp. (Telegraphy)	1750	−345	300	—	—	—	350
															1500	−300	300	—	—	—	300
														Class-B Amp. (Audio)[7]	1750	−215	30/300	—	35[8]	5200	575
HK454H	250	5.0	11	5000	375	85	30	4.6	3.4	1.4	100	J.	2N	Class-C Amp. (Telegraphy)	3500	−275	270	60	28	—	760
HK454-L	250	5.0	11	5000	375	60	12	4.6	3.4	1.4	100	J.	2N	Class-C Amp. (Telephony)	3500	−450	270	45	30	—	760
5867 AX-9901	250	5.25	14.1	—	—	—	25	7.0	5.3	0.15	100	—	—	Class-C Amplifier	3000	−400	363	80	—	—	950
212E 241B 312E	275	14	4.0	3000	350	75	16	14.9	18.8	8.6	1.5	N.	T-2A T-2AA	Class-C Amp. (Telegraphy)	3500	−275	270	60	28	—	760
														Class-C Amp. (Telephony)	3500	−450	270	45	30	—	760
														Class-B Amp. (Audio)[7]	2000	−105	40/300	—	50[8]	8000	650
300T[1]	300	8.0	11.5	3500	350	75	16	4.0	4.0	0.6	—	J.	2N	Class-C Amp. (Telegraphy)	2000	−225	300	—	—	—	400
HK304-L	300	5/10	26/13	3000	1000	150	10	12	9.0	0.8	—	N.	4BC	Class-C Amp. (Telephony)	1500	−200	300	75	—	—	300
527	300	5.5	135.0	—	—	—	38	19.0	12.0	1.4	200	N.	T-4B	Oscillator at 200 Mc.	Approximately 250 watts output						
HK654	300	7.5	15	4000	600	100	22	6.2	5.5	1.5	20	J.	2N	Class-C Amp. (Telegraphy)	2000	−380	500	75	57	—	720
														Class-C Amp. (Telephony)	2000	−365	450	110	70	—	655
														Grid-Modulated Amp.	3500	−210	150	15	15	—	210
3-300A3 304TH	300	5/10	25/12.5	3000	900	170	20	13.5	10.2	0.7	40	N.	4BC	Class-C Amplifier	1500	−125	667	115	25	—	700
														Class-B Amp. (Audio)[7]	3000	−150	134/667	420[9]	6.0[8]	10200	1400
3-300A2 304TL						150	12	8.5	9.1	0.6	40	N.	4BC	Class-C Amplifier	1500	−250	665	90	33	—	700
														Class-B Amp. (Audio)[7]	3000	−260	130/667	650[9]	6.0[8]	10200	1400
833A	350	10	10	3300	500	100	35	12.3	6.3	8.5	30	N.	T-1AB	Class-C Amp. (Telegraphy)	2000	−200	475	65	25	—	740
														Class-C Amp. (Telephony)	2500	−300	335	75	30	—	635
270A	350	10	4.0	3000	375	75	16	18	21	2.0	7.5	N.	T-1A	Class-C Amp. (Telegraphy)	3000	−375	350	—	—	—	700
														Class-C Amp. (Telephony)	2250	−300	300	80	—	—	450
849[1]	400	11	5.0	2500	350	125	19	17	33.5	3.0	3	N.	T-1A	Class-C Amp. (Telegraphy)	2500	−250	300	20	8.0	—	560
														Class-C Amp. (Telephony)	2000	−300	300	30	14	—	425
831[1]	400	11	10	3500	350	75	14.5	3.8	4.0	1.4	—	N.	T-1AA	Class-C Amp. (Telegraphy)	3500	−400	275	40	30	—	590
														Class-C Amp. (Telephony)	3000	−500	200	60	50	—	360

* Cathode resistor in ohms.
** Grid resistor ohms.

[1] Discontinued.
[2] Twin triode. Values, except interelement capacities, are for both sections in push-pull.
[3] Output at 112 Mc.

[4] Grid-leak resistor in ohms.
[5] Peak valves.
[6] Per section.
[7] Values are for two tubes in push-pull.

[8] Max. signal value.
[9] Peak a.f. grid-to-grid volts.
[10] For single tube.
[11] Class-B data in Table I.
[12] Forced-air cooling.

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
3A4	2.0	1.4 2.8	0.2 0.1	150	135	0.9	4.8	0.2	4.2	10	B.	7BB	Class-C Amp. (Telegraphy)	150	135	0	−26	18.3	6.5	0.13	2300	—	—	1.2
3D6	4.5	2.8 1.4	0.11 0.22	180	135	0.9	7.5	0.3	5.5	50	L.	6BB	Class-C Amp. (Telegraphy)	150	135	—	−20	23	6.0	1.0	—	0.25	—	1.4
3B4	3.0	2.5 1.25	0.165 0.33	150	135	—	4.6	0.16	7.6	100	B.	7CY	Class-C Amp.	150	135	—	−75	25	—	—	—	—	—	1.25
HY63[1]	3.0	2.5 1.25	0.1125 0.225	200	100	0.6	8.0	0.1	8.0	60	O.	T-8DB	Class-C Amp. (Telegraphy)	200	100	—	−22.5	20	4.0	2.0	—	0.1	—	3.0
													Class-C Amp. (Telephony)	180	100	—	−35	15	3.0	2.0	—	0.2	—	2.0
6AK6	3.5	6.3	0.15	375	250	1.0	3.6	0.12	4.2	54	B.	7BK	Class-C Amp. (Telegraphy)	375	250	—	−100	15	4.0	3.0	—	—	—	4.0
5A6	5.0	2.5 5.0	0.46 0.23	150	150	2	8.5	0.15	9.5	100	B.	9L	Class-C Amp.	150	150	0	−24	40	11	1.2	—	—	—	3.1
5618	5.0	6.0 3.0	0.23 0.46	300	125	2.0	7.0	0.24	5.0	80	B.	7CU	Class-C Amp. (Telegraphy)	300	75	0	−45	25	7.0	1.5	32000	0.3	—	5.4
RK64[1]	6.0	6.3	0.5	400	100	3.0	10	0.4	9.0	60	M.	5AW	Class-C Amp. (Telegraphy)	400	100	30	−30	35	10	3.0	—	0.18	—	10
													Class-C Amp. (Telephony)	300	—	30	−30	26	8.0	4.0	30000	0.2	—	6.0
1610	6.0	2.5	1.75	400	200	2.0	8.6	1.2	13	20	M.	T-5CA	Class-C Amp. (Telegraphy)	400	150	—	−50	22.5	7.0	1.5	—	0.1	—	5.0
5686	7.5	6.3	0.35	250	250	3.0	6.4	0.11	4.0	160	B.	Fig. 29	Class-C Amp. (Telegraphy)	250	250	—	−50	40	10.5	2.0	—	0.15	—	6.5
														250	180	—	−30	30	6.5	2.0	—	0.10	—	5.0
6AQ5	8.0	6.3	0.45	350	250	2.0	7.6	0.35	6.0	54	B.	7BZ	Class-C Amp. (Telegraphy)	350	250	—	−100	47	7.0	5.0	—	—	—	11
6V6GT	8.0	6.3	0.45	350	250	2.0	9.5	0.7	7.5	10	O.	7AC	Class-C Amp. (Telegraphy)	350	250	—	−100	47	7.0	5.0	—	—	—	11
6AG7	9.0	6.3	0.65	375	250	1.5	13	0.06	7.5	10	O.	8Y	Class-C Amp. (Telegraphy)	375	250	—	−75	30	9.0	5.0	—	—	—	7.5
RK56	8.0	6.3	0.55	300	300	4.5	10	0.2	9.0	60	M.	5AW	Class-C Amp. (Telegraphy)	400	300	—	−40	62	12	1.6	—	0.1	—	12.5
													Class-C Amp. (Telephony)	250	200	—	−40	50	10	1.6	2800	0.28	—	8.5
RK23[1] RK25 RK25B[1]	10	2.5 6.3	2.0 0.9	500	250	8	10	0.2	10	—	M.	6BM	Class-C Amp. (Telegraphy)	500	200	45	−90	55	38	4.0	—	0.5	—	22
													Class-C Amp. (Telephony)	400	150	0	−90	43	30	6.0	8300	0.8	—	13.5
													Suppressor-Modulated Amp.	500	200	−45	−90	31	39	4.0	—	0.5	—	6.0
1613	10	6.3	0.7	350	275	2.5	8.5	0.5	11.5	45	O.	7S	Class-C Amp. (Telegraphy)	350	200	—	−35	50	10	3.5	20000	0.22	—	9
													Class-C Amp. (Telephony)	275	200	—	−35	42	10	2.8	10000	0.16	—	6.0
2E30	10	6.0	0.7	250	250	2.5	10	0.5	4.5	160	B.	7CQ	Class-C Amp. (Telegraphy)	250	200	—	−50	50	10	2.5	—	0.2	—	7.5
													Class-AB_2 Amp. (Audio)[6]	250	250	—	−30	40/120	4/20	2.3[7]	87[8]	0.2	3800	17
5812	10	6.0	0.65	300	250	2.5	9.0	0.2	7.4	165	B.	7CQ	Class-C Amp. (Telegraphy)	300	200	—	−45	55	3.0	0.75	—	1.5	—	7.0
6216	10	6.3	1.2	300	200	1.0	12.3	0.37	6.7	50	B.	Fig. 73	Class-C Amp. (Telegraphy)	300	150	—	−50	63	8.0	2.0	—	0.3	—	8.8
													Class-C Amp.-Doubler	300	—	—	−75	50	6.0	1.0	25000	0.6	—	4.0
837 RK44[1]	12	12.6	0.7	500	300	8	16	0.2	10	20	M.	6BM	Class-C Amp. (Telegraphy)	500	200	40	−70	80	15	4.0	20000	0.4	—	28
													Class-C Amp. (Telephony)	400	140	40	−40	45	20	5.0	13000	0.3	—	11
													Suppressor-Modulated Amp.	500	—	−65	−20	30	23	3.5	14000	0.1	—	5.0
5763 6062	12	6.0	0.75	300	250	2	9.5	0.3	4.5	175	B.	9K	Class-C Amp. (Telegraphy)	300	250	0	−60	50	5.0	3.0	—	0.35	—	8.0
													Doubler to 175 Mc.	300	250	0	−75	40	4.0	1.0	12500	0.6	—	3.6
6F6 6F6G	12.5	6.3	0.7	400	275	3.0	6.5	0.2	13	10	O.	7AC	Class-C Amp. (Telegraphy)	400	275	—	−100	50	11	5.0	—	—	—	14
							8.0	0.5	6.5				Class-C Amp. (Telephony)	275	200	—	−35	42	10	2.8	—	0.16	—	6.0
802	13	6.3	0.9	600	250	6.0	12	0.15	8.5	30	M.	6BM	Class-C Amp. (Telegraphy)	600	250	40	−120	55	16	2.4	22000	0.30	—	23
													Class-C Amp. (Telephony)	500	245	40	−40	40	15	1.5	16300	0.10	—	12
													Suppressor-Modulated Amp.	600	250	−45	−100	30	24	5.0	14500	0.6	—	6.3
2E24	13.5	6.3[5]	0.65	500	200	2.3	8.5	0.11	6.5	125	O.	7CL	Class-C Amp. (Telephony)	400	180	—	−45	50	8.0	2.5	27500	0.15	—	13.5
														500	180	—	−45	54	8.0	2.5	40000	0.16	—	18.0
				600	200	2.5							Class-C Amp. (Telegraphy)	400	200	—	−45	75	10.0	3.0	20000	0.19	—	20
														600	195	—	−50	66	10	3.0	40500	0.21	—	27

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
2E26	13.5	6.3	0.8	600	200	2.5	13	0.2	7.0	125	O.	7CK	Class-C Amp. (Telegraphy)	600	185	—	−45	66	10	3.0	41500	0.17	—	27
													Class-C Amp. (Telephony)	500	180	—	−50	54	9.0	2.5	35500	0.15	—	18
				500	200	2.3							Class-AB$_2$ Amp. (Audio)[6]	500	125	—	−15	22/150	32[7]	—	60[8]	0.36[7]	8000	54
HY6V6-GTX	13	6.3	0.5	350	225	2.5	9.5	0.7	9.5	60	O.	7AC	Class-C Amp. (Telegraphy)	300	200	—	−45	60	7.5	2.5	—	0.3	—	12
													Class-C Amp. (Telephony)	250	200	—	−45	60	6.0	2.0	15000	0.4	—	10
HY60	15	6.3	0.5	425	225	2.5	10	0.2	8.5	60	M.	5AW	Class-C Amp. (Telegraphy)	425	200	—	−62.5	60	8.5	3.0	—	0.3	—	18
													Class-C Amp. (Telephony)	325	200	—	−45	60	7.0	2.5	—	0.2	—	14
HY65[1]	15	6.3	0.85	450	250	4.0	9.1	0.18	7.2	60	O.	T-8DB	Class-C Amp.-Oscillator	450	250	—	−45	75	15	3.0	—	0.5	—	24
													Class-C Amp. (Telephony)	350	200	—	−45	63	12	3.0	—	0.5	—	16
2E25	15	6.0	0.8	450	250	4.0	8.5	0.15	6.7	125	O.	5BJ	Class-C Amp.-Oscillator	450	250	—	−45	75	15	3.0	—	0.4	—	24
													Class-C Amp. (Telephony)	400	200	—	−45	60	12	3.0	—	0.4	—	16
													Class-AB$_2$ Amp. (Audio)[6]	450	250	—	−30	44/150	10/40	3.0	142[8]	0.9[7]	6000	40
306A	15	2.75	2.0	300	300	6.0	13	0.35	13	—	M.	T-5CB	Class-C Amp. (Telephony)	300	180	—	−50	36	15	3.0	8000	—	—	7.0
307A RK-75	15	5.5	1.0	500	250	6.0	15	0.55	12	—	M.	T-5C	Class-C Amp. (Telegraphy)	500	250	0	−35	60	13	1.4	20000	—	—	20
													Suppressor-Modulated Amp.	500	200	−50	−35	40	20	1.5	14000	—	—	6.0
832[3]	15	6.3 12.6	1.6 0.8	500	250	5.0	7.5	0.05	3.8	200	N.	7BP	Class-C Amp. (Telegraphy)	500	200	—	−65	72	14	2.6	21000	0.18	—	26
													Class-C Amp. (Telephony)	425	200	—	−60	52	16	2.4	14000	0.15	—	16
832A[3]	15	6.3 12.6	1.6 0.8	750	250	5.0	7.5	0.05	3.8	200	N.	7BP	Class-C Amp. (Telegraphy)	750	200	—	−65	48	15	2.8	36500	0.19	—	26
													Class-C Amp. (Telephony)	600	200	—	−65	36	16	2.6	25000	0.16	—	17
844[1]	15	2.5	2.5	500	180	3.0	9.5	0.15	7.5	—	M.	5AW	Class-C Amp. (Telegraphy)	500	175	—	−125	25	—	5.0	—	—	—	9.0
													Class-C Amp. (Telephony)	500	150	—	−100	20	—	—	—	—	—	4.0
865	15	7.5	2.0	750	175	.0	8.5	0.1	8.0	15	M.	T-4C	Class-C Amp. (Telegraphy)	750	125	—	−80	40	—	5.5	—	1.0	—	16
													Class-C Amp. (Telephony)	500	125	—	−120	40	—	9.0	—	2.5	—	10
1619	15	2.5	2.0	400	300	3.5	10.5	0.35	12.5	45	O.	T9H	Class-C Amp. (Telegraphy)	400	300	—	−55	75	10.5	5.0	9500	0.36	—	19.5
													Class-C Amp. (Telephony)	325	285	—	−50	62	7.5	2.8	5000	0.18	—	13
													Class-AB$_2$ Amp. (Audio)[6]	400	300	0	−16.5	75/150	6.5/11.5	—	77[8]	0.4[7]	6000	36
5516	15	6.0	0.7	600	250	5.0	8.5	0.12	6.5	80	O.	7CL	Class-C Amp. (Telegraphy)	600	250	—	−60	75	15	5.0	—	0.5	—	32
													Class-C Amp. (Telephony)	475	250	—	−90	63	10	4.0	22500	0.5	—	22
													Class-AB$_2$ (Audio)[6]	600	25	—	−25	36/140	1/24	4[7]	80[8]	0.16	10500	67
AX-9905[3]	16	6.3	0.68	400	250	5	8.5	0.05	3.3	186	O.	Fig. 34	Class-C Amplifier	400	250	—	−80	80	6	3.5	—	0.39	—	20.8
														250	175	—	−70	80	6.5	4.2	—	0.26	—	16.9
254A	20	5.0	3.25	750	175	5.0	4.6	0.1	9.4	—	M.	T-4C	Class-C Amplifier	750	175	—	−90	60	—	—	—	—	—	25
6L6	21	6.3	0.9	400	300	3.5	10	0.4	12	10	O.	7AC	Class-C Amp.-Oscillator	400	300	—	−125	100	12	5.0	—	—	—	28
6L6G							11.5	0.9	9.5				Class-C Amp. (Telephony)	325	250	—	−70	65	—	9.0	—	0.8	—	11
6L6GX	21	6.3	0.9	500	300	3.5	11	1.5	7.0	—	O.	7AC	Class-C Amp. (Telegraphy)	500	250	—	−50	90	9.0	2.0	—	0.25	—	30
													Class-C Amp. (Telephony)	325	225	—	−45	90	9.0	3.0	—	0.25	—	20
HY6L6-GTX	21	6.3	0.9	500	300	3.5	11	0.5	7.0	60	O.	7AC	Class-C Amp.-Oscillator	500	250	—	−50	90	9.0	2.0	—	0.5	—	30
													Class-C Amp. (Telephony)	400	225	—	−45	90	9.0	3.0	16000	0.8	—	20
T21	21	6.3	0.9	400	300	3.5	13	0.7	12	30	M.	6A	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	—	25
													Class-C Amp. (Telephony)	350	200	—	−45	65	17	5.0	—	0.35	—	14
RK49	21	6.3	0.9	400	300	3.5	11.5	1.4	10.6	—	M.	6A	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	—	25
													Class-C Amp. (Telephony)	300	200	—	−45	60	15	5.0	6700	0.34	—	12
5881	23	6.3	0.9	400	300	3	—	—	—	—	O.	7AC	Class-C Amplifier	Same as 6L6										
1614	25	6.3	0.9	450	300	3.5	10	0.4	12.5	80	O.	7AC	Class-C Amp. (Telegraphy)	450	250	—	−45	100	8	2.0	12500	0.15	—	31
													Class-C Amp. (Telephony)	375	250	—	−50	93	7.0	2.0	10000	0.15	—	24.5
													Class-AB$_1$ Amp. (Audio)[6]	530	340	—	−36	60/160	20[7]	—	72[8]	—	7200	50
RK41[1] RK39	25	2.5 6.3	2.4 0.9	600	390	3.5	13	0.2	10	30	M.	5AW	Class-C Amp. (Telegraphy)	600	300	—	−90	93	10	3.0	—	0.38	—	36
													Class-C Amp. (Telephony)	475	250	—	−50	85	9.0	2.5	25000	0.2	—	26

Type	Max. Plate Dissipation Watts	Cathode		Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
		Volts	Amp.				Grid to Fil.	Grid to Plate	Plate to Fil.															
HY61	25	6.3	0.9	600	300	3.5	11	0.2	7.0	60	M.	5AW	Class-C Amp. (Telegraphy)	600	250	—	−50	85	9.0	4.0	39000	0.4	—	40
													Class-C Amp. (Telephony)	475	250	—	−50	100	9.0	3.5	25000	0.2	—	27
													Class-AB₂ Amp. (Audio)[6]	600	300	—	−30	200[7]	10[7]	—	—	0.1[7]	—	80
815[3]	25	12.6 6.3	0.8 1.6	500	200	4.0	13.3	0.2	8.5	125	O.	8BY	Class-C Amp.-Oscillator	500	200	—	−45	150	17	2.5	—	0.13	—	56
													Class-C Amp. (Telephony)	400	175	—	−45	150	15	3.0	—	0.16	—	45
													Class-AB₂ Amp. (Audio)[3]	500	125	—	−15	22/150	32[7]	—	60[8]	0.36[7]	8000	54
254B	25	7.5	3.25	750	150	5.0	11.2	0.085	5.4	—	M.	T-4C	Class-C Amplifier	750	150	—	−135	75	—	—	—	—	—	30
1624	25	2.5	2.0	600	300	3.5	11	0.25	7.5	60	M.	T-5DC	Class-C Amp. (Telegraphy)	600	300	—	−60	90	10	5.0	30000	0.43	—	35
													Class-C Amp. (Telephony)	500	275	—	−50	75	9.0	3.3	25000	0.25	—	24
													Class-AB₂ Amp. (Audio)[6]	600	300	—	−25	42/180	5/15	106[8]	—	1.2[7]	7500	72
3DX3	25	6.3	3.0	1500	200	—	—	—	—	250	S.	Fig. 40	Class-C Amp. (Telegraphy)	1000	200	—	−155	75	—	2.8	—	0.57	—	50
6146 6159	25	6.3 26.5	1.25 0.3	750	250	3.0	13.5	0.22	9.0	60	M.	7CK	Class-C Amp. (C. W. 15 Mc.)	750	160	—	−85	120	14.7	3.0	—	0.3	—	69
													Class-C Amp. (C. W. 175 Mc.)	400	200	—	−54	150	9	1.8	—	3.0	—	35
													Class-C Amp. (Telephony)	600	150	—	−85	112.5	12	3.0	—	0.3	—	52
													Class-AB₂ Amp. (Audio)[6]	750	165	—	−45	35/240	0.6/21	101[8]	—	0.07	8000	130
3E22[3]	30	12.6 6.3	0.8 1.6	560	225	6.0	14	0.22	8.5	200	O.	8BY	Class-C Amp. (Telegraphy)[3]	600	200	—	−55	160	20	7.0	20000	0.45	—	72
													Class-C Amp. (Telephony)[3]	560	200	—	−50	160	20	6.5	18000	0.4	—	67
RK66	30	6.3	1.5	600	300	3.5	12	0.25	10.5	60	M.	T-5C	Class-C Amp.-Oscillator	600	300	—	−60	90	11	5.0	—	0.5	—	40
													Class-C Amp. (Telephony)	500	—	—	−50	75	8.0	3.2	25000	0.23	—	25
807 807W 5933 1625	30	6.3 12.6	0.9 0.45	750	300	3.5	11	0.2	7.0	60	M.	5AW 5AZ	Class-C Amp. (Telegraphy)	750	250	—	−45	100	6	3.5	85000	0.22	—	50
													Class-C Amp. (Telephony)	600	275	—	−90	100	6.5	4.0	50000	0.4	—	42.5
													Class-AB₂ Amp. (Audio)[6]	750	300	—	−32	60/240	5/10	92[8]	—	0.2[7]	6950	120
													Class-B Amp. (Audio)[11]	750	—	—	0	15/240	—	555[8]	—	5.3[7]	6650	120
2E22	30	6.3	1.5	750	250	10	13	0.2	8.0	—	M.	5J	Class-C Amp.-Oscillator	500	250	22.5	−60	100	16	6.0	15000	0.55	—	34
													Class-C Amp.-Oscillator	750	250	22.5	−60	100	16	6.0	30000	0.55	—	53
													Suppressor-Modulated Amp.	750	250	−90	−65	55	29	6.5	17000	0.6	—	16.5
3D23 TB-35	35	6.3	3.0	—	—	—	6.5	0.2	1.8	250	M.	Fig. 54	Class-C Amp. (Telegraphy)	1500	375	—	−300	110	22	15	—	4.5	—	130
													Class-C Amp. (Telephony)	1000	300	—	−200	85	14	10	—	2.0	—	60
AX-9903[3] 5894A	40	6.3 12.6	1.8 0.9	600	250	7	6.7	0.08	2.1	150	N.	Fig. 10	Class-C Amp. (Telegraphy)	600	250	—	−80	200	16	2	—	0.2	—	80
													Class-C Amp. (Telephony)	600	250	—	−100	200	24	8	—	1.2	—	85
RK20[1] RK20A RK46[1]	40	7.5 7.5 12.6	3.0 3.25 2.5	1250	300	15	14	0.01	12	—	M.	T-5C	Class-C Amp. (Telegraphy)	1250	300	45	−100	92	36	11.5	—	1.6	—	84
													Class-C Amp. (Telephony)	1000	300	0	−100	75	30	10	23000	1.3	—	52
													Suppressor-Modulated Amp.	1250	300	−45	−100	48	44	11.5	—	1.5	—	21
													Grid-Modulated Amp.	1250	300	45	−142	40	7.0	1.8	—	1.5	—	20
HY69	40	6.3	1.5	600	300	5.0	15.4	0.23	6.5	60	M.	T-5D	Class-C Amp.-Oscillator	600	250	—	−60	100	12.5	4.0	30000	0.25	—	42
													Class-C Amp. (Telephony)	600	250	—	−60	100	12.5	5.0	30000	0.35	—	42
													Modulated Doubler	600	200	—	−300	90	11.5	6.0	35000	2.8	—	27
													Class-AB₂ Amp. (Audio)[6]	600	300	—	−35	200[7]	18[7]	5.0[7]	—	0.3[7]	—	80
829[1,3]	40	6.3 12.6	2.25 1.12	500	225	6	14.5	0.1	7.0	200	N.	7BP	Class-C Amp. (Telegraphy)	500	200	—	−45	240	32	12	9300	0.7	—	83
													Class-C Amp. (Telephony)	425	200	—	−60	212	35	11	6400	0.8	—	63
													Grid-Modulated Amp.	500	200	—	−38	120	10	2.0	—	0.5	—	23
829A[1,3]	40	6.3 12.6	2.25 1.12	750	240	7.0	14.4	0.1	7.0	200	N.	7BP	Class-C Amp.-Oscillator	750	200	—	−55	160	30	12	18300	0.8	—	87
													Class-C Amp. (Telephony)	600	200	—	−70	150	30	12	13300	0.9	—	70
													Grid-Modulated Amp.	750	200	—	−55	80	5.0	0	—	0.7	—	24
829B[3] 3E29[3]	40	12.6 6.3	1.125 2.25	750	240	6 7 7	14.5	0.12	7.0	200	N.	7BP	Class-C Amp. (Telegraphy)	500	200	—	−45	240	32	12	9300	0.7	—	83
													Class-C Amp. (Telephony)	425	200	—	−60	212	35	11	6400	0.8	—	63
													Class-B Amp. (Audio)[6]	500	200	—	−18	27/230	—	56[8]	—	0.39	4800	76

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amp.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Con-nec-tions	Typical Operation	Plate Volt-age	Screen Volt-age	Sup-pressor Volt-age	Grid Volt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
HY1269	40	6.3 12.6	3.5 1.75	750	300	5.0	16.0	0.25	7.5	6	M.	T-5DB	Class-C Amp.-Oscillator	750	300	—	− 70	120	15	4	—	0.25	—	63
													Class-C Amp. (Telephony)	600	250	—	− 70	100	12.5	5	35000	0.5	—	42
													Grid-Modulated Amp.	750	300	—	—	80	—	—	—	—	—	20
													Class-AB_2 Amp. (Audio)[6]	600	300	—	− 35	200[7]	—	—	—	0.3	—	80
3D24	45	6.3	3.0	2000	400	10	6.5	0.2	2.4	125	L.	T-9J	Class-C Amp.-Oscillator	2000	375	—	−300	90	20	10	—	4.0	—	140
														1500	375	—	−300	90	22	10	—	4.0	—	105
715-B	50	26/28	—	—	—	—	—	—	—	—	—	—	Class-C Amp. (Telegraphy)	1500	300	—	—	125	—	—	—	—	—	—
5562	45	6.3	3.0	2000	400	8	6.5	0.2	1.8	120	M.	Fig. 54	Class-C Amp. (Telegraphy)	1500	375	—	−300	116	21	12	—	3.6	—	135
													Class-C Amp. (Telephony)	1000	300	—	−200	85	14	10	—	2.0	—	60
HK-57	50	5	5	3000	500	25	7.29	0.05	3.13	200	N.	Fig. 64	Class-C Amp. (Telegraphy)	2000	450	+30	−145	110	2	1	—	0.15	—	166
													Class-C Amp. (Telephony)	2000	450	+30	−145	88	2	1.5	—	0.2	—	135
													Suppressor-Modulated Amp.	2000	450	−190	−240	80	14	2.5	110000	0.6	—	90
RK47	50	10	3.25	1250	300	10	13	0.12	10	—	M.	T-5D	Class-C Amp. (Telegraphy)	1250	300	—	− 70	138	14	7.0	—	1.0	—	120
													Class-C Amp. (Telephony)	900	300	—	−150	120	17.5	6.0	—	1.4	—	87
													Grid-Modulated Amp.	1250	300	—	− 30	60	2.0	0.9	—	4.0	—	25
312A	50	10	2.8	1250	500	20	15.5	0.15	12.3	—	M.	T-6C	Class-C Amp. (Telegraphy)	1250	300	20	− 55	100	36	5.5	—	0.7	—	90
													Class-C Amp. (Telephony)	1000	—	40	− 40	95	35	7.0	22000	1.0	—	65
													Suppressor-Modulated Amp.	1250	—	−85	− 50	50	42	5.0	22000	0.55	—	23
804	50	7.5	3.0	1500	300	15	16	0.01	14.5	15	M.	T-5C	Class-C Amp. (Telegraphy)	1500	300	45	−100	100	35	7.0	34000	1.95	—	110
													Class-C Amp. (Telephony)	1250	250	50	− 90	75	20	6.0	50000	0.75	—	65
													Grid-Modulated Amp.	1500	300	45	−130	50	13.5	3.7	—	1.3	—	28
													Suppressor-Modulated Amp.	1500	300	−50	−115	50	32	7.0	—	0.95	—	28
4D22	50	25.2 12.6	0.8 1.6	750	350	14	28	0.27	13	60	N.	Fig. 50	Class-C Amp. (Telegraphy)	750	300	—	−100	240	26	12	—	1.5	—	135
														600	300	—	−100	215	30	10	—	1.25	—	100
4D32		6.3	3.75									Fig. 51	Class-C Amp. (Telephony)	600	—	—	−100	220	28	10	10000	1.25	—	100
														550	—	—	−100	175	17	6	15000	0.6	—	70
													Class-AB_2 Amp. (Audio)[6]	600	250	—	− 25	100/365	26[7]	70[8]	—	0.45[7]	3000	125
305A	60	10	3.1	1000	200	6	10.5	0.14	5.4	—	M.	T-4CE	Class-C Amp. (Telegraphy)	1000	200	—	−200	125	—	—	—	—	—	85
													Class-C Amp. (Telephony)	800	200	—	−270	125	—	—	—	—	—	70
HY67	65	6.3 12.6	4.5 2.25	1250	300	10	—	0.19	14.5	—	M.	T-5DB	Class-C Amp. (Telegraphy)	1250	300	—	− 80	175	22.5	10	—	1.5	—	152
													Class-C Amp. (Telephony)	1000	300	—	−150	145	17.5	14	—	2.0	—	101
													Grid-Modulated Amp.	1250	300	—	—	78	—	—	—	—	—	32.5
814	65	10	3.25	1500	300	10	13.5	0.1	13.5	30	M.	T-5D	Class-C Amp. (Telegraphy)	1500	300	—	− 90	150	24	10	50000	1.5	—	160
													Class-C Amp. (Telephony)	1250	300	—	−150	145	20	10	48000	3.2	—	130
													Grid-Modulated Amp.	1500	250	—	−120	60	3.0	2.5	—	4.2	—	35
4-65A	65	6.0	3.5	3000	400	10	8.0	0.08	2.1	160[9]	N.	Fig. 48	Class-C Amp. (Telegraphy)	3000	250	—	− 90	115	20	10	—	1.7	—	280
				2500	400								Class-C Amp. (Telephony)	2500	250	—	−150	108	16	8	—	1.9	—	225
				3000	600								Class-B Linear Amp.	2500	500	—	−100	20/230	0/35	6[10]	—	1.8[10]	—	325[7]
				3000	600								Class-AB_2 Amp. (Audio)[6]	1800	250	—	− 35	50/220	0/25	180[8]	—	2.2[7]	20000	270
282A	70	10	3.0	1000	250	5	12.2	0.2	6.8	—	M.	T-4C	Class-C Amp. (Telegraphy)	1000	150	—	−160	100	—	—	—	—	—	33
													Class-C Amp. (Telephony)	750	150	—	−180	100	—	50	—	—	—	50
4E27/ 8001	75	5.0	7.5	4000	750	30	12	0.06	6.5	75	J.	7BM	Class-C Amp. (Telegraphy)	2000	500	60	−200	150	11	6	136000	1.4	—	230
													Class-C Amp. (Telephony)	1800	400	60	−130	135	11	8	125000	1.7	—	178
													Suppressor-Modulated Amp.	2000	500	−300	−130	55	27	3.0	—	0.4	—	35

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES— *Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
HK257 HK257B	75	5.0	7.5	4000	750	25	13.8	0.04	6.7	75 120	J.	7BM	Class-C Amp. (Telegraphy)	2000	500	60	−200	150	11	6.0	—	1.4	—	230
													Class-C Amp. (Telephony)	1800	400	60	−130	135	11	8.0	—	1.7	—	178
													Suppressor-Modulated Amp.	2000	500	−300	−130	55	27	3.0	—	0.4	—	35
828	80	10	3.25	2000	750	23	13.5	0.05	14.5	30	M.	5J	Class-C Amp. (Telegraphy)	1500	400	75	−100	180	28	12	40000	2.2	—	200
													Class-C Amp. (Telephony)	1250	400	75	−140	160	28	12	30000	2.7	—	150
													Grid-Modulated Amp.	1500	400	75	−150	80	4.0	1.3	—	1.3	—	41
													Class-AB_1 Amp. (Audio) [6]	2000	750	60	−120	50/270	2/60	240	—	0	18500	385
RK28	100	10	5.0	2000	400	35	15	0.02	15	—	J.	5J	Class-C Amp. (Telegraphy)	2000	400	45	−100	150	55	13	21000	2.0	—	210
													Class-C Amp. (Telephony)	1500	400	45	−100	135	52	13	21000	2.0	—	155
													Suppressor-Modulated Amp.	2000	400	−45	−100	85	65	13	—	1.8	—	60
													Grid-Modulated Amplifier	2000	400	45	−140	80	20	4.0	—	0.9	—	75
RK48 RK48A	100	10	5.0	2000	400	22	17	0.13	13	—	J.	T-5D	Class-C Amp. (Telegraphy)	2000	400	—	−100	180	40	6.5	—	1.0	—	250
													Class-C Amp. (Telephony)	1500	400	—	−100	148	50	6.5	22000	1.0	—	165
													Grid-Modulated Amplifier	1500	400	—	−145	77	10	1.5	—	1.6	—	40
850	100	10	3.25	1250	175	10	17	0.25	25	15	J.	T-3B	Class-C Amp. (Telegraphy)	1250	175	—	−150	160	—	35	—	10	—	130
													Class-C Amp. (Telephony)	1000	140	—	−100	125	—	40	—	10	—	65
													Grid-Modulated Amplifier	1250	175	—	−13	110	—	—	—	—	—	40
860	100	10	3.25	3000	500	10	7.75	0.08	7.5	30	M.	T-4CB	Class-C Amp.-Oscillator	3000	300	—	−150	85	25	15	—	7.0	—	165
													Class-C Amp. (Telephony)	2000	220	—	−200	85	25	38	100000	17	—	105
813	125	10	5.0	2250	400	22	16.3	0.2	14	30	J.	5BA	Class-C Amp. (Telegraphy)	2250	400	0	−155	220	40	15	46000	4.0	—	375
													Class-C Amp. (Telephony)	2000	350	0	−175	200	40	16	41000	4.3	—	300
													Grid-Modulated Amplifier	2250	400	0	−110	85	2.5	—	—	—	—	75
													Class-B Amp. (Audio) [6]	2500	750	0	−95	35/360	1.2/55	—	—	0.35	17000	650
4-125A 4D21 6155	125	5.0	6.2	3000	400	20	10.3	0.03	3.0	120	N.	5BK	Class-C Amp. (Telegraphy)	3000	350	—	−150	167	30	9	—	2.5	—	375
													Class-C Amp. (Telephony)	2500	350	—	−210	152		9	—	3.3	—	300
													Class-AB_2 Amp. (Audio) [6]	2500	350	—	−43	93/260	0/6	178 [8]	—	1.0	22200	400
4E27A/ 5-125B	125	5.0	7.5	4000	750	20	10.5	0.08	4.7	75	J.	7BM	Class-C Amp. (Telegraphy)	3000	500	60	−200	167	5	6	—	1.6	—	375
														1500	500	60	−130	200	11	8	—	1.6	—	215
														1000	750	0	−170	160	21	3	—	0.6	—	115
RK28A	125	10	5.0	2000	400	35	15	0.02	15	—	J.	5J	Class-C Amp. (Telegraphy)	2000	400	45	−100	170	60	10	—	1.6	—	250
													Class-C Amp. (Telegraphy)	1500	400	45	−100	135	54	10	18500	1.6	—	150
													Grid-Modulated Amp.	2000	400	45	−55	80	18	2.0	—	0.5	—	60
													Suppressor-Modulated Amp.	2000	—	−45	−115	90	52	11.5	30000	1.5	—	60
803	125	10	5.0	2000	600	30	17.5	0.15	29	20	J.	5J	Class-C Amp. (Telegraphy)	2000	500	40	−90	160	45	12	—	2.0	—	210
													Class-C Amp. (Telephony)	1600	400	100	−80	150	45	25	27000	5.0	—	155
													Suppressor-Modulated Amp.	2000	—	−110	−100	80	48	15	35000	2.5	—	53
													Grid-Modulated Amplifier	2000	600	40	−80	80	20	4.0	—	2.0	—	53
4X-150A [9]	150	6.0	2.0	1000	300	15	16.1	0.02	4.7	500	N.	T-9J	Class-C Amp. (Telegraphy)	1000	250	—	−80	200	39	7	—	0.69	—	148
														750	250	—	−80	200	37	6.5	—	0.63	—	110
														600	250	—	−75	200	35	6	—	0.52	—	85
4X-150G	150	2.5	6.25	1250	300	15	16.1	0.02	4.7	165	N.	—	Class-C Amp. (Telegraphy)	1250	250	—	−90	200	20	11	—	1.2	—	195
PE340/ 4D23 [9]	150	5.0	7.5	4000	400	—	11.6	0.06	4.35	120	N.	5BK	Class-C Amp. (Telegraphy)	3000	400	—	−290	200	27	7	—	2.6	—	450
													Class-C Amp. (Telephony)	2500	400	—	−425	180	27	9	—	4	—	350
													Class AB_2 Audio [6]	2500	400	—	−95	284 [7]	7 [7]	—	—	1.8 [7]	19100	460

TABLE XVII—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amp.	Max. Plate Voltage	Max. Screen Voltage	Max. Screen Dissipation Watts	Interelectrode Capacitances (μμfd.) Grid to Fil.	Grid to Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Screen Voltage	Suppressor Voltage	Grid Voltage	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistor Ohms	Approx. Grid Driving Power Watts	Class B P-to-P Load Res. Ohms	Approx. Output Power Watts
AT-340	150	5	7.0	4000	400	—	9.04	0.19	4.16	120	J.	5BK	Class-C Amp.-Oscillator	3000	400	—	−500	165	75	—	—	2.4	—	—
RK65	215	5.0	14	3000	500	35	10.5	0.24	4.75	60	J.	T-3BC	Class-C Amp. (Telegraphy)	3000	400	—	−100	240	70	24	—	6.0	—	510
													Class-C Amp. (Telephony)	2500	—	—	−150	200	70	22	30000	6.3	—	380
4-250A 5D22 6156	250	5.0	14.5	4000	600	35	12.7	0.06	4.5	75	N.	5BK	Class-C Amp. (Telegraphy)	3000	500	—	−180	330	60	10	—	2.6	—	800
													Class-C Amp. (Telephony)	3000	400	—	−310	225	30	9	—	3.2	—	510
													Class-AB_2 (Audio)[6]	1500	300	—	−48	100/485	0/34	192[8]	—	4.7[7]	5400	428
GL-5D24	250	5.0	14.1	4000	350	50	12.7	0.06	4.5	85	N.	5BK	Class-C Amp. (Telegraphy)	Same as 4-250A										GL-5D24
4-400A[9]	400	5.0	14.5	4000	600	35	12.5	0.12	4.7	110	N.	5BK	Class-C Teleg. or Telephony	4000	300	—	−170	270	22.5	10	—	10	—	720
861	400	11	10	3500	750	35	14.5	0.1	10.5	20	N.	T-1B	Class-C Amp. (Telegraphy)	3500	500	—	−250	300	40	40	—	30	—	700
													Class-C Amp. (Telephony)	3000	375	—	−200	200	—	55	70000	35	—	400

[1] Discontinued.
[2] Triode connection—screen grid tied to plate.
[3] Dual tube. Values for both sections, in push-pull. Interelectrode capacitances, however, are for each section.
[4] Terminals 3 and 6 must be connected together.
[5] Filament limited to intermittent operation.
[6] Values are for two tubes in push-pull.
[7] Max.-signal value.
[8] Peak grid-to-grid a.f. volts.
[9] Forced-air cooling required.
[10] Average value.
[11] Two tubes triode connected, G_2 to G_1 through 20K Ω. Input to G_2.

TABLE XVIII—KLYSTRONS

Type	Freq. Range-Mc.	Cathode Volts	Cathode Amp.	Base Connections	Typical Operation	Beam Volts	Beam Ma. (Max.)	Beam Watts (Max.)	Control-Electrode Volts	Reflector Volts	Cathode Ma.	R.F. Driving Power Watts[1]	Output Watts
2K22	4240–4910	6.3	0.44	—	Reflex Oscillator	300	—	—	—	−120/−180	—	—	0.75–0.115
2K25/723A-B	8702–9548	6.3	0.44	Fig. 60	Reflex Oscillator	300	32	—	—	−130/−185	25	—	0.033
2K26	6250–7060	6.3	0.50	Fig. 60	Reflex Oscillator	300	25	—	—	−65/−120	—	—	0.120
2K28[3]	1200–3750	6.3	0.65	Fig. 61	Reflex Oscillator	300[7]	45	—	300	−155/−290	30	—	0.140
2K29	3400–3960	6.3	0.44	—	Reflex Oscillator	300	—	—	—	−75/−180	—	—	0.85–0.106
2K33	23500–24500	6.3	0.65	Fig. 62	Reflex Oscillator	1800[7]	—	—	−20/−100	−80/−220	6	—	0.04
2K34	2730–3330	6.3	1.6	Fig. 58	Oscillator-Buffer *	1900	150	450	−45	—	75	—	10–14
2K35	2730–3330	6.3	1.6	Fig. 58	Cascade Amplifier *	1500	150	450	0	—	75	0.005	5
2K39[3]	7500–10300	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0.	−660	30	—	0.46
2K41	2660–3310	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	+24	−510	60	—	0.75
2K42[3]	3300–4200	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−650	45	—	0.75
2K43[3]	4200–5700	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−320	40	—	0 8
2K44[3]	5700–7500	6.3	1.3	Fig. 59	Reflex Oscillator *	1000	60	75	0	−700	43	—	0.9
2K45	8500–9660	6.3	0.762	—	Reflex Oscillator	300	—	—	—	−95/−145	—	—	0.02–0.032
2K46	2730–3330[1] 8190–10000[2]	6.3	1.3	Fig. 58	Frequency Multiplier *	1500	60	60	−90	—	30	0.01/0.07	0.01–0.07
2K47	250–280[1] 2250–3360[2]	6.3	1.3	Fig. 58	Frequency Multiplier *	1000	60	60	−35	—	50	3.5	0.15
2K48	6900–10850	6.3	0.515	—	Reflex Oscillator	1250	—	—	—	−175/−300	—	—	0.025
2K56	3840–4460	6.3	5.0	Fig. 60	Reflex Oscillator	300	25	—	—	−85/−150	—	—	0.090
3K21[3]	2300–2725	6.3	1.6	Fig. 58	Oscillator-Amplifier *	2000	150	450	0	—	125	1–3	10–20
3K22[3]	3320–4000	6.3	1.6	Fig. 58	Oscillator-Amplifier *	2000	150	450	0	—	125	1–3	10–20
3K23[3]	950–1150	6.3	1.6	Fig. 59	Reflex Oscillator *	1000	90	80	0	−300	70	—	1–2
3K27[3]	750–960	6.3	1.6	Fig. 59	Reflex Oscillator *	1000	90	80	0	−300	70	—	1–2
3K30 (410R)[3]	2700–3300	6.3	1.6	Fig. 58	Oscillator-Amplifier *	2000	150	450	0	—	125	1–3	10–20
6BL6	1250–6000	—	—	—	Reflex Oscillator	350	—	—	+1	0/−400	25	—	—

TABLE XVIII—KLYSTRONS—*Continued*

Type	Freq. Range-Mc.	Cathode Volts	Cathode Amp.	Base Connec- tions	Typical Operation	Beam Volts	Beam Ma. (Max.)	Beam Watts (Max.)	Control-Electrode Volts	Reflector Volts	Cathode Ma.	R.F. Driving Power Watts [4]	Output Watts
6BM6	550–3000	—	—	—	Reflex Oscillator	350	—	—	+ 1	0/−600	20	—	—
707B [5]	1200–3750	6.3	0.65	Fig. 61	Reflex Oscillator	300 [7]	45	—	300	−155/−290	30	—	0.140
SD1103	1250–6000	—	—	—	Reflex Oscillator	350	—	—	+10	0/−400	25	—	—
SD1104	550–3000	—	—	—	Reflex Oscillator	350	—	—	+10	0/−600	22	—	—
QK140	29700–33520	6.3	0.65	—	Reflex Oscillator	2250	—	—	−20/−250	−50/−200	—	—	0.01–0.02
QK159	2950–3275	6.3	0.65	Fig. 63	Reflex Oscillator	300	45	—	300	−100/−175	20	—	0.150
QK226	37100–42600	6.3	0.65	—	Reflex Oscillator	2500	—	—	−20/−200	−50/−200	—	—	0.005
QK227	41700–50000	6.3	0.65	—	Reflex Oscillator	3000	—	—	− 20/−200	−50/−200	—	—	0.005
QK289	27270–30000	6.3	0.65	—	Reflex Oscillator	2250	—	—	−20/−250	−50/−200	—	—	0.01–0.02
QK290	29700–33520	6.3	0.65	—	Reflex Oscillator	2250	—	—	−20/−250	−50/−200	—	—	0.01–0.02
QK291	33520–36250	6.3	0.65	—	Reflex Oscillator	2250	—	—	−20/−250	−50/−200	—	—	.005–.018
QK292	35100–39700	6.3	0.65	—	Reflex Oscillator	2500	—	—	−20/−200	−50/−200	—	—	.005–.01
QK293	37100–42600	6.3	0.65	—	Reflex Oscillator	2500	—	—	−20/−200	−50/−200	—	—	.005
QK294	41700–50000	6.3	0.65	—	Reflex Oscillator	3000	—	—	−20/−200	−50/−200	—	—	.005
QK295	50000–60000	6.3	0.65	—	Reflex Oscillator [8]	3500	—	—	− 20/−200	−50/−200	—	—	—
QK306	18000–22000	6.3	0.65	—	Reflex Oscillator	1800	—	—	−20/−100	−80/−220	—	—	.01–.04
RK726C	2700–2960	6.3	0.44	—	Reflex Oscillator	300	—	—	—	−75/−135	—	—	0.085–0.1
RK5721	3500–12000	6.3	0.58	—	Reflex Oscillator	1000	—	—	+4/+18	−60/−625	—	—	0.1 0.125
RK5976	6250–7460	6.3	0.44	—	Reflex Oscillator	300	—	—	—	−78/−158	—	—	0.85–0.11
RK5981	1245–1460	6.3	0.455	—	Reflex Oscillator	225	—	—	—	−30/−330	—	—	0.04–0.1
RK6043	2950–3275	6.3	0.65	—	Reflex Oscillator	300	—	—	+300	−100/−175	—	—	0.15–0.175
RK6115	5100–5900	6.3	0.44	—	Reflex Oscillator	300	—	—	—	−115/−175	—	—	0.07–0.1
Z-668	21900–26100	—	—	—	Reflex Oscillator *	1700	—	15	—	−1700/−2300	—	—	0.02
5836	1250–6000	—	—	—	Reflex Oscillator	350	—	—	+10	0/−400	25	—	—
5837	550–3000	—	—	—	Reflex Oscillator	350	—	—	+10	0/−600	22	—	—

[1] Input frequency.
[2] Output frequency.
[3] Tuner required.
[4] At max. ratings.
[5] Has demountable tuning cavity.
[6] Cathode current specified on each tube.
[7] G2 and G3 voltage.
[8] Two tubes to cover range.
* Forced-air cooling required.

TABLE XIX—CAVITY MAGNETRONS

Type	Class	Band or Range Mc.	Heater Volts	Heater Amps.	Maximum Ratings Anode KV.	Maximum Ratings Anode Amps.	Maximum Ratings Duty Cycle	Maximum Ratings Input Watts	Typical Operation Anode KV.	Typical Operation Anode Amps.	Typical Operation Field Gauss	Typical Operation Pulse µ Sec.	Typical Operation P.P.S.	Peak Pwr. Output KW.
RK2J22	1	3267–3333	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2250	1.0	1000	265
RK2J23	1	3071–3100	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J24	1	3047–3071	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J25	1	3019–3047	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J26	1	2992–3019	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J27	1	2965–2992	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J28	1	2939–2965	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J29	1	2914–2939	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	2400	1.0	1000	275
RK2J30	1	2860–2900	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J31	1	2820–2860	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J32	1	2780–2820	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J33	1	2740–2780	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J34	1	2700–2740	6.3	1.5	22.0	30.0	.002	600	20.0	30.0	1900	1.0	1000	285
RK2J36	1	9003–9168	6.3	1.3	13.5	12.0	.002	200	11.5	10.0	2500	1.0	1000	15.0
RK2J38	1	3249–3263	6.3	1.25	6.0	8.0	.012	200	4.9	3.0	Pkg.	1.0	2000	5.0
RK2J39	1	3267–3333	6.3	1.25	6.0	8.0	.002	200	5.4	5.0	Pkg.	1.0	2000	8.7
2J42	1	9345–9405	6.3	0.5	5.7	6.5	.001	—	—	—	4800	2.5	—	14
2J42A	1	9345–9405	6.3	0.5	8.0	7.0	.001	—	—	—	6500	2.5	—	35
RK2J48	1	9310–9320	6.3	1.0	16.0	16.0	.002	230	12.0	12.0	4850	1.0	1000	50.0
RK2J49	1	9000–9160	6.3	1.0	16.0	16.0	.0012	180	12.0	12.0	5400	1.0	1000	58.0
RK2J50	1	8740–8890	6.3	1.0	16.0	16.0	.0012	180	12.0	12.0	5400	1.0	1000	58.0
RK2J51	2	8500–9600	6.3	1.1	16	16	.0012	230	15	14	Pkg.	1.0	1000	45
RK2J54	2	3123–3259	6.3	1.5	14.0	15.0	.002	250	11.6	12.5	1400	1.0	2000	45.0
RK2J55	1	9345–9405	6.3	1.0	16.0	16.0	.001	180	12.8	12.0	Pkg.	1.0	1000	50.0
RK2J56	1	9215–9275	6.3	1.0	16.0	16.0	.001	180	12.8	12.0	Pkg.	1.0	1000	50.0
RK2J58	2	2992–3100	6.3	1.5	22.0	15.0	.002	600	10.5	12.5	1450	1.0	2000	50.0
RK2J61A	2	3000–3100	6.3	1.5	15.0	15.0	.002	250	10.7	12.5	1300	1.0	2000	35.0
RK2J62A	2	2914–3010	6.3	1.5	15.0	15.0	.002	250	10.2	12.5	1300	1.0	2000	35.0
RK2J66	2	2845–2905	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J67	2	2795–2855	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J68	2	2745–2805	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J69	2	2695–2755	6.3	1.5	20.0	25.0	.001	400	18.0	25.0	1700	1.0	1000	150
RK2J70	1	3030–3110	6.3	1.25	7.5	15	.002	200	7	8	Pkg.	0.5	1000	20
RK2J71	1	3190–3201	6.3	1.25	5.5	8	.002	100	5	5	Pkg.	1.0	2000	6
3J31	1	23744–24224	6.0	1.9	15.0	14.0	.0005	—	—	—	7600	1.0	—	54
RK4J31	1	2860–2900	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J32	1	2820–2860	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J33	1	2780–2820	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J34	1	2740–2780	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J35	1	2700–2740	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J36	1	3650–3700	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J37	1	3600–3650	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J38	1	3550–3600	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J39	1	3500–3550	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J40	1	3450–3500	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J41	1	3400–3450	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2500	1.0	400	750
RK4J43	1	2992–3019	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J44	1	2965–2992	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
4J50	1	9345–9405	13.6	3.5	23.0	27.5	.004	—	—	—	6300	0.5	—	300
4J52	1	9345–9405	12.6	1.9	16.0	15.0	.002	—	—	—	5000	6.0	—	120
RK4J53	1	2793–2813	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK4J54	1	6875–6775	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J55	1	6775–6675	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J56	1	6675–6575	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J57	1	6575–6475	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J58	1	6475–6375	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
RK4J59	1	6375–6275	12.6	3.75	25.0	35.0	.001	650	17.5	30.0	Pkg.	1.0	1000	200
4J78	1	9003–9168	13.6	3.5	23.0	27.5	.004	—	—	—	6300	0.5	—	300
RK5J26	2	1220–1350	23.5	2.2	31.0	60.0	.002	1800	27.5	46.0	1400	4.0	225	400
RK725A	1	9345–9405	6.3	1.0	16.0	16.0	.001	180	12.0	12.0	5400	1.0	1000	50.0
RK730A	1	9345–9405	6.3	1.1	16.0	16.0	.001	180	13.0	12.0	5400	1.0	1000	40.0
RK5586	2	2700–2900	16.0	3.1	30.0	70.0	.001	1200	28.0	70.0	2700	1.0	400	900
RK5609	4	2425–2475	6.3	3.8	1.7	0.15	CW	200	1.5	0.15	Pkg.	—	—	0.125
RK5657	2	2900–3100	16.0	3.4	32.5	70.0	.001	1300	32.5	70.0	2700	1.0	500	800
RK5982	1	9335–9415	6.3	3.2	15.5	14.5	.001	225	15.5	13.4	Pkg.	4.5	200	75.5
QK174C	3	1990–2110	4.0	3.1	2.2	0.18	—	198	1.85	0.15	Pkg.	—	—	0.07
QK312	4	2425–2475	8.5	3.2	7.0	2.5	CW	3600	5.1	0.56	Pkg.	cw.	cw.	1.5

[1] Fixed-frequency—Pulsed. [2] Tunable—Pulsed. [3] Tunable—C.W.-F.M. [4] Fixed-frequency—C.W.

TABLE XX—TRANSISTORS

No.	Type	Maximum Ratings: Collector Diss. M. Watts	Collector Ma.	Collector Volts	Emitter Diss. M. Watts	Emitter Ma.	Characteristics: Current Amp. Factor	Coll. R. KΩ	Emitter R. Ω	Base R. Ω	Use	Typical Operation: Collector Ma.	Collector Volts	Emitter Ma.	Input Resistance Ohms	Output Load R. Ohms	Power Gain Db.	Noise Figure Db.	Base Ma.	Power Output M. Watts
2A	Pt.-Cont.	120	8	50	—	—	2	—	—	—	General	—	−10	1.0	800	15K	20	—	—	—
2B	Pt.-Cont.	120	8	50	—	—	2	—	—	—	General	—	−10	1.0	800	15K	—	—	—	—
2C	Pt.-Cont.	100	8	50	—	—	2	—	—	—	Switching	4.0	0/−2	3.0	—	—	—	—	—	—
2D	Pt.-Cont.	100	8	50	—	—	2	—	—	—	General	1.0	−15	0	—	—	—	—	—	—
2E	Pt.-Cont.	100	8	50	—	—	2	—	—	—	General	1.0	−15	0	500	10K	20	—	—	—
2F	Pt.-Cont.	120	8	100	—	—	2	—	—	—	Switching	5.0	0/−1.2	3.0	—	—	—	—	—	—
2G	Pt.-Cont.	120	8	100	—	—	2	—	—	—	Switching	5.0	0/−1.2	3.0	—	—	—	—	—	—
2N32	Pt.-Cont.	50	8	40	—	3	2.2	—	—	—	Pulse or Switching	—	−25	0.5	400	31K	21	—	—	—
2N33	Pt.-Cont.	30	7	8.5	—	0.8	—	—	—	—	Oscillator 50 Mc.	3.3	−8	0.3	—	—	—	—	—	1.0
2N34	Jct. PNP	50	8	25	—	8.0	0.98	—	—	—	General	10	−6	1.0	—	—	40	—	0.25	—
2N35	Jct. NPN	50	8	25	—	8.0	0.98	—	—	—	General	10	6	1.0	—	—	40	—	0.25	—
2N36	Jct. PNP	50	8	20	—	—	45	—	—	—	General	—	−6.0	1.0	1000	30K	40	—	0.01	—
2N37	Jct. PNP	50	8	20	—	—	30	—	—	—	General	—	−6.0	1.0	1000	30K	36	—	0.02	—
2N38	Jct. PNP	50	8	20	—	—	15	—	—	—	General	—	−6.0	1.0	1000	30K	32	—	0.05	—
A1698	Pt.-Cont.	120	15	100	—	15.0	—	—	—	200	Switching	—	—	—	—	—	—	—	—	—
CK716	Pt.-Cont.	100	4	40	—	10.0	2.5	—	—	—	General	1.5	−10	0.5	250	15K	18	45	—	3.0
CK721	Jct.-PNP	30	5	20	—	5.0	40	—	—	—	General	2.0	−3	—	—	1250	38	22	0.3	2.8
CK722	Jct. PNP	30	5	20	—	5.0	12	—	—	—	General	0.5	−1.5	—	—	—	30	22	0.2	—
G-11	Pt.-Cont.	100	7	30	—	3.0	2.2	—	—	200	Amp. Oscillator	—	—	—	475	20K	17	57	—	—
G-11A	Pt.-Cont.	100	7	30	—	3.0	2.2	—	—	500	Switching	—	−15	1.0	800	20K	—	—	—	—
M1689	Pt.-Cont.	80	40	50	—	40.0	—	—	—	—	Switching	—		—	800	10K	—	—	—	—
M1725	Pt.-Cont.	200	20	50	—	15.0	2.1	—	—	115	Audio and Carrier	4.0	−5	1.5	195	8K	18	48	—	4.5
M1729	Pt.-Cont.	200	20	50	—	15.0	2.5	—	—	75	Audio and Carrier	5/7	−30	1/2	190	15K	20/18	54	—	50.0
M1752	Jct. PNP	50	5	50	—	5.0	0.98	—	—	240	General	—	—	—	25	13K	—	—	—	—
OC50	Pt.-Cont.	120	25	30	25	—	—	—	—	—	Amp. Oscillator	—	−5	1.5	155	6800	—	43	—	—
OC51	Pt.-Cont.	120	15	100	—	15	2.5	—	—	—	Switching	1.6	−40	0	350	26K	—	—	—	—
PT-2A	Pt.-Cont.	100	10	40	—	5	1.5	10	300	500	Audio Amplifier	—	−30	1.0	300	20K	19	57	—	—
PT-2S	Pt.-Cont.	100	10	40	—	5	2.0	—	—	500	Switching	—	−30	1.0	—	—	—	—	—	—
R1734	Pt.-Cont.	120	—	—	—	—	—	—	—	—	Switching	—	—	—	—	—	—	—	—	—
RD2517	Jct. NPN	50	5	30	—	—	0.93	100	35	500	Audio and R.F.	—	4.5	1.0	—	4500	32	22	—	1.9
RD2520	Jct. NPN	50	5	40	—	—	0.95	500	35	100	Audio and R.F.	—	4.5	1.0	—	4500	34	22	—	2.0
RD2521	Jct. NPN	50	10	40	—	10.0	0.975	300	30	100	Amp. Oscillator	—	4.5	1.0	—	—	37	22	—	—
RD2525	Jct. NPN	25	5	25	—	—	125	200	35	500	Amp. Oscillator	—	4.5	—	—	4500	42	22	—	1.9
RR-14	Jct. PNP	50	5	25	—	—	25	700	30	270	Audio Amplifier	0.5	−1.5	—	—	—	36	22	—	—
RR-20	Jct. PNP	50	5	25	—	—	40	700	30	270	Audio Amplifier	0.5	−1.5	—	—	—	40	22	—	—
RR-21	Jct. PNP	50	5	25	—	—	25	—	—	—	Audio Amplifier	3.0	−15.0	—	—	5000	—	—	—	20
RR-34	Jct. PNP	30	5	20	—	—	10	500	30	270	Audio Amplifier	0.5	−1.5	1.0	—	30K	30	—	—	—
T-21A	Pt.-Cont.	50	5	30	20	5	0.95	500	50	500	General	1.0	−4.5	1.0	500	30K	38	25	—	—
TA-161B	Pt.-Cont.	140	—	—	—	—	2	—	—	120	General	3.0	−20	1.5	300	15K	22	55	—	50.0
TP-01	Pt.-Cont.	100	—	35	—	—	2	—	—	—	General	—	−22.5	0.3/0.8	400	10K	18	—	—	—
X-22	Jct. NPN	50	5	40	—	—	0.90	—	—	—	Audio Switching	—	4.5	1.0	35	—	—	—	—	—
X-23	Jct. NPN	50	5	40	—	—	0.95	—	—	—	Audio Switching	—	4.5	1.0	35	—	—	—	—	—

TABLE XXI — GERMANIUM CRYSTAL DIODES

Type	Use	Max. Inverse Volts	Peak Rectif'd Ma.	Max. Surge Ma.	Max. Reverse µ-Amp.	Max. Average Ma.
1N34 1N34A	General	60	150	500	50 @ 10 V. 800 @ 50 V	40
1N35	[1]	50	60	100	10 @ 10 V.	22.5
1N38 1N38A	100-Volt Diode	100	150	500	6 @ 3 V. 625 @ 100 V.	40
1N39	200-Volt Diode	200	150	500	200 @ 100 V. 800 @ 200 V.	40
1N40[2]	Varistor	25	60	100	50 @ 10 V.	22.5
1N41[2]	Varistor	25	60	100	50 @ 10 V.	22.5
1N42[2]	Varistor	50	60	100	6 @ 3 V. 625 @ 100 V.	22.5
1N43	Varistor	60[4]	125	500	850 @ 50 V.	40
1N44	Varistor	115[4]	100	400	1000 @ 50 V.	40
1N45	Varistor	75[4]	100	400	410 @ 50 V.	40
1N46	Varistor	60[4]	125	500	1500 @ 50 V.	40
1N47	Varistor	115[4]	90	350	410 @ 50 V.	30
1N48 G5[3]	General	85	150	400	833 @ 50 V.	50
1N51 G5C[3]	General	50	100	300	1667 @ 50 V.	25
1N52 G5D[3]	General	85	150	400	150 @ 50 V.	50
1N54 1N54A	Hi Back Resistance	35	150	500	10 @ 10 V.	40
1N55 1N55A	150-Volt Diode	150	150	500	300 @ 100 V. 800 @ 150 V.	40
1N56 1N56A	Hi-Conduction	40	200	1000	300 @ 30 V.	50
1N57	Diode	80	150	500	500 @ 75 V.	40

Type	Use	Max. Inverse Volts	Peak Rectif'd Ma.	Max. Surge Ma.	Max. Reverse µ-Amp.	Max. Average Ma.
1N58 1N58A	100-Volt Diode	100	150	500	800 @ 100 V.	40
1N60	Vid. Det.	25	150	500	30 @ 1.5	50
1N61	Diode	130	150	500	300 @ 100 V.	40
1N63 G5E[3]	General	125	150	400	50 @ 50 V.	50
1N64[1] G5F[3]	Vid. Det.	20	—	—	—	—
1N65 G5G[3]	Hi Back Resistance	85	150	400	200 @ 50 V.	50
1N66[2]	General	60	150	500	800 @ 50 V.	50
1N67	Hi Back Resistance	80	100	500	50 @ 50 V.	35
1N68	Restorer	100	100	500	625 @ 100 V.	35
1N69	General	75	125	400	850 @ 50 V.	40
1N70	General	125	90	350	410 @ 50 V.	30
1N71[2]	Varistor	50[4]	200	1000	300 @ 30 V.	60
1N72 G7[3]	U.H.F.	2	75	—	—	25
1N73[2]	Quad	75	60	100	50 @ 10 V.	22.5
1N74[2]	Quad	75	60	100	—	22.5
1N75	General	125	150	400	50 @ 50 V.	50
CK705	General	60	150	500	800 @ 50 V.	50
CK706	Vid. Det.[1]	40	125	300	—	35
CK707	Restorer	80	100	500	100 @ 50 V.	35
CK708	Restorer	100	100	500	625 @ 100 V.	35
CK710	U.H.F. Mix.	5	75	—	500 @ 2 V.	25

Ratings given are for individual diodes. Average life is over 10,000 hours. Ambient temperature range for all types— − 50° C. to + 75° C. Average shunt capacitance − 0.8 µµfd. Units with A suffix are glass types.

[1] Matched dual diode. [2] Unit has four matched diodes. [3] G.E. designation.

[4] Min. reverse volts for zero dynamic resistance.

TABLE XXII — MINIATURE SELENIUM RECTIFIERS

Manufacturer	Type Number	Max. A.C. Volts	Peak Inverse Volts	Peak Current Ma.	Max. R.M.S. Ma.	Max. D.C. Output Ma.	Rectifier Service
Federal Telephone and Radio Corporation	402D3200	117	380	—	—	50	Half-Wave
"	402D2788 # 402D3150A	117	380	900	220	75	Half-Wave
"	403D2625 403D2625A	117	380	1200	325	100	Half-Wave
"	402D3151	18	—	—	—	100	Half-Wave
"	402D3239A	160	—	—	—	75	Doubler
"	403D3240A	160	—	—	—	100	Doubler
General Electric Co.	6RS5GH2	117	380	650	163	65	Half-Wave
"	6RS5GH1	117	380	750	187	75	Half-Wave
Radio Receptor Company, Inc.	5L1	117	380	—	—	75	Half-Wave
"	5M1	117	380	—	—	100	Half-Wave

Circular plates—discontinued.

The Catalog Section

☆ ☆ ☆

In the following pages is a catalog file of products of the principal manufacturers and the principal distributors who serve the radio field: industrial, commercial, amateur. All firms whose advertising has been accepted for this section have met The American Radio Relay League's rigid standards for established integrity; their products and engineering methods have received the League's approval.

31st EDITION 1954

INDEX OF ADVERTISERS

★ CATALOG SECTION ★

The Radio Amateur's Handbook


We're "expecting"...
You'll notice that our parts catalog in this edition of the handbook is not as extensive as in previous years. The reason is that we're hard at work designing new and exciting additions to our components line.
The components you will find on the following pages are so popular any changes in design would be "gilding the lily".
National
NC
EST. 1914
NATIONAL COMPANY, Inc.
MALDEN, MASSACHUSETTS

NC-88 $129.95*

THE ALL NEW *World Master*

8 Big Features

No Comparable

Receiver Can Offer!

No other receiver at anywhere near the low price offers you these eight "high-priced" features: (1) Calibrated bandspread for 80, 40, 20, 15 and 11 meter bands (large 6″ indirectly-lighted lucite scales). (2) Delayed A.V.C. (3) Higher sensitivity. (4) New miniature tubes used exclusively. (5) Antenna trimmer. (6) Better selectivity. (7) An extra tube (total of eight plus rect.). (8) More compact.

Other wanted features include: Range of 540 kcs. to 40 mcs. in 4 bands. Tuned R.F. stage. Two I.F. stages, 2 audio stages with phono input and 2-position tone control. Built-in-speaker. Separate high frequency oscillator. Sensitivity control. Series valve noise limiter. Headphone jack. Standby-receive switch.

CONTROLS: Ant. trimmer, main tuning, bandspread, sensitivity, receive-standby, band selector, ANL-OFF, tone, AM-CW, pitch, vol.-A.C. OFF.

TUBE COMPLEMENT: 6BA6, r.f.; 6BE6, mixer; 6C4, h.f. osc.; 6BD6, 1st i.f.; 6BD6, 2nd i.f.; 6AL5, 2nd det., AGC, ANL: 12AX7, 1st aud. and c.w. osc.; 6AQ5, aud. output; 5Y3/GT rect.

SW-54 $59.95*

THE *Mighty Midget*

Top Novice and Standby Receiver

COVERAGE: Entire frequency range from 540 kc. to 30 mc. in 4 bands. Voice, music or code.

FEATURES: Sensitive and selective superhet circuit, using new minature tubes. Slide rule general coverage dial with police, foreign, amateur and ship bands clearly marked. Unique plastic bandspread dial is adjustable to assure logging accuracy over entire range. Built-in speaker and power supply.

CONTROLS: Main tuning and Bandspread, On-Off and Volume, Receive-Standby, Bandswitch, AM-CW, Speaker, Phones.

TUBE COMPLEMENT: 12BE6, converter; 12BA6, CW osc. IF amp; 12AV6, 2nd det., 1st ud., AVC; 50C5, audio output; 35Z5, rectifier.

SIZE: 11″ wide, 7″ high, 7″ deep.

NATIONAL COMPANY, INC.

HRO-Sixty $533.50* (less speaker)

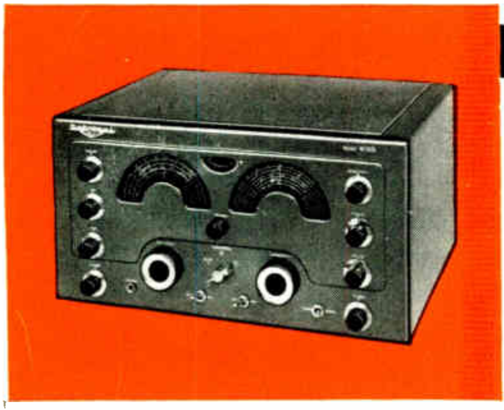
Greatest Tuning Range and Best Sensitivity of Any Commercial Receiver!

COVERAGE: 50–430 kc, 480 kc–35 mc. And 50–54 mc. Voice, CW, NFM (with adapter).

FEATURES: Edge-lighted, direct frequency-reading scale with one range in view at a time. 3 I.F. stages at 456 kcs. employing 12 permeability-tuned circuits on all bands plus one I.F. stage at 2010 kcs. on all frequencies above 7 mcs. Switching is done automatically when coil set is plugged in. Built-in, isolated heavy-duty power supply. Sensitivity of 1 mv. or better at 6 db. sig. noise. Selectivity variable from 8 kc. overall to app. 1200 cps. at 40 db. Current-regulated high frequency oscillator and second converter heaters. Voltage-regulated high frequency oscillator and S-meter amplifier. Negl gible drift after warm-up. Micrometer dial for logging. Provision for crystal calibrator unit. Variable ant. trimmer. Lively S-meter. Min. tubes in front end and high freq. osc. Osc. circuits not disabled when receiver in send position. High-fidelity push-pull audio (± 2db 50 15,000 cps.) with phono jack. BFO switch separated from BFO freq. control. Illumination dimmer control. Accessory socket for Select-O-Ject. Smooth gray finish (table and rack).

CONTROLS: Bandswitch, Oscillator, Tone, Ant. Trimmer, Dimmer, AVC, Limiter, AF Gain, Calibration, CWO, Phasing, Selectivity, On-Off, RF gain, AM-NFM-PHONO.

TUBE COMPLEMENT: 6BA6, 1st r.f.; 6BA6, 2nd r.f.; 6BE6; mixer; 6C4 h.f. oscillator; 6BE6, 2nd high-frequency conv.; 6SG7, 1st i.f.; 6SG7, 2nd i.f.; 6SG7, 3rd i.f.; 6H6 det. & AVC, 6H6, ANL; 6SJ7, 1st audio, 6SN7, phase splitter and S-meter amp.; 6V6GT (2) p.p. audio; 5V4G, rect.; 6SJ7, BFO; OB2, volt reg. 4H4 Osc. Fil. Cur. Reg.



NC-183D $383.50* (less speaker)

'Highest Price' Features Yet Almost $120 Less!

COVERAGE: Continuous from 540 kcs. to 31 mcs. plus 48 to 56 mcs. for 6-meter reception.

FEATURES: Two tuned R.F. stages. Dual conversion on 3 highest bands. Voltage regulated osc. and BFO. Main tuning dial covers range in five bands. Bandspread dial calibrated for amateur 80, 40, 20, 15, 11–10 and 6-meter bands. Bandspread usable over entire range. Six-position crystal filter. New-type noise limiter. High fidelity push-pull audio. Accessory socket for NFM adaptor or other unit, such as crystal calibrator.

CONTROLS: CWO, Switch, CWO pitch, Tone, AF Gain, Main Tuning, Bandspread, Ant. Trimmer, Bandswitch, Send-Receive Phono-Radio, Selectivity, Phasing, Limiter, RF Gain.

TUBE COMPLEMENT: 6BA6, 1st r.f.; 6BA6, 2nd r.f.; 2-6BE6 first and second converter; 3-6BA6 I.F.; 1-6AL5 second det.-AVC; 1-6AH6, AVC amplifier; 1-6SJ7, c.w. osc.; 1-6AL5, limiter; 1-6SJ7 first audio; 1-6J5, phase inverter; 2-6V6GT, audio output; 1-OB2, voltage reg.; 1-5U4G, rect.

NC-125 $199.95* (less speaker)

The Only Receiver With the Famous Select-O-Ject Circuit — Yet Moderately Priced!

COVERAGE: 560 kcs. to 35 mc. in 4 bands. Voice or CW.

FEATURES: Edge-lighted direct-reading scale with amateur, police, foreign, ship frequencies clearly marked. Sensational National Select-O-Ject built-in. Exceptional sensitivity on all bands. S-meter reads S9 to 50 mv. signal. AVC, ANL, jack for phono or NFM adaptor, volt. reg., stabilized osc., audio essentially flat to 10,000 c.p.s.

CONTROLS: Main Tuning, Bandspread, Freq. (SOJ) Boost (SOJ), Send-Receive, Pitch, CWO-MVC-AVC-ANL, AF Gain, Tone, Trimmer, Bandswitch, RF Gain.

TUBE COMPLEMENT: 6SG7 RF amp., 6SB7-Y osc.-mixer, 6SG7 1st IF, 6SG7 2nd IF, 6H6 2nd det-AVC-ANL, 6SL7GT phase shifter, 6SL7GT boost-reject aud. amp., 6SL7GT 1st aud-CWO, 6V6GT aud. output, OD3 VR-150 volt. reg., 5Y3GT rect.

**Slightly higher west of the Rockies*

61 SHERMAN STREET, MALDEN, MASS.

NATIONAL

COIL FORMS

XR-50. These mica-filled bakelite coil forms may be wound as desired to provide a permeability tuned coil. The form winding length is 11/16" and the form winding diameter is ½ inch. The iron slug is ⅜" dia. by ½" long.

XR-51. same but with brass slug.

High-grade ceramic coil forms conforming to JAN specifications. May be wound as desired to provide a permeability-tuned coil. Extra lugs provided.

XR-60 Grooved for #26 wire with iron slug.

XR-61 Grooved for #26 wire with brass slug.

XR-62 Not grooved, winding length 1¼" with iron slug.

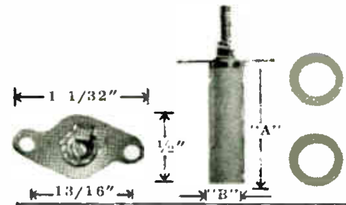
XR-63 Not grooved, winding length 1¼" with brass slug.

XR-70 Grooved for #19 wire with iron slug.

XR-71 Grooved for #19 wire with brass slug.

XR-72 Not grooved, winding length 1" with iron slug.

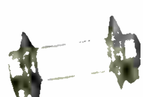
XR-73 Not grooved, winding length 1" with brass slug.



TYPE	CORE	"A" DIM.	"B" DIM.
XR 80	BRASS	1¼"	17/64"
XR 81	IRON	1¼"	17/64"
XR 82	BRASS	1¾"	17/64"
XR 83	IRON	1¾"	17/64"
XR 90	BRASS	1¼"	⅜"
XR 91	IRON	1¼"	⅜"
XR 92	BRASS	1¾"	⅜"
XR 93	IRON	1¾"	⅜"

Small ceramic coil forms designed primarily for high frequency applications and conforming to government specifications. Call form is Grade L4 ceramic (JAN 1-10); base is silver-plated brass; core is brass or iron. Supplied with two nylon rings to separate coils if more than one is wound on same form. Small holes in rings can be used to secure leads.

COUPLINGS



TX-1

TX-1. Leakage path 1".

TX-9

TX-9. This small insulated flexible coupling provides high electrical efficiency when used to isolate circuits. Insulation is steatite. 1⅝" diam. Fits ¼" shaft.

TX-10

TX-10. A very compact insulated coupling free from backlash. Insulation is canvas bakelite. 1-1/16" diam. Fits ¼" shaft.

TX-19

TX-19. A steatite insulated flexible coupling for ¼" shafts. Conservatively rated at 5000 volts peak. Diameter 1⅜", length 1". Length and flashover voltage can be increased by turning collars outboard.

TX-23

TX-23. A deluxe insulated flexible coupling designed for coupling ¼" shafts. Will handle a maximum radial misalignment of 1/16" also 2 degrees maximum angular misalignment.

SAFETY GRID AND PLATE CAPS

SPP-9

SPP-3

SPP-9. Ceramic insulation. Fits 9/16" diameter.

SPP-3. Ceramic insulation. Fits ⅜" diameter. National Safety Grid and Plate Caps have a ceramic body which offers protection against accidental contact with high voltage caps on tubes.

GRID AND PLATE GRIPS

12

24

8

Type 12, for 9/16" Caps.

Type 24, for ⅜" Caps.

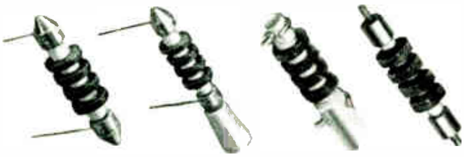
Type 8, for ¼" Caps.

National Grid and Plate Grips provide a secure and positive contact with the tube cap and yet are released easily by a slight pressure on the ear.

Write Dept. 830 for complete catalog or detail drawings of parts advertised

COMPONENTS

PRECISION-WOUND R.F. CHOKES



R-100 R-100U R-100S R-100ST

R-100 R-100U R-100S R-100ST

These RF chokes are identical electrically, but differ in mounting provisions. The R-100 employs pigtail leads; the R-100U has pigtail leads and a removable stand-off insulator; the R-100S has cotter-pin lug terminals and a non-removable stand-off insulator; the R-100ST has a 6-32 threaded stud at each end. These chokes are available in 2.5, 5 and 10 millihenry sizes and are rated at 125 milliamperes.

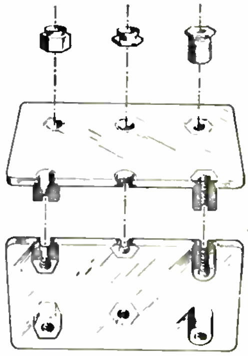
R-300 R-300U R-300S R-300ST

R-300 R-300U R-300S R-300ST

These RF chokes are similar in size to R-100 series but have higher current capacity. The R-300U is provided with a removable stand-off insulator at one end. The R-300S has a non-removable stand-off insulator and cotter-pin lug terminals. The R-300ST has a 6-32 threaded stud at each end. Inductance values of 0.5, 1.0, 2.5 and 5.0 millihenries are available with a current rating of 300 milliamperes. R-300, R-300U, R-300S and R-300ST are identical electrically.

CAPTIVE NUTS

National Captive Nuts of stainless steel may be pressed into aluminum and certain types of brass sheet metal to provide integral flush-mounted tapped holes in a wide variety of sizes. Four basic types have been designed for metal thicknesses of 1/16", 3/32", ⅛", 3/16" and ¼".



R-152

R-33

R-50

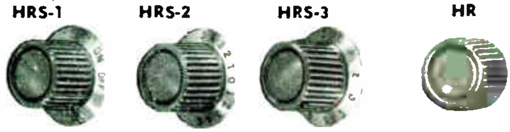
R-33. The R-33 series chokes are 2-section RF chokes available in 10, 50, 100 and 750 microhenry sizes. Also available in this series is a single layer solenoid choke of 1 microhenry inductance. All are rated at 100 milliamperes. The chokes are wound on a ⅝" long form and range in diameter up to 5/16" maximum.

R-50, R-50-1. The R-50 series chokes are 3 and 4-section RF chokes and available in 0.5, 1, 2.5, and 10 millihenry sizes. They are rated at 100 millamperes. The chokes are wound on a 1" long form and have a maximum diameter of 15/32". The 10 millihenry R-50-1 choke is wound on an iron core.

R-152. For use in the range between 2 and 4 Mc. Ideal for high power transmitter stages operated in the 80 meter amateur band. Inductance 4 m.h., DC resistance 10 ohms, DC current 600 ma Coils honeycomb wound on steatite core.

DIALS, MECHANISMS AND KNOBS

HRS-1 HRS-2 HRS-3 HR



HRT-M

AM

N

HRS (gray or black) The HRS series knobs are a popular easy to grip knob. They are molded of high quality plastic and have 1⅜" dia. chrome plated bevel skirts fit ¼" shafts available in the following scales:

HRS-1	ON-OFF	through 30°
HRS-2	5-0-5	through 180°
HRS-3	0-10	through 300°
HRS-4	Single etched line	
HRS-5	0-10	through 180°

HR (gray or black) An HRS type knob without the chrome plated skirt but with a white dot for spotting relative control settings.

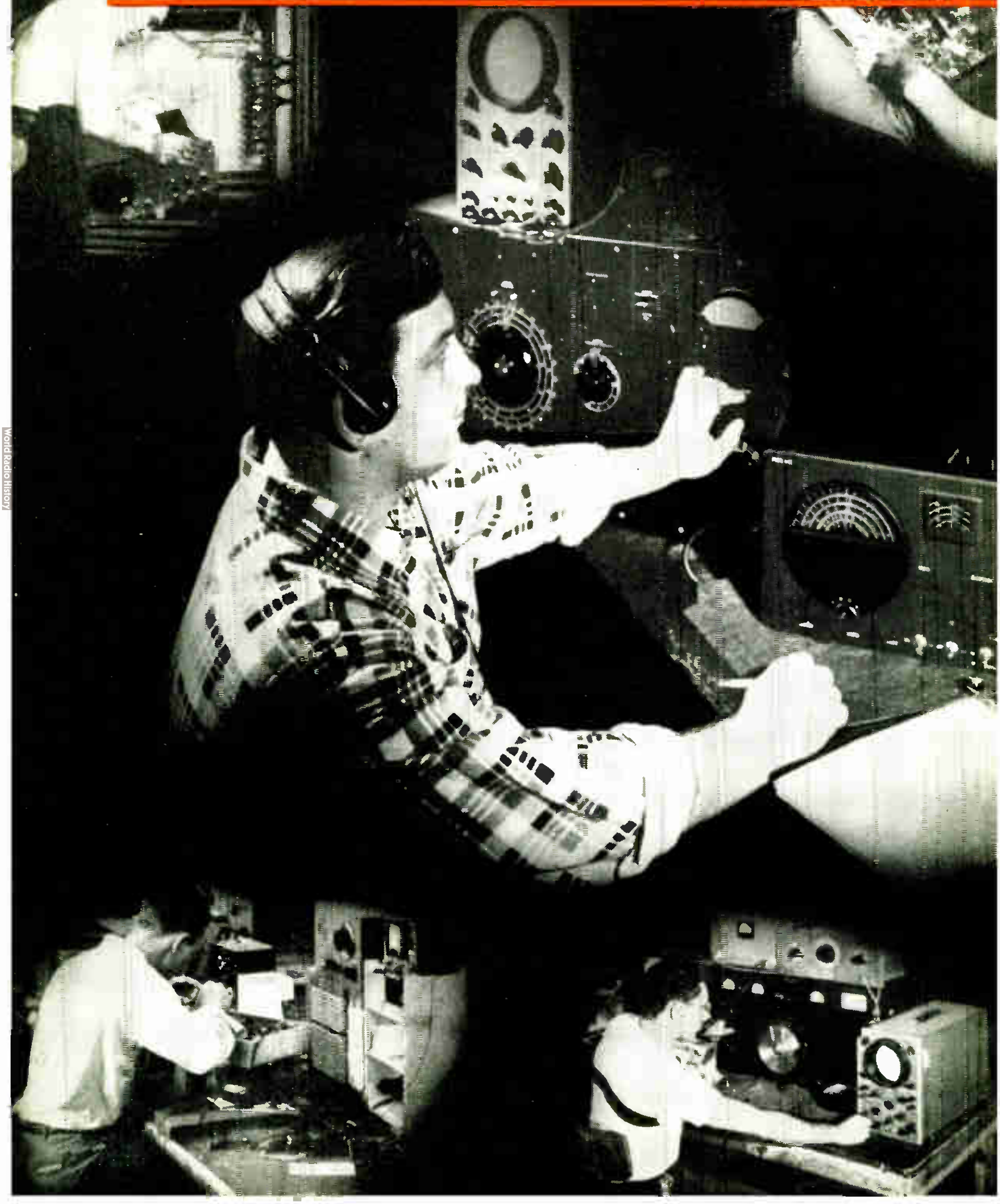
HRT-M. This smaller version of the HRT — now available in choice of gray or black — is 1-7/16" in diameter.

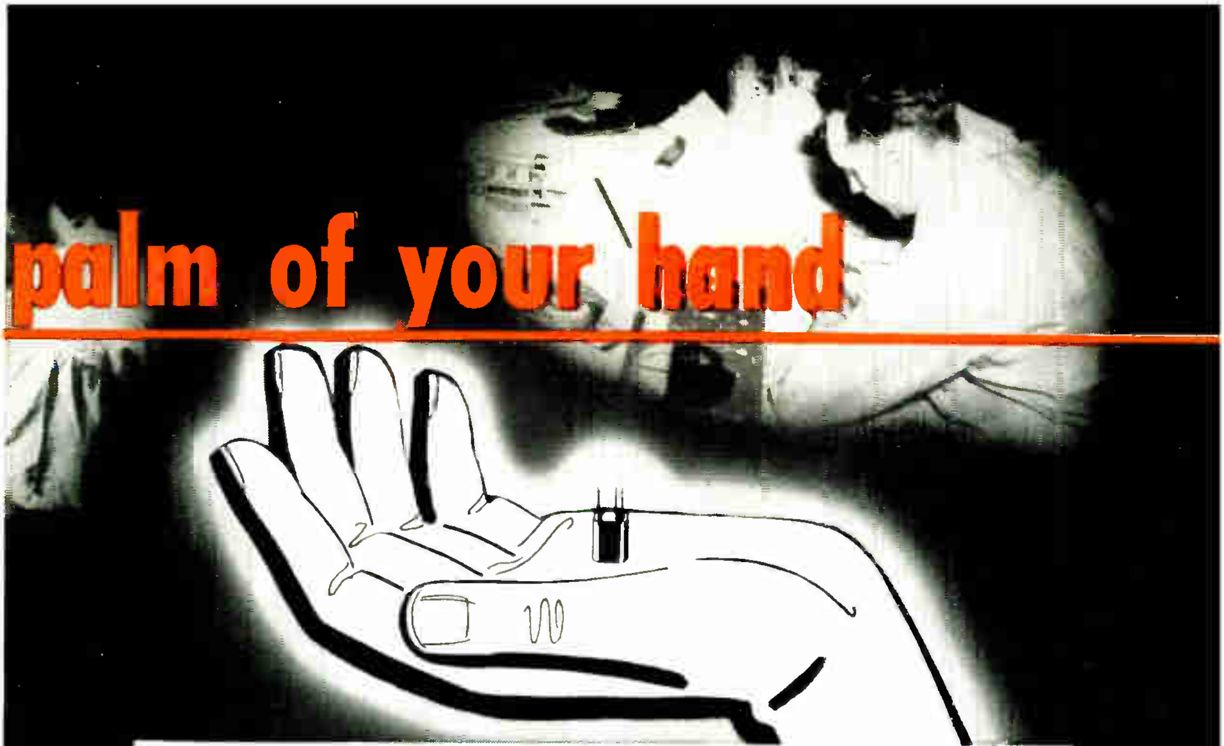
AM Dial. The original "Velvet Vernier" mechanism in a metal skirted dial 3" in dia. ratio 5 to 1. It is available with 2, 3, 4, 5 or 6 scale and fits ¼" shaft. Mechanisms also available separately.

N Dial. The four-inch N and AD Dials have engine divided and die stamped scales respectively. The N Dial has a decimal vernier; the AD Dial employs a pointer. The planetary drive has a ratio of 5 to 1, and is contained within the body of the dial. 2, 3, 4, 5 or black scale. Fits ¼" shaft. **Specify scale.**

HRM KNOB. This straight knurl brass satin chrome finished knob with arrowhead fits ¼" shaft. See catalog for description.

Tomorrow...in the





That tiny metal "spider" may change the very shape of communications in a year or so. It's a transistor, of course, the device that's hailed as the successor to the vacuum tube. And, as you would expect, Hallicrafters is among the leaders in experimentation with this remarkable electronic phenomenon.

Transistors aren't ready yet. To date it has been impossible to design them with the degree of accuracy, electronically speaking, or the sensitivity of the vacuum tube. Probably they will never completely replace tubes, anyway. In most applications their use is not justified. But where space is at a premium, where weight and power must be limited, as in the case of portable equipment or aircraft communications, they will be invaluable.

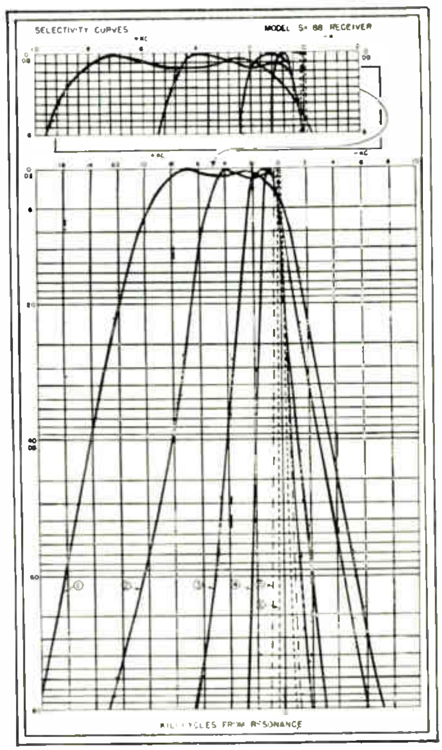
Transistor research is just one more example of a long Hallicrafters tradition. It is a tradition of engineering leadership unmatched in the field. And this is important to you as a radio amateur, because the best of Hallicrafters engineering thinking is lavished on the equipment that goes to you—our first and most important customer.

Take a look at the equipment and specifications on the succeeding pages. You'll see what we mean.

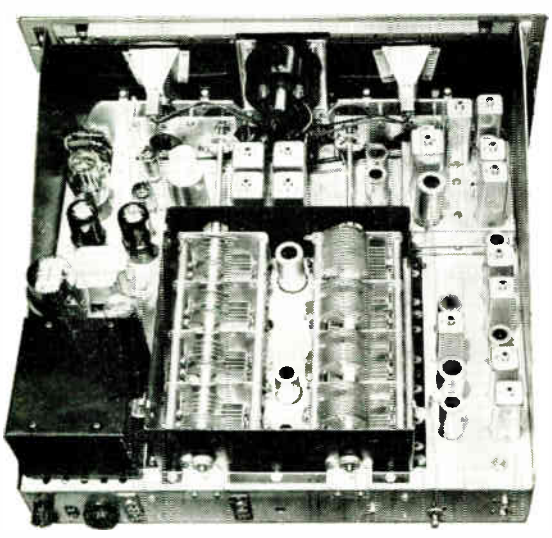
hallicrafters

engineering for tomorrow

Only the
brings you all these features



These Curves Tell the Story
Compare with Any Other Receiver!



hallicrafters SX-88

—everyone a necessity today!

1. Full precision gear drive for main and band spread tuning.
2. Six position Band Width Control (selectivity) from 250 cycles to 10 kc.
3. 10 watt inverse feed back and push-pull audio output.
4. Exhalted B.F.O. for tops in single side band reception.
5. Buffer amplifier in B.F.O. circuit.
6. Antenna trimmer.
7. Amplified and delayed A.V.C.
8. Local oscillator circuits individually temperature compensated for each band.
9. Built-in 100 kc calibration crystal.
10. Second conversion oscillators crystal controlled.
11. A.V.C. operates for CW reception.
12. Inertia tuning (fly wheels both dials).
13. Full frequency coverage from 535 kc to 33.3 Mc.
14. Calibrated electrical bandspread 160, 80, 40, 20, 15, 11 and 10 meters.
15. Logging scales on each tuning shaft.
16. Dial locks on each tuning shaft.
17. Tuning dial indicators resettable from front panel for maximum calibration accuracy.
18. Auxiliary AC socket on rear of chassis.
19. Illuminated band-in-use indicator.
20. Illuminated S meter.
21. Dual S meter calibration S units and microvolts.
22. Auxiliary power socket plus .6 amps at 6.3 volts and 10 ma at 150 volts for accessories.
23. Standard 8¾" by 19" panel for rack mounting if desired.
24. 50 kc i-f output jack via cathode follower for teletype converter, oscilliscopes, etc.
25. Five position response control (tone control).
26. Two r-f stages.
27. 17 tubes plus voltage regulator, current regulators and rectifier.

FRONT PANEL CONTROL

Main tuning.
Bandspread.
Band Selector 6 positions.
Volume: 0-10 and AC/off.
Band width in kc: 10, 5, 2½, 1¼, .5 and .250
Pitch: (B.F.O.) +5-0-5.
Response: Bass Boost, High Fidelity, Normal, Communications.
Antenna trimmer +5-0-5.
Sensitivity 0-10.

FRONT PANEL TOGGLE SWITCH

Noise limiter on/off.
A.V.C. on/off.
Calibrator on/off.
Receiver standby.
C.W.—AM—SSSC (single side band suppressed carrier).

CHASSIS REAR

Speaker terminals 3.2/8/500-600 ohms.
Antenna terminals 52-600 ohms.
AC Accessory socket 117 volts at 250 watts.
Power socket—Octal for external power supply to receiver, such as batteries, and in addition this socket supplies 6.3 volts at 600 ma and 150 dc at 10 ma for future accessories.
Audio Input—phono jack.
Fuse holder for AC power circuit.

FREQUENCY RANGE (Main tuning dial)

Band 1—535 to 1710 kc.
Band 2—1690 to 3080 kc.
Band 3—2980 to 5570 kc.
Band 4—5370 to 10,000 kc.
Band 5—9.8 to 18.3 Mc.
Band 6—17.8 to 33.3 Mc.

SENSITIVITY

Bands 2-6—1 microvolt for ½ watt output.
1 microvolt for 10 db signal to noise ratio.
Band 1 —10 microvolts for ½ watt output.

IMAGE REJECTION

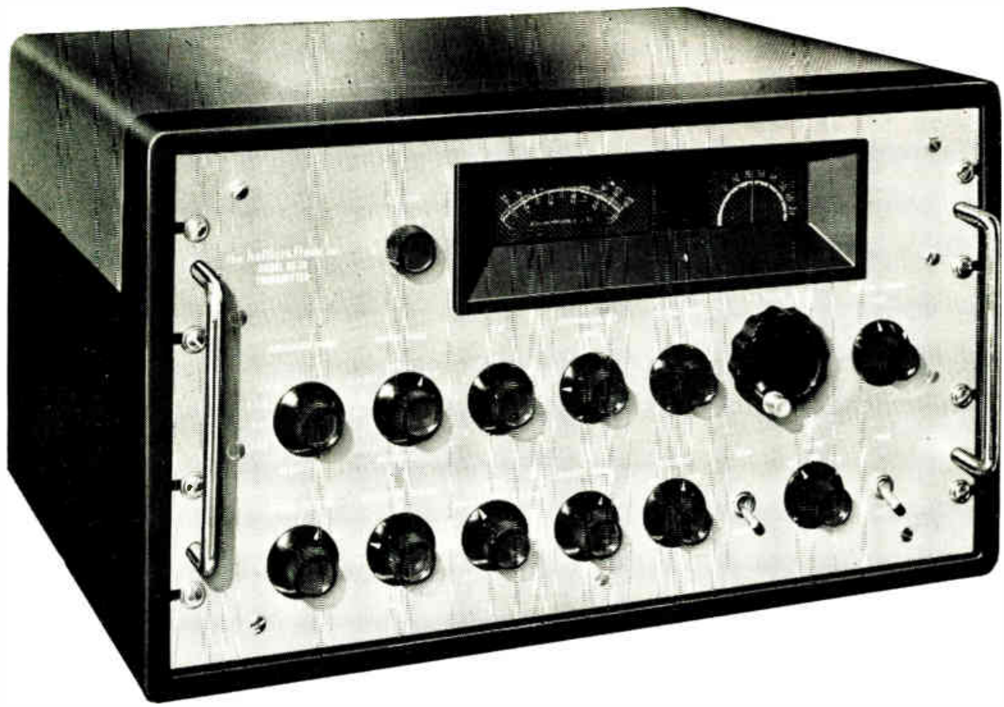
Not less than 80 db on frequencies lower than 20 Mc.
Not less than 60 db on frequencies from 20 to 30 Mc.

SPURIOUS RESPONSES (IF and oscillator tweets)

Not less than 80 db except at 1700 kc where it is not less than 50 db.

BAND WIDTH (Selectivity)

Position	6 db (nose)	60 db (skirts)
10 kc	10 kc	21 kc
5 kc	5 kc	15 kc
2.5 kc	2.5 kc	7.5 kc
1.25 kc	1.25 kc	3.75 kc
.500 kc	500 cps	1.50 kc
.250 kc	250 cps	850 cps



hallicrafters

model HT-20
AM-CW transmitter

This Hallicrafters 100 watt AM-CW Transmitter is the modern successor to the HT-9 known throughout the world for reliability, ruggedness, flexibility and lowest cost for maximum dependable watts per dollars.

Performance: T.V.I. suppressed—completely shielded and filtered rf compartment plus built-in low-pass 52 ohm coaxial line output filter provides 90 db or greater suppression of all frequencies higher than 40 Mc. 100 watt AM phone output.

Frequency Coverage: Continuous coverage from 1.79 to 30 Mc.

Components: Heavy duty commercial type power and modulation transformers. All parts rated for commercial service conditions.

Controls: Full band switching. No plug-in coils—choice of 10 crystals—all controls on front panel.

Tubes: Seven rf and audio tubes plus 5 rectifiers.

Physical Data: Cabinet size—20 inches long, 12½ inches high, 17¼ inches deep—panel size for rack mounting—19 x 10½ inches. Shipping wt. 130 lbs. For 105-125 V. 60 cycle.

Model HT-20 Transmitter $449.50

hallicrafters

model SX-73
the finest in versatility—

Here, from Hallicrafters world-famous short wave laboratories, is a superb communications receiver —the SX-73, proud successor to so many famous top-quality Hallicrafters receivers. Absolutely without equal in its combination of ruggedness, sensitivity, stability, selectivity, resettability, and image and i-f rejection. Based on an original design developed by Hallicrafters for the armed forces for universal use all over the world, this new receiver will surpass all others in versatility, dependability, performance and value.

Performance: Continuous frequency coverage 540 kc to 54.0 Mc. Two r-f, two i-f stages. Dual conversion above 7 Mc; second beat oscillator is crystal controlled. Choice of six pretuned crystal controlled channels in range 1.5 to 30 Mc. Single tuning knob turns main and bandspread dials (6 to 1 ratio between the two); 50 to 1 tuning ratio. Resettability accurate to within 30 cycles per megacycle. Selectivity variable 14.5 kc to 300 cycles at 6 db down. Sensitivity less than 2 microvolts for .5 watts output. Signal to noise ratio 10 db for 2 mv input. Image rejection 80 to 120 db. I-f rejection not less than 60 db. AVC circuit will hold up to one volt without overload. Series type noise limiter. Carrier level meter. Audio response plus or minus 1½ db from 300 to 3500 cycles.

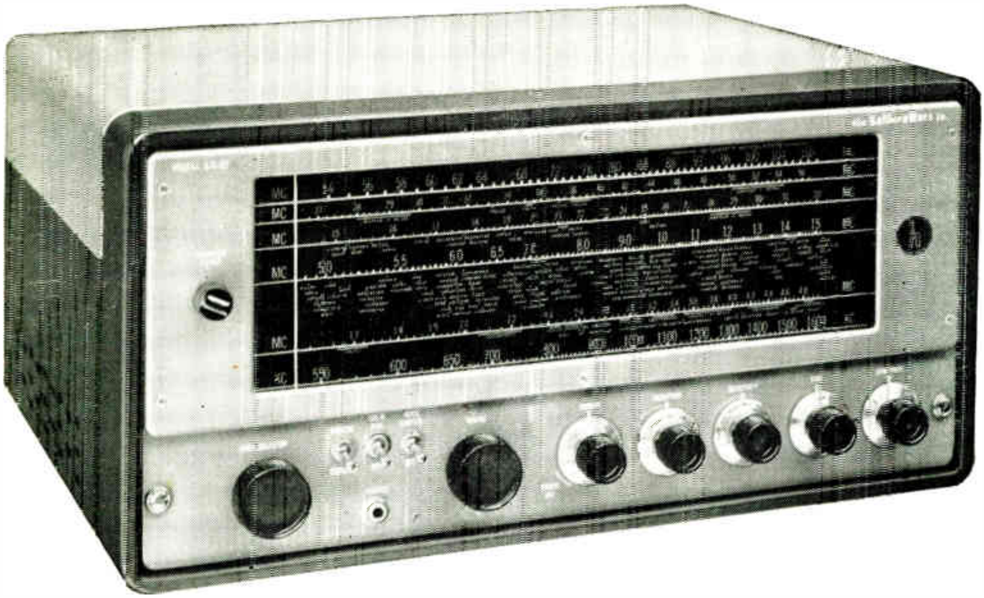
Controls: Tuning knob with dial lock; Band Selector 540-1350 kc, 1.35-3.45 Mc., 3.45-7.00 Mc., 7.00-14.4 Mc., 14.4-29-7 Mc., 29-7-54.0 Mc; r-f Gain and AC on/off BFO Pitch, Xtal Phasing, 6-pos. Xtal Selectivity, 6-pos. Xtal fixed-frequency channel selector, a-f Gain, Xtal tuning Vernier; Rec./Standby, BFO, AVC, and ANL switches; BFO injection control and carrier meter adj. on rear.

Physical Data: Two-tone gray steel cabinet with satin chrome trim. Piano hinge top. Size 20 in. wide, 11 in. high, 18½ in. deep.

External Connections: Antenna Input 50 to 200 ohms throughout tuning range. Output 600 and 50 ohms. For 50/60 cycle current at 75, 105, 117, 130, 190, 210, 234, or 260 volts.

17 tubes plus voltage regulator, current regulator and rectifier.

Model SX-73—Use R-46 Speaker . . $\$975^{00}$



hallicrafters

model SX-62
all-wave high fidelity

The world's finest receiver for the All-Wave listener. Unequalled in coverage and performance on all wave bands—Standard Broadcast, Short-Wave or FM. Continuous coverage from 540 kc to 109 Mc. Having basically the same chassis as a fine communications receiver, the SX-62 provides communications-receiver performance in simplified form. A single tuning control covers the wide-vision dial. Only one band lights up at a time—you always know just where you are tuning. In addition a 500 kc crystal calibration oscillator is built in, enabling you to adjust the dial pointer to show the exact frequency being tuned at any time.

Performance: Continuous AM reception 540 kc to 109 Mc; FM band 27-109 Mc. Temperature compensated, voltage regulated. Two RF, three IF stages; dual IF channels (455 kc and 10.7 Mc). Audio flat 50-15,000 cycles; 10 watt push-pull output.

Controls: Band Selector 540-1620 kc. 1.62-4.9 Mc, 4.9-15 Mc, 15-32 Mc, 27-56 Mc, 54-109 Mc; Receive/Standby, Calibration Osc. On/Off, Noise Limiter, Tuning, AF Gain, Phono/FM/AM/CW, six-position Selectivity, four-position Tone, RF Gain, Calibration Reset.

Physical Data: Satin black steel cabinet with satin chrome trim. Top opens on piano hinge. Cabinet 20" wide by 10¼" high by 16" deep.

External Connections: Doublet or single wire antenna. 500 and 5000-ohm outputs. Phone jack. Phonograph input jack. Socket for external power and Remote control connections. 105-125 V. 50/60 cycle AC line.

14 Tubes plus Voltage Regulator and Rectifier: Two 6AG5 RF Amps., 7F8 Conv., 6SK7 IF Amp., 6SG7 IF Amp., 6SG7 IF Amp., 6SG7 FM Limiter and AM Det., 6H6 FM Det., 6J5 BFO, 6H6 ANL, 6SL7 AF Amp., two 6V6 Push-Pull Output, 6C4 Calibration Osc., VR-150 Regulator, 5U4G Rectifier.

Universal Model SX-62U: Same as above only for 115/250 volts, 25/60 cycle AC.

Model SX-62 or SX-62U $349.95

hallicrafters

model SX-71
command performance

From the Hams at Hallicrafters to Hams everywhere comes this top-performing receiver in the medium price class. Extra sensitivity, selectivity, and stability, definitely superior image rejection with double superheterodyne circuit, plus built-in Narrow Band FM reception. Extra wide dials for main and bandspread tuning. Surpasses in ham performance many receivers priced considerably higher.

Performance: Continuous AM reception from 538 kc to 34 Mc, and 46 to 56 Mc. Built-in limiter and balanced detector stages for hiss-free NBFM reception. Double conversion (2075 and 455 kc i-f channels) gives image rejection of better than 150 to 1 at 28 Mc. Temperature compensated, voltage regulated. One r-f, two conversion, and 3 i-f stages yield high gain for sensitivity of .7 microvolts with 50 milliwatts output. Audio peaked for communications frequencies, with 3 watt output.

Controls: Band Selector 538-1650 kc, 1600-4800 kc, 4.6-13.5 Mc, 12.5-34 Mc, 46-56 Mc. Separate main and Bandspread tuning controls; bandspread dial calibrated for 80, 40, 20, 15, 10, and 6 Meter Bands. BFO Pitch 3-position Selectivity, Crystal Phasing, Tone, a-f Gain and r-f Gain controls. ANL, BFO, and Receive/Send switches. "S" Meter adjustment on rear.

Physical Data: Satin black steel cabinet with chrome trim. Piano hinge top. Size 18½" wide by 8⅞" high by 12" deep. Ship. wt. 33 lbs.

External Connections: Use doublet or single wire antenna. 500 and 3.2 ohm outputs for separate speaker. Phone jack. Socket for external power supply. Connections for remote control. For 105-125 volts 50/60 cycle AC.

11 Tubes plus Voltage Regulator and Rectifier: 6BA6 r-f Amp., 6C4 Osc., 6AU6 Mixer, 6BE6 2nd Conv., three 6SK7 i-f Amps., 6H6 ANL and delayed AVC, 6SC7 BFO and a-f Amp., 6AL5 Det., 6K6GT Output, VR-150 Reg., and 5Y3GT Rect.

Universal Model SX71U: Same as above only for 115/250 volts, 25/60 cycle AC.

Model SX71 or SX71U $249.95

hallicrafters

model S-76 double superhet

Double conversion receiver, double superhet with 50 kc second i-f and 4-inch "S" Meter.

Performance: Continuous coverage 538-1580 kc and 1.72-32 Mc. Double conversion eliminates images. 50 kc second i-f gives excellent "skirt" selectivity with "nose" selectivity variable from 5.6 kc down to 500 cycles. Temperature compensated, voltage regulated. One r-f, two conversion, and two i-f stages. 2½ watts output.

Controls: Band Selector 538-1580 kc, 1.72-49 Mc, 4.6-13 Mc, 12-32 Mc; Separate Main and Band-spread tuning; bandspread calibrated for 80, 40, 20, 15, 11, 10 meters; five-position Selectivity with phono switch built in; BFO Pitch; full-range Tone; AVC, BFO, ANL, Rec./Standby switches. "S" Meter adjustment on rear.

Physical Data: Satin black steel cabinet with plastichrome skirts. Piano hinge top. Size 18½" wide, 8⅞" high, 9½" deep. Ship. wt. appr. 46 lbs.

External Connections: Use doublet or single wire antenna. 500 or 3-2 ohm outputs. Phone jack. Phono input jack. Connections for external power and remote control. Mounting holes provided for coax connector. For 105-125 volts 50/60 cycle AC.

9 Tubes plus Regulator and Rectifier: 6CB6 r-f Amp., 6AU6 1st Conv., 6C4 Osc., 6BA6 1st i-f, 6BE6 2nd Conv., 6BA6 2nd i-f, 6AL5 Det., ANL, 6SC7 BFO, 6K6GT Output, VR-150 Reg., 5Y3GT Rect.

Model S76-AC $199.95

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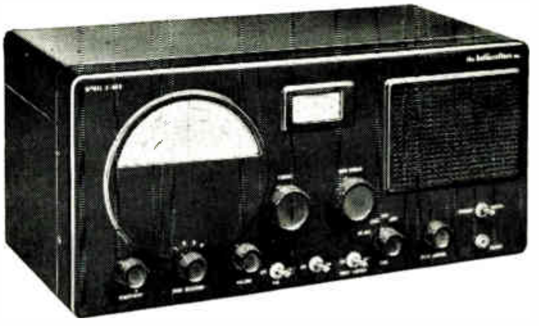
hallicrafters

model S-40B ham favorite



Superior performance. Complete with PM speaker.

Performance: AM reception 540 kc to 43 Mc. Temperature compensated oscillator. One RF and two IF stages. Audio response to 10,000 cycles.

Controls: Band Switch 540-1700 kc, 1700-5300 kc, 5.3-15.7 Mc, 15.7-43.0 Mc. Main tuning in Mc; bandspread dial has arbitrary scale. AF and RF Gain controls; AVC, BFO, and Noise Limiter switches; three-position Tone, BFO Pitch, and Receive/Standby controls.

Physical Data: Satin black steel cabinet. Size 18½" wide by 8⅞" high by 9½" deep. Ship. wt. 32 lbs.

External Connections: Doublet or single wire antenna. Phone jack. S-40 uses 105-125 V. 50/60 cycles AC only. S-77A uses 105-125 V. DC or 50/60 cycle AC.

7 Tubes plus Rectifier: (in S-40B) 6SG7 RF Amp., 6SA7 Conv., two 6SK7 IF Amps., 6H6 ANL and AVC, 6SL7 BFO and Det., 6K6GT Output, 5Y3GT Rectifier.

Model S-40B $129.95

Model S-77A $129.95

hallicrafters

model S-53A
top performance —small size

Unquestionably the finest small communications receiver built. Several steps better than the S-38C but not as good as the S-40B. Complete in itself, with built-in PM speaker.

Performance: Coverage 540-1600 kc, 2.6-31 Mc plus 48-54.5 Mc. Two stages IF amplification.

Controls: Main tuning in Mc; separate bandspread dial with logging scale plus Mc calibration for 48-54.5 Mc band; Receive/Standby switch; Band switch 540-1630 kc; 2.5-6.3 Mc, 6.3-16 Mc, 14-31 Mc, and 48-54.5 Mc; AM/CW; RF Gain, Noise Limiter, AF Gain, two-position Tone; Speaker/Phones switch on rear. Illuminated dial.

Physical Data: Satin black steel cabinet with chrome trim. Top opens on piano hinge. Size 12⅞" wide by 7" high by 7¾" deep. Ship. wt. 19 lbs.

External Connections: Doublet or single wire antenna. Phone tip jacks. Phonograph input jack. 105-125 V. 50/60 cycle AC line.

7 Tubes plus Rectifier: 6C4 Osc., 6BA6 Mixer, two 6BA6 IF Amps., 6H6 Det., AVC and ANL, 6SC7 BFO and AF Amp., 6K6GT Output, 5Y3GT Rectifier.

Mode S-53A . . . . $\$99^{95}$

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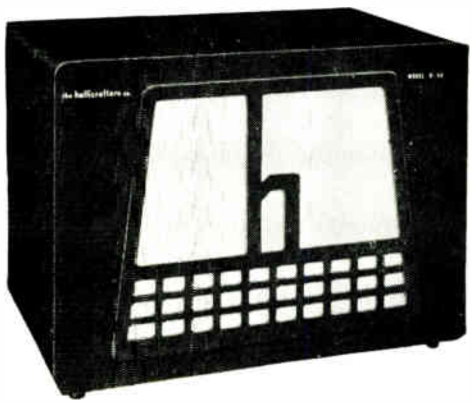
hallicrafters

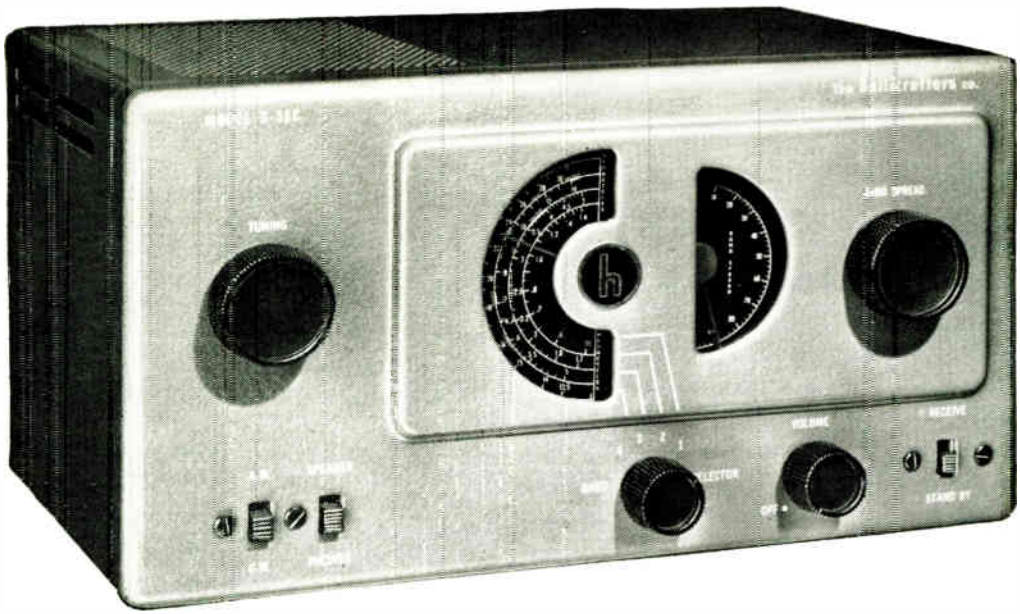


model R-46 communications speaker

Matching 10" PM speaker for use with Hallicrafters Communications receiver SX-71, SX-73, SX-62, or S-76. 80 to 5,000 cycle range. Matching transformer with 500/600-ohm input. Speaker voice coil Impedance, 3.2 ohms.

Black steel cabinet matches SX-71 and other Hallicrafters cabinets. Cloth covered metal grill. 15" x 10⅞" x 10⅞" deep. Ship. wt. 17 lbs.

Model R-46 Speaker . . . . $\$24^{95}$



hallicrafters

famous S-38C–
biggest buy in SW

The lowest priced communications receiver on the market . . . with many features found in much higher priced sets. Standard Broadcast plus three Short-Wave bands. Built-in PM speaker.

Performance: Continuous AM reception 540 kc to 32 Mc. Maximum sensitivity and selectivity from expertly engineered chassis.

Controls: Main tuning in Mc; separate electrical bandspread dial with arbitrary scale; Speaker/Phones, AM/CW switches; Band Switch 540-1650 kc, 1.65-5 Mc, 5-14.5 Mc, 13.5-32 Mc; AF Gain, Receive/Standby.

Physical Data: Steel cabinet in gray hammer-tone finish. Size 12⅞" wide by 7" high by 7¾" deep. Ship. wt. 14 lbs.

External Connections: Doublet or single wire antenna. Phone tip jacks. 105-125 V. DC or 50/60 cycle AC.

4 Tubes plus Rectifier: 12SA7 Conv., 12SK7 IF Amp. and BFO, 12SQ7 Det. and AVC, 50L6GT Output, 35Z5GT Rectifier.

220-Volt Line Cord: Available separately. Works for AC or DC.

Model S-38C $59.95

Line Cord for 220 V. Operation . . . $2.00

hallicrafters

model ST-83 finest hi-fi FM-AM tuner

This AM/FM Super-Fidelity unit carries the UL seal of approval and meets the F.C.C. specifications on oscillator radiation. Phono inputs, built-in pre-amp., accessory inputs for TV, tape recorders, etc. Dual outputs; medium and low impedance, tone controls; bass 12 db, treble 12 db.

Accessory power sockets dual at 200 watt 117 volts each. Tubes 6CB6 FM r-f amplifier, 12AT7 FM osc. converter, 6CD6 AM r-f amplifier, 6BE6 AM osc. converter, 6BA6 1st i-f amplifier 10.7 Mc, 6BA6 2nd i-f amplifier 455 kc and 10.7 Mc, 6BA6 3rd i-f amplifier, 6AL5 FM detector, 6AV6 AM detector and phono pre-amplifier, 6C4 cathode follower, 12AU7 audio tone control amplifier, 6AX5 rectifier.

Black steel with silver finish trim and chrome lite base. 14" x 17½" x 9½" deep. Ship. wt. 18 lbs. Ten tubes plus rectifier.

For 105/125 V. 50/60 cycle AC . . $129.95

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hallicrafters

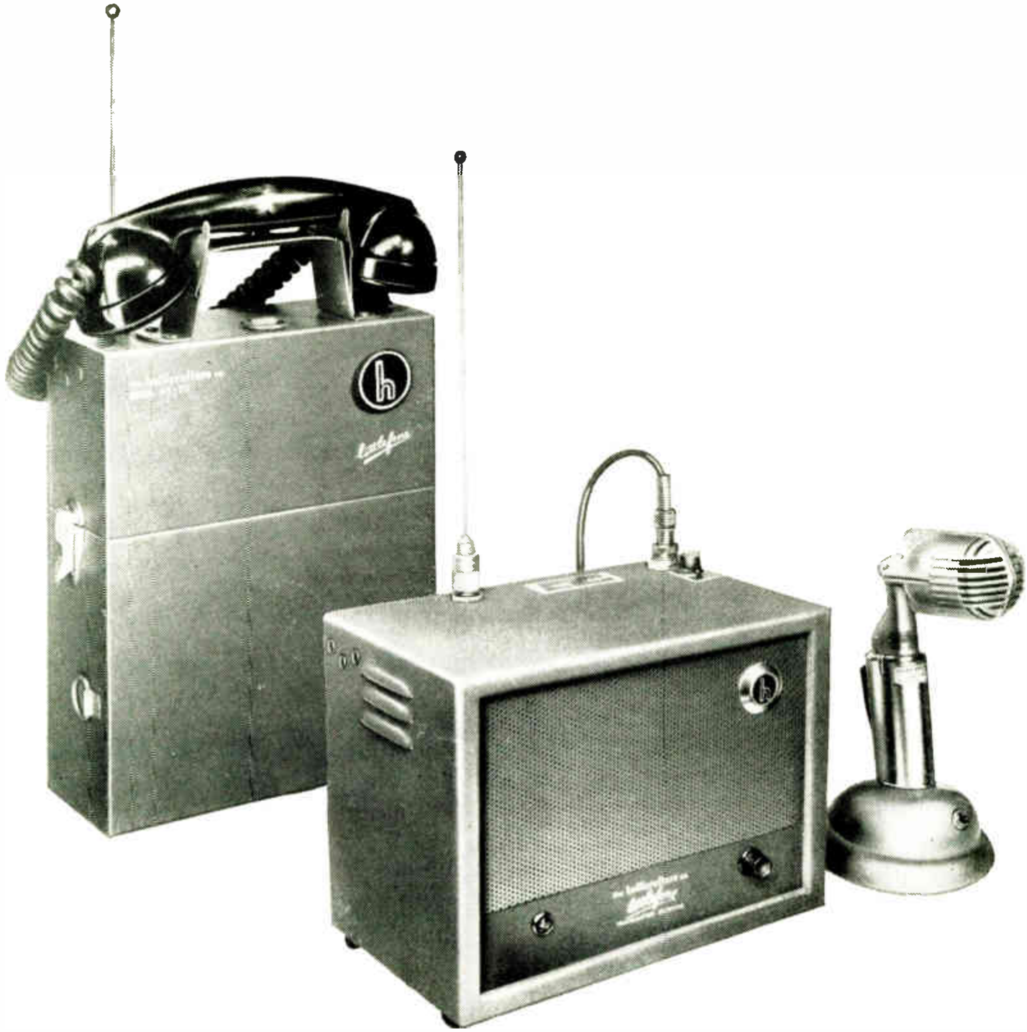
model A-84 widest range hi-fi amplifier

The perfect mate for any AM/FM tuner. Exclusive output transformer giving widest range ever produced. Frequency range, 10 to 100,000 cycles per second at 10 watts (with perfect uniformity) and harmonic distortion of less than 0.25% at 10 watt level. Power output of 15 watts maximum.

Mineral oil impregnated coupling condensers, power supply input condenser oil filled.

Chrome lite chassis base. 13½" x 7⅝" x 13½" deep. Ship. wt. 26 lbs. All five tubes triode.

For 105/125 V. 50/60 cycle AC . . $99.50



hallicrafters

littlefone – portable radio-telephone

The Littlefone series of equipment are FM two-way radio telephone units operating at 25-50 Mc or 144-174 Mc. Both the receiver and transmitter are crystal controlled and a total of 22 sub-miniature tubes are used. The complete portable model with antenna and telephone hand-set weighs only fourteen pounds and will operate for more than eight hours on the self-contained rechargeable storage batteries. Models for AC power line and 6/12 volts DC operation employ the same r-f chassis as the portable units but an audio power output stage is added to drive the loud speaker. Adjustable squelch controls are available on all models. Power outputs 2 watts on 25-50 Mc and 1 watt on 144-174 Mc. Lower powered wet and dry battery models also available.

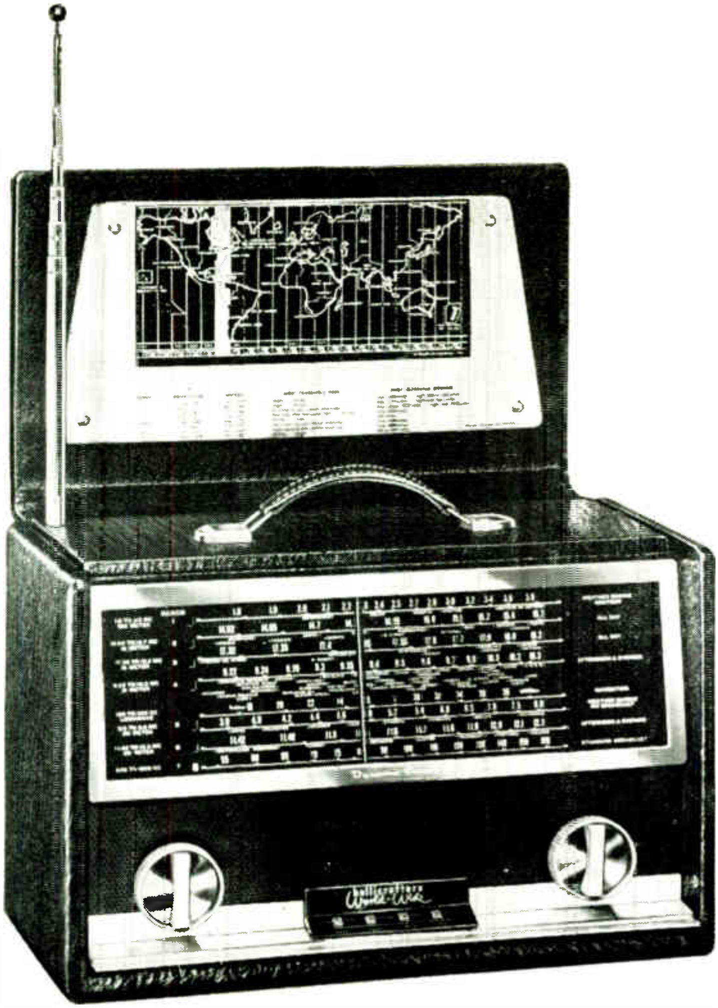
Hand Carry from $324^{95} to $399^{95}

plus $17.12 F.E.T. plus $21.93 F.E.T.

Central Station . . . Same performance and specifications as Hand Carry unit. Audio-amplifier, providing one watt of audio for loud speaker. AC operated with power consumption of 35 watts. Plugs in any AC outlet of 117 V. Hallicrafters S-81 receivers may be used as extra stationary stations.

Central Station . . . . $485^{00}

plus $23.00 F.E.T.



hallicrafters

finest SW and broadcast portable made

The Hallicrafters "World-Wide," Model TW-1000, the finest short-wave and broadcast portable radio made. Superior Standard Broadcast covers 535-1620 kc plus seven other bands covering 1.7-3.9, 3.8-8.2, 9.2-10.4, 11.4-12.4, 14.6-15.7 and 17.3-18.3 Mcs, plus special marine weather band.

Sleek metal trim on smart leatherette cabinet. Full-view, easy to tune, overseas dial—a Hallicrafters exclusive. World-Wide short-wave radio map tells you what's on the air. Red indicator for easy band identification. Four way tone control.

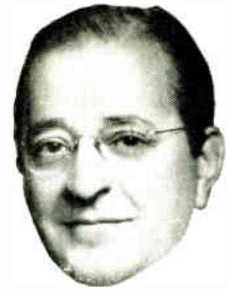
Three antennas for maximum performance—built-in loop, 64" telescope "whip" antenna, and removable "Skyrider" that fastens to car, railroad or airplane windows—lets the "World-Wide" play anywhere. Simplified controls include Dynamic Turret Tuner for accurate band selection. Five tubes plus rectifier. 105-125 V. AC or DC or battery.

Model TW-1000 . . . . $149.95

hallicrafters

the plants

Hallicrafters plants, four of them, are the most modern in the entire field of electronics. Here skilled craftsmen on modern assembly lines produce the Hallicrafters equipment that is known for highest quality in 89 countries — that is first choice of 33 governments — that is by long odds the overwhelming choice of all of our own armed services. And, most exacting test of all, Hallicrafters is the choice of the most critical expert in the world — the American ham operator.



William J. Halligan
President

R. W. Durst
Executive Vice President

Richard A. Graver
Vice President in charge of Marketing

W. J. Halligan, Jr.
Sales Mgr. Radio & Communications

Fritz Franke
Asst. Sales Mgr. Communications

hallicrafters

the people

Companies are only as good as the people that work for them. Hallicrafters has, for years, been fortunate in the people that have made the company great, and have kept it that way. One thing makes them unusual — they bring an attitude and an interest to their jobs that other men reserve for their hobbies. Hallicrafters men are hams at heart — and most of them are hams by license. Here's what Bill Halligan, Senior, says about his job:

"The radio ham market," expounds Bill Halligan, "today is the most challenging and the most thrilling in all radio. The ham is never fooled by expensive cabinets — he wants every nickel's worth of performance in the chassis. And he wants the absolute latest in circuit design. In working with him and pioneering equipment for him, we feel we are building a background for **future developments.**"

Jack Halligan
Communication Sales

John S. Mahoney
Director of Advertising

Norm Cooper
Service Mgr.

Have you seen it?

hallicrafters

a new dimension in picture realism—
a new height in viewing enjoyment!

Here is television only Hallicrafters could produce. Over twenty years of dealing with the most exacting of high frequency electronics have provided the experience, the know-how, the precision manufacturing techniques it took to produce these sets. And, just as no manufacturer can match this Hallicrafters heritage of experience—no manufacturer can match this Hallicrafters TV picture.

In short, these television sets are exactly like what they are—Hallicrafters equipment. If you are thinking of a television set—or know anyone that is—remember Hallicrafters. There is a difference that you can see—a superiority that is the envy of the industry.

hallicrafters

World's Leading Exclusive Manufacturer of Communications and High Fidelity Equipment, Radio and Television

4401 West Fifth Avenue, Chicago 24, Illinois

Hallicrafters Ltd., 51 Camden Street, Toronto, Canada

JAMES MILLEN
MALDEN · MASSACHUSETTS

ONE INCH INSTRUMENTATION OSCILLOSCOPE

Miniaturized, packaged panel mounting cathode ray oscilloscope designed for use in instrumentation in place of the conventional "pointer type" moving coil meters uses the 1" 1CP1 tube. Panel bezel matches in size and type the standard 2" square meters. Magnitude, phase displacement, wave shape, etc. are constantly visible on scope screen.

No. 90901, less tube . . . . $

INSTRUMENT DIAL

The No. 10030 is an extremely sturdy instrument type indicator. Control shaft has 1 to 1 ratio. Veeder type counter is direct reading in 99 revolutions and vernier scale permits readings to 1 part in 100 of a single revolution. Has built-in dial lock and ¼" drive shaft coupling. May be used with multi revolution transmitter controls, etc., or through gear reduction mechanism for control of fractional revolution capacitors, etc., in receivers or laboratory instruments.

No. 10030 . . . . $

GRID DIP METER

The No. 90651 MILLEN GRID DIP METER is compact and completely self contained. The AC power supply is of the "transformer" type. The drum dial has seven calibrated uniform length scales from 1.7 MC to 300 MC with generous over laps plus an arbitrary scale for use with special application inductors. Internal terminal strip permits battery operation for antenna measurement.

No. 90651, with tube . . . . $

Additional Inductors for Lower Frequencies

No. 46702—925 to 2000 KC . . . . $
No. 46703—500 to 1050 KC . . . .
No. 46704—325 to 600 KC . . . .
No. 46705—220 to 350 KC . . . .

LABORATORY SYNCHROSCOPES

The 5" laboratory synchroscopes are available with and without detector-video strips.

Model P-4-2, with tubes . . . . $
Model P-4E-2, with tubes . . . .

MINIATURE SYNCHROSCOPE

The compact design of the No. 90952, measuring only 7½" x 5⅝" x 13", and weighing only 17 lbs., makes available for the first time a truly DESIGNED FOR APPLICATION "field service" Synchroscope.

No. 90952, with tubes . . . . $

CATHODE RAY OSCILLOSCOPES

The No. 90902, No. 90903 and No. 90905 Rack Panel Oscilloscopes, for two, three and five inch tubes, respectively, are inexpensive basic units comprising power supply, brilliancy and centering controls, safety features, magnetic shielding, switches, etc. As a transmitter monitor, no additional equipment or accessories are required. The well known trapezoidal monitoring patterns are secured by feeding modulated carrier voltage from a pickup loop directly to vertical plates of the cathode ray tube and audio modulating voltage to horizontal plates. By the addition of such units as sweeps, pulse generators, amplifiers, servo sweeps, etc., all of which can be conveniently and neatly constructed on companion rack panels, the original basic 'scope unit may be expanded to serve any conceivable industrial or laboratory application.

No. 90902, less tubes . . . . $
No. 90903, less tubes . . . .
No. 90905, less tubes . . . .

'SCOPE AMPLIFIER—SWEEP UNIT

Vertical and horizontal amplifiers along with hard tube, saw tooth sweep generator. Complete with power supply mounted on a standard 5¼" rack panel

No. 90921, with tubes . . . . $

REGULATED POWER SUPPLIES

A compact, uncased, regulated power supply, either for table use in the laboratory or for incorporation as an integral part of larger equipments. Regulated, unregulated, bias and filament voltages provided.

Model 90201, less tubes . . . . $

JAMES MILLEN
MALDEN · MASSACHUSETTS

STANDING WAVE RATIO BRIDGE

The Millen S.W.R. bridge provides easy and inexpensive measurement of standing wave ratio on antennas using co-ax cable. As assembled the bridge is set up for 52 ohm line. A calibrated 75 ohm resistor is mounted inside the case for substitution in the circuit when 75 ohm line is used.
No. 90671 . . . . $

PHASE-SHIFT NETWORK

A complete and laboratory aligned pair of phase-shift networks in a single compact 2" x 1 7/16" x 4" case with characteristics so as to provide a phase shift between the two networks of 90° ± 1.3° over a frequency range of 225 cycles to 2750 cycles. This unit is equally well adapted for use in either single sideband transmitting or receiving equipment. When used in a suitably designed transmitter it is possible to obtain a 40 db suppression of the unwanted sideband. The No. 75012 precision adjusted phase-shift network makes possible the building of single sideband equipment without the necessity of complicated laboratory equipment for network adjustment.
No. 75012 . . . . $

R9'er MATCHING PREAMPLIFIER

The Millen 92101 is an electronic impedance matching device and a broad-band preamplifier combined into a single unit, designed primarily for operation on 6 and 10 meters. Coils for 20 meter band also available.
No. 92101, less tubes . . . . $

50 WATT EXCITER-TRANSMITTER

Modern design includes features and shielding for TVI reduction, bandswitching for 4–7–14–21–28 megacycle bands, circuit metering. Conservatively rated for use either as a transmitter or exciter for high power PA stages. 5763 oscillator-buffer-multiplier and 6146 power amplifier. Rack mounted.
No. 90801, less tubes . . . . $

VARIABLE FREQUENCY OSCILLATOR

The No. 90711 is a complete transmitter control unit with 6SK7 temperature-compensated, electron coupled oscillator of exceptional stability and low drift, a 6SK7 broad-band buffer or frequency doubler, a 6A67 tuned amplifier which tracks with the oscillator tuning, and a regulated power supply. Output sufficient to drive an 807 is available on 160, 80 and 40 meters and reduced output is available on 20 meters. Since the output is isolated from the oscillator by two stages, zero frequency shift occurs when the output load is varied from open circuit to short circuit. The entire unit is unusually solidly built so that no frequency shift occurs due to vibration. The keying is clean and free from all annoying chirp, quick drift, jump, and similar difficulties often encountered in keying variable frequency oscillators.
No. 90711, with tubes . . . . $

HIGH VOLTAGE POWER SUPPLY

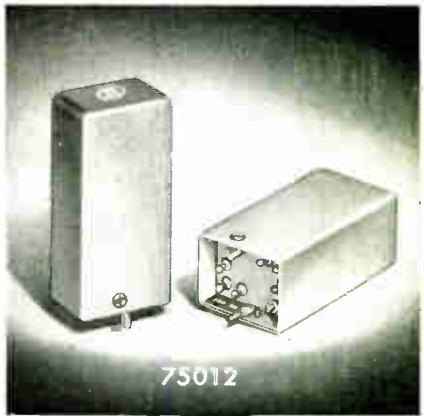
The No. 90281 high voltage power supply has a d.c. output of 700 volts, with maximum current of 235 ma. In addition, a.c. filament power of 6.3 volts at 4 amperes is also available so that this power supply is an ideal unit for use with transmitters, such as the Millen No. 90801, as well as general laboratory purposes. The power supply uses two No. 816 rectifiers. The panel is standard 8¾" x 19" rack mounting.
No. 90281, less tubes . . . . $

HIGH FREQUENCY RF AMPLIFIER

A physically small unit capable of a power output of 70 to 85 watts on 'phone or 87 to 110 watts on C-W on 20, 15, 11, 10, 6 or 2 meter amateur bands. Provision is made for quick band shift by means of the new No. 48000 series VHF plug-in coils. The No. 90811 unit uses either an 829-B or 3E29.
No. 90811 with 10 meter band coils, less tube . . . . $.

RF POWER AMPLIFIER

This 500 watt amplifier may be used as the basis of a high power amateur transmitter. The No. 90881 RF power amplifier is wired for use with the popular "812A" type tubes. Other popular tubes may be used. The amplifier is of unusually sturdy mechanical construction, on a 10½" relay rack panel. Plug-in inductors are furnished for operation on 10, 20, 40 or 80 meter amateur bands. The standard Millen No. 90801 exciter unit is an ideal driver for the new No. 90881 RF power amplifier.
No. 90881, with one set of coils, but less tubes . . . . $



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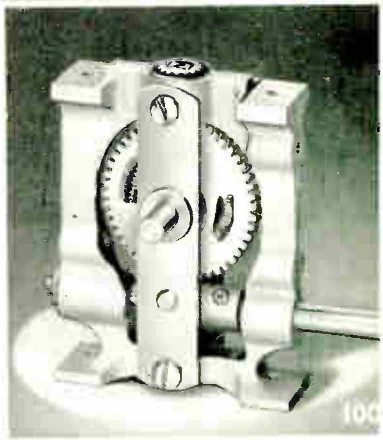


PANEL DIALS

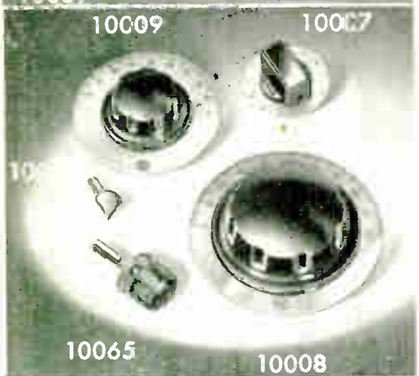
The No. 10035 illuminated panel dial has 12 to 1 ratio; size, 8½" x 6½". Small No. 10039 has 8 to 1 ratio, size, 4" x 3¼". Both are of compact mechanical design, easy to mount and have totally self-contained mechanism, thus eliminating back of panel interference. Provision for mounting and marking auxiliary controls, such as switches, potentiometers, etc., provided on the No. 10035. Standard finish, either size, flat black art metal.

No. 10039 . . . . $
No. 10035 . . . .

WORM DRIVE UNIT

Cast aluminum frame may be panel or base mounted. Spring loaded split gears to minimize back lash.
Standard ratio 16 1. Also in 48 1 on request.

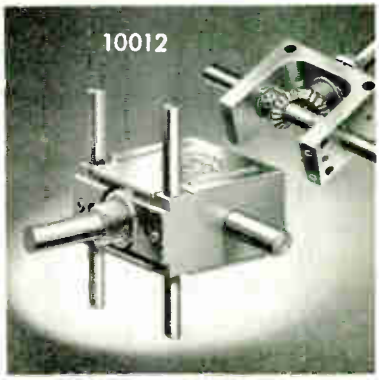
No. 10000—(state ratio) . . . . $



DIALS AND KNOBS

Just a few of the many stock types of small dials and knobs are illustrated herewith. 10007 is 1⅝" diameter, 10009 is 2½" and 10008 is 3½".

No. 10002 . . . . $
No. 10007 . . . .
No. 10008 . . . .
No. 10009 . . . .
No. 10015 . . . .
No. 10018 . . . .
No. 10021 . . . .
No. 10065 . . . .



RIGHT ANGLE DRIVE

Extremely compact, with provisions for many methods of mounting. Ideal for operating potentiometers, switches, etc., that must be located, for short leads, in remote parts of chassis.

No. 10012 . . . . $

HIGH VOLTAGE INSULATED SHAFT EXTENSION

No. 10061 shaft locks and the No. 39023 insulated high voltage potentiometer extension mountings are available as a single integrated unit—the No. 39024. The proper shaft length is independent of the panel thickness. The standard shaft has provision for screw driver adjustment. Special shaft arrangements are available for industrial applications. Extension shaft and insulated coupling are molded as a single unit to provide accuracy of alignment and ease of installation.

No. 39023, non locking type . . . . $
No. 39024, locking type . . . .

SHAFT LOCKS

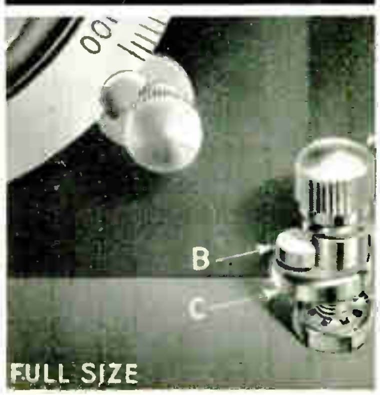
In addition to the original No. 10060 and No. 10061 "DESIGNED FOR APPLICATION" shaft locks, we can also furnish such variations as the No. 10062 and No. 10063 for [illegible] thumb operation as illustrated above. The No. 10061 instantly converts any plain "¼ shaft" volume control, condenser, etc. from "plain" to "shaft locked" type. Easy to mount in place of regular mounting nut.

No. 10060 . . . . $
No. 10061 . . . .
No. 10062 . . . .
No. 10063 . . . .

TRANSMISSION LINE PLUG

An inexpensive, compact, and efficient polystyrene unit for use with the 300 ohm ribbon type polyethylene transmission lines. Fits into standard Millen No. 33102 (crystal) socket. Pin spacing ½", diameter .095".

No. 37412 . . . . $



DIAL LOCK

Compact, easy to mount, positive in action, does not alter dial setting in operation! Rotation of knob "A" depresses finger "B" and "C" without imparting any rotary motion to Dial. Single hole mounted.

No. 10050 . . . . $

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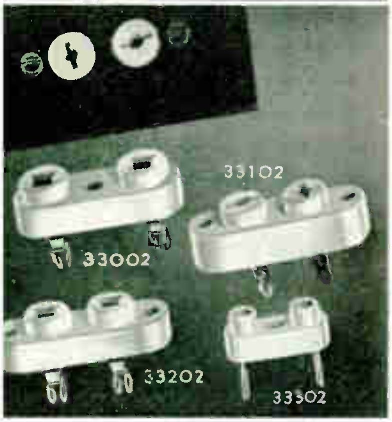

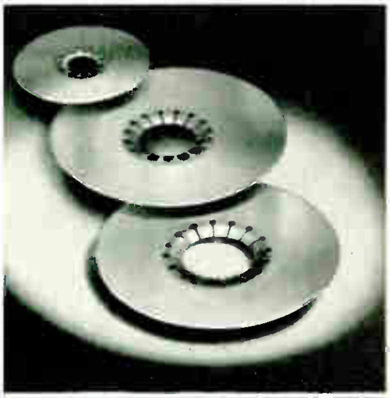

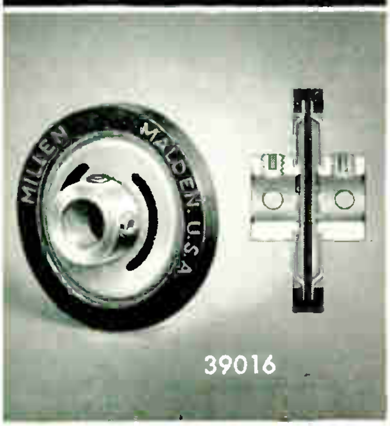

TUBE SOCKETS DESIGNED FOR APPLICATION

MODERN SOCKETS for MODERN TUBES! Long Flashover path to chassis permits use with transmitting tubes, 866 rectifiers, etc. Long leakage path between contacts. Contacts are type proven by hundreds of millions already in government, commercial and broadcast service, to be extremely dependable. Sockets may be mounted either with or without metal flange. Mounts in standard size chassis hole. All types have barrier between contacts and chassis. All but octal and crystal sockets also have barriers between individual contacts in addition.

The No. 33888 shield is for use with the 33008 octal socket. By its use, the electrostatic isolation of the grid and plate circuits of single-ended metal tubes can be increased to secure greater stability and gain.

The 33087 tube clamp is easy to use, easy to install, effective in function. Available in special sizes for all types of tubes. Single hole mounting. Spring steel, cadmium plated.

Cavity Socket Contact Discs, 33446 are for use with the "Lighthouse" ultra high frequency tube. This set consists of three different size unhardened beryllium copper multifinger contact discs. Heat treating instructions forwarded with each kit for hardening after spinning or forming to frequency requirements.

Voltage regulator dual contact bayonet socket, 33991 black phenolic insulation and 33992 with low loss high leakage mica filled phenolic insulation.

No. 33004 .... $
No 33005 ....
No. 33006 ....
No. 33007 ....
No. 33008 ....
No. 33888 ....
No. 33087 ....
No. 33002 ....
No. 33102 ....
No. 33202 ....
No. 33302 ....
No. 33446* ....
No. 33991 ....
No. 33992 ....

*For set of 3. Single discs $0.00 each.

FLEXIBLE COUPLINGS

The No. 39000 series of Millen "Designed for Application" flexible coupling units include, in addition to improved versions of the conventional types, also such exclusive original designs as the No. 39001 insulated universal joint and the No. 39006 "slide-action" coupling (in both steatite and bakelite insulation).

The No. 39006 "slide-action" coupling permits longitudinal shaft motion, eccentric shaft motion and out-of-line operation, as well as angular drive without backlash.

The No. 39005 is similar to the No. 39001, but is not insulated and is designed for applications where relatively high torque is required. The steatite insulated No. 39001 has a special anti-backlash pivot and socket grip feature. All of the above illustrated units are for ¼" shaft and are standard production type units. The No. 39016 incorporates features which have long been desired in a flexible coupling. No Back Lash—Higher Flexibility—Higher Breakdown Voltage—Smaller Diameter—Shorter Length—Higher Alignment Accuracy—Higher Resistance to Mechanical Shock—Solid Insulating Barrier Diaphragm—Molded as a Single Unit.

No. 39001 .... $
No. 39002 ....
No. 39003 ....
No. 39005 ....
No. 39006 ....
No. 39016 ....

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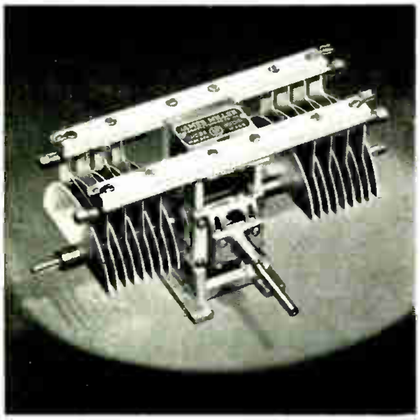
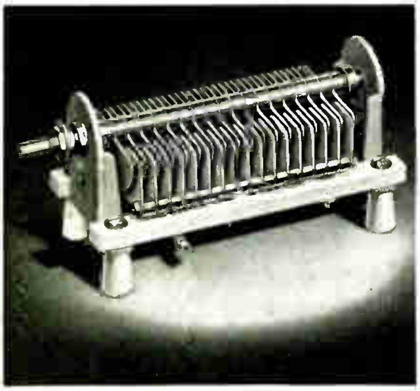


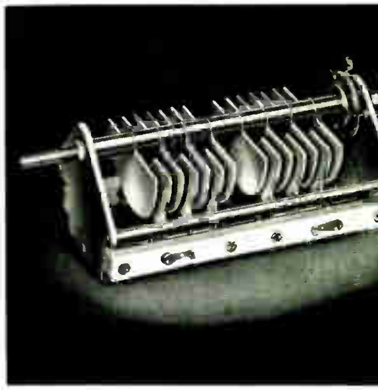
04000 and 11000 SERIES TRANSMITTING CONDENSERS

A new member of the "Designed for Application" series of transmitting variable air capacitors is the 04000 series with peak voltage ratings of 3000, 6000, and 9000 volts. Right angle drive, 1–1 ratio. Adjustable drive shaft angle for either vertical or sloping panels. Sturdy construction, thick, round-edged, polished aluminum plates with 1¾″ radius. Constant impedance, heavy current, multiple finger rotor contactor of new design. Available in all normal capacities.

The 11000 series has 16 1 ratio center drive and fixed angle drive shaft.

Code	Volts	Capacity	Price
11035	3000	35	$
11050	3000	50	
11070	3000	70	
04050	6000	50	
04060	9000	60	
04100	6000	90	
04200	3000	205	

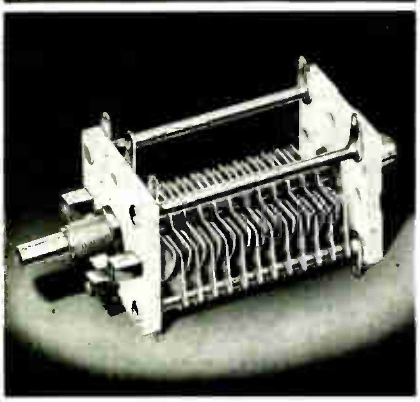


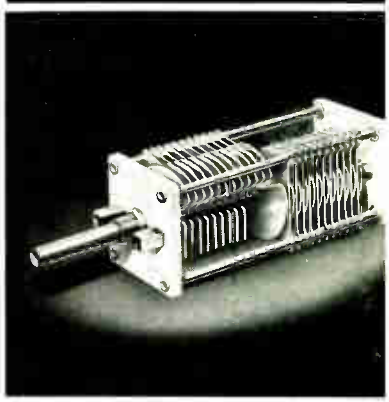


12000 and 16000 SERIES TRANSMITTING CONDENSERS

Rigid heavy channeled aluminum end plates. Isolantite insulation, polished or plain edges. One piece rotor contact spring and connection lug. Compact, easy to mount with connector lugs in convenient locations. Same plate sizes as 11000 series above.

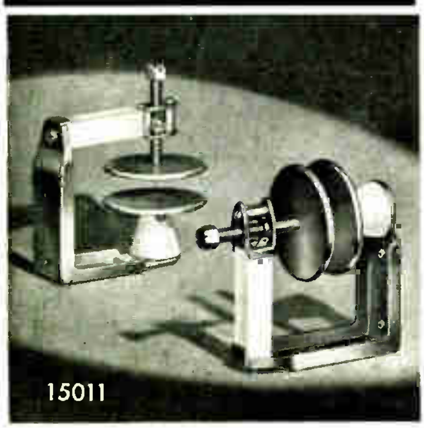
The 16000 series has some plate sizes as 04000 series. Also has constant impedance, heavy current, multiple finger rotor contactor of new design. Both 12000 and 16000 series available in single and double sections and many capacities and plate spacing.





THE 28000–29000 SERIES VARIABLE AIR CAPACITORS

"Designed for Application," double bearings, steatite end plates, cadmium or silver plated brass plates. Single or double section .022″ or .066″ air gap. End plate size: 19 16″ x 11 16″. Rotor plate radius: ¾″. Shaft lock, rear shaft extension, special mounting brackets, etc., to meet your requirements. The 28000 series has semi-circular rotor plate shape. The 29000 series has approximately straight frequency line rotor plate shape. Prices quoted on request. Many stock sizes.



NEUTRALIZING CAPACITOR

Designed originally for use in our own No. 90881 Power Amplifier, the No. 15011 disc neutralizing capacitor has such unique features as rigid channel frame, horizontal or vertical mounting, fine thread over-size lead screw with stop to prevent shorting and rotor lock. Heavy rounded-edged polished aluminum plates are 2″ diameter. Glazed Steatite insulation.

No. 15011 .... $

THRU-BUSHING

Efficient, compact, easy to use and neat appearing. Fits ¼″ hole in chassis. Held in place with a drop of solder or a "nick" from a crimping tool.

No. 32150 .... $

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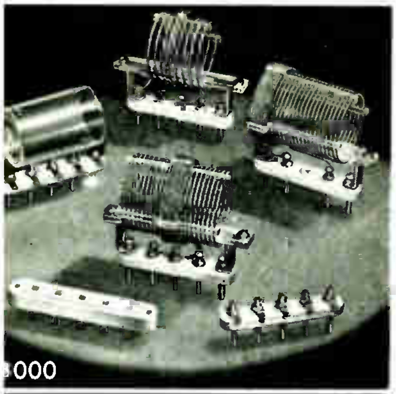
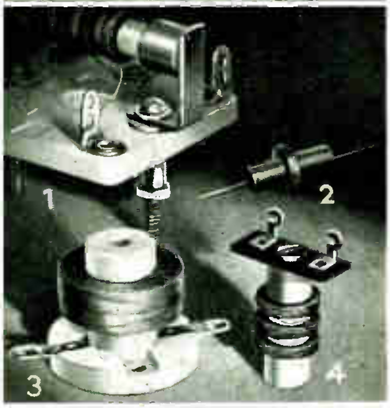
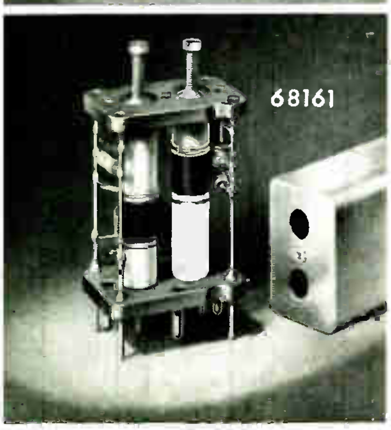






TRANSMITTING TANK COILS

A full line—all popular wattages for all bands. Send for special catalog sheet.

TUNABLE COIL FORM

Standard octal base of low loss mica-filled bakelite, polystyrene ½″ diameter coil form, heavy aluminum shield, iron tuning slug of high frequency type, suitable for use up to 35 mc. Adjusting screw protrudes through center hole of standard octal socket.

No. 74001, with iron core........ $
No. 74002, less iron core.........

RF CHOKES

Many have copied, few have equalled, and none have surpassed the genuine original design Millen Designed for Application series of midget RF Chokes. The more popular styles now in constant production are illustrated herewith. Special styles and variations to meet unusual requirements quickly furnished.

Figures 1 and 4 illustrate special types of RF chokes available on order. The popular 34300 and 34200 series are shown in figures 2 and 3 respectively.

General Specifications: 2.5 mH, 250 mA for types 34100, 34101, 34102, 34103, 34104, and 1 mH, 300 mA for types 34105, 34106, 34107, 34108, 34109.

No. 34100.................... $
No. 34101....................
No. 34102....................
No. 34103....................
No. 34104....................

MIDGET COIL FORMS

Made of low loss mica filled brown bakelite. Guide funnel makes for easy threading of leads through pins.

No. 45000.................... $
No. 45004....................
No. 45005....................

OCTAL BASE AND SHIELD

Low loss phenolic base with octal socket plug and aluminum shield can 1 7/16 x 1 7/8 x 3 15/16.

No. 74400......................... $

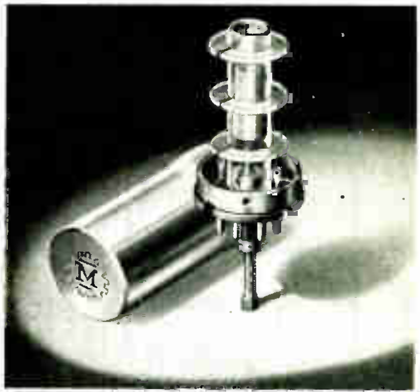
I.F. TRANSFORMERS

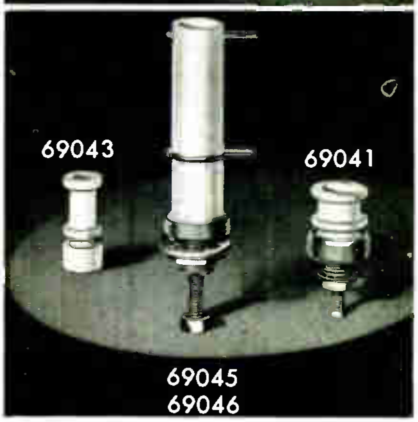
The Millen "Designed for Application" line of I.F. transformers includes air condenser tuned, and permeability tuned types for all applications. Standard stock units are for 456, 1600 and 5000 kc. B.F.O. also available.

PERMEABILITY TUNED CERAMIC FORMS

In addition to the popular shielded plug-in permeability tuned forms, 74000 series, the 69040 series of ceramic permeability tuned unshielded forms are available as standard stock items. Winding diameters and lengths of winding space are 13/32 x 7/32 for 69041-2; ¼ x ⅜ for 69043-7-8; ½ x 1 1/16 for 69045-6; 3/16 x 3/16 for 69044.

No. 69041—(Copper Slug)............ $
No. 69042—(Iron Core)..............
No. 69043—(Iron Core)..............
No. 69044—(Copper Slug)............
No. 69045—(Copper Slug)............
No. 69046—(Iron Core)..............
No. 69047—(Copper Slug)............
No. 69048—(Iron Core)..............





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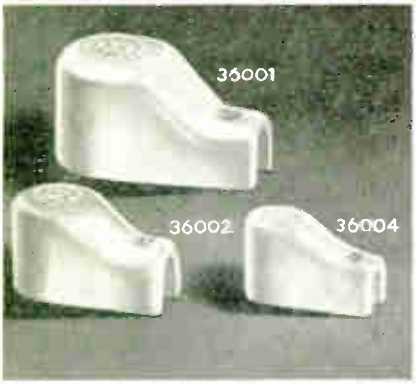
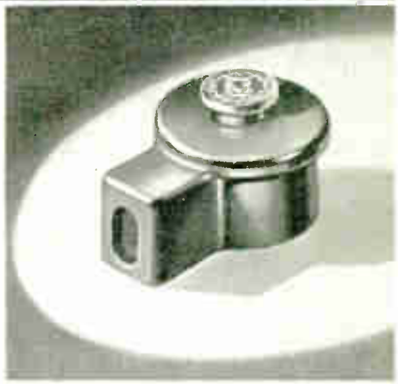




CERAMIC PLATE OR GRID CAPS

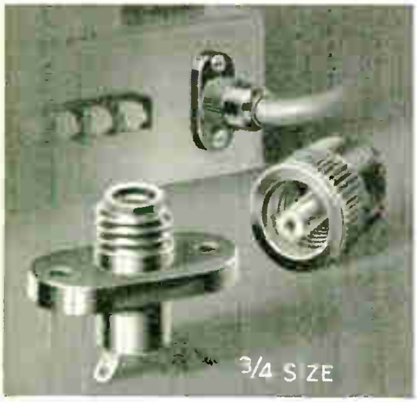
Soldering lug and contact one-piece. Lug ears annealed and solder dipped to facilitate easy combination "mechanical plus soldered" connection of cable

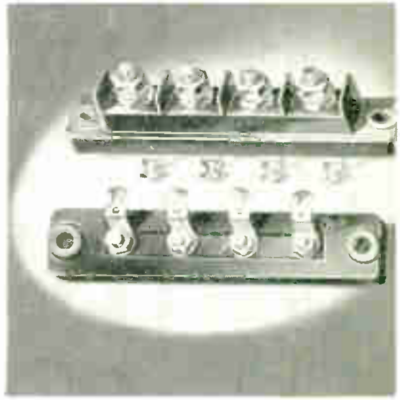
No. 36001—9/16″ . . . . $
No. 36002—3/8″ . . . .
No. 36004—1/4″ . . . .

SNAP LOCK PLATE CAP

For Mobile, Industrial and other applications where tighter than normal grip with multiple finger 360° low resistance contact is required. Contact self-locking when cap is pressed into position. Insulated snap button at top releases contact grip for easy removal without damage to tube.

No. 36011—9/16″ . . . . $
No. 36012—3/8″ . . . .





SAFETY TERMINAL

Combination high voltage terminal and thru-bushing. Tapered contact pin fits firmly into conical socket providing large area, low resistance connection. Pin is swivel mounted in cap to prevent twisting of lead wire.

No. 37001, Black or Red . . . . $
No. 37501, Low loss . . . .

TERMINAL STRIP

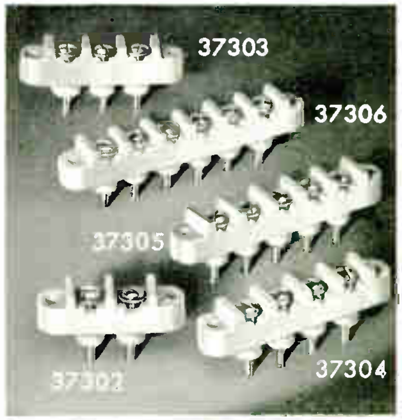
A sturdy four-terminal strip of molded black Textolite. Barriers between contacts. "Non turning" studs, threaded 8-32 each end.

No. 37104 . . . . $

POSTS, PLATES and PLUGS

Designed for Application! Compact, easy to use. Made in black and red regular bakelite as well as low loss brown mica filled bakelite or steatite for R.F. uses. Posts have captive head.

No. 37202 Plates (pr.) . . . . $
No. 37212 Plugs . . . .
No. 37222 Posts (pr.) . . . .



STEATITE TERMINAL STRIPS

Terminal and lug are one piece. Lugs are Navy turret type and are free floating so as not to strain steatite during wide temperature variations. Easy to mount with series of round holes for integral chassis bushings.

No. 37302 . . . . $
No. 37303 . . . .
No. 37304 . . . .
No. 37305 . . . .
No. 37306 . . . .

CATHODE RAY TUBE SHIELDS

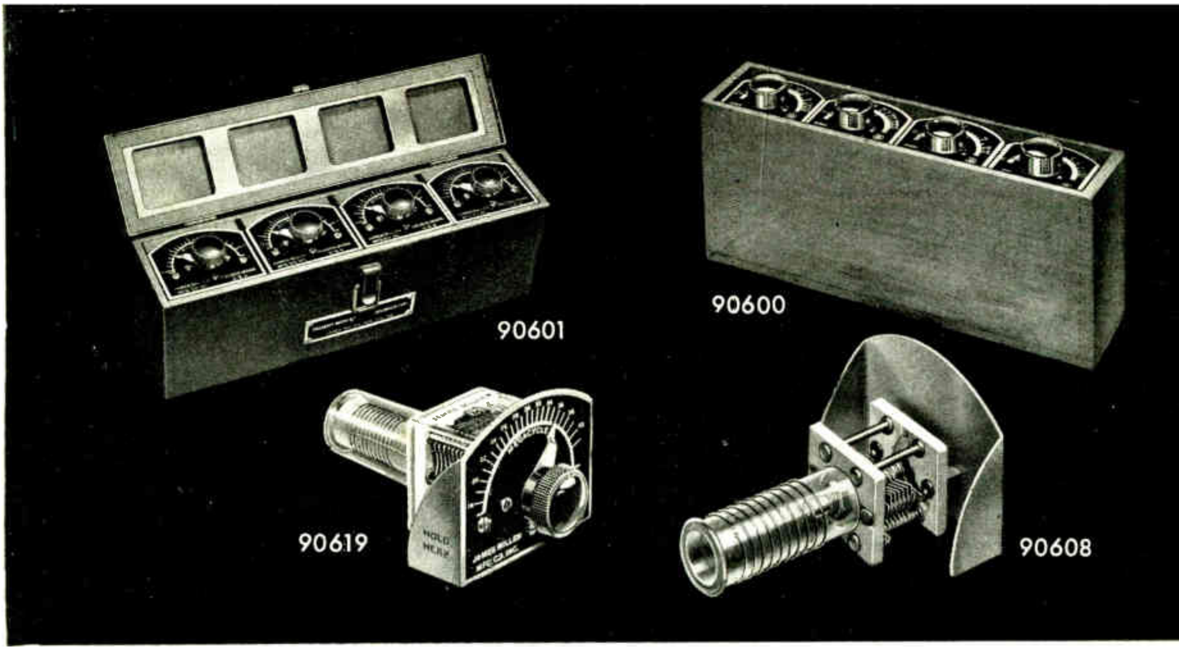
For many years we have specialized in the design and manufacture of magnetic metal shields of nicoloi and mumetal for cathode ray tubes in our own complete equipment, as well as for applications of all other principal complete equipment manufacturers. Stock types as well as special designs to customers' specifications promptly available.

No. 80045—Nicoloi for 5BP1 . . . . $
No. 80055—Nicoloi for 5CP1 . . . .
No. 80043—Nicoloi for 3″ tube . . . .
No. 80042—Nicoloi for 2″ tube . . . .

BEZELS FOR CATHODE RAY TUBES

Standard types are of satin finish black plastic. 5″ size has neoprene support cushion and green lucite filter. 3″ and 2″ sizes have integral cushioning.

No. 80075—5″ . . . . $
No. 80073—3″ . . . .
No. 80072—2″ . . . .
No. 80071—1″ . . . .



Midget Absorption Frequency Meters

Many amateurs and experimenters do not realize that one of the most useful "tools" of the commercial transmitter designer is a series of very small absorption type frequency meters. These handy instruments can be poked into small shield compartments, coil cans, corners of chassis, etc., to check harmonics; parasitics; oscillator-doubler, etc., tank tuning; and a host of other such applications. Quickly enables the design engineer to find out what is really "going on" in a circuit.

Types 90605 thru 90609 are extremely small and designed primarily for engineering laboratory use where they will be handled with reasonable care. The most useful combination being the group of four under code No. 90600 and covering the total range of from 3.0 to 140 megacycles. When purchased in sets of four under code No. 90600 a convenient carrying and storage case is included. Series 90601 are slightly larger and very much more rugged. They are further protected by a contour fitting transparent polystyrene case to protect against damage and dirt. This latter series is designed primarily for field use and are not quite as convenient for laboratory use as the 90605 thru 90608 types. All types have dials directly calibrated in frequency.

Code	Description	Net Price
90604	Range 160 to 210 mc.	$
90605	Range 3.0 to 10 mc.	
90606	Range 9.0 to 23 mc.	
90607	Range 23 to 60 mc.	
90608	Range 50 to 140 mc.	
90609	Range 130 to 170 mc.	
90610	Range 105 to 150 mc.	
90619	Range 350 to 1000 kc.—Neon Indicator	
90620	Range 150 to 350 kc.—Neon Indicator	
90625	Range 2 to 6 mc.—Neon Indicator	
90626	Range 5.5 to 15 mc.—Neon Indicator	
90600	Complete set of 90605 thru 90608, in case	
90601	Complete set Field type Frequency Meters in metal carrying case 1.5 to 40 mc.	

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V. MacNabb
915 Westfield Blvd.

SAN FRANCISCO
Moulthrop & Hunter
160 Tenth Street

CHICAGO
G. G. Ryan
549 W. Washington Blvd.

BOSTON
Harry Gerber
739 Boylston Street

DALLAS
Jack Yount
1431 Pleasant Drive

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Technical Equipment Co.
Main & Milton Sts.
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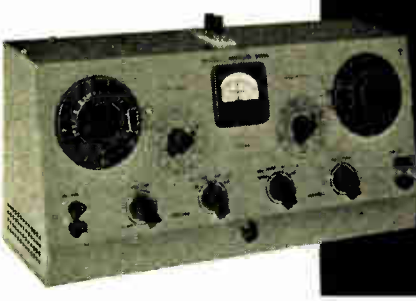
JAMES MILLEN MFG. CO., INC.

MAIN OFFICE AND FACTORY

150 EXCHANGE ST., MALDEN, MASSACHUSETTS, U. S. A.



Heathkit IMPEDANCE BRIDGE KIT

MODEL IB-2

$59.50

SHIPPING WT. 15 LBS.

Features

- Simpson 100-0-100 microampere meter.
- Completely AC operated.
- Built-in phase shift generator and amplifier.
- Battery type tubes, no warm-up required.
- Newly designed two section CRL dial.
- Single knob D, Q, and DQ functions.
- Special impedance matching transformer.
- New modern cabinet styling.
- ½% precision resistors and silver mica condensers.

Another new, outstanding instrument design so typically characteristic of Heathkit operation in producing high quality instrument kits at the lowest possible price. A new, improved model Impedance Bridge kit featuring modern cabinet styling, with slanted panel for convenience of operation and interpretation of scales at a $10.00 price reduction over the preceding model. Built-in adjustable phase shift oscillator and amplifier with all tubes of the battery operated type completely eliminates warm-up time. The instrument is entirely AC line operated. No bothersome battery replacements.

The Heathkit IB-2 Impedance Bridge Kit actually represents four instruments in one compact unit. The Wheatstone Bridge for resistance measurements, the Capacity Comparison Bridge for capacity measurements, Maxwell Bridge for low Q, and Hay Bridge for high Q inductance measurements. Read Q, D, DQ all on one dial thereby eliminating possible confusion due to the incorrect dial reference or adjustment. Only one set of instrument terminals necessary for any measurement function. Panel provisions provided for external generator use.

A newly designed two section CRL dial provides ten separate "units" switch settings with an accuracy of .5%. Fractions of units are read on a continuously variable calibrated wire-wound control. A special minimum capacity, shielded, balanced impedance matching transformer between the generator and the bridge. The correct impedance match is automatically switch selected to provide constant load operation of the generator circuit. The instrument uses ½% precision resistors and condensers in all measurement circuits.

The new Heathkit IB-2 provides outstanding design features not found in any other kit instrument. The single low price includes the power supply, generator, and amplifier stages. No need to purchase separate instrument accessories in order to obtain the type of operation desired.

Heathkit AUDIO WATTMETER KIT

MODEL AW-1

$29.50

SHIPPING WT. 6 LBS.

A new Heathkit design for the audio engineer, serious hi fi enthusiast, recording studio, or broadcast station; the Heathkit Audio Wattmeter Kit. This specialized instrument instantly indicates the output level of the equipment under test without requiring the use of external load resistors. All readings are taken directly from the calibrated scales of a 4½" 200 microampere Simpson meter.

The Heathkit Audio Wattmeter features five full scale power measurement ranges from 5 milliwatts up to 50 watts with db ranges of −15 db to +48 db. The instrument has a power measurement rating of 25 watts continuous and 50 watts maximum for intermittent operation. Non-inductive resistance load impedances of 4, 8, 16, and 600 ohms are provided through a panel impedance selector switch. Frequency effect is negligible from 10 cycles to 250 kc. A conventional VTVM circuit utilizes a 12AU7 twin triode tube. The meter bridge circuit uses four germanium diodes for good linearity.

With the Heathkit AW-1 desired information can be obtained instantly and conveniently without bothering with the irksome setups and calculations usually required. Useful for power curve measurements, frequency response checks, monitoring indicator, etc. Convenient calibration directly from 110 volt AC line source. This new instrument will help to supply the answers to your audio operating or power output problems.

Heathkit LABORATORY GENERATOR KIT

MODEL LG-1

$39.50

SHIP. WT. 16 LBS.

Another welcome new addition to the popular line of Heathkit instruments, the Heathkit Laboratory Generator. Specifically designed for flexibility of operation, accuracy and versatility beyond the performance level provided by the conventional service type generator. Frequency coverage of the Colpitts oscillator is 150kc to 30mc in five convenient ranges with provisions for internal or external modulation up to 50%, and .1 volt RF output throughout the frequency range. Panel mounted 200 microampere Simpson meter for RF "set reference level" to provide relative indication of RF output. Individually shielded oscillator and shielded variable and step attenuator provide flexible control of RF output.

The circuit features a 6AF4 high frequency oscillator, a 6AV5 amplifier with grid modulation, 12AU7 400 cycle oscillator and modulator, OB2 voltage regulator tube, and a selenium rectifier for the transformer operated power supply. The smart professional instrument appearance and over-all flexibility of operation will prove a decided asset to any industrial or educational laboratory. The Heathkit Laboratory Generator sets a new level of operation, far superior to any instrument in this price classification.

HEATH COMPANY • Benton Harbor 26, Mich.

NEW 5" *Heathkit* OSCILLOSCOPE KIT

MODEL O-9

$59.50

SHIPPING WT. 28 LBS.

CHECK THESE *Features*

- New 5UP1 CR tube
- Re-trace blanking
- Voltage regulation
- Extended band width
- Peak-to-peak calibrating provisions
- Good square wave response
- Astigmatism control
- New heavy duty shielded power transformer

Announcing the latest addition to a brilliant series of Heathkit Oscilloscopes, the new Model O-9. This outstanding instrument incorporates all of the features developed and proven in the production of well over 50,000 kits, in addition to a host of many new design features for truly outstanding performance. This new scope features a brand new (no surplus) commercially available 5UP1 cathode ray tube for fine focusing, high intensity, and freedom from halation. The 5" CR tube is the standard size for design and industrial laboratories, development engineers, and service men. The only size CR tube offering a wide range of types, colors, phosphors, and persistence. The answer to good oscilloscope performance lies in improved basic design and operating characteristics, and not in the use of larger CR tubes.

VERTICAL AMPLIFIER — New extended band width vertical amplifier with sensitivity of .025 volts per inch, down 3 db at 2 mc, down only 5½ db at 3 mc. Three step vertical input attenuator, quality ceramic variable capacitors for proper input compensation, provisions for calibrated 1 volt peak-to-peak reference, with calibrated screen for direct reading of TV pulses.

HORIZONTAL AMPLIFIER — New input selector switch provides choice of horizontal input, 60 cycle sweep input, line sync, internal sync, and external sync. Expanded horizontal sweep produces sweep width several times the cathode ray tube diameter. New blanking amplifier for complete retrace blanking and new phasing control.

POWER SUPPLY — New high voltage power supply and filtering circuit for really fine hairline focusing. New heavy duty power transformer with adequate operating reserve. Voltage regulated supply for both vertical and horizontal amplifiers for absolutely rock steady traces and complete freedom from bounce and jitter due to line variations.

The acid test of any oscilloscope operation is the ability to reproduce high frequency square waves and the new Heathkit O-9 will faithfully reproduce square waves up to 500 kc. This is the ideal all around, general purpose oscilloscope for educational and industrial use, radio and TV servicing, and any other type of work requiring the instantaneous reproduction and observation of actual wave forms and other electrical phenomena.

Heathkit LOW CAPACITY PROBE KIT

NO. 342

$3.50 SHIP. WT. 1 LB.

Oscilloscope investigation of high frequency, high impedance, or broad bandwidth circuits encountered in television work requires the use of a low capacity probe to prevent loss of gain, distortion, or false service information. The Heathkit Low Capacity Probe features a variable capacitor to provide the necessary degree of instrument impedance matching. New probe styling with bright polished aluminum housing and polystyrene probe ends.

Heathkit SCOPE DEMODULATOR PROBE KIT

NO. 337-B

$3.50

SHIP. WT. 1 LB.

In applications such as trouble shooting or aligning TV, RF, IF, and video stages, the frequency ranges encountered require demodulation of signals before oscilloscope presentation. The newly-styled Heathkit Demodulator Probe in polished aluminum housing will fulfill this function and readily prove its value as an oscilloscope service accessory. Detailed assembly sheet provided, including instructions for probe operation.

Heathkit VOLTAGE CALIBRATOR KIT

MODEL VC-2

$11.50

SHIPPING WT. 4 LBS.

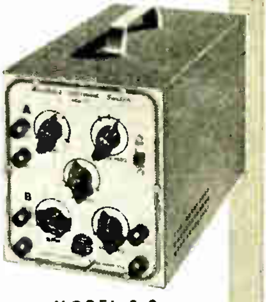
The Heathkit Voltage Calibrator provides a convenient method of making peak-to-peak voltage measurements with an oscilloscope by establishing a relationship on a comparison basis between the amplitude of an unknown wave shape and the known output of the voltage calibrator. Peak-to-peak voltage values are read directly on the calibrated panel scales. To offset line voltage supply irregularities, the instrument features a voltage regulator tube.

With the Heathkit Voltage Calibrator, it is possible to measure all types of complex wave forms within a voltage range of .01 to 100 volts peak-to-peak. A convenient "signal" position on the panel switch by-passes the calibrator completely and the signal is applied to the oscilloscope input thereby eliminating the necessity for transferring test leads.

Heathkit ELECTRONIC SWITCH KIT

The basic function of the Heathkit S-2 Electronic Switch Kit is to permit simultaneous oscilloscope observation of two separate traces which can be either separated or superimposed for individual study. A typical example would be observation of a signal as it appears at both the input and output stages of an amplifier. It will also serve as a square wave generator over the range of switching frequencies often providing the necessary wave form response information without incurring the expense of an additional instrument.

Continuously variable switching rates in three ranges from less than 10 cps to over 2,000 cps. Individual controls for each input channel and a positioning control. The five tube transformer operated circuit utilizes two 6SJ7, two 6SN7, and one 6X5 tubes. Buy this kit and enjoy increased versatility of operation from your oscilloscope.



MODEL S-2

$23.50

SHIP. WT. 11 LBS.

HEATH COMPANY • Benton Harbor 26, Mich.

Heathkit VACUUM TUBE VOLTMETER KIT

MODEL V-6

$24.50

SHIPPING WT. 6 LBS.

Features

- New 1½ volt full scale low range
- 1,500 volt upper limit DC range
- Increased accuracy through 50% greater scale coverage
- High impedance 11 megohm input
- Center scale zero adjust
- Polarity reversal switch
- 1% precision resistors
- Clearly marked db scales

The beautiful Heathkit Model V-6 VTVM, the world's largest selling kit instrument, now offers many outstanding new features in addition to retaining all of the refinements developed and proven in the production of over 100,000 VTVM's. This is the basic measuring instrument for every branch of electronics. Easily meets all requirements for accuracy, stability, sensitivity, convenience of ranges, meter readability, and modern styling. It will accurately measure DC voltages, AC voltages, offers tremendous ohmmeter range coverage, and a complete db scale for a total of 35 meter ranges.

New 1½ volt full scale low range provides well over 2¼" of scale length per volt. Upper DC scale limit 1,500 volts. DC ranges 0-1.5, 5, 15, 50, 150, 500, 1,500 volts full scale. AC ranges 0-1.5, 5, 15, 50, 150, 500, 1,500 (1,000 volts maximum). Seven ohmmeter ranges from .1 ohm to 1,000 megohms. For added convenience a DC polarity reversing switch and a center scale zero adjustment for FM alignment.

The smartly styled, compact, sturdy, formed aluminum cabinet is finished in an attractive gray crackle exterior. The beautiful two-color, durable, infra-red, baked enamel panel further adds to the over-all professional appearance.

Top quality components used throughout. 1% precision resistors — silver contact range and selector switches — selenium rectifier — transformer operated power supply. Individual calibration on both AC and DC for maximum accuracy. DB scale printed in red for easy identification, all other scales a sharp, crisp black for easy reading. A variety of accessory probes shown on this page still add further to over-all instrument usefulness.

Heathkit 30,000 VOLT DC PROBE KIT

For TV service work or any similar application where the measurement of high DC voltage is required, the Heathkit Model 336 High Voltage Probe Kit will prove invaluable. A precision multiplier resistor mounted inside the two-color, sleek, plastic probe body provides a multiplication factor of 100 on the DC ranges of the Heathkit 11 megohm VTVM. The entire kit includes precision resistor, two-color plastic probe, tip connector spring, test lead, phone plug panel connector, and complete assembly instructions.

No. 336

$4.50

SHIP. WT. 2 LBS.

Heathkit PEAK-TO-PEAK PROBE KIT

No. 338-B

$5.50

SHIP. WT. 2 LBS.

Now read peak-to-peak voltages on the DC scales of the Heathkit 11 megohm VTVM. Readings can be directly made from the VTVM scale without involved calculations. Measurements over the frequency range of 5 kc to 5 mc. Use this probe to extend the usefulness of your VTVM in radio and TV service work. The Peak-to-Peak Probe Kit features the new polished aluminum housing with two-color polystyrene probe ends. Detailed assembly sheet including instructions for probe operation.

Heathkit RF PROBE KIT

The Heathkit RF Probe used in conjunction with any 11 megohm VTVM will permit RF measurements up to 250 mc, ± 10%. A useful, convenient accessory for those occasions when RF measurements are desired. The RF probe body is housed in the new, smartly-styled polished aluminum probe body featuring two-color polystyrene probe ends and a low capacity flexible shielded test lead. The kit is complete with all necessary material and a detailed assembly sheet as well as instructions for probe operation.

No. 309-B

$3.50

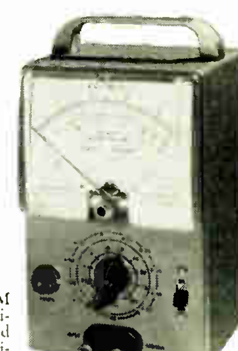
SHIP. WT. 2 LBS.

Heathkit AC VACUUM TUBE VOLTMETER KIT

MODEL AV-2

$29.50

SHIPPING WT. 5 LBS.



The new Heathkit AC VTVM that makes possible those sensitive AC measurements required by laboratories, audio enthusiasts, and experimenters. Especially useful for hum investigation, sensitive null detection, phono pick-up output measurements, making frequency response runs, gain measurements, ripple voltage checks, etc. Low level measurements are easy to make because of the complete voltage coverage of the instrument and the one knob operation.

The large 200 microampere Simpson meter has clearly marked and easy to read meter scales. Ten voltage ranges covering from .01 rms full scale to 300 volts rms full scale, with frequency response ± 1 db from 20 cycles to 50,000 cycles. Instrument input impedance 1 megohm, ten db ranges from −52 db to +52 db. For stability and good linearity characteristics the meter bridge circuit features 4 germanium diodes. Attractive instrument styling, a companion piece for the popular Heathkit VTVM and the new AW-1 Audio Wattmeter.

HEATH COMPANY • Benton Harbor 26, Mich.

NEW *Heathkit*

MULTIMETER KIT

MODEL MM-1

$26.50

SHIPPING WT. 6 LBS.

CHECK THESE *Features*

- 20,000 ohms per volt DC sensitivity, 5,000 ohms per volt on AC
- Polarity reversal switch
- 1% precision multiplier resistors
- 50 microampere 4½" Simpson meter
- Meter ranges for service convenience
- New resistor ring-switch assembly
- Total of 35 meter ranges
- New Modern cabinet styling

The most important Heathkit announcement of the year, the new 20,000 ohms per volt Heathkit Multimeter, Model MM-1. The universal service measuring instrument, accurate, sensitive, portable, and completely independent of AC line supply. Particularly designed for service use incorporating many desirable features for the convenience of the service man. Full 20,000 ohms per volt sensitivity on DC ranges — 5,000 ohms per volt sensitivity on AC — polarity reversal switch, no bothersome transferring of test leads — 1% precision multiplier resistors — large 4½" recessed non-glare 50 microampre Simpson meter — conveniently slanted control panel — recessed safety type banana jacks — standard universally available batteries — rugged practical sized cabinet with plastic carrying handle, and a total of 35 calibrated meter ranges.

RANGES

Voltage ranges selected entirely for service convenience. For example 1½ volt full scale low range for measuring portable radio filament voltages, bias voltages, etc., 150 volt full scale range for AC-DC service work, 500 volt full scale range for conventional transformer operated power supply systems. Complete voltage ranges AC and DC, 0-1.5—5—50—150—500—1,500—5,000 volts. DC current ranges, 0-150 microamperes—15 milliamperes—150 milliamperes—500 milliamperes—15 amperes. Resistance measurements from .2 ohms to 20 megohms x 1 x 1,000 x 10,000. DB coverage from —10 db to +65 db.

CONSTRUCTION

Entirely new design permits assembly, mounting and wiring of precision resistors on a ring-switch assembly unit. The major portion of instrument wiring is completed before mounting the ring-switch assembly to the panel. No calibration procedure is required, all precision resistors readily accessible in event of replacement.

CABINET

Strikingly modern cabinet styling featuring two piece construction, durable black Bakelite cabinet, with easy to read panel designations. Cabinet size 5½" wide x 4" deep x 7½" high. Good cabinet physical stability when operated in vertical position.

The Heathkit MM-1 represents a terrific instrument value for a high quality 20,000 ohms per volt unit using all 1% deposited carbon type precision resistors. Here is quality, performance, functional design, and attractive appearance, all combined in one low priced package.

Heathkit

BATTERY TESTER KIT

MODEL BT-1

$8.50

SHIP. WT. 2 LBS.

The Heathkit Battery Tester measures all types of dry batteries between 1½ volts and 150 volts under actual load conditions. Readings are made directly on a three color Good-Weak-Replace scale. Operation is extremely simple and merely requires that the test leads be connected to the battery under test. Only one control to adjust in addition to a panel switch for "A" or "B" battery types. The Heathkit Battery Tester features compact assembly, accurate meter movement, and a three deck wire-wound control, all mounted in a portable rugged plastic cabinet. Checks portable radio batteries, hearing aid batteries, lantern batteries, etc.

Heathkit

HANDITESTER KIT

MODEL M-1

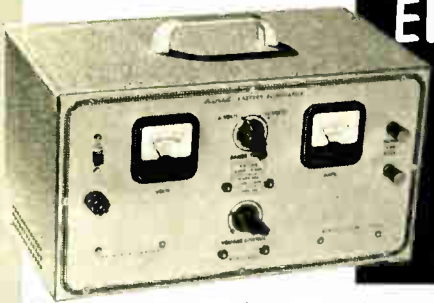
$14.50

SHIPPING WT. 3 LBS.

The Heathkit Model M-1 Handitester readily fulfills major requirements for a compact, portable volt ohm milliammeter. Despite its compact size, the Handitester is packed with every desirable feature required in an instrument of this type. AC or DC voltage ranges full scale. 0-10—30—300—1,000—5,000 volts. Two ohmmeter ranges, 0-3,000 and 0-300,000. Two DC current measurement ranges, 0-10 milliamperes and 0-100 milliamperes. The instrument uses a Simpson 400 microampere meter movement, which is shunted with resistors to provide a uniform 1 milliampere load on both AC and DC ranges. Special type, easily accessible, battery mounting bracket — 1% deposited carbon type precision resistors — hearing aid type ohms adjust control. The Handitester is easily assembled from complete instructions and pictorial diagrams. Necessary test leads are included in the price of this popular kit.

HEATH COMPANY • Benton Harbor 26, Mich.

New *Heathkit* 12 Volt BATTERY ELIMINATOR KIT



MODEL BE-4

$31.50

SHIPPING WT. 18 LBS.

CHECK THESE *Features*

- Either 6 or 12 volt operation
- Continuously variable voltage output
- Constant ammeter and voltmeter monitoring
- Automatic overload relay — self-resetting
- Two 10,000 mf condensers
- New 18 disc split type heavy duty rectifier unit
- Fuse protection

Here is the new Heathkit Battery Eliminator necessary for modern, up-to-date operation of your service shop. The Heathkit Model BE-4 furnishes either 6 volts or 12 volts output which can be selected at the flick of a panel switch. Use the BE-4 to service the new 12 volt car radios in addition to the conventional 6 volt radios.

This new Battery Eliminator provides two continuously variable output ranges, 0-8 volts DC at 10 amperes continuously, or 15 amperes maximum intermittent; 0-16 volts DC at 5 amperes continuously or 7.5 amperes maximum intermittent. The output voltage is clean and well filtered as the circuit uses two 10,000 mf condensers. The continuously variable voltage output feature is a definite aid in determining the starting point of vibrators, the voltage operating range of oscillator circuits, etc. Panel mounted meters constantly monitor voltage and current output and will quickly indicate the presence of a major circuit fault in the equipment under test. The power transformer primary winding is fuse protected and for additional safety an automatic relay of the self-resetting type is incorporated in the DC output circuit. The heavy duty rectifier is a split type 18 plate magnesium copper sulfide unit used either as a full wave rectifier or voltage doubler according to the position of the panel range switch.

Here is the ideal battery eliminator for all of your service problems and as an additional feature, it can also be used as a battery charger. Another new application for the Heathkit Battery Eliminator is a variable source of DC filament supply in audio development and research. More than adequate variable voltage and current range for normal applications.

Heathkit VIBRATOR TESTER KIT

Your repair time is valuable, and service use of the Heathkit Vibrator Tester will save you many hours of work. This tester will instantly tell you the condition of the vibrator being checked. Checks vibrators for proper starting and the easy to read meter indicates quality of output on a large Bad-?-Good scale. The Heathkit VT-1 checks both interrupter and self rectifier types of vibrators. Five different sockets for checking hundreds of vibrator types.

The Heathkit Vibrator Tester operates from any battery eliminator capable of delivering continuously variable voltage from 4 to 6 volts DC at 4 amperes. The new Heathkit Model BE-4 Battery Eliminator would be an ideal source of supply.

MODEL VT-1

$14.50

SHIPPING WT. 6 LBS.

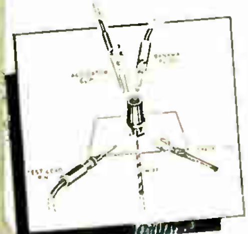
NEW *Heathkit* VARIABLE VOLTAGE ISOLATION TRANSFORMER KIT

The new Heathkit Isolation Transformer Kit provides line isolation for AC-DC radios (not an auto transformer), thereby eliminating shock hazard, hum problems, alignment difficulties, etc. The output voltage is variable from 90 to 130 volts AC and is constantly monitored by a panel mounted AC volt meter. Use it to increase AC supply voltage in order to induce breakdown of faulty components in circuits thereby saving service time. Use it also to simulate varying line voltage conditions and to determine the line voltage level at which oscillator circuits cease functioning, particularly in three-way portable radios. Rated at 100 watts continuous operation and up to 200 watts maximum intermittent operation. A useful radio and TV service tool.

MODEL IT-1

$16.50

SHIP. WT. 9 LBS.



Heathkit BINDING POST

Binding post kit now available so that standardization of all instrument connectors is possible. This new, five-way binding post will accommodate an alligator clip, banana plug, test lead pin, spade lug, or hook-up wire. Sold in units of 20 binding post assemblies. Each assembly includes binding post, flat and shoulder fiber washers, solder lug, and nut. 120 pieces in all. Kit 362, $4.00.

Heathkit TECHNICAL APPLICATION BULLETINS

An exclusive Heathkit service. Technical application bulletins prepared by recognized instrument authorities outlining various combinations of instrument applications. Available now with 40 four-page illustrated bulletins and an attractive flexible loose-leaf binder. Only $2.00. (No c.o.d. on this item, please.)

HEATH COMPANY • Benton Harbor 26, Mich.

CHECK THESE *Features*

- INCREDUCTOR controllable inductor sweep
- TV and IF sweep deviation 12-30 mc
- 4 mc- 220 mc continuous frequency coverage
- Oscillator operation entirely on fundamentals
- Output in excess of 100,000 microvolts
- Automatic amplitude circuit
- Voltage regulation
- Simplified operation

NEW *Heathkit*

TV ALIGNMENT GENERATOR KIT

MODEL TS-3

$44.50

SHIPPING WEIGHT 18 POUNDS

Proudly announcing an entirely new advanced model TV and FM Sweep Generator, the Heathkit Model TS-3. This new design provides features and combinations of functions not found in any other service type instrument. Every design consideration has been given to the requirements of the TV service man to provide a flexible, variable sweep source with more than adequate RF output and complete frequency coverage throughout the TV and FM spectrum.

The frequency range of the TS-3 is from 4 mc to 220 mc in four switch selected ranges. All frequency ranges are overlapping for complete coverage. A particularly important feature of the instrument is that the oscillator operates entirely on fundamentals, thereby providing complete freedom from spurious oscillation and parasitics normally encountered in beat frequency type oscillators. This circuity assures a much higher total RF output level and simplifies attenuation problems.

The new TS-3 features an entirely new principle of sweep operation. Sweep action is entirely electronic with no moving parts or electro-mechanical devices so commonly used. The heart of the sweep system is a newly-developed INCREDUCTOR controllable inductor. With this system, the value of inductance of each oscillator coil is electrically varied with an AC control current, and the inductance variation is achieved by a change in the magnetic state of the core on which the oscillator coils are wound. This system provides a sweep deviation of not less than 12 mc on all TV frequencies, and up to a maximum of 30 mc on TV IF frequencies. The high RF output level throughout the instrument frequency range overcomes the most common complaint of the older type sweep generators. A new, automatic amplitude control circuit maintains the output level flat to ± 2 db throughout the instrument range. For convenience of operation a low impedance 50 ohm output is used.

Operation of the instrument has been simplified through the reduction of panel controls and separate panel terminals provide for external synchronization if desired. The circuit uses a voltage regulator tube to maintain stable instrument operation. A built-in variable oscillator marker further adds to flexibility of instrument operation. Provisions are also made for the use of an external marker, such as your service type signal generator, if desired. Use the Heathkit TS-3 for rapid, accurate TV alignment work, and let it help you solve those time consuming, irksome problems so frequently encountered.

NEW *Heathkit*

SIGNAL GENERATOR KIT

MODEL SG-8

$19.50

SHIPPING WEIGHT 8 POUNDS

Announcing the new Heathkit Model SG-8 service type Signal Generator, incorporating many design features not usually found in an instrument in this price range. The RF output is from 160 kc to 100 mc in five ranges, all on fundamentals, with useful harmonics up to 200 mc. The RF output level is in excess of 100,000 microvolts throughout the frequency range.

The oscillator circuit consists of a 12AT7 twin triode tube. One half is used as a Colpitts oscillator, and the other half as a cathode follower output which acts as a buffer between the oscillator and external load. This circuity eliminates oscillator frequency shift usually caused by external circuit loading.

All coils are factory wound and adjusted, thereby completely eliminating the need for calibration and the use of additional calibrating equipment. The stable low impedance output features a step and variable attenuator for complete control of RF level. A 6C4 triode acts as a 400 cycle sine wave oscillator and a panel switching system permits a choice of either external or internal modulation.

The transformer operated circuit is easy to assemble, requires no calibration, and meets every service requirement for an adjustable level variable frequency signal source, either modulated or un-modulated.

NEW *Heathkit*

BAR GENERATOR KIT

MODEL BG-1

$14.50

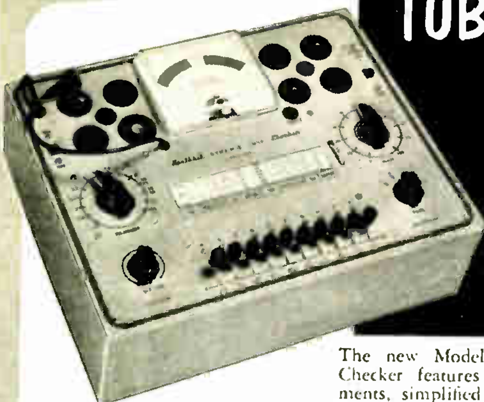
SHIPPING WEIGHT 6 POUNDS

The Heathkit BG-1 Bar Generator represents another welcome addition to the fast growing line of popular Heathkits. The station transmitted test pattern is rapidly disappearing, and the bar generator is the logical answer to the TV service man's problem in obtaining quick, accurate adjustment information without waiting for test patterns.

The Heathkit BG-1 produces a series of horizontal or vertical bars on a TV screen. Since these bars are equally spaced, they will quickly indicate picture linearity of the receiver under test. Panel switch provides "stand-by position" — "horizontal position" — "vertical position." The oscillator unit utilizes a 12AT7 twin triode for the RF oscillator and video carrier frequencies. A neon relaxation oscillator provides low frequency for vertical linearity tests. The instrument will not only produce bar patterns but will also provide an indication of horizontal and vertical sync circuit stability, as well as overall picture size.

Instrument operation is extremely simple, and merely requires connection to the TV receiver antenna terminal. The unit is transformer operated for safety when used in conjunction with universal or transformerless type TV circuits.

HEATH COMPANY • Benton Harbor 26, Mich.



NEW *Heathkit*

TUBE CHECKER KIT

MODEL TC-2

$29.50

SHIP. WT. 12 LBS.

CHECK THESE NEW *Features*

- Simplified harness wiring
- Improved, smooth, anti-backlash roll chart action
- Optional roll chart illumination
- Individual element switches
- Portable or counter style cabinet
- Spare blank socket
- Contact type pilot light test socket
- Simplified test set-up procedure
- Line adjust control
- 4½" three-color meter

The new Model TC-2 Heathkit Tube Checker features many circuit improvements, simplified wiring, new roll chart drive and illumination of roll chart. The instrument is primarily designed for the convenience of the radio and TV service man and will check the operating quality of tubes commonly encountered in this type of work. Test set-up procedure is simplified, rapid, and flexible. Panel sockets accommodate 4, 5, 6, and 7 pin tubes, octal and loctal, 7 and 9 pin miniatures, 5 pin Hytron and a blank socket for new tubes. Built-in neon short indicator, individual three-position lever switch for each tube element, spring return test switch, 14 filament voltage ranges, and line set control to compensate for supply voltage variations, all represent important design features of the TC-2. Results of tube tests are read directly from a large 4½" Simpson three-color meter, calibrated in terms of Bad-?-Good. Information that your customer can readily understand. Checks emission, shorted elements, open elements, and continuity.

The use of closer tolerance resistors in critical circuits assures correct test information and eliminates the possibility of inaccurate test interpretation. Improvement has been made in the mechanical roll chart drive system, completely eliminating diagonal running, erratic operation, and backlash. The thumb wheel gear driven action is smooth, positive, and free running. As an additional feature, the roll chart is illuminated for easier reading, particularly when the tube checker is used on radio or TV home service calls.

Wiring procedure has been simplified through the extended use of multi-cable, color coded wires, providing a harness type installation between tube sockets and lever switches. This procedure insures standard assembly and imparts that "factory built" appearance to instrument construction. Completely detailed information is furnished in the new step-by-step construction manual, regarding the set-up procedure for testing of new or unlisted tube types. No delay necessary for release of factory data.

The new Heathkit Tube Checker will prove its value in building service prestige through usefulness—simplified operation—attractive professional appearance. Don't overlook the fact that the kit price represents a savings of $40.00 to $50.00 over the price of a comparable commercially built instrument. At this low price, no service man need be without the advantages offered by the Heathkit Tube Checker.

New

HEATHKIT

PORTABLE TUBE CHECKER KIT

MODEL TC-2P

$34.50

SHIP. WT. 14 LBS.

The portable model is supplied with a strikingly attractive two-tone cabinet finished in rich maroon, proxylin impregnated, fabric covering with a contrasting gray on the inside cover. Detachable cover, brass-plated hardware, sturdy plastic handle help to impart a truly professional appearance to the instrument.

PORTABLE TUBE CHECKER CABINET as described above will fit all earlier Heathkit TC-1 Tube Checkers. Shipping weight 7 lbs. Cabinet only, 91-8, $7.50.

Heathkit TV PICTURE TUBE TEST ADAPTER

No. 355 Ship. Wt. 1 Lb. **$4.50**

The Heathkit TV Picture Tube Test Adapter used with the Heathkit Tube Checker will quickly check for emission, shorts, etc., and determine picture tube quality. Consists of standard 12 pin TV tube socket, four feet of cable, octal socket connector, and data sheet.

Heathkit POWER SUPPLY KIT

MODEL PS-2

$33.50

SHIPPING WT. 17 LBS.

The Heathkit Laboratory Power Supply features continuously variable, regulated, voltage output with good stability under wide load variations. A 4½" Simpson plastic enclosed panel mounted meter provides accurate meter output information of voltage or current. All panel terminals completely isolated from the cabinet. Separate 6.3 volt AC supply at 4 amperes for filament requirements. Ripple component exceptionally low, stand-by switch provided to eliminate warm-up time of the five tube circuit.

LABORATORY AND SERVICE SHOP BOOKLETS

"Planning Your Service Business" by John T. Frye, and "Establishing the Industrial Electronics Laboratory" by Louis B. Garner, Jr., are booklets available to Heathkit customers at no charge. These booklets, written by nationally recognized authorities, outline the various requirements and considerations for establishing your own service business or for setting up an industrial electronics laboratory. Full attention is given to various details that are frequently overlooked when projects of this nature are undertaken. Just write in to the Heath Company requesting your free copy, or attach a memo to your next order.

HEATH COMPANY • Benton Harbor 26, Mich.

CHECK THESE *Features*

- Visual and aural signal tracing
- Two channel input
- High RF sensitivity
- Unique noise locater circuit
- Calibrated wattmeter
- Substitution test speaker
- Utility amplifier
- RF, audio probes and test leads included

Heathkit VISUAL-AURAL SIGNAL TRACER KIT

MODEL T-3

$23.50

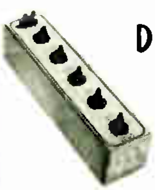
SHIPPING WEIGHT 10 POUNDS

An entirely new type of signal tracer incorporating a combination of features not found in any other instrument. Designed expressly for the radio and TV service man, particularly for the servicing of AM, FM, and TV circuits. Here in a five tube, transformer operated instrument are all of the useful functions so necessary for speedy, accurate isolation of service difficulty.

This new signal tracer features a special high gain RF input channel, used in conjunction with a newly-designed wide frequency range demodulator probe. High RF sensitivity permits signal tracing at the receiver antenna input. A separate low gain channel and probe available for audio circuit exploration. Both input channels are constantly monitored by an electron ray beam indicator, so that visual as well as aural signal indications may be observed. The instrument can also be used for comparative estimation of gain per stage.

A decidedly unusual feature is a noise localizer circuit in conjunction with the audio probe. With this system, a DC potential is applied to a suspected circuit component and the action of the voltage in the component can be seen as well as heard. Invaluable for ferreting out noisy or intermittent condensers, noisy resistors, controls, coils, IF and power transformers, etc. A built-in calibrated wattmeter circuit is very useful for a quick preliminary check of the total wattage consumption of the equipment under test. Separate panel terminals provide external use of the speaker or output transformer for substitution purposes. Saves valuable service time by eliminating the necessity for speaker removal on every service job. The terminals also permit the utilization of other shop equipment, such as your oscilloscope or VTVM. The T-3 Signal Tracer can be used as a high gain amplifier for checking tuners, record changers, microphones, phono crystals, etc.

Don't overlook the interesting service possibilities provided through the use of this new instrument and let it work for you by saving time and money. The kit is supplied complete with all tubes, circuit components, demodulator probe, audio probe, and additional test leads.



Heathkit DECADE RESISTANCE KIT

MODEL DR-1 **$19.50** SHIP. WT. 4 LBS.

The Decade Resistance Kit provides individual switch selection of resistance values using twenty 1% resistors providing a choice of 1 to 99,999 ohms in 1 ohm steps. Ceramic wafer switches, silver-plated contacts, smooth, positive detent action, baked enamel panel, and handsome, polished birch cabinet.

Heathkit DECADE CONDENSER KIT

MODEL DC-1 **$16.50** SHIP WT 4 LBS.

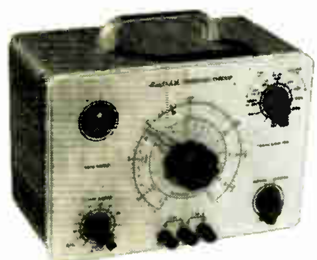
The Heathkit Decade Condenser Kit features silver mica, precision condensers with a rated accuracy of ± 1%. Capacity values are arranged in three decades from 100 mmf to .111 mf in steps of 100 mmf. Ceramic wafer switches with silver-plated contacts and smooth detent action. Useful in laboratory work, for circuit development.

Heathkit RESISTANCE SUBSTITUTION BOX KIT

MODEL RS-1 **$5.50** SHIP. WT. 2 LBS.

The Heathkit Resistance Substitution Box provides individual switch selection of any one of 36 RTMA 1 watt 10% standard value resistors, ranging from 15 ohms to 10 meghoms. Many applications in circuit development work, and also in radio and TV service work. Ideal for experimentally determining resistance values and for quickly altering circuit operating characteristics. Entire unit housed in attractive Bakelite cabinet, featuring the new universal type Heathkit binding posts to simplify circuit connections.

Heathkit CONDENSER CHECKER KIT



MODEL C-3

$19.50

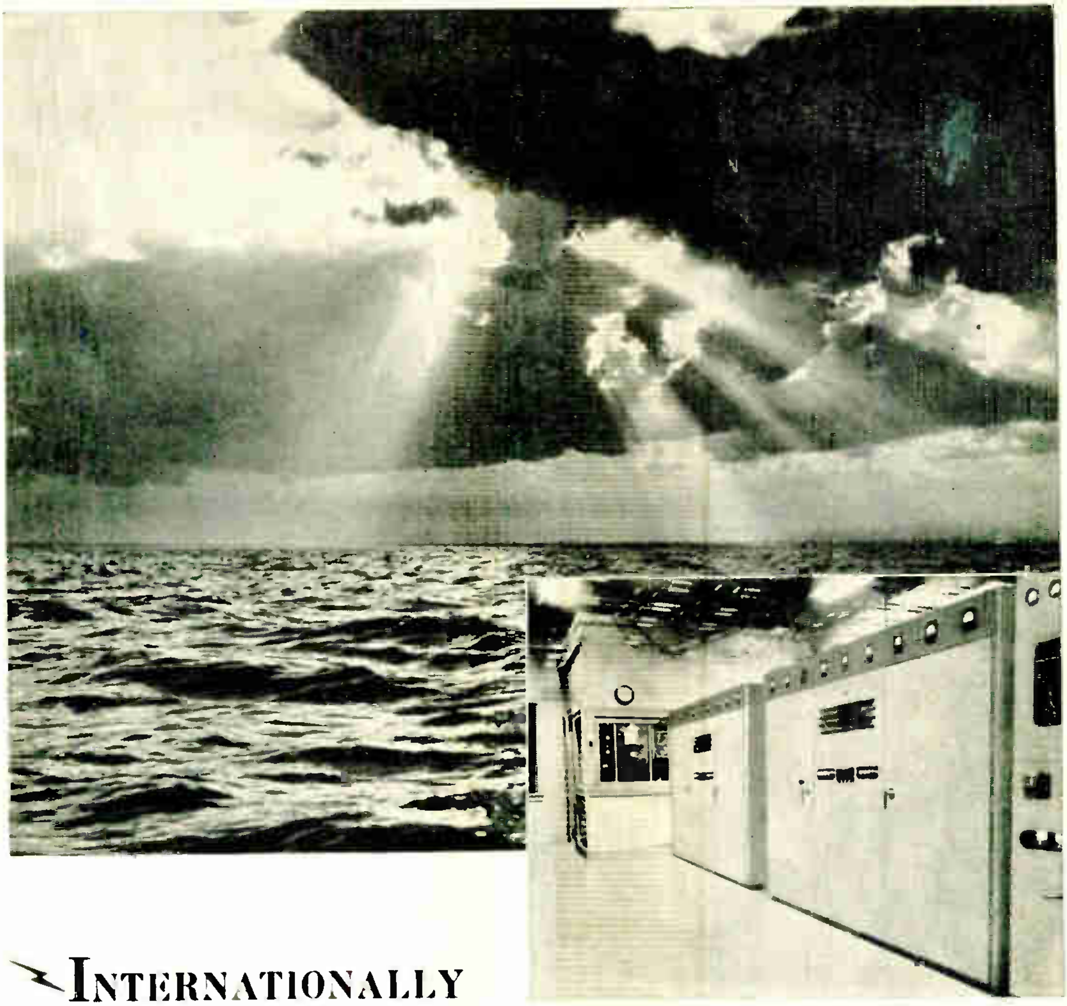
SHIPPING WT. 8 POUNDS

Use the Heathkit C-3 Condenser Checker to quickly and accurately measure those unknown condenser and resistor values. All readings are taken directly from the calibrated panel scales without requiring any involved calculation. Capacity measurements in four ranges from .00001 mf to 1,000 mf. Checks paper, mica, ceramic, and electrolytic condensers. A power factor control is available for accurate indication of electrolytic condenser measurements. A leakage test switch with switch selection of five polarizing voltages, 25 volts to 450 volts DC, will indicate condenser operating quality under actual load condition. The spring return leakage test switch automatically discharges the condenser under test and eliminates shock hazard to the operator.

Resistance measurements can be made in the range from 100 ohms to 5 megohms. Here again all values are read directly on the calibrated scale. Increased circuit sensitivity coupled with an electron beam null indicator increases overall instrument usefulness.

For safety of operation the circuit is entirely transformer operated and the instrument is housed in the attractive, newly-styled Heathkit cabinet, featuring rounded corners, and drawn aluminum panel. The outstanding low kit price for this surprisingly accurate instrument includes necessary test leads. Good service shop operation requires the use of this specialized instrument, designed for the express purpose of determining unknown condenser values and operating characteristics.

HEATH COMPANY • Benton Harbor 26, Mich.



INTERNATIONALLY SPEAKING

Two Gates 10,000 watt transmitters installed at Press Wireless, Inc., Hicksville, L. I., N. Y., one of the world's foremost companies in international communications.

When it comes to handling reams of press copy, commercial messages and facsimile on an around the clock basis—more and more, engineers are agreed that the rugged, straightforward, easy to service and dependable Gates high powered transmitters are the answer.

Date line—Bangkok, Thailand, Egypt, Norway, Dakar, Alaska, San Salvador, Korea, New York, or you name it — the best guess is — copy via Gates-made international short wave transmitters. Any Gates office will gladly participate in your planning. Overseas customers will also find a Gates representative close at hand.

GATES

GATES RADIO COMPANY

Manufacturing Engineers Since 1922

QUINCY, ILLINOIS . . . U. S. A.

INTERNATIONAL—13 East 40th St., New York City
(Cable - - - ARLAB)

DOMESTIC

51 East 42nd St.
New York City

7501 Sunset Blvd.
Los Angeles, Calif.

Warner Bldg.
Washington, D. C.

13th & Spring Sts.
Atlanta, Georgia

2700 Polk Avenue
Houston, Texas

Canadian Marconi Co.
Montreal, Canada

VIKING II TRANSMITTER KIT

JOHNSON VIKING II TRANSMITTER KIT

TVI suppressed, bandswitching and completely self-contained, the Viking II transmitter delivers 130 watts CW output and 100 watts phone 160 through 10 meter amateur bands.

Output Range

Band	Frequency Range	Band	Frequency Range
160	1.8 mc. — 2.4 mc.	20	9.8 mc. — 15.0 mc.
80	2.9 mc. — 4.4 mc.	15	15.0 mc. — 21.8 mc.
40	5.2 mc. — 8.0 mc.	10-11	21 mc. — 30 mc.

Effectively TVI suppressed, features include a heavily copper plated cabinet serving as a simple RF shield. Lid is bonded to cabinet with silver plated phosphor bronze contact fingers yet is easily opened by removing just three thumb screws. Cabinet lid and bottom plate are perforated for effective air cooling.

Special shields provided for meter, dial window and shaft openings. Choke-capacitor type filters for power line leads, VFO receptacle, antenna relay terminals as well as key and mike jacks, plus internal filters at RF tubes effectively eliminate spurious radiation. Wide range pi-network output tuning system provides up to 30 DB second harmonic attenuation before any external filtering. Shielded coaxial connectors are used for RF output and VFO input terminals.

RF section consists of a 6AU6 oscillator, a 6AQ5 buffer doubler and parallel 6146 output amplifier. Modulator; pp 807's operating class ABI with 6AU6 speech amplifier and 6AU6 driver. Parallel 5R4GY HV rectifiers. 5V4G low voltage rectifier with 6AL5 bias rectifier and 6AQ5 clamper screen voltage regulator. Fixed bias applied to buffer and output amplifier for break-in CW operation. Audio response limited to center of speech range.

All parts furnished including a complete set of tubes, cabinet, punched chassis, wiring harness, wire, terminals, grommets and all other hardware. Carefully detailed and illustrated instructions for assembly, test, and operation are also included.

Supplied for 115 volt 50/60 cycle operation only. Cabinet dimensions 20″ wide, 10⅝″ high, 13″ deep. Net weight assembled, 65 pounds.

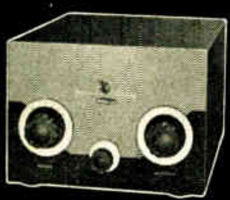
Cat. No.		Amateur Net
240-102	Viking II Transmitter Kit with tubes, less crystals, key, and mike	**$279.50**
240-102-2	Viking II Transmitter, wired and tested	**337.00**

VIKING VFO KIT

VIKING VFO KIT

Variable frequency oscillator with 160 and 40 meter output for frequency multiplying transmitters. Accurately calibrated for all amateur bands from 160 thru 10 meters. 6AU6 electron coupled oscillator, OA2 voltage regulator. Excellent stability is assured by temperature compensated padders and rigid construction. 6-1 vernier tuning with high reset accuracy. Power requirements 6.3 volts, 3 amperes, 250-300 volts 15 ma, DC unregulated. (Power and input connections provided on every Viking transmitter.) Kit furnished complete, less tubes. All parts, assembly and calibration instructions included. When used with a Viking II or other shielded transmitter having filtered power leads, no TVI suppression measures are required.

Cat. No.		Amateur Net
240-122	Viking VFO Kit, less tubes	**$42.75**
240-122-2	Viking VFO Kit, wired and tested	**63.75**



MATCHBOX ANTENNA COUPLER

JOHNSON "MATCHBOX"

"Matchbox" loads an almost infinite variety of antennas from 3.5 to 30.0 mcs. Matches balanced antennas from 25 to 1200 ohms resistance. Loads unbalanced, or single wire antennas of approximately 25 to 3000 ohms resistance. Tunes out large amounts of reactance. Self-contained antenna changeover relay. Bandswitching—front panel controls—no internal adjustment required to change bands.

Cat. No.		Amateur Net
250-23	Johnson "Matchbox", assembled, wired, tested	**$49.85**

LOW PASS FILTER

LOW PASS FILTER

The Johnson Low Pass Filter consists of four individually shielded sections. Handles more than 1000 watts RF, provides 75DB or more attenuation above 54 mc. Insertion loss, less than .25DB. Teflon insulated fixed capacitors are replaceable. SO-239 coaxial connectors at each end. Furnished completely wired and pre-tuned.

Cat. No.		Amateur Net
250-20	Low Pass Filter	**$13.50**

SIGNAL SENTRY

JOHNSON SIGNAL SENTRY

CW OR PHONE MONITOR—ALL BANDS, NO TUNING

A non frequency sensitive RF actuated monitor, the Johnson Signal Sentry monitors CW or phone on all amateur bands to 50 mc. Triggered directly by transmitter RF, unit acts as "On the Air" indicator because transmitter failure is immediately apparent by lack of monitor output. Circuit designed to mute receiver audio when energized by transmitter enabling break-in operation. Powered from receiver or other available supply, the Signal Sentry is easily installed by plugging into receiver headphone jack—plugging headphones into monitor and coupling RF probe to transmitter. Controls: CW TONE and audio VOLUME.

Maroon cabinet only 3⅞″ x 3⅝″ x 3¾″. Unit is supplied wired and tested including instructions and all cables and connectors for installation. Tubes required but not furnished: one 12AX7, one 12AU7. Shipping weight 3 lbs.

Cat. No.		Amateur Net
250-25	Johnson Signal Sentry, less tubes and power supply	**$14.70**

S.W.R. BRIDGE

SWR BRIDGE

Provides accurate measurement of standing wave ratios to insure most effective use of a low pass filter and antenna coupler for ultimate in TVI suppression. 52 ohms impedance can be changed to 70 ohms or other desired value. Equipped with SO-239 connectors and polarized meter jacks.

Cat. No.		Amateur Net
250-24	SWR Bridge	**$9.75**

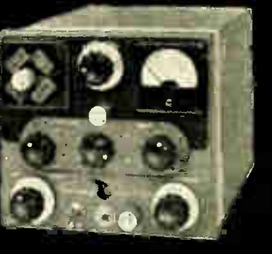
6 BAND VIKING MOBILE TRANSMITTER

Power packed mobile kit designed especially for advanced amateurs who demand peak performance. Rated at 60 watts maximum PA input, the Viking Mobile Transmitter features **instant bandswitching** (75, 40, 20, 15, 11, and 10 meters) and **gang-tuning for maximum operating convenience.** 100% AM modulation engineered for communications audio pass band. Adjustable, single control coupling system simplifies loading. Ganged coupling circuits for each band are pre-tuned for efficient antenna matching. Separate PA trimmer permits unusually large frequency excursions without antenna loading coil adjustments. RF section, 6BH6 oscillator, 6AQ5 buffer doubler and 807 power amplifier. Powerful PP 807 modulator, operates well within ratings and is designed for extra audio punch necessary for solid mobile communications. 6BH6 speech amplifier and 6BH6 driver has sufficient gain for either high impedence or carbon microphone.

Unique RF bias system protects RF tubes and modulators and eliminates power-wasting cathode resistor bias. Special "Tune, Receive, Transmit" switch enables "non-swish" VFO tuning and receiver muting. Other controls: Audio Gain, PA Exitation, Meter Switch, Crystal selector and filament On-Off.

Unit may be wired for either 6 or 12 volts. Designed for maximum efficiency using power supplies delivering 300 volts (30 watts PA input) to 600 volts (60 watts PA input) at 200 MA, the Viking Mobile is furnished in kit form with all parts, punched chassis, hardware and connectors. Step by step instructions completely illustrated for simplified assembly by the experienced amateur. Complete control wiring specifications and antenna suggestions included. Dimensions 6 7/16"x7 1/8"x10 5/16". Weight 16 lbs. Less tubes, crystals, microphone and power supply.

Cat. No.		Amateur Net
240-141	Viking Mobile Transmitter Kit	**$99.50**
240-141-2	Wired and Tested Viking Mobile available on special order.	



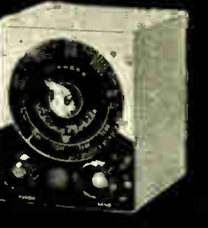
VIKING MOBILE TRANSMITTER KIT

VIKING MOBILE VFO

A diminutive variable frequency oscillator designed specifically for mobile use, the Viking Mobile VFO measures only 4"x4 1/4"x5". Double bearing ceramic insulated tuning capacitor . . . ceramic air dielectric trimmers . . . ceramic switches . . . ceramic coil form and heavy aluminum cabinet minimize frequency shift due to road shock and vibration. Split colpitts oscillator, compensated for an extremely wide range of ambient temperatures, is voltage regulated. Separate amplifier-multiplier stage, operating essentially class A, provides isolation for constant oscillator loading.

Edge lighted lucite dial with vernier tuning designed for maximum visibility and accurate reset; VFO is calibrated for 75, 40, 20, 15, 11 and 10 meter bands. Output 3.5 to 4 mc. for 75 meter band and 7.05 to 7.45 for 40, 20, 15 and 10. 10.5 mc. output also available for straight doubling to 15 meters. Tube lineup, 6BH6 oscillator, 6BH6 buffer-multiplier and OA2 regulator. Adequate output will drive any straight pentode crystal stage. Cables and connectors for use with Viking Mobile supplied; easily adapted for use with virtually any mobile transmitter. Complete kit includes all parts, hardware and assembly and operating instructions. Weight approximately 3 lbs.

Cat. No.		Amateur Net
240-152	Viking Mobile VFO Kit, less tubes	**$29.45**
240-152-2	Viking Mobile VFO wired and tested, less tubes	**44.95**



VIKING MOBILE VFO

"JOHNSON BI-NET"

Fully Automatic — Dual Band Antenna Resonator — Patents Pending

Dual mobile antenna loading network for 10 and 20 meters. Mounts in center of standard mobile whip antenna for completely automatic band change while in motion. No relays or mechanical control required. Consists of two adjustable silver plated inductors and ceramic insulated fixed capacitor enclosed in a streamlined plastic housing. 3/8" x 24 female threads for antenna mounting. Size, 4 7/16" high, 5 5/16" long, 2 3/8" wide. Weight, 14 oz.

Cat. No.	Amateur Net
250-22	**$10.95**

BI-NET ANTENNA RESONATOR

DYNAMOTOR POWER SUPPLIES

Supplies plate voltages for all stages of JOHNSON VIKING MOBILE and VFO; PA input approximately 50 watts. Base contains contactor, fuses, filter and 50 watt adjustable dropping resistor. Supplied with connectors for Viking Mobile. Completely wired and assembled. Dimensions 6 1/8" x 7 1/4" x 7 1/8". Rated 500 volts, 200 ma. intermittent. Weight 13 3/4 lbs.

Cat. No.		Amateur Net
239-102	Dynamotor Power Supply 6 volt primary	**$89.50**
239-104	Dynamotor Power Supply 12 volt primary	**92.50**

DYNAMOTOR BASE KITS

Complete kit with all parts as above less dynamotor. Supplied with receptacles and plugs for Viking Mobile power and control cables.

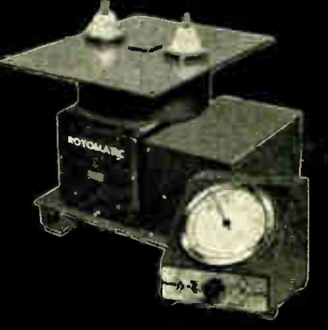
Cat. No.		Amateur Net
239-101	6 volt base kit	**$16.50**
239-103	12 volt base kit	**17.40**

IMPROVED ROTOMATIC ANTENNA ROTATOR

Designed for rigorous service, the Johnson Rotomatic Rotator supports beam antennas weighing up to 175 pounds even under heavy icing conditions or high wind loading. Rotates 1 1/4 RPM — full 360° either direction—over-all gear reduction, 1200 to 1.

Heavily chrome plated RF slip rings provide smooth, noise free operation and low contact resistance. Auxiliary slip rings provided for antenna switching relay control. Rotator housing is cast aluminum; with 5/16 steel rotating table. Unit hinged to tilt 90° for antenna adjustments; can be rotated in tilted position. Complete assembly includes attractive desk top control box with selsyn indicator. Azimuth bearings continuously presented on illuminated dial. Controls include reversible rotation switch, power switch, and antenna relay switch. Weight 76 pounds.

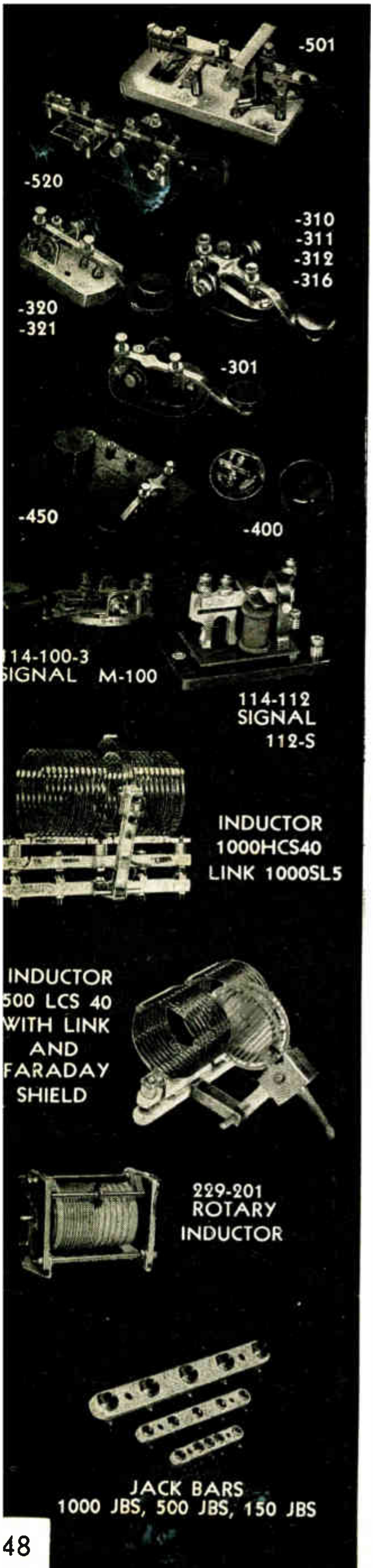
Cat. No.		Amateur Net
138-112	Rotomatic Rotator	**$324.00**
144-16	8 Conductor Control Cable	**.26 ft.**



ROTOMATIC ROTATOR AND DIRECTION INDICATOR

PARASITIC BEAM ANTENNAS—ELEMENT KITS—BOOMS

Descriptive literature and prices available on request.



E. F. JOHNSON COMPANY

224 SECOND AVENUE SOUTHWEST

SPEED-X KEYS, PRACTICE SETS, BUZZERS

Standard Semi-Automatic Keys

Improved model, heavy steel base, rubber feet. Chrome plated vibrator and hardware. Ten adjustments, lowest and highest speeds. Circuit closing switch. Adjustable paddles.

114-500 ⅛" contacts; black wrinkle.. $13.50
114-501 ¼" contacts, polished chrome . 19.20
114-501L Lefthanded 21.00

New Special Semi-Automatic Key

Combines best features of former amateur and professional models. Heavy cast metal base, 6¼" x 3" x ½", finished in black wrinkle enamel. Vibrator same as on deluxe key—easy action and speed adjustment from lowest to highest speeds. Hardware and vibrator heavily chrome plated. ⅛" coin silver contacts. Lock nuts. Rubber mounting feet. Circuit closing switch.

114-520 Sp. Model, Semi-Automatic . . $11.50

Heavy Duty Keys

Chrome plated key arm. ¼" coin silver contacts. Navy knob.

114-320 Black wrinkle enamel base . . $3.60
114-321 Polished chrome plated base 5.10

Standard Keys

High quality, low cost. Provision for plugging in semi-automatic key. ⅛" coin silver contacts.

114-310 Black wrinkle, less switch . . . $ 3.00
114-310S Black wrinkle, with switch . 3.25
114-311 Chrome plated, less switch . . 4.90
114-311S Chrome plated, with switch 5.15
114-316 Brass wrinkle, less switch 3.00

Molded Base Keys

Black phenolic base. ⅛" coin silver contacts. Metal parts nickel plated.

114-301 Less switch. $1.95

Practice Keys

For beginners. ⅛" coin silver contacts.

114-300 Molded brown phenolic base. $1.85

Practice Set

Constant frequency buzzer & key mounted on 4" x 6" phenolic base.

114-450 Code practice set. $3.95

Constant Frequency Buzzer

Fully adjustable, holds frequency. Uses 2 dry cells or "C" battery.

114-400 Constant frequency buzzer . $1.50

Telegraph Sounder

Formerly manufactured by the Signal Electric Company, Menominee, Michigan, this improved telegraph sounder is designed for instant response. Brass sounder gives clear resonant tone—steel bar frame with black enamel finish. Brass bridge and adjustment screws, with instrument lacquer finish—black lacquered steel sounder plate. Instrument mounted on mahogany finished wood base, brass binding posts and rubber feet.
Net weight 2 lbs., shipping weight 3 lbs.

114-112 Signal No. 112-S (4 ohms resistance)
114-113 Signal No. 113-S (20 ohms resistance)

Telegraph Key M100

Another outstanding signal key, now manufactured by E. F. Johnson, this professional telegrapher's key is equipped with circuit closing switch. Base, binding posts and switch lever are brass with instrument lacquer finish. Polished nickel plated key lever. Platinor contacts .072" diameter. Net Weight 1 lb.

114-100-3 Signal No. M-100

HI-Q AIR WOUND HAM INDUCTORS

Swinging link inductors for amateur bands 160 thru 6 meters; 150, 500 and 1000 watt sizes. Two inductance values for each band permit choice of appropriate L/C ratio dictated by amplifier plate voltage and plate current. Polystrene insulation, Steatite bases and heavier wire sizes insure highest efficiency. HCS—Inductors match high voltage, low current tubes. LCS—Inductors match low voltage, high current tubes.

Swinging Link Coils

1000 Watt Type No.	Max. Cap.*	Net Price	500 Watt Type No.	Max. Cap.*	Net Price	150 Watt Type No.	Max. Cap.*	Net Price
1000HCS160	99	$7.50	500HCS160	100	$3.75	150HCS160	102	$3.30
1000LCS160	140	7.50	500LCS160	148	3.75	150LCS160	151	3.30
1000HCS80	46	6.70	500HCS80	45	3.45	150HCS80	51	3.00
1000LCS80	73	6.70	500LCS80	76	3.45	150LCS80	68	3.00
1000HCS40	24	6.05	500HCS40	27	3.15	150HCS40	28	2.70
1000LCS40	55	6.05	500LCS40	50	3.15	150LCS40	57	2.70
1000HCS20	19	5.55	500HCS20	25	2.70	150HCS20	21	2.40
1000LCS20	26	5.55	500LCS20	37	2.70	150LCS20	32	2.40
1000H/LCS14	19	5.20	500H/LCS14	19	2.25	150H/LCS14	19	2.10
1000H/LCS10	18	4.90	500H/LCS10	19	2.10	150H/LCS10	19	1.95
			500H/LCS6	18	2.10	150H/LCS6	16	1.95

Digits at the end of catalog numbers indicate the amateur band inductors tune.
*Maximum capacity required for resonance at the low frequency end of band.

Jack Bar Assemblies

Watts	Type No.	Net Price
150	150JBS	$ .90
500	500JBS	1.20
1000	1000JBS	1.80

Rotary Inductor

Type No. **229-201** (not illustrated) Variable pitch winding of No. 14 tinned copper wire. Maximum inductance 10 microhenries. Form and end plates Steatite, roller contact. Overall size: 2½" wide x 4½" long x 3" high. Net Price 8.85

Swinging Link Arm Assemblies

150/500SLA For 150/500 Watt Inductors. $1.15
1000SLA For 1000 Watt Inductors . . . 1.40

"Plug-In" Swinging Links

Type No.	No. Turns	Net Price
150/500SL12	12	$1.80
150/500SL5	5	1.15
150/500SL2	2	1.00
1000SL10	10	1.90
1000SL5	5	1.45
1000SL2	2	1.10

FARADAY SHIELD

Cat. No.	Description	Net Price
238-303	150/500 watt swinging link shield, hood and lead assembly	$2.60
238-304	Same as above, for 1000 watts	2.75
238-301	105/500 watt link shield only	1.45
238-302	1000 watt link shield only	1.60

Metallic plated screen on polystyrene discs.

CAPACITORS, INDUCTORS, SOCKETS, INSULATORS, PLUGS AND JACKS, KNOBS AND DIALS, AND PILOT LIGHTS

WASECA MINNESOTA

VARIABLE CAPACITORS

Partial Listing

This is a partial listing of the large JOHNSON line of quality capacitors. Several types are not shown, likewise many additional sizes are available in most types. All types employ steatite insulation. Approximate flashover voltage is 100x final numerals in catalog numbers (except Type N). "L" dimension is overall length less shaft extension.

TYPES C and D

Sturdy, rigid construction at low cost! Aluminum plates .051" thick, rounded edges. Panel space Type C, 5½" wide x 5⅜" high, Type D, 4¼" wide and 4" high.

Type C—Single Section

Type No.	Max. Cap.	Air Gap	No. Plates	L	Net Price
250C70	252	.175"	24	6 13/16	14.25
500C70	496	.175"	47	12 3/16	17.40
350C90	337	.250"	43	14 27/32	17.00
50C110	51	.350"	8	4 25/32	8.75
100C130	102	.500"	21	13 11/32	12.20

TYPE D—Single Section

Type No.	Max. Cap.	Air Gap	No. Plates	L	Net Price
500D35	496	.080"	39	6 25/32	13.50
150D45	146	.125"	17	4 25/32	8.35
70D70	72	.175"	11	4 25/32	7.50
100D70	98	.175"	15	4 25/32	8.10
250D70	244	.175"	37	10 5/16	12.00
100D90	99	.250"	19	7 11/16	9.15
150D90	149	.250"	29	10 5/16	11.00

Type C—Dual Section

Type No.	Max. Cap. Per Sect.	Air Gap	No. Plates	L	Net Price
300CD70	305	.175"	29	16 25/32	22.80
150CD90	147	.250"	19	14 27/32	18.00
50CD110	50	.350"	8	10 5/16	12.35
100CD110	103	.350"	17	16 25/32	18.40
50CD130	51	.500"	10	14 27/32	14.70

TYPE D—Dual Section

Type No.	Max. Cap. Per Sect.	Air Gap	No. Plates	L	Net Price
500DD35	496	.080"	39	13 11/32	20.70
150DD45	155	.125"	18	9 15/32	13.30
50DD70	52	.175"	8	5 13/16	9.50
70DD70	72	.175"	11	7 11/16	11.00
100DD70	97	.175"	15	9 15/32	12.40
150DD70	151	.175"	23	13 11/32	16.00
50DD90	52	.250"	10	9 15/32	11.00

TYPES E and F

Rugged compact units for low and medium power transmitters. Aluminum plates .032 thick, rounded edges. Stainless steel shafts. Panel space, Type E, 2⅝" square, Type F, 2 1/16" square.

TYPE E—Single Section

Type No.	Max. Cap.	Air Gap	No. Plates	L	Net Price
350E20	353	.045"	33	3 17/32	6.10
500E20	488	.045"	45	4 15/32	7.05
100E30	100	.075"	15	2 9/16	4.95
250E30	251	.075"	37	4 15/16	6.55
50E45	53	.125"	12	2 31/32	4.75
150E45	145	.125"	33	6 3/32	6.30

TYPE F—Single Section

Type No.	Max. Cap.	Air Gap	No. Plates	L	Net Price
100F20	106	.045"	17	2¼	5.30
150F20	154	.045"	25	2⅞	5.90
100F30	99	.075"	25	3 19/32	5.90
150F30	148	.075"	37	4⅞	6.75

TYPE E—Dual Section

Type No.	Max. Cap. Per Sect.	Air Gap	No. Plates	L	Net Price
300ED20	312	.045"	29	6 21/32	9.90
100ED30	99	.075"	15	5⅜	7.85
150ED30	153	.075"	23	7 1/16	9.20
200ED30	196	.075"	29	8⅜	10.80
50ED45	52	.125"	12	6 5/32	7.20
100ED45	100	.125"	23	9 9/32	9.45

TYPE F—Dual Section

Type No.	Max. Cap. Per Sect.	Air Gap	No. Plates	L	Net Price
100FD20	104	.045"	17	4 23/32	7.60
150FD20	153	.045"	25	6	8.80
70FD30	66	.075"	17	5 23/32	7.70
100FD30	99	.075"	25	7 7/16	8.90

TYPE M MINIATURE

Smallest ever built, yet tops in performance. Ideal for VHF, miniature test equipment, etc. Pane space ⅝" x ¾". Air gap .017. Mounts in ¼" hole.

Single

Type No.	Capacity Max.	Capacity Min.	Net Price
5M11	5.0	1.5	.95
9M11	8.7	1.8	1.00
15M11	14.2	2.3	1.15
20M11	19.6	2.7	1.30

Differential

Type No.	Capacity Max.	Capacity Min.	Net Price
6MA11	5.0	1.5	1.40
9MA11	8.7	1.8	1.55
15MA11	14.2	2.3	1.75
19MA11	19.6	2.7	2.00

Butterfly

Type No.	Capacity Max.	Capacity Min.	Net Price
3MB11	3.1	1.5	1.35
5MB11	5.1	1.8	1.50
9MB11	8.0	2.2	1.70
11MB11	10.8	2.7	1.90

TYPE L

Ceramic soldered—no eyelets or rivets to loosen. All brass, soldered construction. "Bright alloy" plated. Ideal for rough service. Panel space 1⅜" square. Air gap .030". In addition to those listed, also available in Differential types.

Single End Plate

Type No.	Cap. per Sect. Max.	Cap. per Sect. Min.	No. Plates	Net Price
10L15.........	11	2.8	3	1.85
25L15.........	27	3.5	7	1.95
50L15.........	51	4.6	13	2.15
75L15.........	75	5.7	19	2.35

Single End Plate

Type No.	Cap. per Sect. Max.	Cap. per Sect. Min.	No. Plates	Net Price
100L15........	99	6.8	21	3.60
200L15........	202	11.6	51	4.50

Dual Section

Type No.	Cap. per Sect. Max.	Cap. per Sect. Min.	No. Plates	Net Price
25LD15........	27	3.5	7	3.90
50LD15........	51	4.6	13	4.30
100LD15.......	99	6.8	25	5.05

Butterfly

Type No.	Cap. per Sect. Max.	Cap. per Sect. Min.	No. Plates	Net Price
10LB15.........	10.5	2.8	5	2.20
25LB15.........	26	4.3	12	2.50
50LB15.........	51	6.8	23	2.90

TYPE R CAPACITOR

JOHNSON version of a highly popular standardized capacitor widely used in compact portable and mobile equipment. End plates are of extra heavy nickel-plated brass, Steatite insulating bars. All soldered and riveted construction.

Cat. No.	Type No.	Net Price	Plates per Sect.	Length
149-1	**20R12**	**$2.05**	3	1 7/32
149-2	**35R12**	**2.10**	5	1 7/32
149-3	**50R12**	**2.15**	7	1 7/32
149-4	**75R12**	**$2.30**	11	1 7/32
149-5	**100R12**	**2.45**	14	1 13/32
149-6	**140R12**	**2.65**	19	1 19/32

TYPE N

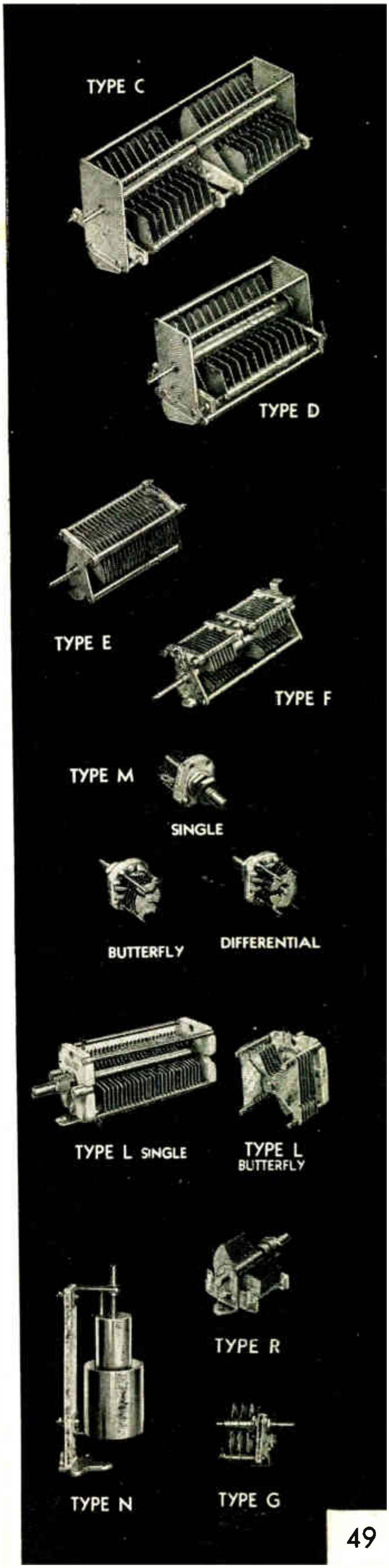
Small mounting space requirements, extremely high voltage rating and fine adjustment make these neutralizing capacitors ideal.

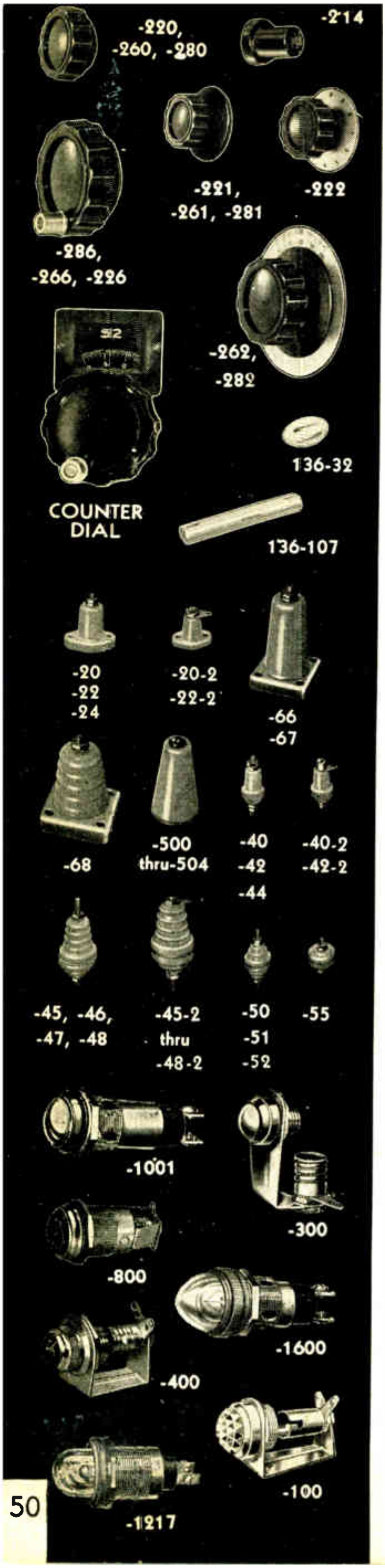
Type No.	Capacity Max.	Capacity Min.	Spacing	Net Price
N125........	11.0	1.1	.125"	5.70
N250........	10.6	1.4	.250"	6.75
N375........	10.7	1.7	.375"	8.60

TYPE G

Extremely popular as neutralizing capacitors for medium and low power stages. Also widely used for grid and plate tuning at high frequencies.

Type No.	Cap. per Sect. Max.	Cap. per Sect. Min.	Spacing	Net Price
13G45.......	13	4.7	.125"	2.75
6G70........	5.7	3.5	.225"	3.00
12G70.......	12	6	.225"	3.50





E. F. JOHNSON COMPANY

224 SECOND AVENUE SOUTHWES

JOHNSON KNOBS & DIALS

Featuring fresh, advanced styling, these new JOHNSON models will enhance the appearance of your equipment. Molded phenolic knobs have 12 well defined flutes, large gripping area. Knob faces slightly convex, sides slightly tapered to contribute to pleasing appearance. Beautiful satin chrome scales that will retain their new appearance indefinitely. Each knob and dial has brass set screw insert molded in place. Standard shaft size ¼", "specials" available for 3/16" shafts. Special models available or quantity orders.

Cat. No.	Description	Net Price
	KNOBS	
116-220	Knob, 1⅛" dia., black phenolic	$.29
116-260	Knob, 1⅝" dia., black phenolic	.39
116-280	Knob, 2⅜" dia., black phenolic	.59
	SKIRTED KNOBS	
116-221	Knob, 1⅛" dia. with 1½" black phenolic skirt	$.54
116-261	Knob, 1⅝" dia. with 2x1/16" black phenolic skirt	.67
116-281	Knob, 2⅜" dia. with 3" black phenolic skirt	.82
	VERNIER DIALS	
116-265	Dial, 2¾" satin chrome scale 0-100 over 180 degrees with 3 to 1 friction vernier drive and single line indicator	3.00
116-285	Dial, 4" chrome scale 0-100 over 180 degrees with 5 to 1 friction vernier drive and single line indicator	3.60
	COUNTER DIAL	
116-208-1	Counter dial with lock, escutcheon and 2⅜" spinner knob	$11.10
116-208-4	Same as above without dial lock	9.80

Cat. No.	Description	Net Price
	DIALS	
116-222	Dial, 1½" bevelled satin chrome scale with 1⅛" knob. Scales as follows:	
116-222-1	100-0 over 180 degrees	$ .56
116-222-2	10-0 over 270 degrees	.56
116-222-3	7-1 over 180 degrees	.56
116-222-4	ON-OFF over 60 degrees	.56
116-222-5	Single line	.56
116-262	Dial, 2¾" satin chrome scale with single line indicator 0-100 over 180 deg.	1.40
116-282	Dial, 4" satin chrome scale with single line indicator 0-100 over 180 degrees	2.00
	SPINNER KNOBS	
	For multi-turn devices such as variable inductors, potentiometers etc.	
116-226	Spinner knob 1⅛" black phenolic	$.56
116-266	Spinner knob 1⅝" black phenolic	.67
116-286	Spinner knob 2⅜" black phenolic	.82
	INSTRUMENT KNOB	
116-214-1	—for ¼" shaft	$ .33
116-214-2	—for 3/16" shaft	.33

FEEDER INSULATORS

Nos. 136-122, -124 and -126 are conventional feeder spreaders of high grade low absorption porcelain, Silicone impregnated. No. 136-122 has notches for 1½" line spacing. All have ⅜x½" cross section.

Cat. No.	Lgth.	Net Price
136-122	2"	$.12
136-124	4"	.17
136-126	6"	.22

ANTENNA INSULATORS

136-107-112, are wet process 1" in dia. 136-104 is dry process ⅝" square. 136-32 is dry process compression strain type, 1½" long. All antenna insulators glazed.

Cat. No.	Br'k'g Strength	Length	Net Price
136-104	400 lbs.	4"	$.15
136-107	800 lbs.	7"	.75
136-112	800 lbs.	12"	.85
136-32	800 lbs.	1½"	.10

INSULATORS AND BUSHINGS

Stand-Off Insulators

Cat. No.	H	Hardware	Net Price	Cat. No.	H	Hardware	Net Price
			STEATITE				
135-20	1 3/16"	10-32	.15	135-22-2	1"	74-Jack	.16
135-20-2	1 3/16"	74-Jack	.20	135-24	⅝"	6-32	.10
135-22	1"	8-32	.13				
			PORCELAIN				
135-60	4½"	¼-20	.68	135-62	2¾"	¼-20	.36
			Metal Base Types				
135-65	1⅜"	10-32	.21	135-67	4½"	¼-20	.72
135-66	2¾"	¼-20	.50	135-68	2"	10-32	.34
			Steatite Cone Insulators				
135-500	⅝"	6-32	.20	135-503	2"	10-32	.49
135-501	1"	8-32	.24	135-504	3"	10-32	.85
135-502	1½"	8-32	.43				
			Thru-Panel Insulators				
			STEATITE				
135-40	1¼"	10-32	.25	135-42-2	⅞"	74-Jack	.32
135-40-2	1¼"	74-Jack	.33	135-44	⅝"	6-32	.18
135-42	⅞"	10-32	.24				
			PORCELAIN				
135-45	1⅜"	10-32	.33	135-47	4½"	¼-20	1.24
135-45-2	1⅜"	74-Jack	.41	135-47-2	4½"	76-Jack	1.46
135-46	2¾"	¼-20	.84	135-48	2"	10-32	.48
135-46-2	2¾"	76-Jack	1.06	135-48-2	2"	74-Jack	.60
			Lead-In Bushings				
			STEATITE				
135-50	½"	6-32	.24	135-52	1⅛"	¼-20	.59
135-51	13/16"	10-32	.40	135-55	¼"	6-32	.23

PILOT LIGHTS

A partial listing of the basic JOHNSON pilot light types in greatest demand. Jewel colors available are red, green, blue, amber, opal and clear.

Cat. No.	Jewel	Socket	Net Price	Cat. No.	Jewel	Socket	Net Price
147-100	1" Faceted	Min. Scr.	.60	147-802	1" Faceted	Cand. Scr.	.71
147-101	1" Smooth	Min. Scr.	.60	147-803	1" Smooth	Cand. Scr.	.71
147-103	1" Faceted	Cand. Scr.	.61	147-804	1" Faceted	Min. Bay.	.70
147-104	1" Smooth	Cand. Scr.	.61	147-805	1" Smooth	Min. Bay.	.70
147-106	1" Faceted	Min. Bay	.63	147-808	1" Color Disc	Min. Bay.	.75
147-107	1" Smooth	Min. Bay.	.63	147-1000*	1" Faceted	Cand. Scr.	1.02
147-300	½" Faceted	Min. Scr.	.30	140-1001*	1" Smooth	Cand. Scr.	1.02
147-301	½" Smooth	Min. Scr.	.30	147-1002*	1" Color Disc	Cand. Scr.	1.07
147-303	½" Faceted	Cand. Scr.	.31	147-1003**	1" Faceted	Cand. Scr.	.99
147-304	½" Smooth	Cand. Scr.	.31	147-1004**	1" Smooth	Cand. Scr.	.99
147-306	½" Faceted	Min. Bay.	.33	147-1005**	1" Color Disc	Cand. Scr.	1.04
147-307	½" Smooth	Min. Bay.	.33	147-1217	1" Lucite	Cand. Scr.	1.16
147-400	½" Faceted	Min. Scr.	.47	147-1218	1" Lucite	Min. Bay.	1.10
147-401	½" Smooth	Min. Scr.	.47	147-1219	1" Lucite	D.C. Bay.	1.46
147-403	½" Faceted	Min. Bay.	.50	147-1600	1" Bullseye	Cand. Scr.	1.54
147-404	½" Smooth	Min. Bay.	.50	147-1604	1" Bullseye	S.C. Bay.	1.46
147-800	1" Faceted	Min. Scr.	.69	147-1605	1" Bullseye	D.C. Bay.	1.59
147-801	1" Smooth	Min. Scr.	.69				

* For S-6 bulb ** For NE-45 neon bulb

CAPACITORS, INDUCTORS, SOCKETS, INSULATORS, PLUGS AND JACKS, KNOBS AND DIALS, AND PILOT LIGHTS

WASECA MINNESOTA

TUBE SOCKETS

Highest Quality Sockets for Every Application

Cat. No.	Description	Net Price
123-206	Industrial bayonet, Steatite, silver plated beryllium copper contacts. Base is 4 pin super jumbo. Tension springs in shell	$2.25
123-209	Medium 4 pin bayonet, white glazed porcelain base, metal shell, heavy phosphor bronze side wiping contacts. 2 13/16" Dia.	1.10
123-209SB	Same as 209 but with Steatite base and beryllium copper contacts	2.30
123-210	Same as 209 except contact to shell spacing not as great. 2 1/2" Dia.	1.10
123-211	Standard 50 watt type. Similar to -209 but double filament contacts. 3 3/8" Dia.	1.40
123-211SB	Same as 211 but with Steatite base and beryllium copper contacts	2.90
124-212	Steatite socket for RCA833 or 833A. 5 1/8" plate leads	6.55
123-216	Giant 5 pin Bayonet. For tubes such as 803, RK28. 3 3/4" Dia.	2.60
123-216SB	Same as -216 but with Steatite base and beryllium copper contacts	4.00
124-213	For Eimac 152TL and 304TL. Contacts arranged for either series or parallel filaments	1.45
124-214	For Eimac 1500TH, with ventilating hole for cooling	1.80
124-215	For 250 watt tubes such as 204A, 849, etc. The plate terminal has a "safety cup" which prevents accidental dislodgement	3.60

Wafer Types

Steatite top and sides glazed. Brass contacts with steel springs cadmium plated.

Cat. No.	Description	Net Price
122-237	Giant 7 pin Steatite wafer. For transmitting tubes such as HK257 and RCA813. With 3/4" diam. ventilating hole in base	.72
122-247	7 pin Steatite for tubes such as 826. Etched aluminum shield	.85
122-244	4 pin Steatite. Super jumbo base tubes such as 8008	1.50
122-101	7 pin Steatite wafer with shield, retainer springs and provision for mounting button mica by-pass capacitors. Designed for VHF use with tubes such as 839	2.00
122-275	Giant 5 pin Steatite wafer socket for 4-125A, RK48 tubes. Ventilation holes in base	1.31

	Net Price		Net Price		Net Price
122-217 7 pin small	$.52	122-225 5 pin	$.42	122-227 7 pin medium	$.49
122-224 4 pin	.39	122-226 6 pin	.46	122-228 Octal socket	.52

Miniature Sockets

120-267	all ceramic, 7 pin	$.33
120-277B	with shield base, 7 pin	.49
133-277S	shield base only	.12

JAN Miniature Sockets

Top mounting, saddle type sockets per JAN spec. S-28A.

120-177	7 Pin	.59
120-199	9 Pin	.79

Miniature Shields

Cat. No.	Description	For Socket	
133-278-6	1 3/8" High, N.P. Brass	177,277	$.15
133-278-7	1 3/4" High, N.P. Brass	177,277	.19
133-278-8	2 1/4" High, N.P. Brass	177,277	.22
133-278-9	1 1/2" High, N.P. Brass	199	.22
133-278-10	1 15/16" High, N.P. Brass	199	.24
133-278-11	2 3/4" High, N.P. Brass	199	.29

PLUGS AND JACKS

Banana Spring Type

Accurately turned from brass, with milled nuts and tinned terminals. Nickel plated. Nickel-silver springs. Beryllium copper springs and other platings available in production quantities. Low contact resistance, high current capacity.

-75 series plugs fit —74 series jacks, -77 series plugs fit -76 jack. -7451 and -7452 have molded phenolic heads.

JACKS

		Net Price
108-74	1/4—28 x 11/32 thread	$.10
108-7451	1/4—28 x 1/2 thread, red	.20
108-7452	1/4—28 x 1/2 thread, black	.20
108-76	3/8—24 x 15/16 thread	.26

PLUGS

108-75	6—32 x 3/8 thread	.11
108-75A	6—32 x 3/4 thread	.12
108-75BB	3/8 x 1 3/8 handle, black	.28
108-75BR	3/8 x 1 3/8 handle, red	.28
108-75C	6—32 x 5/16 screw	.11
108-77	10—32 x 3/8 thread	.24
108-77A	10—32 x 3/4 screw	.27
108-77BB	5/8 x 1 3/4 handle, black	.39
108-77BR	5/8 x 1 3/4 handle, red	.39

Tip Jacks and Plugs

PLASTIC HEAD TIP JACKS

Attractively colored strong Plaskon heads, accurately threaded 1/4-32 with milled hex nut and insulating washers for 3/8 hole.

Cat. No.	Color	Net Price
105-520	Red	$.14
105-521	Black	.14
150-522	Dk. Green	.14
105-524	Brown	.14
105-525	Lt. Blue	.14
105-526	Orange	.14
105-527	Yellow	.14
105-528	Lt. Green	.14
105-529	Dk. Blue	.14
105-530	Ivory	.14

Molded Tip Jacks

Heavy duty type. Nickel plated brass, body molded into phenolic head. 5/16-40 thread, and insulating washers for 3/8 hole.

No. 105-418	Red	$.22
No. 105-419	Black	.22

All Metal Tip Jack

Nickel plated brass, 5/16 hex head, 1/4-32 thread, with insulating washers for 3/8 hole. 105-1 similar but headless, no nut or washers, for mounting in 1/4-32 tapped panel hole.

No. 105-417 ... $.14 No. 105-1 ..... $.10

Solderless Tip Plugs

No. 105-15 13/16 prong	$.16
No. 105-415 9/16 prong	.15

NYLON TIP JACKS

Patents Pending

Completely insulated jack, body molded from low-loss Nylon. Threaded 1/4-32, jack mounts with single nut. Overall dimensions, diameter 3/8", length 29/32". Available with beryllium copper contacts.

B.C. Cont. Cat. No.	Color	Net Price	B.C. Cont. Cat. No.	Color	Net Price
105-601-1	White	$.30	105-606-1	Orange	$.30
105-602-1	Red	.30	105-607-1	Yellow	.30
105-603-1	Black	.30	105-608-1	Brown	.30
105-604-1	Dark Green	.30	105-609-1	Light Green	.30
105-605-1	Light Blue	.30	105-610-1	Dark Blue	.30
105-611-1	Ivory	.30			

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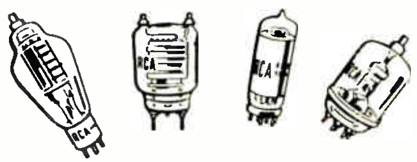

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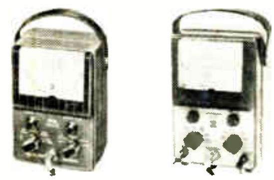

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WR-49A	RF Signal Generator	59.50
WV-77A	Junior *Volt Ohmyst**	47.50
WV-97A	Senior *Volt Ohmyst**	67.50
WV-87A	Master *Volt Ohmyst**	112.50

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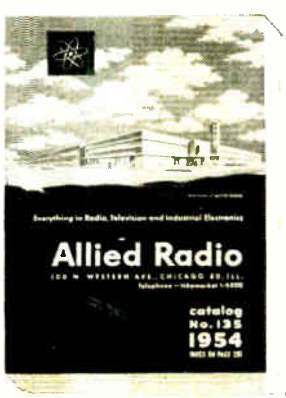

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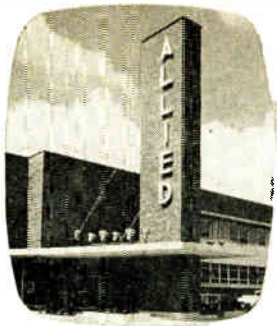

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S-76	Dual-Conversion Receiver	199.95
S-72A	4-Band Portable Receiver	109.95
S-72L	Portable with Long Wave Band	119.95
SX-73	Ultra-Deluxe Receiver	975.00
SX-62	All-Wave Receiver	349.95
SX-71	Dual IF Receiver	249.95
S-38C	Low-Cost All-Wave Receiver	59.95
S-81	"Civic Patrol" Receiver (153-173 mc)	59.95
S-82	"Civic Patrol" Receiver (30-50 mc)	59.95
R-46	Matching Speakers for Hallicrafters Receivers SX-62, SX-88, SX-71, SX-73, S-76, etc.	24.95

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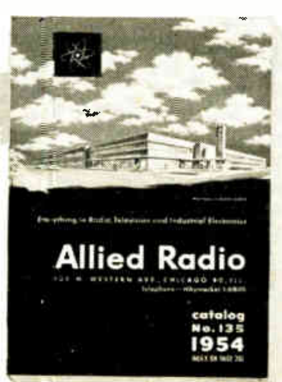

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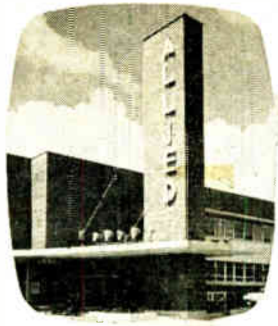

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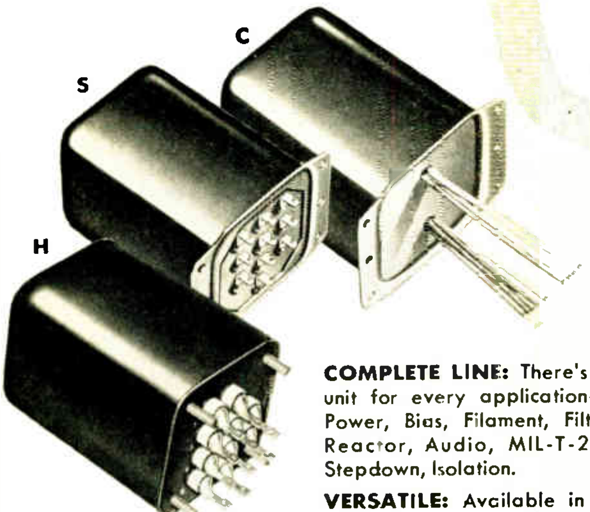
AMATEUR & INDUSTRIAL
"Sealed-in-Steel" TRANSFORMERS



COMPLETE LINE: There's a unit for every application—Power, Bias, Filament, Filter Reactor, Audio, MIL-T-27, Stepdown, Isolation.

VERSATILE: Available in 3 constructions to meet all needs.

H-TYPE. Steel base cover is deep-seal soldered into case. Terminals hermetically sealed. Ceramic bushings. Stud-mounted unit. Meets all MIL-T-27 specs.

S-TYPE. Steel base cover fitted with phenolic terminal board. Convenient numbered solder lug terminals. Flange-mounted unit.

C-TYPE. With 10″ color-coded stripped and tinned leads brought out through fibre board base cover. Flange-mounted unit.

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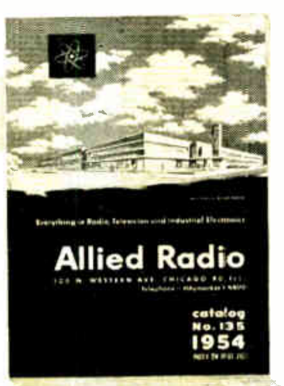

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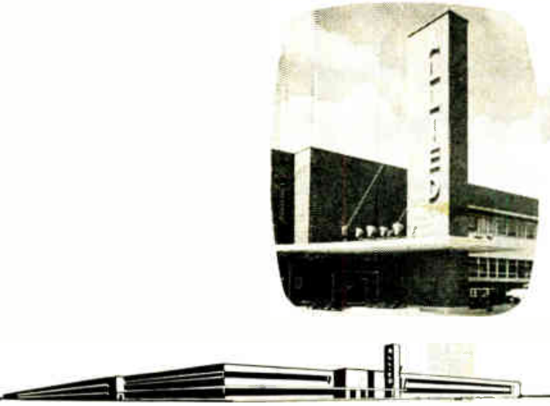

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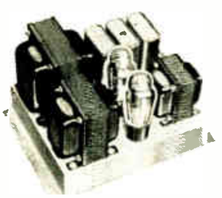


Fig. A. Bandmaster DeLuxe. A complete 50-watt phone-CW transmitter with instant bandswitching 80 through 2 meters (8 bands). Includes new crystal-oscillator-vfo switching circuit. Tubes: 6AQ5 osc., 6AQ5 mult., 807 final. Speech amplifier for crystal mike uses 2-6AU6, 1-12AU7 phase inv., 2-6L6 mod. Requires APS-50 or DPS-50 power supply below. Size, 8 x 12 x 8". Complete with tubes, less crystal and mike. Shpg. wt., 20 lbs.
97 SX 792. Bandmaster DeLuxe Transmitter.......... **$137.50**

Bandmaster Senior Transmitter. 50-watt phone-CW as above, but for use with single-button carbon microphone. Shpg. wt. 20 lbs.
97 SX 791. Bandmaster Senior Transmitter............ **$111.50**

Fig. B. Bandmaster VFO. Covers 80/75, 40, 20, 15, 11/10 meters. Requires: 6.3 v. at 0.65 amp., 300 v. at 30 ma. Sloping front panel; slide-rule dial. Uses 6AG7 and OB2 reg. 9 x 11½ x 4". Wt., 9 lbs.
98 SX 043. Bandmaster VFO **$47.50**

APS-50 AC Power Pack. For use with above transmitters. Delivers 425 v. at 275 ma., and 6.3 v. at 4 amps. With 2-5U4G rect. For 110 v. A.C. 50-60 cycles. 11 x 6⅞ x 8¾". Shpg. wt., 27 lbs.
97 SX 698. APS-50 AC Power Pack.................. **$39.50**

DPS-50-6 Dynomotor. For portable operation of above x-mitters, from 6 volt storage battery. Output: 300 v. at 250 ma. 10⅛ x 5¼ x 5⅞". Shpg. wt., 16 lbs.
97 S 697. DPS-50-6 Dynamotor...................... **$87.50**

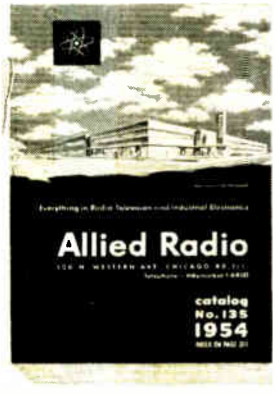
DPS-50-12 Dynomotor. For use with any 12 volt storage battery. Delivers 400 v. DC at 250 ma. Shpg. wt., 16 lbs.
98 SX 776. DPS-50-12 Dynamotor.................. **$54.50**

Prices subject to change

Order your HARVEY-WELLS equipment from ALLIED. Get quick, expert shipment straight from stock. And take advantage of our easiest-pay terms and 15-day trial offer.

268-Page ALLIED Buying Guide

Write for your FREE copy. It's your *complete* guide to *everything in* Amateur and Industrial electronic supplies. Lists everything you need—includes full selection of equipment described above. Be sure to get and use your 268-page ALLIED Catalog.

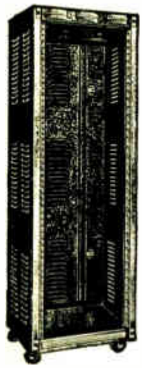


KEEP IT HANDY

ALLIED RADIO

100 N. Western Ave.
Chicago 80, Illinois

DE LUXE RELAY RACKS



These relay racks are made of 16 gauge steel with 1/8" panel supports. The panel mounting supports are recessed so that no edges of the panel will be exposed.

The front and back of the top, the two sides and the door are well louvered to provide adequate ventilation. Snap catches are positioned on the door. A streamlined appearance is achieved by the use of rounded corners and red-lined chrome trim. The relay rack is shipped knocked-down and complete with all necessary hardware for assembly. All standard 19" panels will fit these racks.

A SPECIAL FEATURE IS THE USE OF FOUR STURDY SUPPORTS ON THE BOTTOM SO THAT CASTERS CAN BE FASTENED DIRECTLY TO THE BASE, THEREBY ACHIEVING READY MOBILITY. Bud RC-7756 casters will fit this unit. Casters are not included in price of cabinet. These relay racks are supplied in either black or grey wrinkle or grey hammertone finish. The overall width is 22" and the depth is 17 1/4" on all sizes listed.

Catalog No.	Overall Height	Panel Space	Shipping Wt.
CR-1774	42 1/16"	36 3/4"	90 lbs.
CR-1771	47 5/16"	42"	100 lbs.
CR-1772	66 9/16"	61 1/4"	135 lbs.
CR-1773	82 5/16"	77"	155 lbs.

ADD-a-RACK SERIES

It has always been necessary to buy special racks without louvers on one side to obtain a maximum of panel space with a minimum of floor space. Now, you no longer need to buy a whole new cabinet when you want additional panel space. Through our new and exclusive Add-a-Rack series, BUD not only offers additional racks at a lower cost, but provides you with a sturdier, better looking assembly.

The illustration at top shows two Add-a-Rack cabinets assembled together. The illustration below shows the unique and ingenious method of adding a unit to your present equipment. Instead of buying an entire new outfit, you purchase only four parts: (1) a door (2) a top (3) a bottom and (4) an Add-a-Rack coupling-unit. The right (or left) hand side of your present relay rack is removed and replaced by the Add-a-Rack coupling-unit; next, a top and bottom is fastened into place, and the side taken from the first rack is fastened onto the second rack which has been added. Place the additional door into position and you have two racks properly and efficiently coupled together. In the same simple way, more racks can be added at any time and every one will be in a CONTINUOUS ONE-PIECE assembly.

This series is available in two ways. (1) a double unit consisting of two racks and the Add-a-Rack coupling unit, (2) Add-a-Rack unit, consisting of a door, a top, a bottom and an Add-a-Rack coupling-unit. These units are furnished with all necessary assembling and panel mounting hardware. Choice of finish same as racks above.

Add-a-Rack Unit	Used to Add-a-Rack to	Shipping Weight
AR-1778	CR-1774	70 lbs.
AR-1775	CR-1771	75 lbs.
AR-1776	CR-1772	100 lbs.
AR-1777	CR-1773	127 lbs.

Complete unit consisting of the knocked-down parts necessary for two relay racks coupled together.

CR-1779 two coupled relay racks same size as CR-1774

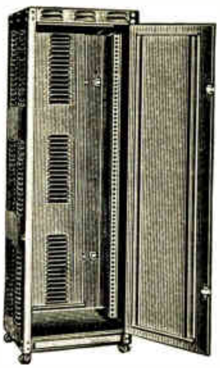
CR-1780 two coupled relay racks same size as CR-1771

CR-1786 two coupled relay racks same size as CR-1772

CR-1799 two coupled relay racks same size as CR-1773

Bud RC-7756 Casters will fit this unit. Casters are not included in price of cabinet.

SUPER DE LUXE RACKS (2 door)



This new Relay Rack is made of 16 gauge steel with 1/8" panel supports. The construction is similar to the series of Bud de luxe Relay Racks shown above. The panel mounting supports are recessed, so that no edges of the panel will be exposed, and they are also adjustable from front to back at various stopping points. This enables you to utilize the space in front and behind the panel to any degree. When placed as far back as the knock-outs provide, the panel is 6" from the front of the Rack.

These Racks have both front and rear doors; the rear door to cover any of the equipment behind the panel, providing easy access. The front door provides a means of concealing dials, knobs, etc., that may be in the front of the panel.

These relay racks also have the exclusive Bud feature of supports on the bottom, so that the casters may be fixed directly to the base. AVAILABLE IN BLACK OR GREY WRINKLE OR LIGHT GREY HAMMERTONE FINISH AT NO EXTRA CHARGE.

Catalog No.	Overall Height	Panel Space	Shipping Wt.
CR-2174	43 1/16"	36 3/4"	110 lbs.
CR-2171	47 5/16"	42"	122 lbs.
CR-2172	66 9/16"	61 1/4"	165 lbs.
CR-2173	82 5/16"	77"	190 lbs.

DE LUXE CABINET RACKS

These cabinet racks have rounded corners and attractive red-lined chrome trim. There is a recessed, hinged door on the top with a snap catch. These racks are made of heavy gauge steel and are of sturdy construction. The five large sizes have a hinged rear door, while the small sizes have a welded panel in the rear.

Adequate ventilation is assured by means of louvered sides and a two inch opening in the bottom of the back extends the entire width. "NO-SCRATCH" EXTENDED METAL FEET ARE EMBOSSED ON THE BOTTOM TO MINIMIZE MARRING OF A TABLE TOP. Racks are furnished in either black or grey wrinkle or grey hammertone finish. Depth 14 3/4", width 22". Will fit standard 19" panels.

Catalog No.	Overall Height	Panel Space	Shipping Wt.
CR-1741	10 9/16"	8 3/4"	29 lbs.
CR-1740	12 5/16"	10 1/2"	31 lbs.
CR-1742	14 1/16"	12 1/4"	32 lbs.
CR-1739	15 13/16"	14"	36 lbs.
CR-1743	19 5/16"	17 1/2"	40 lbs.
CR-1727	22 13/16"	21"	45 lbs.
CR-1744	28 3/16"	26 1/4"	50 lbs.
CR-1728	33 9/16"	31 1/2"	55 lbs.
CR-1745	36 13/16"	35"	60 lbs.

STANDARD RELAY RACK PANELS

Made of Steel or Aluminum. Steel Panels are made of high grade steel 1/8" thick. Aluminum Panels are made of 1/8" thick Aluminum. All Panels are 19" wide. Furnished in either black or grey wrinkle or grey hammertone. Aluminum panels 3/16" thick may be had if desired at 60% increase in cost over 1/8".

STEEL Catalog No.	STEEL Height	ALUMINUM Catalog No.	ALUMINUM Height
PS-1250	1 3/4"	PA-1101	1 3/4"
PS-1251	3 1/2"	PA-1102	3 1/2"
PS-1252	5 1/4"	PA-1103	5 1/4"
PS-1253	7"	PA-1104	7"
PS-1254	8 3/4"	PA-1105	8 3/4"
PS-1255	10 1/2"	PA-1106	10 1/2"
PS-1256	12 1/4"	PA-1107	12 1/4"
PS-1257	14"	PA-1108	14"
PS-1258	15 3/4"	PA-1109	15 3/4"
PS-1259	17 1/2"	PA-1110	17 1/2"
PS-1260	19 1/4"	PA-1111	19 1/4"
PS-1261	21"	PA-1112	21"

Illustrated are only a few of the many types and sizes of Bud Products. For complete catalog write Dept. R-54

BUD RADIO, INC. 2118 East 55th Street, Cleveland 3, Ohio

Precision Made Sheet Metal Products for Every Purpose

STEEL CHASSIS BASES

These chassis are made from one piece of steel, all corners are reinforced and spot welded. The four sides are folded on bottom for additional strength — this also permits a bottom plate to be attached if desired. Furnished in either Black Wrinkle or Electro-Zinc plated.

Black Wrinkle Cat. No.	Zinc Plated Cat. No.	Depth	Width	Height	Gauge
CB-628	CB-629	5"	7"	2"	22
CB-790	CB-1192	7"	9"	2"	22
CB-650	CB-774	8"	17"	2"	20
CB-636*	CB-637	10"	17"	3"	20
CB-658	CB-771	11"	17"	3"	18
CB-660*	CB-773	13"	17"	3"	18
CB-642*	CB-643	13"	17"	4"	18

* Indicates chassis which are punched to accommodate Chassis Mounting Brackets.

For additional sizes consult Bud Catalog

ALUMINUM CHASSIS

The construction and design of these chassis is exactly the same as our steel chassis. The aluminum chassis are welded on government approved spot welders that are the same as used in the welding of aluminum airplane parts. As a result, you can depend on BUD Aluminum Chassis to do a perfect job. Etched Aluminum finish. The gauges in table below are aluminum gauges.

Cat. No.	Depth	Width	Height	Gauge
AC-430	4"	6"	3"	18
AC-402	5"	7"	2"	18
AC-423	7"	17"	3"	16
AC-425	8"	17"	2"	16
AC-420	13"	17"	3"	14
AC-426	11"	17"	2"	14
AC-416	10"	17"	3"	16

For additional sizes consult Bud Catalog

INSTRUMENT AND RECEIVER CABINETS

Each cabinet has an evenly recessed hinged cover with convenient finger lift. The panel on front of cabinet is readily attached with self-tapping screws. Louvers provide ample ventilation. These Cabinets are finished in black wrinkle only.

Cat. No.	Height	Width	Depth
C-973	7"	8"	8"
C-993	7"	10"	8"
C-994	7"	12"	8"
C-995	7"	14"	8"
C-1190	8"	16"	8"
C-975	9"	15"	11"

METAL UTILITY CABINETS

A large number of sizes available makes this line useful for all types of electronic equipment. These cabinets have two removable panels for easy accessibility and are finished in black wrinkle finish only. Those units prefixed by CU are made from steel, those prefixed by AU are made from high grade sheet aluminum.

Cat. No.	Depth	Width	Height
CU-883	2"	4"	4"
CU-728	3"	5"	4"
CU-729	4"	5"	6"
CU-1098	6"	6"	6"
CU-1099	5"	6"	9"
CU-879	7"	8"	10"
CU-1124	6"	7"	12"
CU-880	8"	10"	10"
CU-881	8"	11"	12"
CU-882	7"	9"	15"
AU-1083	2"	4"	4"
AU-1028	3"	5"	4"
AU-1029	4"	5"	6"
AU-1039	6"	6"	6"
AU-1040	5"	6"	9"

NEW BUD FILTERS TO REDUCE OR ELIMINATE TELEVISION INTERFERENCE

The sources of television interference are most often short wave broadcasting stations, amateur radio transmitting stations, diathermy equipment, X-ray equipment, automotive ignition noises or similar sources. The basic problem of eliminating this interference is that of rejection of the signals received from these sources.

LF-601 LOW PASS FILTER

Interference to television receiver reception caused by transmissions from an amateur station can be caused by harmonics or by shock from the transmitter. The shock from the transmitter fundamental can be cured at the television receiver with a Bud HF-600 high pass filter. Harmonics can be greatly reduced or eliminated at the transmitter by use of a Bud LF-601 low pass filter.

The LF-601 high attenuation low pass filter has the following characteristics:

• Minimum attenuation of 85 decibels on all frequencies above 54 megacycles and a minimum of 93 decibels above 70 megacycles. • Maximum rejection is adjustable from 55 to 90 megacycles. This tunable feature provides two slots at least 100 decibels down • The cut-off frequency is 42 megacycles • The unit will easily handle a full kilowatt modulated on a reasonably flat line • The insertion loss is less than one DB • Since the design of this filter provides an adjustable feature, the unit can be used with either 52 ohm or 72 ohm coax • Each inductance is in an individually shielded compartment • All capacitors used are variable • Size 12" x 2½" x 2¼".

HF-600 HIGH PASS FILTER

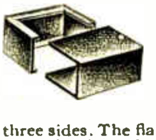
The HF-600 high pass filter has a cut off frequency at 42 megacycles, thus this filter rejects signals from 0 to 42 megacycles. It is within this range that the majority of signals causing interference are received. Since there is no attenuation above 42 megacycles, picture strength or quality is not affected. This unit is easily installed and complete installation instructions are included. The filter is housed in an attractive aluminum case 3¼" x 2⅛" x 1⅛".

NEW PANEL CHASSIS

This new series of Panel Mounting Chassis is for professional type installations primarily. These units are constructed from .062 aluminum with an etched finish. The holes on the front flange are suitable for mounting to a Rack, and are dimensioned to fit standard panel notchings.

Cat. No.	Width	Depth	Height
CB-1370	19"	5 9/32"	1¾"
CB-1371	19"	5 9/32"	3½"
CB-1372	19"	5 9/32"	5¼"
CB-1373	19"	5 9/32"	7"
CB-1374	19"	5 9/32"	8¾"
CB-1375	19"	5 9/32"	10½"
CB-1376	19"	5 9/32"	12¼"
CB-1377	19"	5 9/32"	14"

MINIBOXES



There are thousands of uses in the fields of radio and electronics for these new boxes. They are made from heavy gauge aluminum. The design of the box permits installation of more components than would be possible in the conventionally designed box of the same size. It is of two piece construction, each half forming three sides. The flange type construction assures adequate shielding. Available in etched aluminum finish and gray hammerloid finish.

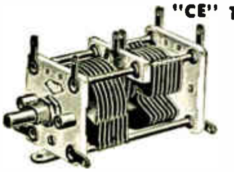
Catalog Numbers Grey	Etched	Length	Width	Height
CU-2100	CU-3000	2¾"	2⅛"	1⅝"
CU-2105	CU-3005	5"	4"	3"
CU-2108	CU-3008	7"	5"	3"
CU-2111	CU-3011	12"	7"	4"
CU-2115	CU-3015	4"	2"	2¾"

For additional sizes consult Bud Catalog

Illustrated are only a few of the many types and sizes of Bud Products. For complete catalog write Dept. R-54

BUD RADIO, INC., 2118 East 55th Street, Cleveland 3, Ohio

BUD Products are made to work better . . . last longer

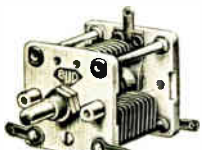


"CE" TYPE DUAL MIDGET CONDENSERS

These Midget Condensers were designed to meet the rigid requirements in design of efficient high frequency electronic devices and precision laboratory equipment. The large front and rear bearings provide for smooth rotation. They feature a rotor wiping contact placed at center of the rotor assembly to assure maximum efficiency at high frequencies. Opposed rotor construction assures perfect counterbalance and provides even torque at any position of rotation. Steatite insulation eliminates closed induction loop in frame. All metal parts cadmium plated.

	PER SECTION				
Catalog Number	Max. Cap.	Min. Cap.	No. of Plates	Air Gap	Distance Behind Panel
CE-2032	35	6	7	.030"	3 1/32"
CE-2033	50	7	9	.030"	3 1/4"
CE-2035	100	9	18	.030"	4 3/32"
CE-2036	150	10	27	.030"	5 3/16"
CE-2041	50	8	15	.060"	4 23/32"

For additional sizes consult Bud Catalog



"CE" MIDGET CONDENSERS SINGLE SECTION DOUBLE BEARING

These Midget Condensers were designed to meet the rigid requirements in design of efficient high frequency electronic devices and precision laboratory equipment. Brass rotor and stator plate stacks are assembled into permanent units by means of electro-soldering, which assures long life and accurate plate spacing. End-plates of Steatite insulate the mounting bushings and angles from the rotor and stator assembles. The large front and rear bearings provide for smooth rotation. Special wiper contact provides noise-free tuning. All metal parts are cadmium plated. Rotor plates semi-circular shaped. Provision for either panel or base mounting.

Catalog Number	Max. Cap. MMFD.	Min. Cap. MMFD.	Air Gap	No. of Plates	Over-all Length
CE-2000	15	4	.030"	3	2 1/2"
CE-2001	35	6	.030"	7	2 23/32"
CE-2002	50	7	.030"	9	2 27/32"
CE-2003	75	8	.030"	14	3 5/32"
CE-2004	100	9	.030"	18	3 11/32"
CE-2005	150	10	.030"	27	3 13/32"
CE-2008	300	15	.030"	52	5 3/16"

For additional sizes consult Bud Catalog

TINY MITE TUNING CONDENSER SINGLE SECTION

This series of condensers has been designed for applications where space or weight are limiting factors and for tuning of high frequency circuits. Rigid construction, close fitting bearing, positive rotor contact and Steatite insulation are the outstanding features. Cadmium plated, soldered, brass plates and rods insure high frequency efficiency.

Catalog Number	Max. Cap. MMFD.	Min. Cap. MMFD	Air Gap	No. of Plates
LC-1640	8	2.5	.017"	3
LC-1644	50	6	.017"	19
LC-1646	100	9	.017"	37
LC-1652*	50	8	.037"	35
LC-1654	15	5.5	.073"	15
LC-1655*	25	9	.073"	27

* Denote double bearing.

For additional sizes consult Bud Catalog

THREE-GANG TINY MITE CONDENSERS

Hams, Radio Constructors and Experimenters can find many uses for these compact, three-gang condensers. Designed particularly for high frequency use, they are adaptable for use in converters, preselectors and receivers covering the Amateur, Television and F.M. bands. Well constructed with soldered brass plates and ceramic brackets. Rotor shaft extended 1/4" at rear. Height 1 5/16". Width 1 3/16". Length behind panel 3 3/8". Mounting holes 2 3/16" apart.

Catalog Number	Cap. Per Section Max.	Min.	No. of Plates Per Section
LC-1845	11	5	3
LC-1846	17	5	4
LC-1847	25	6	5

MIDGET CONDENSERS

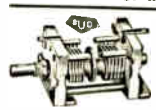
Small size, sturdy construction and high mechanical and electrical efficiency are the outstanding features. Insulation used is Steatite. Rotor and Stator plates are brass and are electro-soldered to their respective rods. All metal parts are cadmium plated.

These condensers have both front and rear bearings and are furnished in either mid-line type plates (straight line wave length), or semi-circular plates (straight line capacity.)

SEMI-CIRCULAR TYPE—DOUBLE BEARING

Catalog Number	Cap. in MMFD. Max.	Min.	Air Gap	Number Plates
MC-1850	15	3	.024"	3
MC-1853	50	5	.024"	7
MC-1855	100	7	.024"	14
MC-1863	50	7	.060"	15
MC-1865	100	12	.060"	31
MC-1867	50	10	.095"	23

For additional sizes consult Bud Catalog



BUD TINY MITE DUAL CONDENSERS

The construction of these units is similar to the regular Tiny Mite Tuning Condensers. The two end pieces are held together firmly with three tie-rods.

A separate round plate is soldered on rotor rod to shield the two stator sections. Large surface front and rear sleeve bearings, provide smooth rotation.

Catalog Number	CAP. PER SECTION Max. MMFD.	Min. MMFD.	Air Gap	No. Plates Per Section	Over-all Length
LC-1659	8	2.5	.017"	3	1 15/16"
LC-1660	15	3	.017"	5	2 1/16"
LC-1661	25	4	.017"	9	2 11/32"
LC-1662	50	6	.017"	19	2 1/32"
LC-1663	100	9	.017"	37	4 1/4"
LC-1664	10	4	.037"	7	2 15/32"
LC-1665	15	5	.037"	11	2 15/16"
LC-1666	25	5.5	.037"	17	3 7/8"
LC-1667	35	6	.037"	21	4"

NEUTRALIZING AND HIGH FREQUENCY TUNING CONDENSERS

This line of condensers will fill every neutralizing and high frequency tuning requirement that modern circuits pose. The two-pillar construction makes this unit unusually sturdy and eliminates any possibility of capacity variation due to vibration. The movable plate is adjusted by means of the threaded shaft to which it is attached, and it is permanently locked in any position by the lock-nut provided. Any loose thread is taken up by a special nut and locked to give smooth operation. All metal parts are of aluminum or brass. Plates have rounded edges. Steatite insulation is used.

Catalog Number	Plate Diameter	MMFD. Capacity Max.	Min.
NC-1000	1 27/32"	11	1
NC-1001	2 13/16"	24	2
NC-1002	4 3/4"	27	6

IRON CORE R. F. CHOKES

The efficiency of any circuit requiring an R. F choke will be definitely improved by utilizing one of these chokes with a finely divided molded metallic core. The improved "Q" possible with this construction results from the D. C. resistance of these chokes being from 40 to 50% less for a given inductance than for regular air-core types. Thus, the D. C. voltage drop through the choke is considerably less, yet the choking action is equally as good. Windings are made with silk-covered enameled wire terminated on convenient soldering lugs, and the chokes are mounted in small square shield cans measuring 1 3/8" x 1 3/8" x 1 7/16".

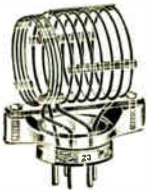
Catalog Number	Inductance mh.	D. C. Resistance Ohms	Current ma.
CH-1277	1.5	11.5	125
CH-1278	2.5	16.	125
CH-1279	3.4	19.5	125
CH-1280	5.5	27.5	125
CH-1281	8.	36.	125
CH-1282	10.	42.5	125
CH-1283	16.	53.	125
CH-1284	30.	82.	100
CH-1285	60.	131.	100
CH-1286	80.	163.	90
CH-1287	125.	221.	90
CH-294	Shield Can Only		

Also available Pie wound *and* Lattice wound Ceramic Coil . . . .

Illustrated are only a few of the many types and sizes of Bud Products. For complete catalog write Dept. R-54

BUD RADIO, INC., 2118 East 55th Street, Cleveland 3, Ohio

BUD Products for high quality and best results



75-WATT TRANSMITTER COILS

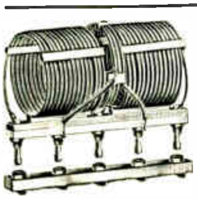
These coils are distinguished by their rigid construction, attractive appearance and conservative power rating. The polystyrene mounting base keeps the coil a safe distance from the chassis — it also permits easy coil removal without disturbing the winding. All coils are air-wound and mount in 5 prong tube sockets.

OEP and OCP Coils are designed for use in circuits using Pentode tubes with high output capacity such as 6L6, 807, etc.

OEL coils have fixed end link and are *not* tapped.
OCL have fixed center link with main winding center tapped.
OLS have adjustable center link, main winding center tapped.
OES have adjustable end link and are *not* tapped.
OEP have adjustable end link and are *not* tapped.
OCP have adjustable center link main winding center tapped.

Catalog No. Fixed End Link	Catalog No. Fixed Center Link	Cat. No. Adjustable Center Link	Cat. No. Adjustable End Link	Band	Capacity*
....	....	OLS-160		160 Meter	100 MMFD
....	....		OES-160	160 Meter	86 MMFD
OEL-80	OCL-80	OLS-80	OES-80	80 Meter	75 MMFD
OEL-40	OCL-40	OLS-40	OES-40	40 Meter	52 MMFD
OEL-20	OCL-20	OLS-20	OES-20	20 Meter	40 MMFD
OEL-15	OCL-15	OLS-15	OES-15	15 Meter	30 MMFD
OEL-10	OCL-10	OLS-10	OES-10	10 Meter	25 MMFD
OEL-6	OCL- 6	.	.	6 Meter	17 MMFD
....	....	OCP-10	OEP-10	10 Meter	45 MMFD
....	....	OCP-20	OEP-20	20 Meter	50 MMFD

AM-8673 Coil Base only



ADJUSTABLE LINK TRANSMITTER COILS

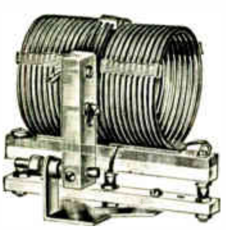
Listed are two types of Coils. CL type of coil has an adjustable CENTER link. ES type of coil has an adjustable END link. The CL and ES can be used where fixed links are specified. No additional cost is involved and more efficient coupling is assured because of this special adjustable link, an *exclusive* BUD *feature*.

150 WATT RATING

Catalog No. Center Link Adjustable	Catalog No. End link Adjustable	Band	Capacity*
RCL-160	RES-160	160 Meters	110 MMFD
RCL-80	RES-80	80 Meters	68 MMFD
RCL-40	RES-40	40 Meters	36 MMFD
RCL-20	RES-20	20 Meters	27 MMFD
RCL-15	RES-15	15 Meters	27 MMFD
RCL-10	RES-10	10 Meters	25 MMFD

AM-1932 — Mounting Base for RCL and RES Coils

Also available in 500W *and* KW *sizes*



VARIABLE LINK TRANSMITTER COILS

The most effective method of varying the loading of an R. F. Stage is by the use of a variable link to the plate tank, a feature incorporated in all Bud Variable Link Coils. The link winding is connected to the jack bar into which the coils are plugged, and this link may be used with any of the coils regardless of the band being worked. The link winding is so arranged that it may be readily controlled from the panel by means of an extension shaft if required.

500 WATT COILS

Catalog Number	Band	Capacity*	Length Mounting Strip Dim.	Mounting Hole Dim.
VLS-160	160 Meter	85 MMFD	5½"	5"
VLS-80	80 M	70 MMFD	5½"	5"
VLS-40	40 M	36 MMFD	5½"	5"
VLS-20	20 M	28 MMFD	5½"	5"
VLS-15	15 M	25 MMFD	5½"	5"
VLS-10	10 M	25 MMFD	5½"	5"

AM-1352 — Base and Link Assembly for 500 Watt Coils

Also avaliable in 150W *and* KW *sizes*

*Denotes tube plus circuit plus tank plus output coupling capacity required to resonate coil at low frequency end of band.

SHIELDED COIL-LINK

These links are made to fit RLS, VLS, and MLS series of coils. This link will prevent capacity coupling between the tank coil and the link and will reduce TVI by greatly attenuating harmonics. The links can be used on co-ax or balanced lines.

Catalog No.	Description
AM-1300	Used with RLS coils (150 W)
AM-1301	Used with VLS coils (500 W)
AM-1302	Used with MLS coils (Kilowatt)

ADD-A-LINK

When the circuit that you are using requires a different number of turns on the coil link than is furnished with the standard coil, the links listed below can be used to replace the standard link.

Cat. No.	Used With	No. of Turns
AM-1303	RLS	3½
AM-1304	RLS	4½
AM-1305	RLS	5½
AM-1307	VLS	3½
AM-1308	VLS	4½
AM-1309	VLS	5½
AM-1310	VLS	6½
AM-1311	MLS	3½
AM-1312	MLS	4½
AM-1313	MLS	5½
AM-1314	MLS	6½

CODE PRACTICE OSCILLATOR AND MONITOR CPO-128

The BUD Codemaster is a real money-saver. No longer do you have to consider your code practice oscillator useless after you have learned the code. A flip of the switch and you have a good CW monitor. This is a really versatile instrument.

It has a 4" built-in permanent magnetic dynamic speaker and will operate up to twenty earphones.

A volume control and pitch control permit adjustments to suit individual requirements. Any number of keys can be connected in parallel to the oscillator for group practice.

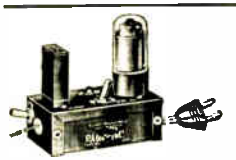
This unit will operate on 110 volts A.C. or D.C. An external speaker may be plugged in without the use of an output transformer. All controls are placed on the front of the unit and all jacks are in the rear. The unit is 6½" high, 5½" wide and 3½" deep. It is finished in Grey Hammertone enamel with red lettering.

MODEL CPO-130

This unit is similar to the CPO-128. The difference is that the 4" speaker is not included. The monitor feature, however, is included. A phone jack is provided for the output and as many as 20 pairs of phones and keys can be operated at one time for class-room operation. This model will also operate a permanent magnetic dynamic speaker. Size is 5½" wide, 4½" high, 3½" deep.

GIMIX GX-79

The BUD Gimix is a multipurpose unit requiring no batteries or power supply. It is calibrated for use on the 10, 15, 20, 40 and 80 meter amateur bands. No additional coils are needed as the one coil does the work on all bands. It can be used as a *Wave-Meter, a Monitor, a Field Strength Indicator, a Carrier Shift Indicator* and a sensitive *Neutralizing Instrument*. Operating instructions supplied with each unit.



FREQUENCY CALIBRATOR FCC-90

To comply with federal regulations, some means of accurately checking transmitter frequency must be available at every "ham" station. The BUD FCC-90 consists of a 100 kc. crystal oscillator that is *Completely Self-Powered*. It will give 100 kc. check points on all bands up to 30 megacycles. This enables the operator to determine exact band edges.

No extra wiring is required to install this unit. Plug the FCC-90 into a 110 volt receptacle, connect the pick-up lead to the antenna binding post of the receiver and the unit is ready for operation. An ON-OFF switch and a STANDBY switch are provided.

Illustrated are only a few of the many types and sizes of Bud Products. For complete catalog write Dept. R-54

BUD RADIO, INC., 2118 East 55th Street, Cleveland 3, Ohio

FOR SALE! These and 100,001 other

1 CONTROLS

— to meet your service requirements. You'll find a type and model for all popular radio, TV sets, plus other electronic gear.


Model B16 Radiohm® Miniature

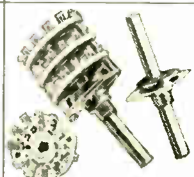

Twin and dual concentric Radiohms

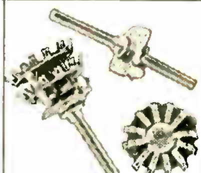

Model B Blue Shaft† Radiohms

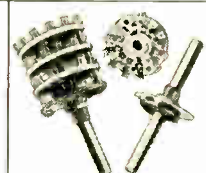

Wirewound Radiohm three and four watt

2 SWITCHES

— hundreds of variations possible. Engineered to give you widest flexibility for all types of applications.


PA-2000 Miniature Rotary, Steatite Insulation


1400 Series, Standard Rotary, Phenolic Insulation


2500 Series Standard Rotary, Steatite Insulation


Small General Purpose and Tone


PCH-4 TV Attenuator

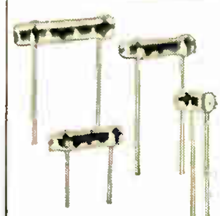

Dual Speaker Switch


Complete Switch Kits

3 CAPACITORS

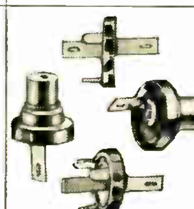
— the world's greatest line of ceramic capacitors.

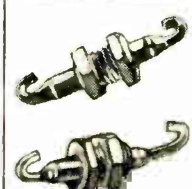
For jobs that demand the best in guaranteed TV-AM-FM servicing.

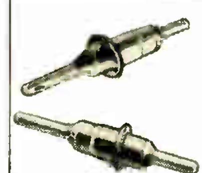



TV HI-VO-KAPS®

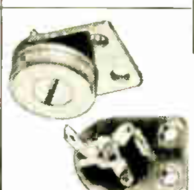

Stand-off


Button-Style


Feed-Thru HI-KAPS®

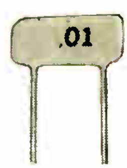

Miniature Feed-Thru HI-KAPS®


Ceramic Trimmers

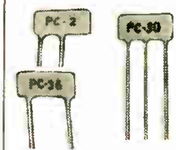

Miniature Ceramic Trimmers

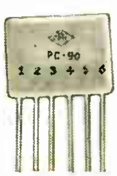
4 PRINTED ELECTRONIC CIRCUITS

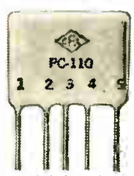
— offering you a new, proved way to reduce extra parts, time, errors and costs—a Centralab "first."



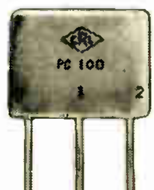
Ceramic Min-Kaps®


Miniature Resistor and Resistor-Capacitor Units

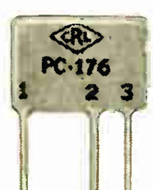

Pentode Couplates®



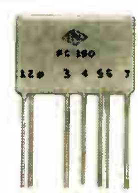
Filplate†
(by-pass and filter application)



Vertical Integrator



Special Plates

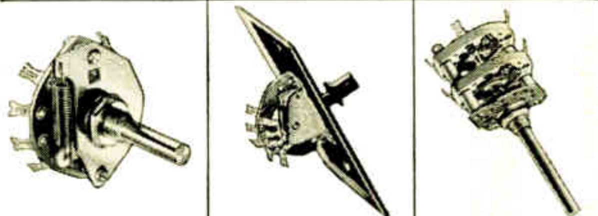

Audet® (audio-detector plate)

Centralab quality electronic components

Type AB Adashaft® Radiohms

Compentrol† Infinitely variable loudness control

Series MX Delta T Type



Rotary Action, Flat

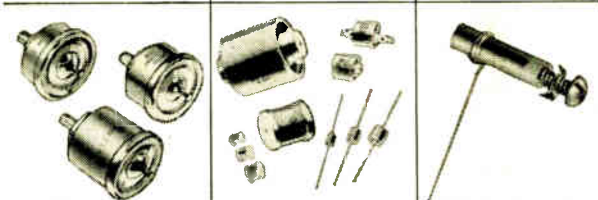
Lever Switch

Ham Type

BC Discs

TC Tubular

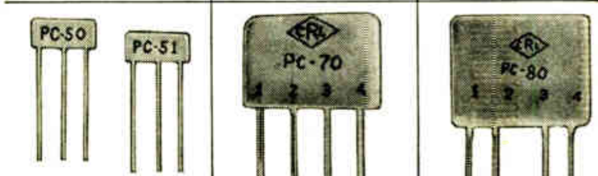
Type MD Molded Disc



High Accuracy Capacitors

Transmitting Capacitors

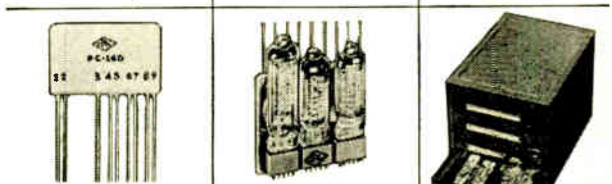
Tubular Ceramic Trimmer



Filpec® (balanced load diode filter)

Standard Triode Couplate®

Midget Triode Couplate®



Pendet† – (Trademark) (pentode detector coupler)

Ampec® standard and miniature 3-stage amplifiers

P.E.C. Kits

HERE'S a perfect check list to determine your electronic needs. And there are hundreds of variations available to you. Whether you need single items or dozens — no other single manufacturer can offer you such a wide choice of standard fastest-for-servicing parts. That's why Centralab is the industry's No. 1 electronic component source . . . and keeps your distributor supplied with the latest developments in the fast-changing electronic field.

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So when you need parts, look for the familiar blue and white CRL package — the product inside is the result of more than 30 years of electronic experience. This experience will continue to pioneer many more profit-building "firsts" for you!

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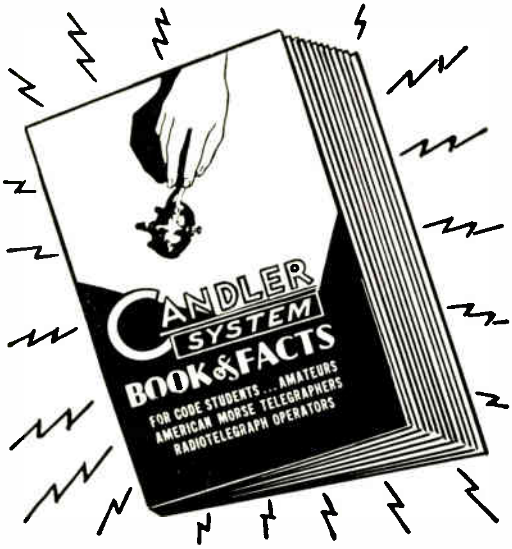
THIS VALUABLE BOOK

ALL ABOUT CODE FREE



HERE'S WHAT WORLD CHAMPION TED McELROY SAYS ABOUT THE FAMOUS CANDLER SYSTEM

"My skill and speed are the result of the exclusive, scientific training Walter Candler gave me. Practice is necessary, but without proper training to develop Concentration, Co-ordination and a keen Perceptive Sense, practice is of little value. One is likely to practice the wrong way."

McElroy is the Official Champion Radio Operator, 75.2 W.P.M. at Asheville Code Tournament.

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Here's your chance to learn code quickly and easily . . . to get maximum speed and proficiency in code . . . right in your own home. The 42-YEAR RECORD of the CANDLER SYSTEM in training beginners to develop and apply their talents to top speed and efficiency is sound proof of what the CANDLER SYSTEM *offers you.*

LEARN CODE THE QUICK, EASY WAY

Think of the thrills and excitement you can have communicating by code with people all over the world . . . when you become a good Amateur Operator. Or, you may have your sights set on becoming a highly-paid, expert Commercial Operator. In either case, you can *save time and money* learning the fundamentals of code sending and receiving and the principles of fast, efficient operating with the world-famous CANDLER SYSTEM.

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Are you hoping for a nice promotion . . . a bigger salary . . . as an expert Operator? Then, now is the time to improve your speed and skill in code sending and receiving. The CANDLER SYSTEM OF CHAMPIONS shows you how to improve your code technique — increase your speed — develop your skill to the highest point — eliminate nervous tension — and pave the way to a top-bracket position as an expert Operator.

INCREASE SPEED and SKILL THE *CANDLER* WAY

Besides experience, you must have special training to develop the extra skill and speed you need to be a successful Commercial Operator. The CANDLER SYSTEM offers you that training in an easy, thorough and interesting way. It is the outstanding method that has developed thousands of first-class operators — even Champions — from ordinary operators. There is a course to meet *your* requirements, and *your* budget, in the CANDLER SYSTEM.

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Gentlemen: Yes . . . I'm interested in the **CANDLER Code Training System.** Please rush my FREE copy of the **CANDLER Book of Facts** today.

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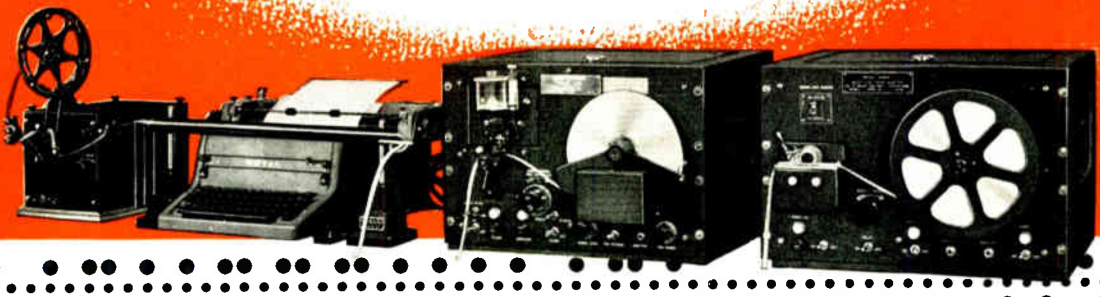
McELROY

THE WORLD'S LARGEST MANUFACTURER OF MORSE TELECOMMUNICATION EQUIPMENT



We are presenting in the following pages a complete system of high-speed automatic Morse telegraph and radio telegraph operation. This system is capable of operating under conditions that would make impossible the use of ordinary five-unit printing telegraph systems. We are quite sure that the many thousands of friends we have made during the past 25 years will be pleased to know of our licensing agreement with the well-known Creed & Company, Ltd. of London, England, which makes possible the production here in Littleton, Massachusetts of much of the equipment described in the following pages. As one communications man to the many in the field who will see these pages, be assured that McElroy equipment is built for a minimum of 20 years of uninterrupted service. That is the way we build all of our equipment. We welcome your inquiries and shall be pleased to send complete technical manuals upon request.

We call your attention to the page devoted to our latest development — Type "S" Telephone Carrier.

McElroy Manufacturing Corporation

T. R. McElroy

McELROY

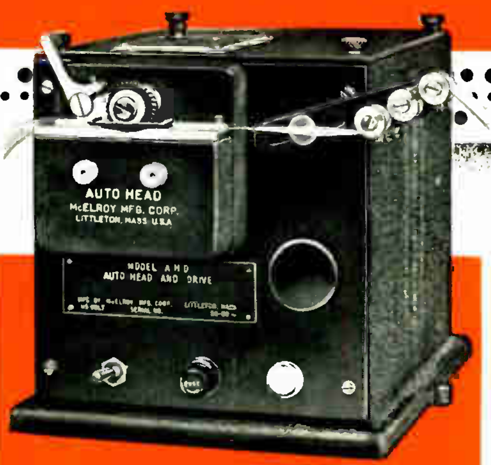
McElroy-Creed Morse Keyboard Perforator Model MC-9W

The McElroy-Creed Morse Keyboard Perforator MC-9W is the heart of this completely automatic high-speed Morse system. This system is sometimes called the Wheatstone code system. The Wheatstone code perforator shown here is the answer to the hope we have had and many other communications people have had (especially the United States Government departments) for a Wheatstone code tape perforator that could be depended upon to do a job without the necessity for having a complete machine shop and highly skilled mechanical engineers to adjust the perforator and keep it in operation. The McElroy-Creed MC-9W gives you dependable operation at all times.

I remember as an operator for R.C.A., we had one of these in our Boston office. It required practically no maintenance whatever during the more than two years that I watched the thing work. The unit is complete as shown . . . no silencing cover is required; no special power supply necessary. We are building these units here in Littleton and can make shipment out of stock.

McElROY

Keyer ADK



Keyer AHD

Using the same high-speed keying head and drive as the McElroy ADK, our AHD functions as a single-pole, double-throw switching mechanism in producing unpolarized "mark" and "space" pulses as indicated on the Wheatstone perforated tape. As with the ADK, this unit offers a selection of speed ranges up to 500 words per minute.

This is the automatic code transmitter that has come within the past few years to be recognized throughout the world as the most dependable Morse transmitter that has ever been built. The McElroy ADK (Autohead-Drive-Keyer) accepts perforated tape from either the keyboard perforator or our three-button perforator and provides a keying voltage and/or high-speed relay keying at any desired speed from 10 to 500 words per minute. Our ADK features a built-in keyed tone for monitoring. These are delivered "off the shelf".

McELROY

Recorder RAPC

At the receiving end, the incoming Morse dots and dashes are fed into the McElroy RAPC (Recorder-Amplifier-Puller-Combination). This quality ink recorder provides maintenance-free, dependable reception under conditions where not even a good operator could get the signal through the hash.

The normal speed of this Morse high-speed system is about 150 words per minute. At this speed none of the equipment is overworked, and this . . . at a speed on circuits where no one would dream of attempting to use ordinary printing telegraph systems. The inked slip may be either transcribed by operators or may be used for monitoring the circuit. This cannot be done with five unit systems.

The RAPC Ink Recorder, as produced on certain Government contracts, is equipped with an automatic tape feed mechanism, which starts the tape puller instantaneously with the first incoming dots and dashes, and stops the tape puller at any interruption or stoppage of code.

Recorder RPC

The RPC Recorder is designed for those installations where the ink recording and tape pulling mechanisms can be remotely located from the amplifier. With its recording and pulling apparatus identical to the McElroy RAPC Recorder, the RPC offers dependable high-speed recording. This unit may be connected to the amplifier section of the RAPC whose sharply peaked output is of sufficient magnitude to drive up to four ink recorders reliably.

McELROY

McElroy-Creed
Morse Reperforator
Model MC-7W

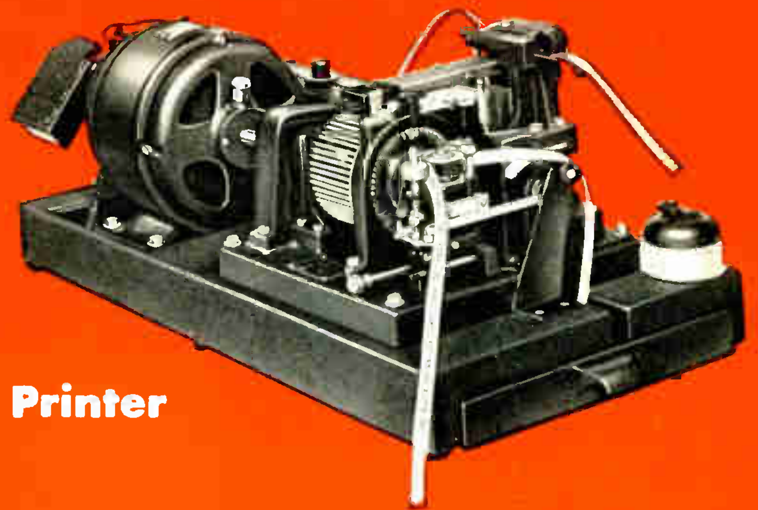
This is the McElroy-Creed Morse Reperforator MC-7W. Incoming Morse dots and dashes are fed into this reperforator at speeds up to 150 words per minute with the MC-7W producing perforated slip identical to that produced on the MC-9W perforator. This is simultaneous with the ink recording which makes it so easy to monitor the circuit. We will be tooled-up and producing these units early in 1954 at which time delivery can be made from stock.

McElroy-Creed Morse Relay
Model MC-27

The McElroy-Creed high-speed Relay MC-27 (shown with cover removed) accepts Morse characters from any standard communications receiver and delivers these code signals to the Morse Reperforator.

McELROY

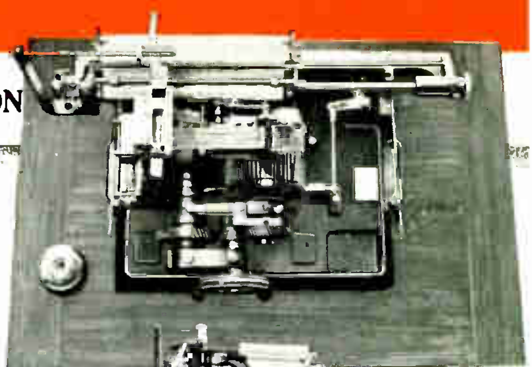
McELROY



McElroy-Creed
Morse Tape Printer
Model MC-1T

We believe this equipment to be the most dependable unit of printing telegraph apparatus we have ever seen in a lifetime of communications experience. Everyone in the Communications field understands only too well the high cost of maintenance and replacement parts on ordinary printing telegraph systems. The MC-1T Morse Tape Printer is so dependable that we old-timers always refer to the unit as the "iron-horse". They just do not fail! The perforated slip from the Morse Reperforator is fed into the Morse Tape Printer at speeds up to 100 words per minute. Upon receipt of this coded slip, the MC-1T Tape Printer translates these signals in the form of Roman characters on the slip. There you have a completely automatic, high-speed, dependable telegraph system with all of the advantages that Morse code possesses over other means of code communication.

McELROY MANUFACTURING CORPORATION

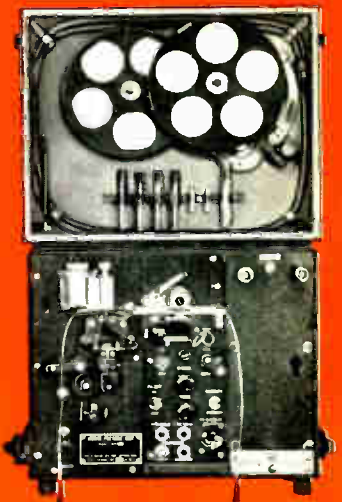


McElroy-Creed
Morse Page Printer
Model MC-1P

This unit is of the same basic design as the Morse Tape Printer shown above except that provision is made for two additional Morse signals thereby allowing for line feed and carriage return. Because the greatest value in our Morse system lies in its almost total absence of maintenance, we are somewhat partial to the Tape Printer MC-1T where a guy doesn't even need a screwdriver to insure operation year after year. Pasting up slip however, can be most time consuming, and for Central Office use, the Morse Page Printer has a very definite application.

McELROY

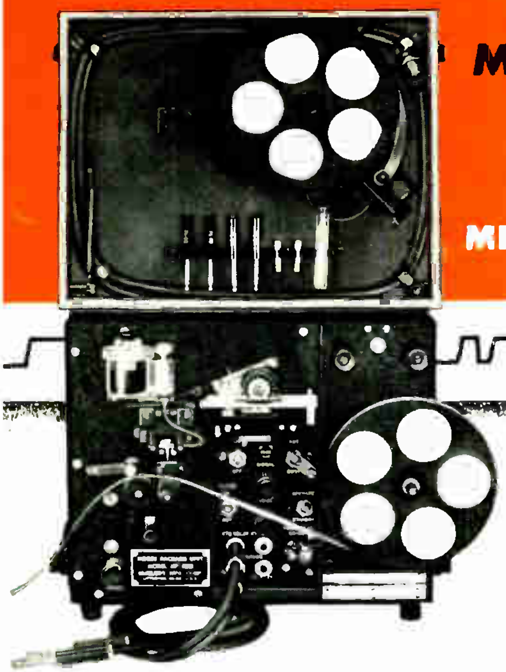
MP-1A Perforating



MP-1A Keying

Morse Package

Model MP-1A



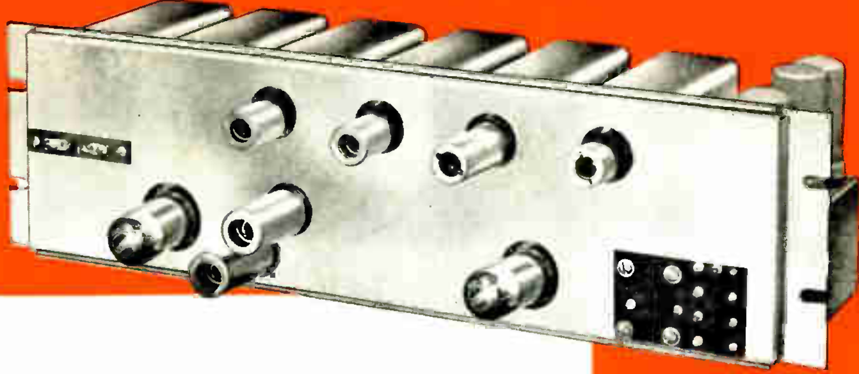
MP-1A Recording

Morse Package MP-1A, shown here, employs our standard three button perforator. Its simplicity of operation will warrant its continued production for those applications where either traffic conditions are light or personnel are not keyboard conscious. The MP-1A keying head is identical in every respect to the McElroy ADK and also provides for high-speed relay and voltage keying with built-in keying tone for monitoring. The MP-1A recorder section has the same basic trouble-free design as all our recording equipment. The tape pulling mechanism provides for a speed range of from 7 to 100 words per minute, or any other specified 7 to 1 ratio.

Under our new licensing agreement with Creed & Company, Ltd., we are now making plans to produce in the very near future, the Morse Package MP-1A unit complete with the McElroy-Creed Keyboard Perforator. The perforator section of this modified MP-1A will be of the same dependable design as the Morse Keyboard Perforator MC-9W, but with the elimination of certain frame castings, etc. This means dependable keyboard perforating, as well as keying and recording, in a package about the size and weight of a portable typewriter . . . to be sure, a completely self-contained and automatic Morse Central Office.

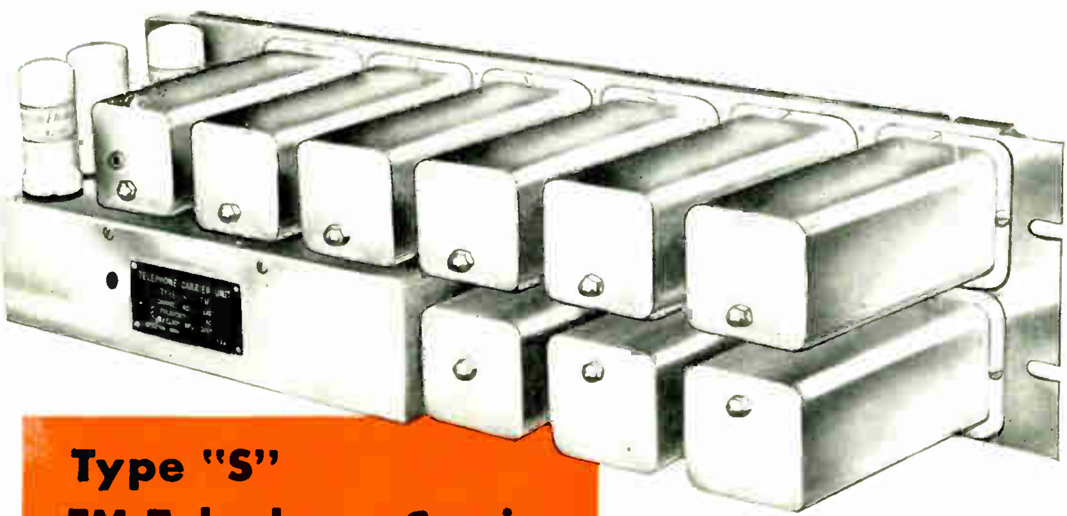
McELROY

McELROY



We are confident that our friends throughout the world will be pleased to know that the same skill and the same people who produce the world's greatest volume of telegraph communications equipment are now producing telephone carrier systems under a license from a group of independent engineers and research men who have developed what we believe to be the lowest priced telephone carrier that has ever been offered.

With each channel of Type S carrier exactly replacing a pair of wires, our unusually low price will allow this unit to "prove-in" under the cost of laying additional wire on distances as low as five miles! Stackable to five channels, Type S provides for five additional talking circuits for toll applications or will bring-in, with full selective ringing up to 50 additional subscribers. The first open-wire telephone carrier to make full use of the FM (Frequency Modulation) principal, Type S carrier provides for toll quality transmission regardless of weather or line variations.



Type "S" FM Telephone Carrier

McELROY *Manufacturing Corporation*

LITTLETON, MASSACHUSETTS, U.S.A. ★ Cables: TEDMAC, Boston, U.S.A. ★ Telephone: Boston Liberty 2-6960

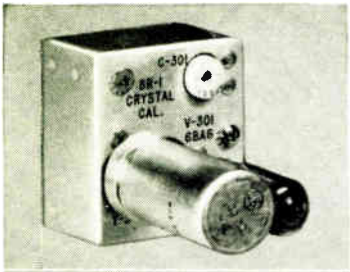
To My Friends in Communications:

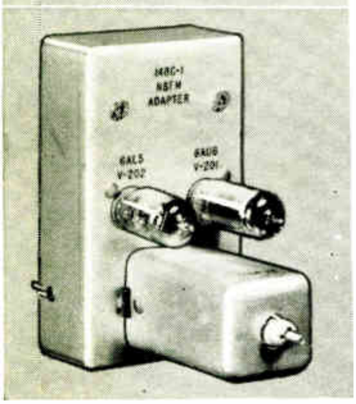
I have tried my best to describe our equipment and tell you what it will do and continue to talk like a telegraph operator rather than a huckster. I wish I could show you some of the beautiful work these excellent facilities of ours have produced and continue to produce in radar cavity work. You just don't show pictures of that kind of stuff, however.

May I emphasize that the greatest thrill I receive out of the mail we receive is not the checks, which are the necessary life blood of the place, but the friendly letters from people all over the world who have bought and have used our equipment. Such encouragement and friendly letters have contributed in no small way to make possible our growth from the days 25 years ago when I made telegraph keys and code practice oscillators in the basement of my home up to now when we are the world's largest producers of Morse equipment.

In conclusion, may I say that this place of ours is no longer a one man outfit. We have a very competent staff of engineering personnel, skilled tool and die makers, first-class machinists and close to 100 girls in our assembly shop, each a Rembrandt with the soldering iron. We will be here with maintenance parts and more and more communications equipment long after I have passed out of the picture. So, at this point, on behalf of about 300 people, I express sincere thanks to you who have found time to read about our equipment.

T. R. McElroy





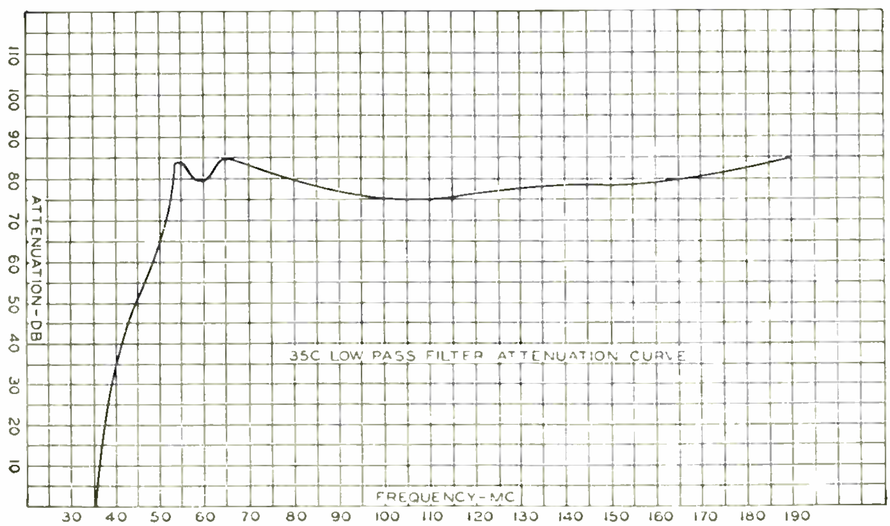
THE Collins 8R-1 CRYSTAL CALIBRATOR

Plugs into a completely wired socket on the 75A-2 or 75A-3 chassis. A set of contacts on the noise limiter-calibrate switch turn the 8R-1 crystal calibrator on and off. When harmonics of the 100 kc crystal have been checked against WWV, an accurate crystal check-point is available every 100 kc throughout the receiver's range. Use of the 8R-1 crystal calibrator in conjunction with the receiver's accurately calibrated linear dial permits interpolation to a fraction of a kilocycle. *Net domestic price $25.00.*

THE Collins 148C-1 NBFM ADAPTER

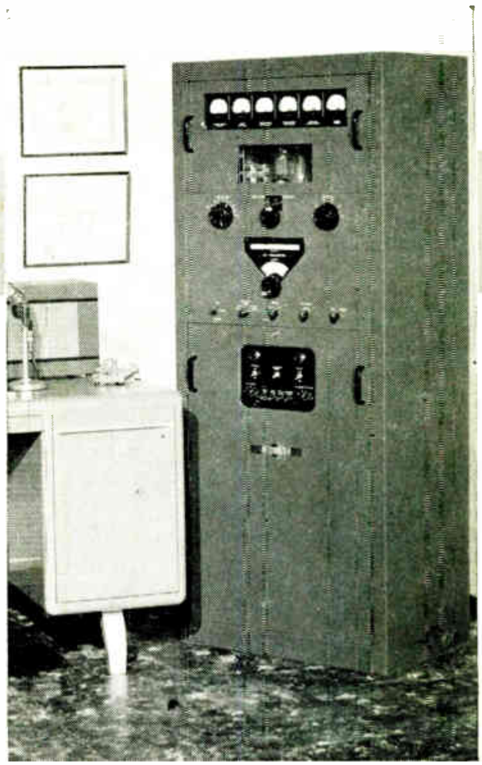
Plugs into a completely wired socket on the 75A-2 or 75A-3 chassis and is controlled by the CW-AM-FM switch on the front panel. Use of the 148C-1 adapter gives true discriminator reception of NBFM signals. *Net domestic price $22.50.*

35C-2 LOW-PASS RF FILTER Although designed as an accessory for the 32V-3 and as standard equipment in the KW-1, coaxial fittings make the 35C-2 readily adaptable to any amateur transmitter having 52 ohm output. It has an insertion loss of only 0.25 db at frequencies below 30 mc, but its three individually shielded filter sections provide about 75 db attenuation at television frequencies. *Net domestic price $40.00.*



COLLINS

kw-1 for maximum OPERATING CONVENIENCE AND POWER



The transmitter and all power supplies are contained in one attractive cabinet.

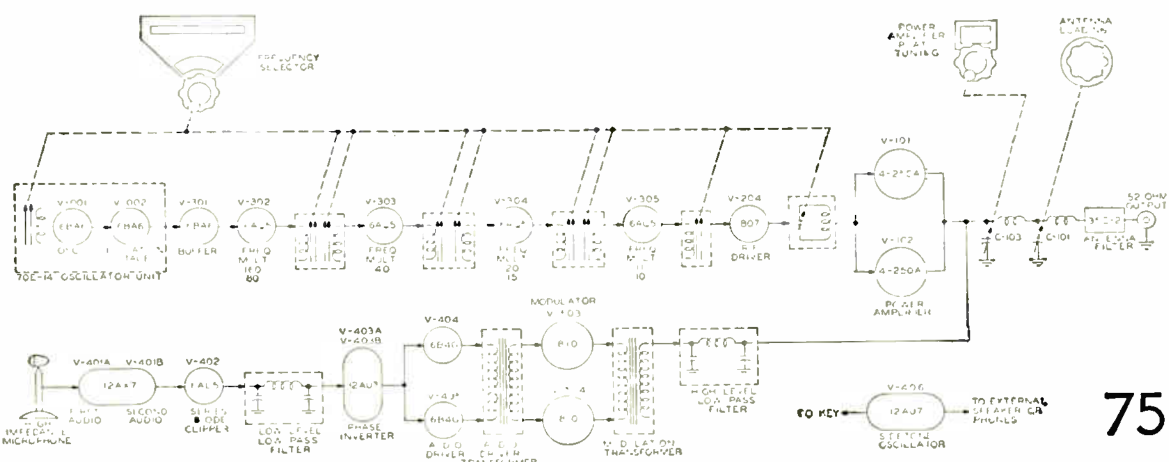
The Collins KW-1 runs a cool 1000 watts but tunes as easily as a 32V-3. It operates on phone or CW in the 160, 80, 40, 20, 15, 11, and 10 meter bands. A single control on the front panel bandswitches the entire transmitter. All RF tuning controls except the final amplifier are ganged. To tune the KW-1, just pick your band, set the built-in VFO to the desired spot, and adjust the final amplifier tuning and loading. That's all there is to it!

The KW-1 design effectively reduces spurious radiation to a very low value, particularly at television frequencies. A metal box inside the main cabinet completely shields the RF section. A large number of tuned circuits at the operating frequency reduce harmonic output from the exciter. The final amplifier pi-L network, developed by Collins, provides an effective TVI filter in the RF output circuit. To this is added the 75 db television-frequency attenuation of the built-in 35C-2 low pass filter. The final amplifier is tuned by a variable vacuum capacitor which provides a very short, low impedance path from the final amplifier plate circuit to ground at television frequencies.

The KW-1's reputation for getting through the QRM is due not only to its high power but also to its audio clipper followed by low-level and high-level low-pass filters which permit a high average percentage of modulation without splatter.

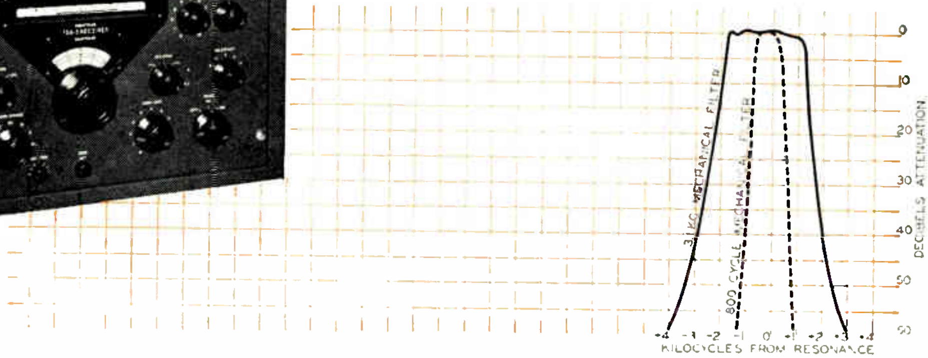
KW-1 FEATURES:

1000 watts input on phone or CW (500 watts on 160); covers 160, 80, 40, 20, 15, 11, and 10 meter bands; 52 ohm coaxial RF output; frequency control by means of the highly stable Collins 70E-14 oscillator; high impedance input for crystal or dynamic microphone; overload relay, fuses and high voltage arc gaps for circuit protection; operates from 115/230 volts 50/60 cycle single phase grounded neutral; completely self-contained in attractive metal cabinet 66½" x 28" x 18"; controls include bandswitching, frequency selector, final amplifier tuning, final amplifier loading, filament switch, filament voltage adjustment, plate switch, overload reset switch, overload relay adjustment, send-standby-calibrate switch, emission selector switch, tune-operate switch, meter switch, power amplifier excitation control, modulator bias control, audio driver bias control, clipping level, audio gain control, dial zero set.
Net domestic price *$3,850.00*



COLLINS

75A-3 RECEIVER with MECHANICAL FILTER



As can be seen in the block diagram, the Collins 75A-3 double conversion superheterodyne receiver, with its crystal-controlled front-end and highly stable low frequency VFO, is like a high frequency crystal-controlled converter working into a very stable low frequency receiver. The high stability and 3.1 kc bandwidth of the 75A-3 make it ideal for AM or single sideband — and an 800 cycle mechanical filter is available as an optional accessory for CW.

All coils are permeability tuned and have a straight-line frequency characteristic allowing linear dial calibration. Only the band in use is visible on the slide rule dial. On the vernier dial each division represents one kc except on the 10 and 11 meter bands, where each dial division represents two kc. This accurate calibration is made possible by the highly stable oscillators in the 75A-3.

The 75A-3 covers the 160, 80, 40, 20, 15, 11, and 10 meter amateur bands. Sensitivity on all bands is 2.5 mv or better for a 10 db signal-to-noise ratio. Image rejection is at least 50 db. AVC is applied to RF as well as IF stages. Separate noise limiters are included for phone and CW. The S-meter is calibrated from 1 to 9 in steps of approximately 6 db, and for 20, 40 and 60 db over S9. S9 corresponds to a signal input of 100 microvolts. Antenna input impedance is 50 to 150 ohms, balanced or unbalanced. A phone jack and 4 ohm and 500 ohm audio output terminals are provided. Sockets and front-panel controls are included for the 8R-1 100 kc crystal calibrator and 148C-1 NBFM adapter which are available as optional accessories. The following controls are on the 75A-3 front panel: tuning, zero set, bandswitch, RF gain, audio gain, BFO pitch, CW limiter, antenna trimmer, crystal selectivity, crystal phasing, mechanical filter selector, CW-AM-FM switch, noise limiter-calibrate switch, on-off-standby switch. The 75A-3 operates from a 115 volt 50/60 cycle ac power source. Cabinet dimensions are: 21-1/8" wide, 12-1/2" high, and 13-1/16" deep. The 19" panel fits a standard relay rack. The 75A-3 weighs approximately 50 pounds.

Net domestic prices:

75A-3 receiver complete with 3 kc mechanical filter: $530.00

10-inch speaker in matching cabinet: $20.00

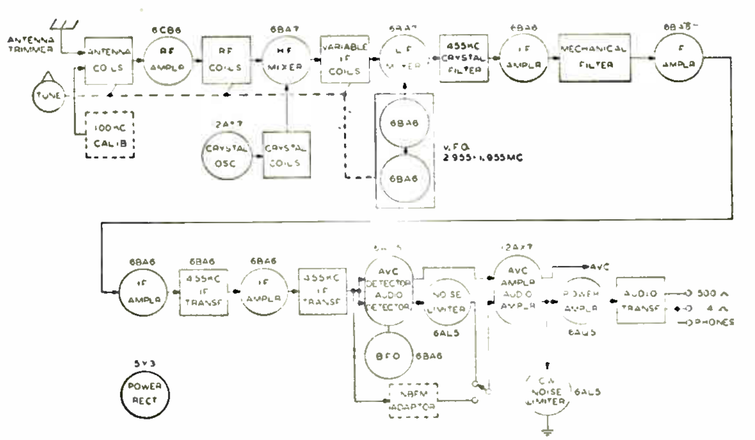
8R-1 crystal calibrator: $25.00

148C-1 NBFM adapter: $22.50

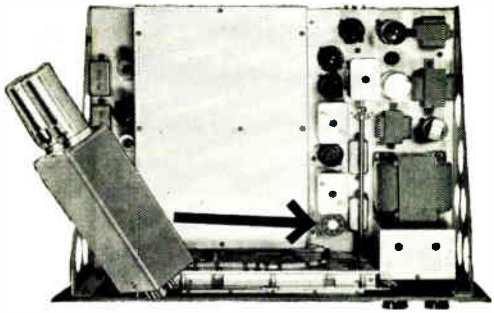
F455B-08...800 cycle mechanical filter: $55.00

F455B-31...3.1 kc mechanical filter: $55.00

F455B-60...6.0 kc mechanical filter: $55.00



MECHANICAL FILTER ADAPTERS FOR Collins 75A-1

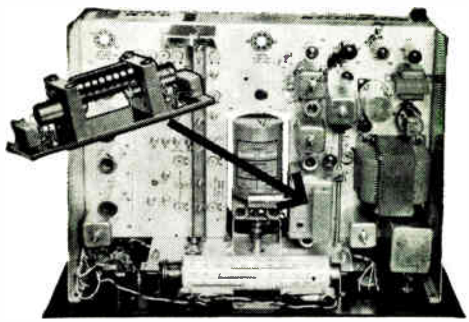


With the mechanical filter adapter shown here it takes only a few seconds to convert your 75A-1 receiver to include the Collins mechanical filter. Just remove the first 500 kc IF tube and plug in the adapter. That's all there is to it. Either a 1400 cycle CW filter or a 3.1 kc phone filter is permanently installed in each adapter.

Type 353C-14 plug-in adapter, complete with 1400 cycle filter, for 75A-1: $75.00

Type 353C-31 plug-in adapter, complete with 3.1 kc filter for 75A-1: $75.00.

MECHANICAL FILTER CONVERSION KIT FOR Collins 75A-2



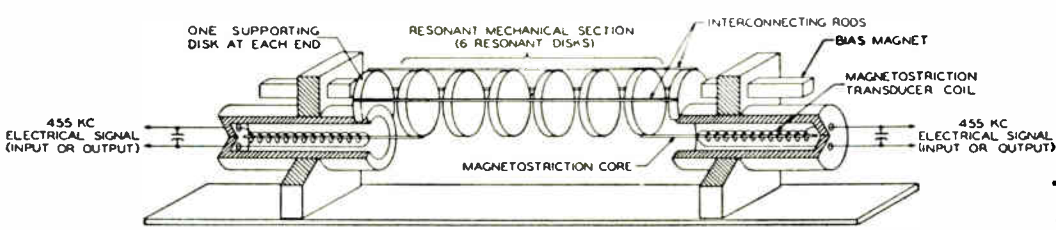
The 75A-2 mechanical filter conversion kit will enable you to convert your 75A-2 to include two mechanical filters. A front-panel switch selects the desired filter. A F455B-31 3.1 kc phone filter is included with each kit and a F455B-08 800 cycle CW filter may be added at any time. You can install the complete kit in one evening or, if you prefer, your distributor will make arrangements for you to return your 75A-2 to Collins. The factory will make minor repairs and completely realign your 75A-2 in addition to installing the conversion kit.

75A-2 conversion kit complete with F455B-31 3.1 kc mechanical filter $80.00.

Factory installation of 75A-2 conversion kit including F455B-31 mechanical filter, minor repairs, and complete realignment of the 75A-2; *F.O.B. Cedar Rapids $105.00.*

F455B-08 800 cycle mechanical filter (plug-in) for 75A-3's and modified 75A-2's $55.00.
F455A-31 3.1 kc mechanical filter (solder terminals) $55.00.
F455A-08 800 cycle mechanical filter (solder terminals) $55.00.

The mechanical filter is a resonant mechanical device in the receiver's i.f. strip. Unlike the crystal filter, the mechanical filter remains in the circuit at all times. As shown here, it consists of three general sections: an input transducer, a mechanically resonant section consisting of a number of metal disks, and an output transducer. A 455 kc. electrical signal applied to the input terminals is converted into a 455 kc. mechanical vibration at the input transducer by means of magnetostriction. This mechanical vibration travels through the resonant mechanical section to the output transducer, and is converted, by magnetostriction, to a 455 kc. electrical signal which appears at the output terminals. There is no mechanical motion except for the imperceptible vibration of the metal disks. The mechanical filter requires no adjustment.

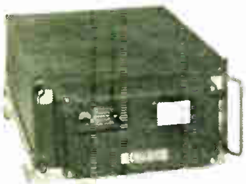


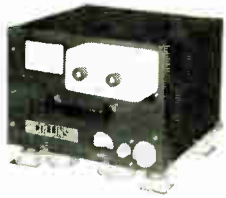
OTHER COLLINS PRODUCTS

AIRCRAFT


Course Indicator


Approach Horizon


618S 144-Channel HF Transceiver

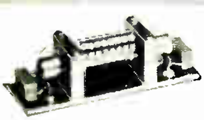

180L Automatic Antenna Tuner


51V Glideslope Receiver


51X 360-Channel VHF Receiver

17M 360-Channel VHF Transmitter


INDUSTRIAL COMPONENTS

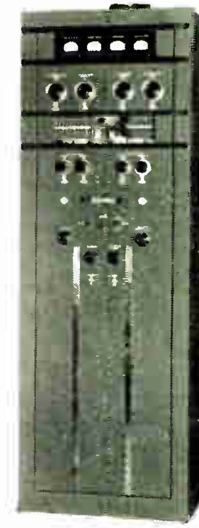

Mechanical Filter


Oscillator

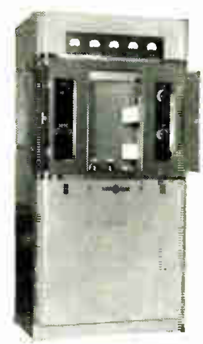
Autotune


COMMUNICATIONS AND BROADCAST

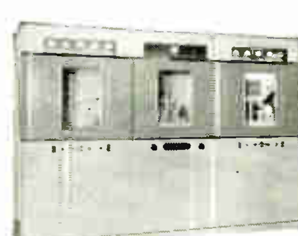

Collins 430 Series Communications Transmitter


Collins 30K-5 2-Channel Communications Transmitter


51J Communications Receiver


20V Broadcast Transmitter


300J 250-Watt Broadcast Transmitter


21E/21M 5,000 Watt 10,000 Watt Broadcast Transmitters

In requesting information on any Collins product, please specify equipment type number.

For Engineering Excellence in Radio Equipment, it's . . .

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2700 W. Olive Avenue
BURBANK

1930 Hi-Line Drive
DALLAS 2

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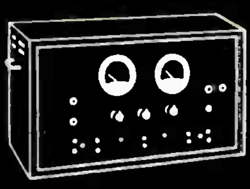
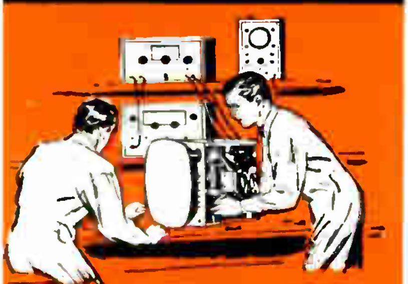
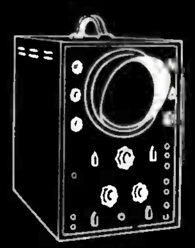


POWER SUPPLY



LABORATORY



OSCILLOSCOPE

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ST-2A — General purpose use.

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● For full information call your nearest G-E Test Equipment Distributor or write: *General Electric Company, Section 564, Electronics Park, Syracuse, New York.*

GENERAL ELECTRIC

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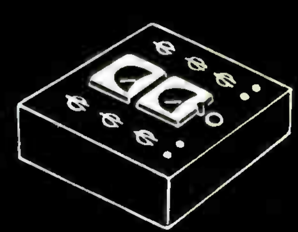

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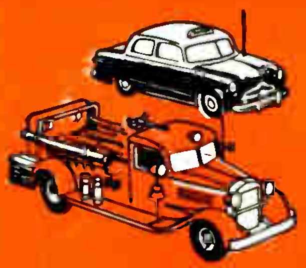
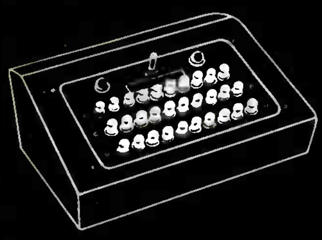


POLICE & FIRE DEPTS.



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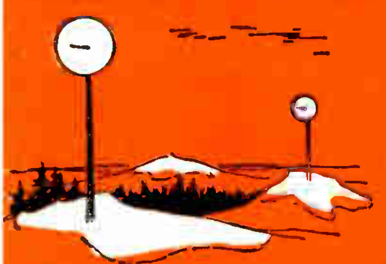


24 CHANNEL COMMUNICATION

G-E Communications Equipment Covers the Range

50 kc to 2,000,000 kc • 1 watt to 3,000 watts

● G-E offers a complete line of communications equipment—from audio to microwave—for police, fire, oil, lumber, industrial and civil defense applications. Typical are:

Selective Tone Signaling Equipment—Duplex dispatching combinations, single tone, two-way, and group calling equipment. Provides up to 900 private lines on one radio channel!

Microwave—G-E microwave equipment offers dependable communication over long distances and in difficult terrain areas. Up to 24 channels available for heavy traffic use.

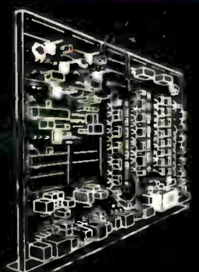
Industrial 2-Way Radio Communication—G-E 2-way radio steps up production—increases profits. With it, equipment can handle more material—personnel do a better job on co-ordinated activities.

G-E EQUIPMENT FEATURES:

- *Narrow* or *wide* band operation
- Low battery drain—cooler running equipment
- Superior design, minimum tubes for efficient operation
- Quality components—G-E makes more of its 2-way radio components than any other manufacturer

For full information on G-E communications equipment call the General Electric office near you or write: *General Electric Company, Section 564, Electronics Park, Syracuse, New York.*

GENERAL ELECTRIC



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SIGNAL SLICER
G-E Ham News, July-Aug., 1951

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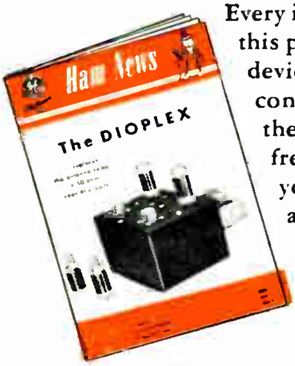
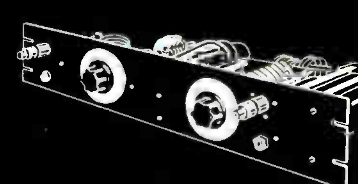


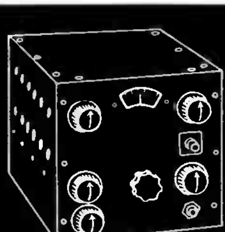


POWER PEAKER
G-E Ham News, Sept.-Oct., 1952

TVR HIGH-PASS FILTER
G-E Ham News, Mar.-Apr., 1951

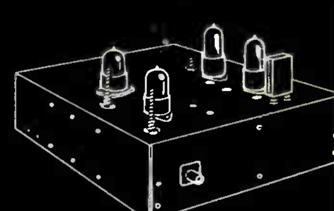
GENERAL ELECTRIC

166-IC1

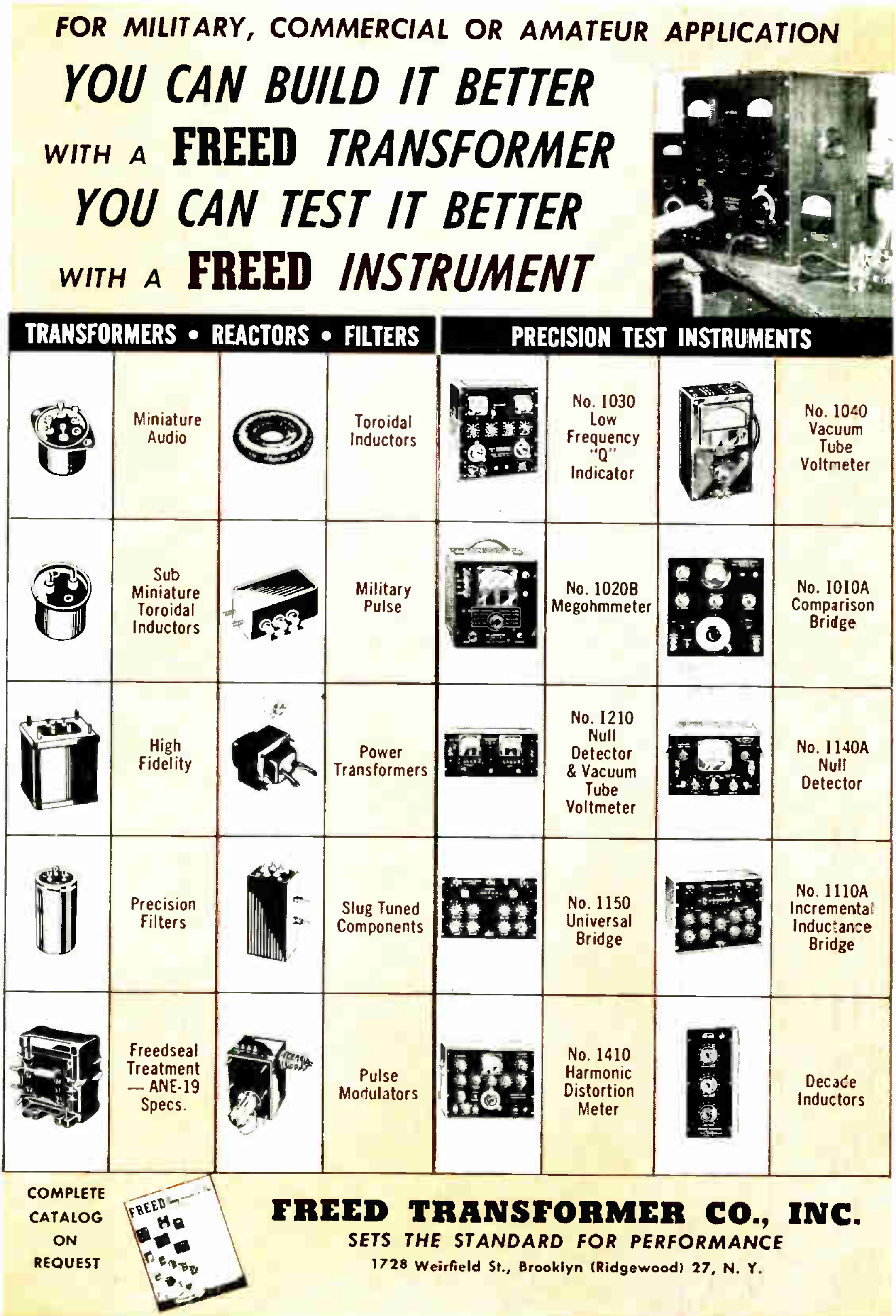


MOBILE-MARINE RECEIVER
G-E Ham News, July-Aug., 1952

S/N-6
CASCODE 2-METER PRE-AMPLIFIER
G-E Ham News, Nov.-Dec., 1952



SUPER 430 CONVERTER
G-E Ham News, July-Aug., 1953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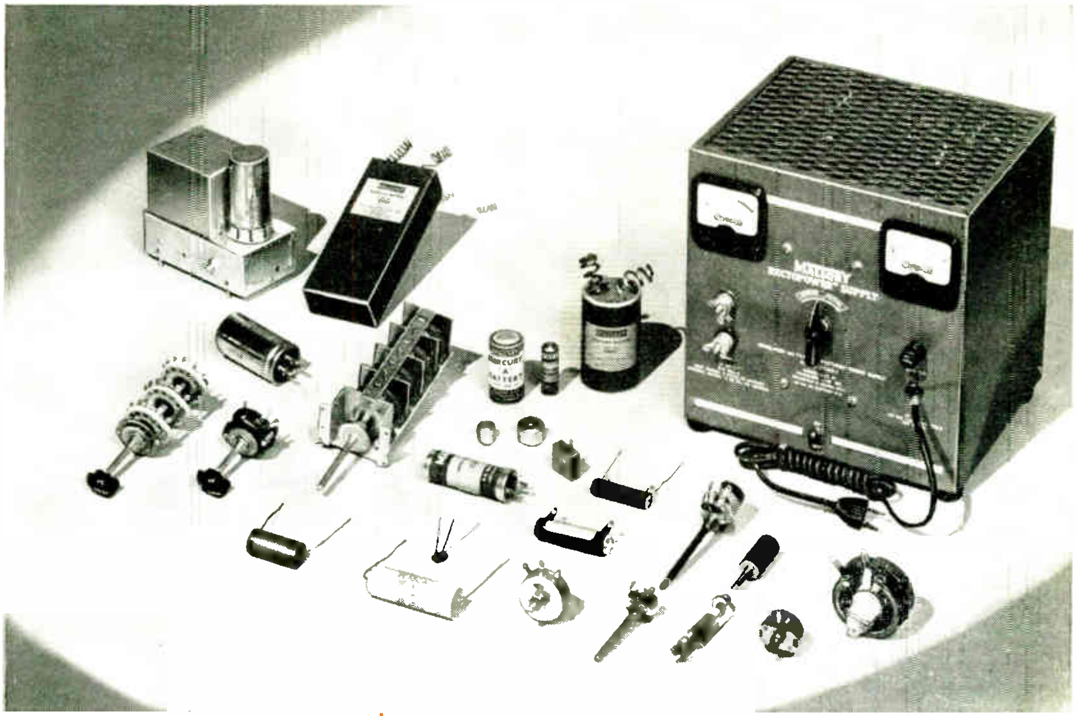
SETS THE STANDARD FOR PERFORMANCE
1728 Weirfield St., Brooklyn (Ridgewood) 27, N. Y.



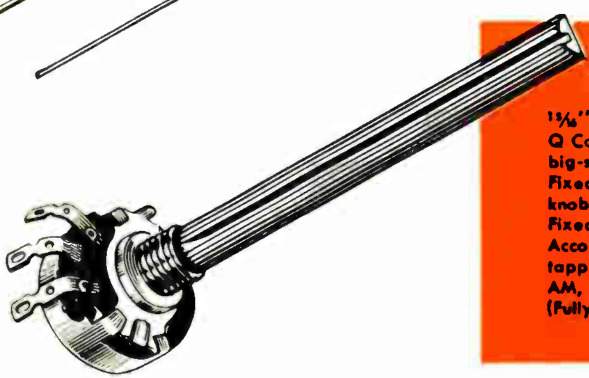
WHEREVER THE CIRCUIT SAYS -\/\/\/-

ADVANCED TYPE BT RESISTORS

Type BT Insulated Composition Resistors—meet JAN-R-11 Specifications at ⅓, ½, 1 and 2 watts. Small size BTB specially designed for miniature 2 watt requirements. Type BT's are suited to television and similar exacting circuits. Extremely low operating temperature. Excellent power dissipation. 10 ohms to 22 megohms in RMA ranges. (Fully described in Catalog RDC8.)

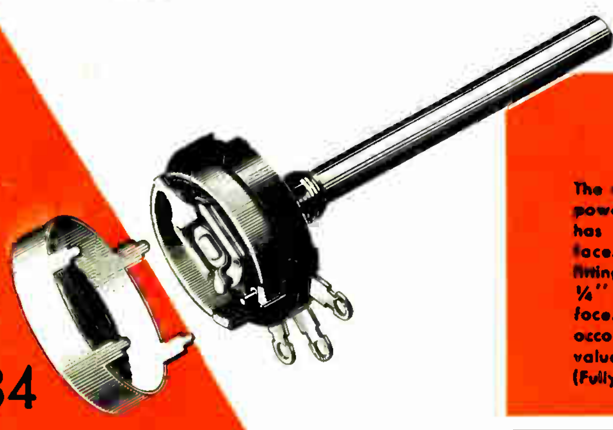
BW INSULATED WIRE WOUND RESISTORS

Exceptionally stable, inexpensive low wattage wire wound resistors. ½, 1 and 2 watts—0.24 ohms to 8,200 ohms in RMA ranges. 50% to 100% overloads can be applied with negligible change, and return to initial value.
(Fully described in Catalog RDC8.)



TYPE Q VOLUME CONTROLS

1⁵⁄₁₆" diameter and ¼" long bushing suit the Q Control to the smallest chassis, yet it handles big-set requirements with ease. Knob Master Fixed Shaft is standard and fits most push-on knobs without alteration. 15 Interchangeable Fixed Shafts provide adaptability to "specials". Accommodates Type 76 Switch. 86 plain and tapped Q Controls give wide coverage of AM, FM and TV needs.
(Fully described in Catalog RDC1.)



2-WATT WIRE WOUND CONTROLS

The most dependable wire wound controls for power requirements up to 2 watts. Type W has ¼" full round shaft 3" long from control face. Type WK has Knob Master Shaft for fitting to knurled and flatted knobs, bushing is ¼" long, and shaft 3" long from mounting face. Both types have 1¼" diameter and accommodate Type W Switches. Resistance values: 2 ohms to 10,000 ohms.
(Fully described in Catalog RDC1.)

POWER WIRE WOUND RESISTORS

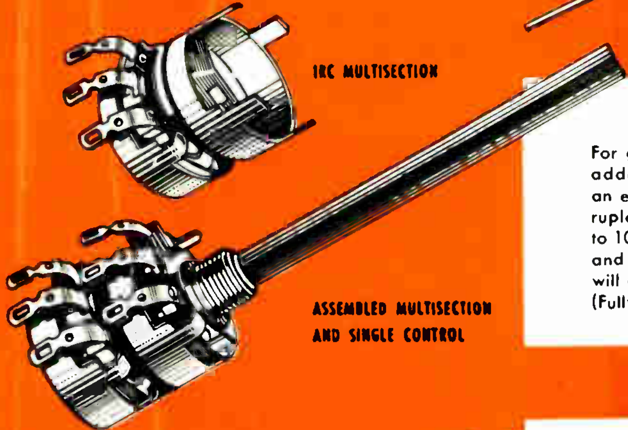
IRC Fixed and Adjustable Power Wire Wounds are rugged resistors engineered for heavy-duty service. They are supplied in a wide variety of power ratings, resistance values, sizes and terminal types.

TUBULAR POWER WIRE WOUNDS—fixed and adjustable, 10 to 200 watts, require no derating in high ranges. 10 and 20 watt fixed types have combination lead and lug terminals.

TYPE PW-4—Four watt power resistors with axial leads. Completely insulated in high temperature plastic. Body dimensions 1¾'' long by ²¹⁄₆₄'' diameter 1 ohm to 8200 ohms.

TYPES PW-7 and PW-10—Seven and ten watt high temperature resistors of practical rectangular design with axial leads.

(Fully described in Catalog RDC5.)



IRC MULTISECTION

ASSEMBLED MULTISECTION AND SINGLE CONTROL

Multisections

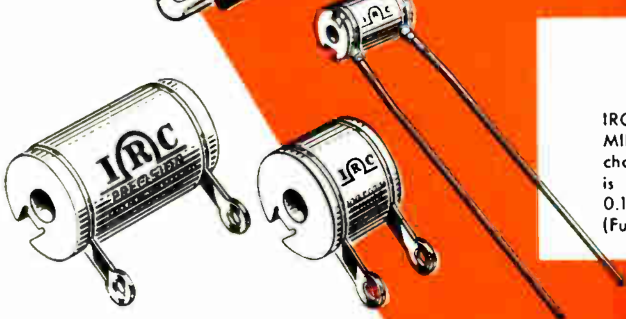
For ganged controls, IRC MULTISECTIONS are added to Q Controls like switches to provide an endless variety of duals, triples and quadruples. Available in 20 values from 500 ohms to 10 megohms. MULTISECTIONS are as easily and quickly attached as switches—and duals will accommodate Type 76 switches.

(Fully described in Catalog RDC1.)

FLAT INSULATED WIRE WOUND RESISTORS

Unsurpassed for adaptability to an extremely wide variety of design requirements. Radical design features impervious phenolic compound casing, special metal mounting bracket that actually speeds transfer of heat from inside chassis. Space-saving MW's afford unusual flexibility in providing taps for voltage dividing applications.

(Fully described in Catalog RB-2.)



MIL TYPE PRECISION WIRE WOUND RESISTORS

IRC's improved Precision Wire Wounds surpass MIL-R-93A Specifications and are fully interchangeable with earlier types. ±1% tolerance is standard. 7 Sizes. Resistance values from 0.1 ohm to 6 megohms.

(Fully described in Catalog DC6.)

Other Products in IRC's complete resistor line are described on the following pages.

INTERNATIONAL RESISTANCE COMPANY

401 N. Broad Street Philadelphia 8, Pa.

In Canada: International Resistance Co., Ltd., Toronto, Licensee

IRC

WHEREVER THE CIRCUIT SAYS —\/\/\/—

CLOSE TOLERANCE DEPOSITED CARBON PRECISTORS

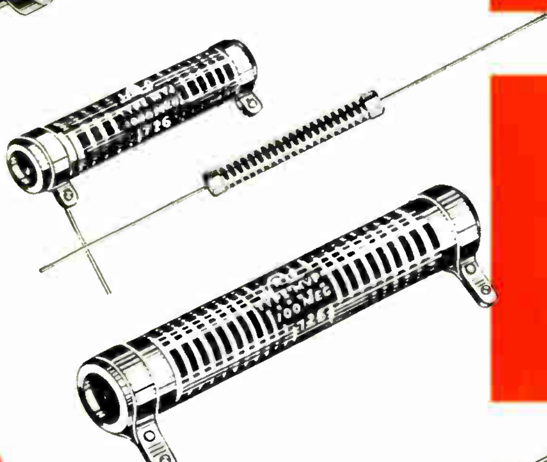
PRECISTORS offer a unique combination of close tolerance, stability and economy. Pure crystalline carbon bonded to selected ceramic cores overcomes limitations of carbon composition resistors and higher cost of precision wire wounds. PRECISTORS offer wide range of values, guaranteed accuracy, high stability, low voltage coefficient, excellent frequency characteristics, predictable temperature coefficient.

(Fully described in Catalog RDC-2.)

HIGH FREQUENCY RESISTORS

Type MP Resistors are designed for frequencies above those of conventional resistors. 2 watts to 90 watts. Special construction, with resistance film bonded to steatite ceramic form, provides stable resistors of low inductance and capacity. Type MPM's are miniature ½ watt units for small-space, high frequency receiver applications.

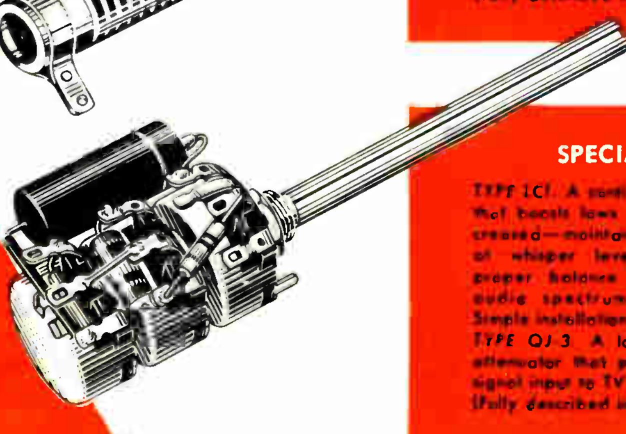
(Fully described in Catalog RF-1.)



HIGH VOLTAGE RESISTORS

Type MV's meet high resistance and power requirements in high voltage applications. Resistance coating in helical turns on ceramic tube provides a conducting path of long effective length. 2 watts to 90 watts. Variety of terminal types. Type MVX's meet requirements for small, high range unit with axial leads. 3" x ½" construction identical with Type MV's, except for terminal.

(Fully described in Catalogs RG-1 and RG-2.)



SPECIAL CONTROLS

TYPE 1CI. A continuously compensated control that boosts lows and highs as volume is decreased—maintains depth and brilliance even at whisper level. Automatically maintains proper balance of all frequencies in the audio spectrum at any listening level. Simple installation.

TYPE QJ 3. A low cost, easily installed TV attenuator that permits ready adjustment of signal input to TV sets.

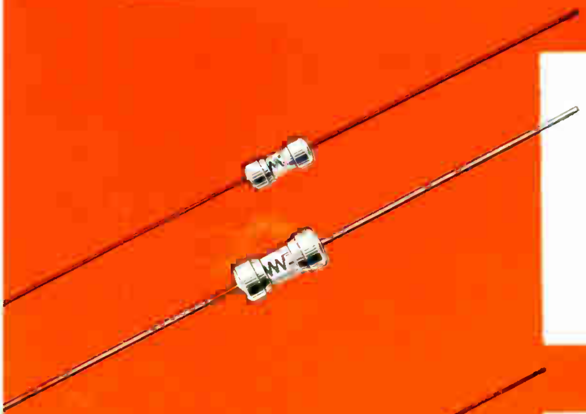
(Fully described in Catalog DC1.)

◁ Other products in IRC's complete resistor line are described on the preceding pages.

SEALED VOLTMETER MULTIPLIERS

Dependable multipliers for use under the most severe humidity conditions, Type MF Resistors consist of a number of IRC Precisions interconnected and hermetically sealed in a glazed ceramic tube. Compact, rugged, stable, fully moisture-proof and easy to install. Maximum current: 1.0 M.A.; 0.5 megohms to 6 megohms.

(Fully described in Catalog RD-2.)

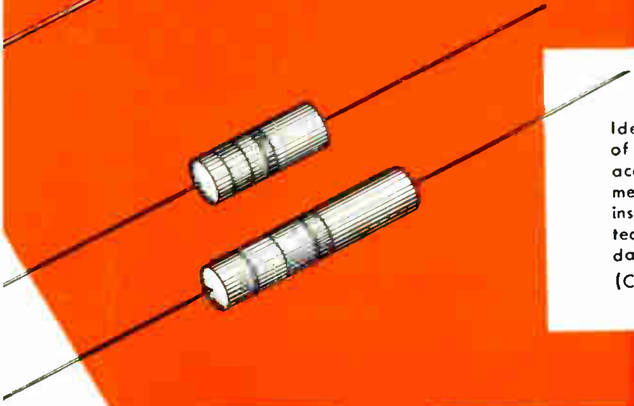


MICROSTAK SELENIUM DIODES

TYPE GA Diodes are IRC engineered for use in low current circuits where very high back resistance and low forward resistance are required. They are small size, hermetically sealed, and ideal for circuit applications up to 1 megacycle.

IRC VARISTORS are non-linear resistors. They are voltage sensitive and provide sharp variation of resistance with applied voltage.

(Fully described in Catalogs RDC11 and RSR-3.)



INSULATED CHOKES

Ideal for TV and similar circuits. Wide range of size and characteristic combinations permit accurate specification to individual requirements. Types CLA and CL-1 Chokes are fully insulated in molded phenolic housings—protected from high humidity, abrasion, physical damage or shorting to chassis.

(Completely described in Catalog RDC7.)

OTHER IRC PRODUCTS

IRC manufactures a wide line of resistors, controls and related electronic components for equipment manufacturers, service technicians and amateurs.

In addition to the products described on these pages, IRC also furnishes—

BORON-CARBON PRECISTORS • MOLDED BORON-CARBON PRECISTORS • INDUSTRIAL CONTROLS • 4 WATT TV CONTROLS • FLAT TYPE POWER WIRE WOUNDS • FEED-THRU TERMINALS.

SEND FOR LITERATURE

◀ For full information on any IRC product visit your local IRC Distributor or write for the Catalog Bulletin in which you are interested.

INTERNATIONAL RESISTANCE COMPANY

401 N. Broad Street Philadelphia 8, Pa.

In Canada: International Resistance Co., Ltd., Toronto, Licensee

IRC

For those who appreciate PROFESSIONAL STANDARDS!

Communications Receivers for finest Performance

SP-600-JX

HQ-140-X

The "SP-600-JX"

The "SP-600-JX", a masterpiece of receiver design, is a 20 tube dual conversion superheterodyne covering the range of 540 Kc to 54 Mc in 6 bands. Operation on any of 6 crystal-controlled fixed frequency channels is immediately available. The power supply is an integral part of this world-famous receiver.

The "SP-600" represents today's ultimate in receiver performance. Stability is .001 to .01 percent, image rejection is 80 db to 120 db down, and spurious responses are at least 100 db down. Sensitivity is 1 microvolt CW and 2 microvolts AM. Selectivity for the 3 calibrated crystal and 3 non-crystal ranges is from 200 cycles to 13 Kc.

The "HQ-140-X"

The "HQ-140-X" was designed to give years of reliable, quality performance. Its many out-standing features are evidence that it was built for those who appreciate professional standards. Extremely accurate frequency setting is achieved because of its carefully calibrated bandspread dial. The Hammarlund patented 455 Kc crystal filter and phasing network makes possible bandwidth changes without the slightest detuning. The separate oscillator (6C4) and mixer (6BE6) contribute to the high degree of oscillator stability.

Low-loss tube sockets, ceramic bandswitches, temperature compensating capacitors, zero temperature coefficient ceramic trimmers, and a bi-metallic compensating plate, all keep frequency drift to less than 0.01%, from the lowest frequency (540 Kc) to the highest (31 Mc).

WANT TO KNOW MORE ABOUT HAMMARLUND RECEIVERS?

Write immediately to have your name placed on our Receiver mailing list.

HAMMARLUND

SIGNALING & CONTROL EQUIPMENT

for dependable operations!

Economical Systems of Modern Design

SELECTIVE CALLING OR SIGNALING

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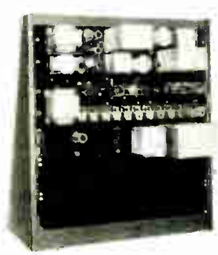
Push-Button Selective Calling



Aircraft Communications System Control

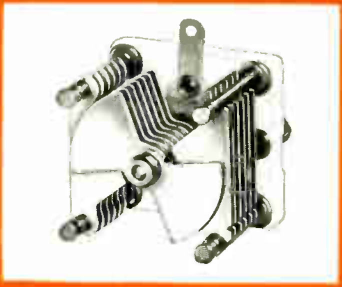
Multi-Gate for Remote Control

THE HAMMARLUND MANUFACTURING CO., INC.
460 WEST 34th STREET • NEW YORK 1, N. Y.

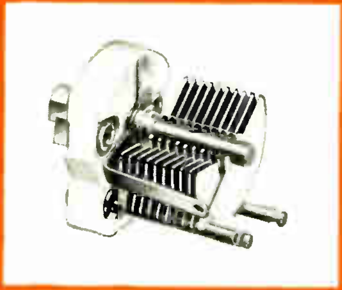
When Your Design Considerations are Critical—Put in

Reliable Component

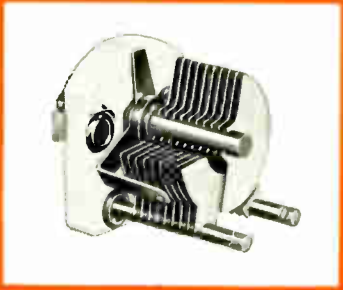
"MAC"



'BFC"



"APC"



"MAPC"

The "MAC"

The "MAC" provides the low minimum capacity essential for use as a trimmer in the VHF range. It was engineered to achieve the smallest dimensions practical to meet the requirements of a miniaturized component. Its silicone treated steatite base is only ¾" x ⅝".

Rotor and stator are soldered assemblies and are of brass, nickel plated. Capacities range from 1.4 to 19.6 mmf.

The "BFC"

The "BFC" "butterfly" type capacitor has very low minimum capacity, low inductance and isolated rotor for use in VHF applications as a series capacitor with no rotor contact. Brass rotors and stators are soldered and nickel-plated. The contact wiper is heavily silver-plated beryllium-copper.

Capacities range from 3.4 to 40.1 mmf.

The "APC"

The "APC" trimmer capacitor originated by Hammarlund over twenty years ago is still widely recognized as the standard capacitor of its type.

Its use is indicated in all classes of equipment where a compact, high quality air dielectric trimmer is required. It was designed to resist effects of temperature, moisture and vibration.

Capacities range from 3.0 to 140 mmf.

The "MAPC"

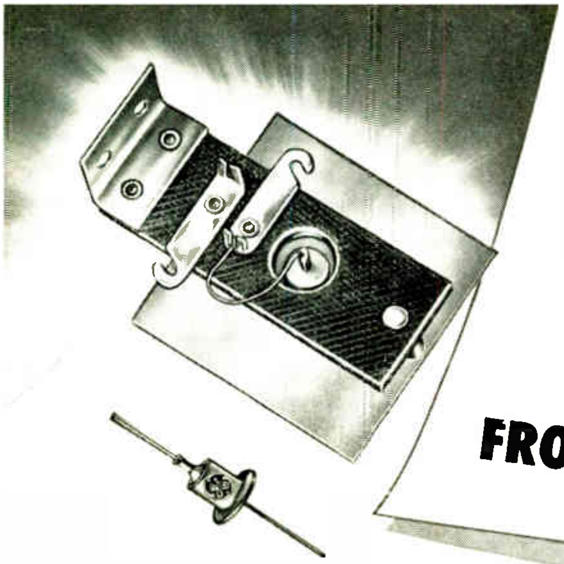
The "MAPC" capacitor is a scaled-down version of the popular "APC" with everything reduced except the quality and performance characteristics.

Lower minimum capacities and low inductance make the "MAPC" suitable for VHF use. Capacities range from 2.3 to 100 mmf.

The "MC"

The "MC" is a versatile single section tuning capacitor designed to give a choice of mountings, connections and capacity characteristics. "MC-S" capacitors have a straight line capacity characteristic. "MC-M" units have offset plates resulting in a "Midline" characteristic which more equally spaces frequencies. Capacities range from 5.5 to 320 mmf.

HAMMARLUND



HOW TO GET THE *Greatest Efficiency* FROM YOUR HAM GEAR

DIFFUSED JUNCTION GERMANIUM RECTIFIERS

ABSOLUTE MAXIMUM RATINGS—T-55°C—RESISTIVE LOAD							
DIFFUSED JUNCTION RECTIFIER	1N91	1N92	1N93	1N151	1N152	1N153	1N158
RMS INPUT VOLTAGE (Volts)	—	—	—	35	70	105	130
PEAK INVERSE VOLTAGE* (Volts)	100	200	300	100	200	300	380
PEAK FORWARD CURRENT (Amps.)	0.47	0.31	0.25	1.57	1.57	1.57	1.57
D.C. OUTPUT CURRENT* (Ma.)	150	100	75	500	500	500	500
D.C. OUTPUT CURRENT—CAPACITIVE LOAD (Ma.)	—	—	—	350	350	350	350
D.C. SURGE CURRENT (Amps.)	25	25	25	25	25	25	25
FULL LOAD VOLTAGE DROP (volts peak)	0.5	0.5	0.5	0.7	0.7	0.7	1.4
LEAKAGE CURRENT (Ma., @ rated P.I.V.)	2.7	1.9	1.2	2.4	1.9	1.2	0.8
CONTINUOUS REVERSE WORKING VOLTAGE (Volts D.C.)	30	65	100	30	65	100	185
OPERATING FREQUENCY (KC)	50	50	50	50	50	50	50
STORAGE TEMPERATURE (°C)	85	85	85	85	85	85	85

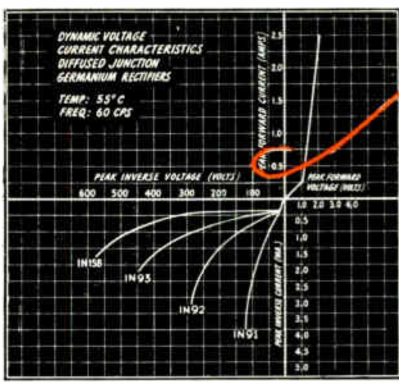
*Typical absolute maximum ratings. For other combinations refer to Fig. 1

- **VERY LOW LOSSES** when used for power supplies, battery chargers, filament supplies and switching circuits.
- **HERMETICALLY SEALED** against deteriorating elements. Glass-to-metal seals throughout.
- **MINIATURE SIZE** to facilitate use in all electronic equipments, yet heat losses are dissipated efficiently.
- **DESIGNED** to meet all military humidity tests and shock and vibration requirements.
- **MULTIPLE ARRANGEMENTS** can be made for full wave or bridge circuits up to tens of amperes. Send us your requirements—our engineers will furnish a complete recommendation.

Send for complete specifications: *General Electric Co., Section 564. Electronics Park, Syracuse, New York.*

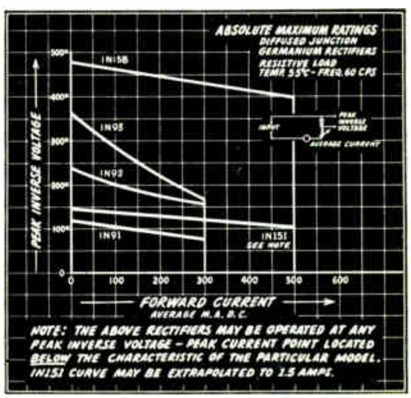
A SINGLE 1N91 RECTIFIER

- operated from a 6.3 volt transformer makes an ideal trickle charger for your storage battery!
- delivers up to .5 amps ...and is an inexpensive way to get D.C. voltage for your tube filaments!



Note: THIS IS ONLY ONE OHM!

FIG. 1

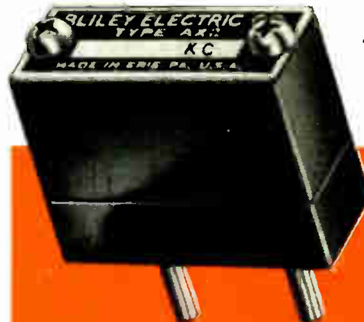


GENERAL ELECTRIC

Bliley TYPE AX2

RANGE (kc)	TOLERANCE (kc)	PRICE
1803-1822 1878-1897 1903-1922 1978-1997	±1	$3.75
3500-3997	±5	$2.95
7000-7425 8000-8222	±5	$2.95
12500-13615 14000-14850	±30	$3.95

On crystals supplied to the tolerance above, the nameplate frequency is calibrated to ± .002% in factory test equipment. The drift is less than .0002% per °C.



Bliley TYPE AX3

RANGE (kc)	TOLERANCE (kc)	PRICE
24000-24333 25000-25500	±5	$3.95

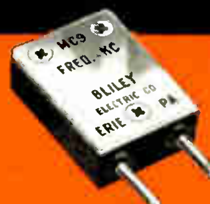
Specially designed third overtone crystal produced for the Bliley CCO-2A oscillator. On crystals supplied to the tolerance above, the nameplate frequency is calibrated to ± .003% in factory test equipment. The drift is less than .0002% per °C.

Bliley Crystals FOR 23 YEARS TOP AMATEUR CHOICE . . .

Bliley TYPE CCO-2A

This famous packaged oscillator unit was designed and engineered to utilize the many advantages of crystal control on 2-6-10-11 meters. With the CCO-2A, output is obtained directly on 6-10-11 meters; operation on 2 meters requires only a tripler stage. Specified for 10 meters and 11 meters is the Bliley type AX2. For 6 meter operation, use Bliley type AX3. On 2 meters, select an AX3 crystal which will triple to the desired transmitting frequency.

PRICE: $11.95 (Less Tube and Crystal)



Bliley TYPE MC9

SPOT FREQUENCIES for NET OPERATION

RANGE (kc)	TOLERANCE (kc)	PRICE
ANY SPECIFIED FREQUENCY BETWEEN 3000-10000	±.03%	$4.80

On crystals supplied to the tolerance above, the nameplate frequency is calibrated to ± .002% in factory test equipment. The drift is less than .0002% per °C.

You will find these famous Bliley amateur products described in Bulletin 44-A—now available at your favorite distributor of amateur equipment.

Bliley CRYSTALS

BLILEY ELECTRIC COMPANY

UNION STATION BLDG., ERIE, PENNSYLVANIA

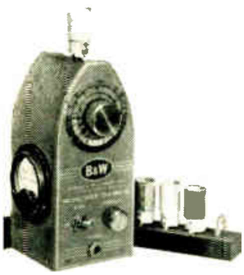
B&W PRODUCTS of the YEAR

Radio Transmitter Model 5100

- OUTPUT—100 watts phone, 125 watts CW.
- OPERATION—VFO or 80 meter crystal. VFO is built-in accurately calibrated, highly stable.
- COVERAGE—80-40-20-15-11-10 Meter bands.
- R-F AMPLIFIER—Pi-network tank circuit for matching impedances from 40 to 125 ohms.
- Television interference suppressed.

This B&W transmitter has been designed for the most discriminating operator. It is built for consistent, dependable service, and a high quality signal output, be it telephone or telegraph. Though highly compact, its layout is straight-forward with all components easily accessible. The heavy gauge steel cabinet is finished in a pleasing shade of blue-gray Hammertone. The richly finished front panel clearly indicates all switch and knob designations. Weighing only 83 lbs., 21½" wide x 15¼" high x 15⅛" deep, this B&W Transmitter is an attractive piece of equipment that will catch the eye of the most critical amateur.

Grid Dip Meter

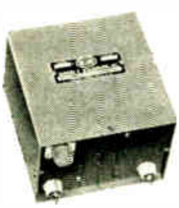


This indispensable measuring instrument enables you to get the most out of your equipment. Its amazing performance will save you time in initial transmitter tuning, neutralizing, antenna loading, and dozens of other jobs. Extremely versatile, it can be used as a sensitive grid dip meter, a signal generator, an absorption wave meter, or as a signal monitor from 1.75 to 260 mc. The calibrated color coded dial is divided in 5 bands matching each of the 5 coils supplied.

1-KW Multi-Band Balun

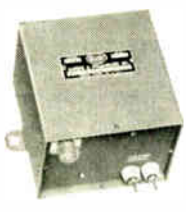
This deluxe unit provides selection of an appropriate Balun by the twist of a selector knob when the operator desires to change operation from one band to another. It may be used with any one of five separate half wave folded dipoles or any multi-band antenna system whose feedline reflects a constant impedance of 300 ohms. Intended for indoor use, within easy reach from the operating position, its design provides for operation within the amateur bands 80-40-20-15 and 10 meters. Impedance characteristics are 75 ohms unbalanced input to 300 ohms balanced output.

1-KW Single Band Beam Balun



These 1-KW Baluns are ideal for use with beam antennas employing the popular "T" matching section. Available in models for either 20, 15, or 10 meter operation, the input of each Balun is designed to match a 75 ohm unbalanced feed line system while the output, in a combination connection with a "T" section, matches 100 ohms. The metal case is weatherproof and fitted with a coaxial input connector and ceramic feedthru output terminals.

1-KW Single Band Balun



These 1-KW Baluns are designed for half-wave folded dipole antennas. Input characteristics are 75 ohms unbalanced to an output of 300 ohms balanced. They are available in five models for operation at either 80-40-20-15 or 10 meters. Housed in sturdy, weatherproof Hammertone metal cases, all are equipped with coaxial input connectors and ceramic feedthru output terminals.

R-F Power Meter

This handy instrument is indispensable for measuring the output power of transmitters, and as a dummy load in tune up procedures when no radiation is desired. The large, fan-type, power output scale is invaluable for indicating the final performance of your equipment. Power rating is 125 watts intermittent, 100 watts continuous. Max. SWR is 1.5:1. Models for 52 or 75 ohm input impedance are available for 0 to 40 mc or 35 to 100 mc.

TURN PAGE FOR MORE B&W DEPENDABLE PRODUCTS

SIGNALING & CONTROL EQUIPMENT

for dependable operations!

Economical Systems of Modern Design

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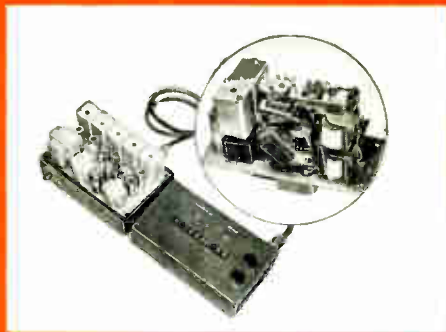
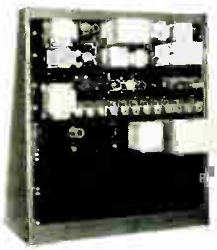


Push-Button Selective Calling



Aircraft Communications System Control

Multi-Gate for Remote Control

THE HAMMARLUND MANUFACTURING CO., INC.
460 WEST 34th STREET • NEW YORK 1, N. Y.

When Your Design Considerations are Critical—Put in

Reliable Components

"MAC"

"BFC"

"APC"

"MAPC"

The "MAC"

The "MAC" provides the low minimum capacity essential for use as a trimmer in the VHF range. It was engineered to achieve the smallest dimensions practical to meet the requirements of a miniaturized component. Its silicone treated steatite base is only ¾" x ⅝".

Rotor and stator are soldered assemblies and are of brass, nickel plated. Capacities range from 1.4 to 19.6 mmf.

The "BFC"

The "BFC" "butterfly" type capacitor has very low minimum capacity, low inductance and isolated rotor for use in VHF applications as a series capacitor with no rotor contact. Brass rotors and stators are soldered and nickel-plated. The contact wiper is heavily silver-plated beryllium-copper.

Capacities range from 3.4 to 40.1 mmf.

The "APC"

The "APC" trimmer capacitor originated by Hammarlund over twenty years ago is still widely recognized as the standard capacitor of its type.

Its use is indicated in all classes of equipment where a compact, high quality air dielectric trimmer is required. It was designed to resist effects of temperature, moisture and vibration.

Capacities range from 3.0 to 140 mmf.

The "MAPC"

The "MAPC" capacitor is a scaled-down version of the popular "APC" with everything reduced except the quality and performance characteristics.

Lower minimum capacities and low inductance make the "MAPC" suitable for VHF use. Capacities range from 2.3 to 100 mmf.

The "MC"

The "MC" is a versatile single section tuning capacitor designed to give a choice of mountings, connections and capacity characteristics. "MC-S" capacitors have a straight line capacity characteristic. "MC-M" units have offset plates resulting in a "Midline" characteristic which more equally spaces frequencies. Capacities range from 5.5 to 320 mmf.

HAMMARLUND

HAMMARLUND CAPACITORS

for Your Equipment

"MC"

The "RMC"

The "RMC" was designed specifically for applications requiring an "MC" type tuning capacitor with very rigid construction, such as mobile operation. Its sturdy frame consists of heavy gauge aluminum end panels held together by three aluminum tie rods.

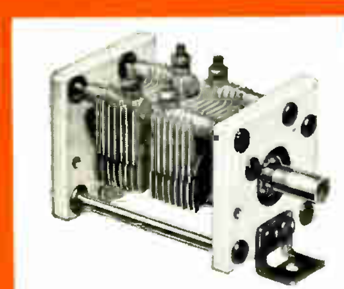
It has a brass sleeve front bearing and a single ball thrust rear bearing for smooth tuning and a high degree of resetability. Capacities range from 7.3 to 327 mmf.

"RMC"

The "VU"

A uniquely designed UHF tuning capacitor using completely original concepts. With it, conventional "lumped constant" circuits, rather than tuned cavity techniques, can be efficiently used up to 500 Mc. Capacitor sections are in series to eliminate rotor wiper and the design utilizes Pyrex balls to form precision bearings and to completely isolate the rotor.

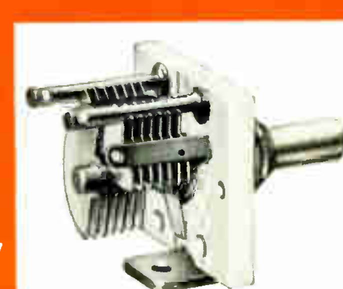
The vacuum tube and inductor can be mounted adjacent to and on opposite sides of the capacitor to minimize circuit inductance. Capacities range from 3.35 to 45 mmf.



"VU"

The "HF" Series

The "HF" is a single section tuning capacitor employing "APC" rotor and stator design. Extra-long sleeve bearing and positive contact nickel-plated phosphor-bronze wiper make this unit ideally suited to high frequency applications. Also available as a dual "HFD". Capacities range from 2.8 to 142 mmf.



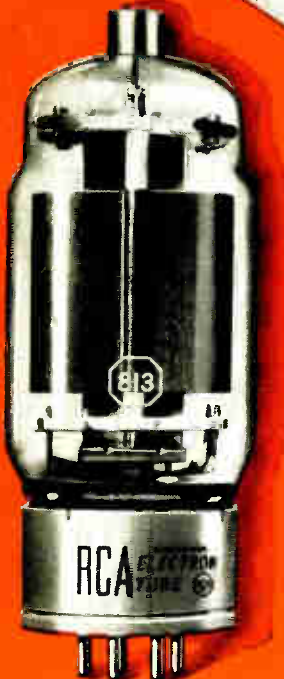
"HF"

Write today for our latest Capacitor Catalog →

THE HAMMARLUND MANUFACTURING CO., INC.
460 WEST 34th STREET • NEW YORK 1, N. Y.

Do it Better

CW and phone power values listed are maximum rated plate-input values for amateur service.



500 watts CW; 400 watts phone
(Full input up to 30 Mc)
This high-power rf amplifier delivers more watts per dollar than any amateur beam power tube in its class. A 6L6 drives it. Operates efficiently over a wide range of plate voltage.

120 watts CW; 90 watts phone
(Full input up to 200 Mc)
Twin Beam Power Tube for VHF. One 2E26 drives it; two 807's modulate it. Maximum plate voltage: 750v on CW, 600v on phone.

17 watts CW; 15 watts phone
(Full input up to 50 Mc
Reduced input up to 175 Mc)
RF amplifier, frequency multiplier, oscillator—handles more power than any "miniature" in amateur radio. For mobile, portable, or fixed service. It delivers full power with only a 350v supply.

75 watts CW; 60 watts phone
(Full input up to 60 Mc)
Versatile beam power tube for rf amplifiers, modulators, and frequency multipliers. A 6AG7 drives it; a pair of 807's modulates it.

40 watts CW; 27 watts phone
(Full input up to 125 Mc
Reduced input up to 175 Mc)
RF power amplifier, freq. multiplier, oscillator. Well-suited for mobile and emergency services. A 6AG7 drives it; a pair of 6V6-GT's or a 6N7 modulates it.

345 watts CW; 270 watts phone
(Full input up to 50 Mc)
Medium-power rf amplifier. A 6AG7 drives it. Maximum plate voltage: 3000v for CW, 2500v for phone.

RCA—Headquarters for Beam Power Tubes, Power Triodes,

with Beam Power

For that modern, high-efficiency, low-cost transmitter, always design around RCA Beam Power Tubes

Beam power tubes—an original RCA development—enable you to design and build high-efficiency rigs at less cost. Because of the low power required to drive beam power tubes, you can operate your transmitter with low-power-level exciters—minimize your TVI problems. With beam power tubes, you get peak performance with fewer stages—which makes for easier band switching—fewer tubes, and fewer components.

Put the advantages of beam power to work for you when you design. Specify RCA. There's an RCA Beam Power Tube for every input up to 1000 watts—and every frequency up to 225 Mc. For additional data on RCA beam power types see Table 17 in Chapter 27 of this Handbook.

RCA HAM TIPS. Written by radio amateurs for radio amateurs, this up-to-the-minute publication is jam-full of how-to-make-it articles, as well as non-constructional articles on TVI. Free from your RCA Tube Distributor. Or write RCA, Commercial Engineering, Section A11M, Harrison, N. J.

RCA®

RADIO CORPORATION of AMERICA

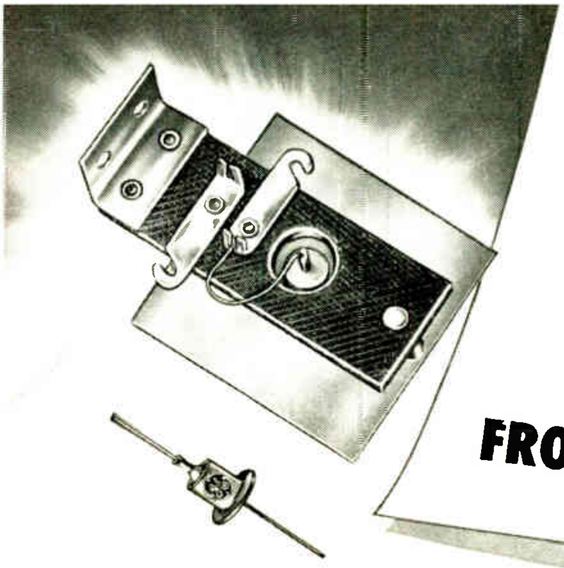
ELECTRON TUBES ***HARRISON, N. J.***

1000 watts and more, CW
675 watts phone
(Full input up to 75 Mc)
High-power "final." A 2E26 drives it. A pair of 810's modulates it. Maximum plate voltage is 4000 volts for CW and 3200 volts for phone.

500 watts CW; 375 watts phone
(Full input up to 120 Mc)
Two 4-125A/4D21's take 1 KW on CW, 750 watts on phone. One 2E26 drives a pair. Two 811-A's modulate them. Maximum plate voltage is 3000v for CW, 2500v for phone.

90 watts CW; 67.5 watts phone
(Full input up to 60 Mc
Reduced input up to 175 Mc)
Circuit versatility of this tube is matched by few other types. Excellent for 2-meter work—handles 48 watts input on phone at 150 Mc. In a class AB_2 modulator, two 6146's can deliver up to 120 watts.

Thyratrons, Rectifiers, Voltage Regulator Tubes



HOW TO GET THE *Greatest Efficiency* FROM YOUR HAM GEAR

DIFFUSED JUNCTION GERMANIUM RECTIFIERS

ABSOLUTE MAXIMUM RATINGS—T-55°C—RESISTIVE LOAD							
DIFFUSED JUNCTION RECTIFIER	1N91	1N92	1N93	1N151	1N152	1N153	1N158
RMS INPUT VOLTAGE (Volts)	—	—	—	35	70	105	130
PEAK INVERSE VOLTAGE* (Volts)	100	200	300	100	200	300	380
PEAK FORWARD CURRENT (Amps.)	0.47	0.31	0.25	1.57	1.57	1.57	1.57
D.C. OUTPUT CURRENT* (Ma.)	150	100	75	500	500	500	500
D.C. OUTPUT CURRENT—CAPACITIVE LOAD (Ma.)	—	—	—	350	350	350	350
D.C. SURGE CURRENT (Amps.)	25	25	25	25	25	25	25
FULL LOAD VOLTAGE DROP (volts peak)	0.5	0.5	0.5	0.7	0.7	0.7	1.4
LEAKAGE CURRENT (Ma., @ rated P.I.V.)	2.7	1.9	1.2	2.4	1.9	1.2	0.8
CONTINUOUS REVERSE WORKING VOLTAGE (Volts D.C.)	30	65	100	30	65	100	185
OPERATING FREQUENCY (KC)	50	50	50	50	50	50	50
STORAGE TEMPERATURE (°C)	85	85	85	85	85	85	85

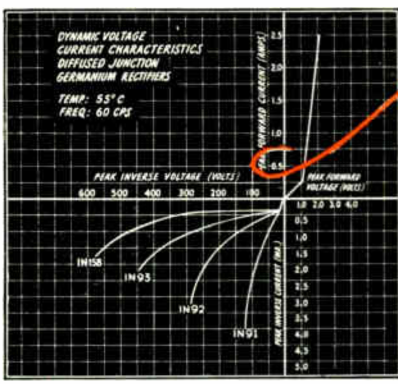
*Typical absolute maximum ratings. For other combinations refer to Fig. 1

- **VERY LOW LOSSES** when used for power supplies, battery chargers, filament supplies and switching circuits.
- **HERMETICALLY SEALED** against deteriorating elements. Glass-to-metal seals throughout.
- **MINIATURE SIZE** to facilitate use in all electronic equipments, yet heat losses are dissipated efficiently.
- **DESIGNED** to meet all military humidity tests and shock and vibration requirements.
- **MULTIPLE ARRANGEMENTS** can be made for full wave or bridge circuits up to tens of amperes. Send us your requirements—our engineers will furnish a complete recommendation.

Send for complete specifications: *General Electric Co., Section 564, Electronics Park, Syracuse, New York.*

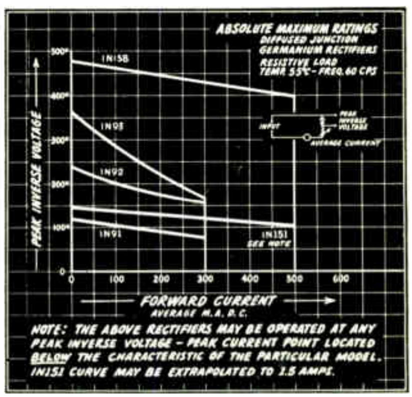
A SINGLE 1N91 RECTIFIER

- operated from a 6.3 volt transformer makes an ideal trickle charger for your storage battery!
- delivers up to .5 amps ...and is an inexpensive way to get D.C. voltage for your tube filaments!



Note: **THIS IS ONLY ONE OHM!**

FIG. 1



GENERAL ELECTRIC

Bliley TYPE AX2

RANGE (kc)	TOLERANCE (kc)	PRICE
1803-1822 1878-1897 1903-1922 1978-1997	±1	$3.75
3500-3997	±5	$2.95
7000-7425 8000-8222	±5	$2.95
12500-13615 14000-14850	±30	$3.95

On crystals supplied to the tolerance above, the nameplate frequency is calibrated to ± .002% in factory test equipment. The drift is less than .0002% per °C.

Bliley TYPE AX3

RANGE (kc)	TOLERANCE (kc)	PRICE
24000-24333 25000-25500	±5	$3.95

Specially designed third overtone crystal produced for the Bliley CCO-2A oscillator. On crystals supplied to the tolerance above, the nameplate frequency is calibrated to ± .003% in factory test equipment. The drift is less than .0002% per °C.

Bliley Crystals FOR 23 YEARS TOP AMATEUR CHOICE . . .

Bliley TYPE CCO-2A

This famous packaged oscillator unit was designed and engineered to utilize the many advantages of crystal control on 2-6-10-11 meters. With the CCO-2A, output is obtained directly on 6-10-11 meters; operation on 2 meters requires only a tripler stage.

Specified for 10 meters and 11 meters is the Bliley type AX2. For 6 meter operation, use Bliley type AX3. On 2 meters, select an AX3 crystal which will triple to the desired transmitting frequency.

PRICE: $11.95 (Less Tube and Crystal)

Bliley TYPE MC9

SPOT FREQUENCIES *for* NET OPERATION

RANGE (kc)	TOLERANCE (kc)	PRICE
ANY SPECIFIED FREQUENCY BETWEEN 3000-10000	±.03%	$4.80

On crystals supplied to the tolerance above, the nameplate frequency is calibrated to ± .002% in factory test equipment. The drift is less than .0002% per °C.

You will find these famous Bliley amateur products described in Bulletin 44-A—now available at your favorite distributor of amateur equipment.

Bliley CRYSTALS

BLILEY ELECTRIC COMPANY

UNION STATION BLDG., ERIE, PENNSYLVANIA

B&W PRODUCTS of the YEAR

Radio Transmitter Model 5100

- **OUTPUT—100 watts phone, 125 watts CW.**
- **OPERATION—VFO or 80 meter crystal. VFO is built-in accurately calibrated, highly stable.**
- **COVERAGE—80-40-20-15-11-10 Meter bands.**
- **R-F AMPLIFIER—Pi-network tank circuit for matching impedances from 40 to 125 ohms.**
- **Television interference suppressed.**

This B&W transmitter has been designed for the most discriminating operator. It is built for consistent, dependable service, and a high quality signal output, be it telephone or telegraph. Though highly compact, its layout is straight-forward with all components easily accessible. The heavy gauge steel cabinet is finished in a pleasing shade of blue-gray Hammertone. The richly finished front panel clearly indicates all switch and knob designations. Weighing only 83 lbs., 21½" wide x 15¼" high x 15⅛" deep, this B&W Transmitter is an attractive piece of equipment that will catch the eye of the most critical amateur.

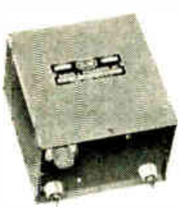
Grid Dip Meter

This indispensable measuring instrument enables you to get the most out of your equipment. Its amazing performance will save you time in initial transmitter tuning, neutralizing, antenna loading, and dozens of other jobs Extremely versatile, it can be used as a sensitive grid dip meter, a signal generator, an absorption wave meter, or as a signal monitor from 1.75 to 260 mc. The calibrated color coded dial is divided in 5 bands matching each of the 5 coils supplied.

1-KW Multi-Band Balun

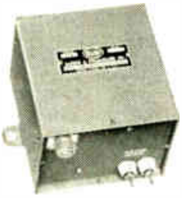
This deluxe unit provides selection of an appropriate Balun by the twist of a selector knob when the operator desires to change operation from one band to another. It may be used with any one of five separate half wave folded dipoles or any multi-band antenna system whose feedline reflects a constant impedance of 300 ohms. Intended for indoor use, within easy reach from the operating position, its design provides for operation within the amateur bands 80-40-20-15 and 10 meters. Impedance characteristics are 75 ohms unbalanced input to 300 ohms balanced output.

1-KW Single Band Beam Balun



These 1-KW Baluns are ideal for use with beam antennas employing the popular "T" matching section. Available in models for either 20, 15, or 10 meter operation, the input of each Balun is designed to match a 75 ohm unbalanced feed line system while the output, in a combination connection with a "T" section, matches 100 ohms. The metal case is weatherproof and fitted with a coaxial input connector and ceramic feedthru output terminals.

1-KW Single Band Balun



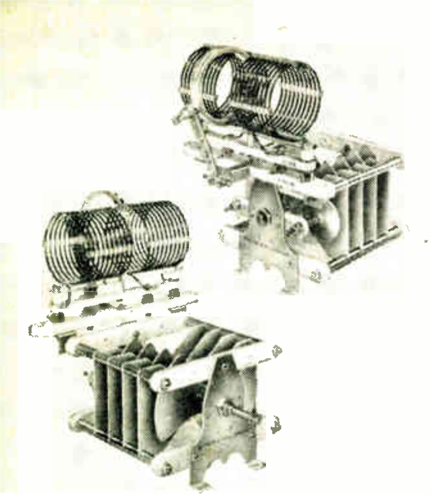
These 1-KW Baluns are designed for half-wave folded dipole antennas. Input characteristics are 75 ohms unbalanced to an output of 300 ohms balanced. They are available in five models for operation at either 80-40-20-15 or 10 meters. Housed in sturdy, weatherproof Hammertone metal cases, all are equipped with coaxial input connectors and ceramic feed-thru output terminals.

R-F Power Meter

This handy instrument is indispensable for measuring the output power of transmitters, and as a dummy load in tune up procedures when no radiation is desired. The large, fan-type, power output scale is invaluable for indicating the final performance of your equipment. Power rating is 125 watts intermittent, 100 watts continuous. Max. SWR is 1.5:1. Models for 52 or 75 ohm input impedance are available for 0 to 40 mc or 35 to 100 mc.

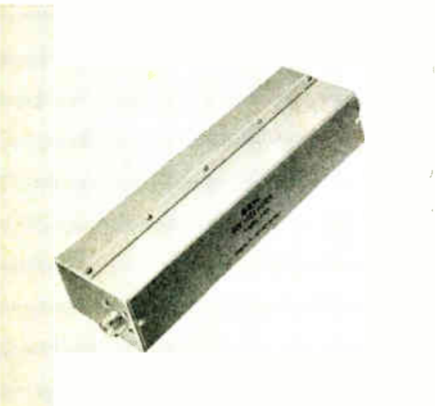
TURN PAGE FOR MORE B&W DEPENDABLE PRODUCTS

B&W PARTS and



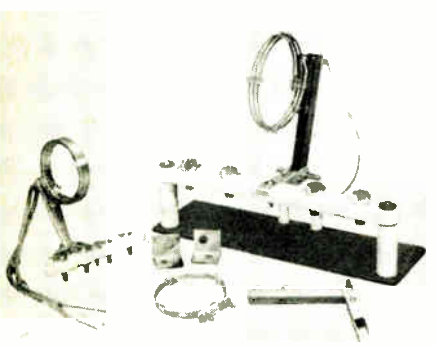
Heavy Duty Butterfly Variable Capacitors

B & W heavy duty butterfly variable capacitors with coils integrally mounted pave the way for increased efficiency in single-ended and push-pull circuits. Better L. C. ratios at high frequencies, with beam power tubes as well as a host of other desirable features, are a reality with these husky units. These include: compact assembly, shorter tuned circuit leads, shorter R. F. paths and optional built-in neutralizing condensers.



Low Pass Filters

B&W Low Pass Filters are highly effective in attenuating harmonics causing TVI. Attenuation is 85db or more thru the TV band; insertion loss less than .25db thru the pass band to 30mc. They can easily handle more than 1 KW of r-f power. Consisting of four "K" sections plus two "M" derived end sections, these filters are constructed of heavy gauge copper, employ the highest grade electrical Teflon insulation, in a compact, screw-assembled case.



Bases and Mounting Assemblies

These accessories permit compact assemblies with companion units such as capacitors, jack bars, plug-in coils, and links. Two groups are available, one for open wire plug-in swinging links, and another for Faraday Shielded links. Assemblies include a jack bar, arm and hinge, link (open wire or shielded), and either a metal bottom plate or capacitor mounting bracket. Individual parts may be purchased.

B&W TEST

Audio Oscillator

Freq. Range: 30 to 30,000 cycles.
Freq. Response: Better than ± 1 DB. 30 to 15,000 cycles with 500 ohm load.
Stability: Better than 1%.

Distortion Meter

Freq. Range: Fundamentals from 30 to 15,000 cycles. Measures harmonics to 45,000 cycles.
Sensitivity: .3 volts minimum input required.

Frequency Meter

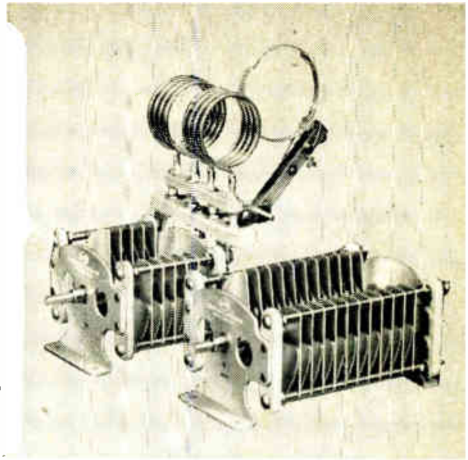
Freq. Range: 0 to 30,000 cycles.
Sensitivity: 0.25 volts minimum input required.
Wave Form: Any form with peak ratios less than 8:1.

BARKER & WILLIAMSON, INC.

EQUIPMENT

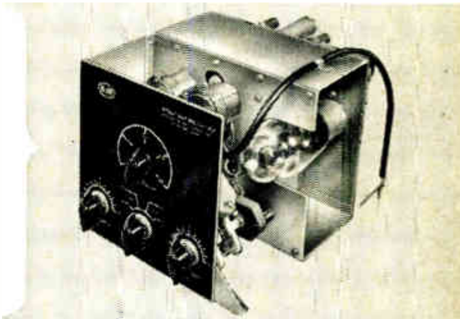
Having 25% of the frontal area of the Heavy Duty Type, these split-stator variable capacitors are ideal for medium power triode or tetrode stage plate circuits and many other applications. Heavy rounded edge plates permit ratings up to 2500 volts dc unmodulated and 1500 volts dc in modulated final amplifier circuits. Design provides peak efficiency and more power in less space.

Junior Butterfly Variable Capacitors



This compact, versatile unit is in keeping with modern trends toward miniaturization. Operated with either crystal or VFO, it serves as an exciter for a high powered rig or as a low powered transmitter with a full 30 watt of output on the amateur bands including 80- 40- 20- 15- 11 and 10 meters. It avoids the most laborious and time consuming part of the job during construction of a new transmitter.

504 Multiplier

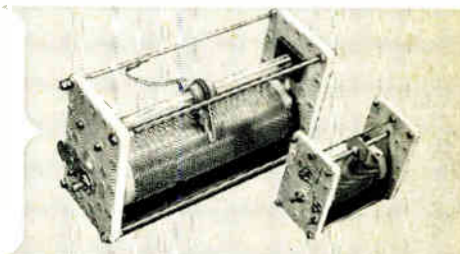


Provides an efficient watertight insulated connector for center-feed antenna systems using coaxial cable for feed lines.
Light in weight, it will withstand pulling strains up to 500 lbs.

CC 50 Coaxial Connector

B & W Rotary Coils are available for all medium and high power requirements of pi-network, final circuits, and antenna coupling and loading units. 500-watt units are supplied with inductances of 1.6, 6.2, 15, and 72 micro-henries, 1000-watt types with 60 or 96 micro-henries.

Rotary Coils



INSTRUMENTS

Sine Wave Clipper

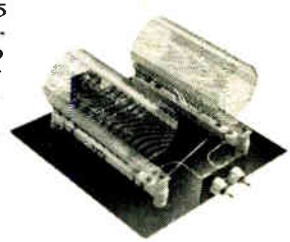
Does the work of a square wave generator costing many times more. Speeds accurate circuit analysis.

Linear Detector

Provides R-F detection and audio bridging circuits. It is an invaluable accessory for distortion meters lacking these features.

BALUN INDUCTORS . . .

Match 75 ohm unbalanced outputs to 75 and 300 ohm balanced feed lines. Two of these sturdy bifilar air-wound balun inductors serve as a compact, highly efficient multi-band (80 to 10 meters) unit for matching feed line systems to both transmitters and receivers.



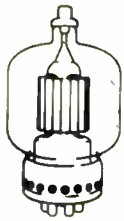
WRITE FOR FREE CATALOG

237 Fairfield Ave., Upper Darby, Pa.

follow the leaders

TRIODES

2C39A
3W5000A3
3W5000F3
3W10000A3
3X2500A3
3X2500F3
3X3000A1
3X3000F1
6C21
25T
35T
35TG
75TH
75TL
100TH
100TL
152TH
152TL
250TH
250TL
304TH
304TL
450TH
450TL
592/3-200A3
750TL
1000T
1500T
2000T



TETRODES

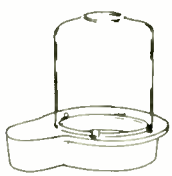
4-65A
4-125A
4-250A
4-400A
4-1000A
4PR60A
4W20000A
4X150A
4X150D
4X150G
4X500A
4X500F

PENTODE

4E27A/5-125B

RECTIFIERS

2-01C
2-25A
2-50A
2-150D
2-240A
2-2000A
KY21A
RX21A
250R
253
866A
872A
8020 (100R)



KLYSTRONS

3K20,000LA
3K20,000LF
3K20,000LK
3K50,000LA
3K50,000LF
3K50,000LK

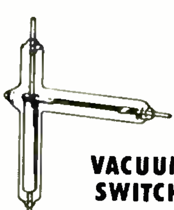
AIR SYSTEM SOCKETS

4-400A/4000
4-400A/4006*
4-1000A/4000
4-1000A/4006*
4X150A/4000
4X150A/4006*
4X150A/4010

*Replacement Chimneys

ACCESSORIES

HR Heat dissipating connectors
Preformed Contact Finger Stock



VACUUM SWITCH

VS-2
12V Coil
24V Coil

ION GAUGE

100 IG ion gauge

VARIABLE VACUUM CAPACITORS

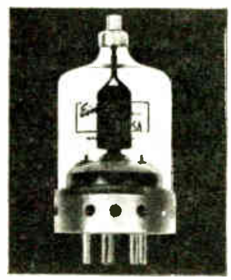
VVC60-20
VVC2-60-20
VVC4-60-20

VACUUM CAPACITORS

VC6-20
VC6-32
VC12-20
VC12-32
VC25-20
VC25-32
VC50-20
VC50-32

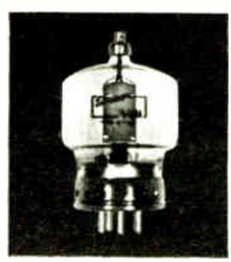
2C39A

This small, rugged triode is designed for use as a power amplifier, oscillator or frequency multiplier to frequencies above 2500 mc. It is particularly suitable for compact fixed or mobile equipment.



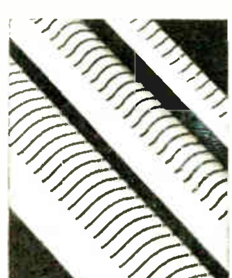
4-125A

The radial-beam power tetrode that made transmitting screen-grid tubes popular. This tube will take a plate input of 500 watts for CW or 380 watts for fone. Driving power is less than two watts. A pair of these tetrodes make an ideal high power fone or CW final for the amateur.



4-250A

A pair of these radial-beam power tetrodes will easily handle a kilowatt for fone. In CW service, one tube will take a kilowatt input. Driving power is only two to three watts per tube. As modulators a pair will deliver as much as 750 watts audio with simple resistance coupled driver stages.



FINGER STOCK

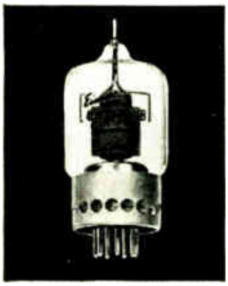
Preformed Contact Finger Stock is a useful electrical "weather strip" around accesses to equipment cabinets as well as providing good circuit continuity between adjustable components. It is ideally designed for making connections to coaxially constructed and external anode tubes.

Export Agents: Frazar & Hansen, 301 Clay St., San Francisco

to EIMAC TUBES!

4X150A

This small external anode radial-beam power tetrode operates efficiently at all frequencies into the UHF range with a driving power of only a few watts. Its small size and ruggedness make it ideal for compact equipment such as mobile.



4E27A

With simple circuits and less than two watts driving power this radial-beam power pentode gives dependable operation and high output. It is capable of an easy 500 watts input in Class-C service — or when suppressor modulated will deliver 75 watts output at carrier conditions.

250T

A tried, proven and continually improved 250 watt triode. The ideal triode for one KW CW input. Will handle 825 watts input on fone. With plate voltage as low as 1500 volts in Class-B audio service a pair will modulate a KW RF stage.

4W20,000A

In pulse service and TV operation the Eimac 4W20,000A is the only time-proved tetrode in its power class. Its rugged construction includes a ceramic envelope that minimizes losses and increases operational life. In VHF-TV operation, it gives 25kw peak sync power output with only 500 watts drive.

KLYSTRONS

Eimac six kilowatt klystrons, 3K20,000L (A,F,K), and 15 kilowatt klystrons, 3K50,000L (A,F,K), are outstanding for high power CW and TV at UHF. Only three tubes of either series are required to span the entire UHF-TV spectrum. Externally tuned circuitry and ceramic envelopes are among other features.

VVC60-20

This is but one type in the Eimac line of variable and fixed vacuum capacitors for plate tank circuits. It is variable over a range of 10 mmfd to 60 mmfd. Maximum rf voltage is 20 kv at 40 amperes.

- Write for 28-page booklet, "Care and Feeding of Power Tetrodes." Available free upon request.

Eimac maintains an Amateurs' Service Bureau for amateur radio operators. Free information may be obtained by writing. Available for engineering consultation and information is the Eimac Application Engineering department.

Eimac TUBES

EITEL - McCULLOUGH, INC.

SAN BRUNO, CALIFORNIA

ESICO

REG. U. S. PAT. OFF.

INDUSTRIAL SOLDERING IRONS

are the result of 25 years of specializing in the manufacture of high quality electric soldering irons and they are used today in a great majority of the country's electrical, radio and electronic plants.

No. 61

A lightweight (2½ ounce), low cost unit for pin-point accuracy in the most delicate soldering operations. Element construction is of same type used in ESICO industrial irons. Handle temperature is never higher than body temperature. Diameter of handle ¾". Tips available in 3 shapes: type B—¼" dia., pyramid point; type A—⅛" dia., straight pencil point; type C—⅛" dia., bent 90 degrees, with pencil-like point. Regularly wound to 25 watts at 105–120 volts. May be obtained in higher wattages at no extra cost when purchased in quantities. Iron, as illustrated, is ½ actual size.

No. 61A

The #61A iron is very similar to the #61 except that the case enclosing the element is slightly longer. Where an iron which can be held as a pencil is required, it is possible to have this iron in wattages as high as 75 watts for fast soldering of moderate size parts. The iron, as illustrated, is ½ actual size.

No. 62B

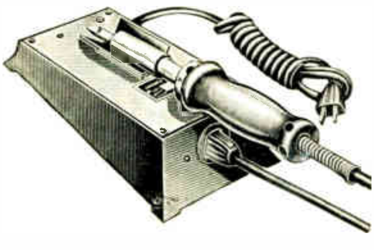
This iron is available in 100 watts for either 105/120 volts or 220/240 volts. It is intended for 100 watt capacity work, but where a small diameter case is required in order to get to the connections to be soldered. The iron is extremely light, and due to special construction has an extremely low handle temperature. Iron, as illustrated, is ⅓ actual size.

No. 38

This is the iron that is so widely used in the large radio plants and, in many instances, is used in a 150 watt capacity, though the standard wattage of the iron is 100 watts. It is recommended that 150 watts be used only where there is fairly continuous soldering. The iron is of a type in which the element can be easily replaced by loosening a knurled nut at the back of the case, and after lead wires are loosened, by pulling out the old element and inserting the new one in a few minutes. The iron is ruggedly constructed and requires a minimum of attention and is completely serviceable within one's own plant. This iron is fast becoming the most popular iron in use in the electronic industry. The iron, as illustrated, is ⅓ actual size.

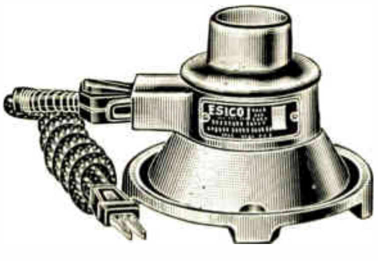
Temperature Control Stand

A practical, time and money saving device which accurately regulates and maintains soldering iron temperature between jobs. Lengthens iron life by reducing tip oxidation and amalgamation of tip with solder which increases with over-heating. When placed on stand, iron rests in a copper cradle which conducts heat of iron and actuates a bimetal to open or close a switch. Temperature is easily regulated by an adjusting slide at bottom of stand. As iron is removed from stand, full current is instantly supplied. Stem rest is adjustable to accommodate various lengths of irons. Stand is a heavy gray iron casting—stays firmly fixed without being fastened.



Solder Pots

Designed to meet rigorous production requirements, ESICO solder pots are made from high quality gray iron castings. They are fitted with heater plate type elements which can be easily and quickly replaced. Elements wound from highest quality nickel chrome resistance wire. Elements of the three pots are interchangeable for greater economy and flexibility.



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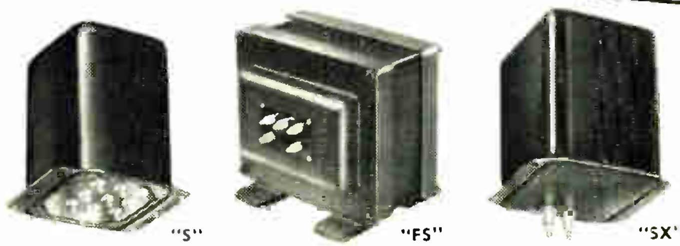
CHICAGO
TRANSFORMERS



Designed for Modern Circuit Applications: Amateur Communications, Broadcast, TV, & Audio

FILAMENT TRANSFORMERS

Primary: 115/230 volts, 50/60 cycles.

Chicago filament transformers are rated to provide voltages and currents for heating a wide range of receiving and transmitting tubes. All units are S-type mountings (see illustration above). Those with secondaries rated at less than 6 amps. have solder-lug terminals, those over 6 amps. have screw-type terminals. Units marked * have ceramic-insulated secondary terminals, especially suited for high voltage rectifier supply.

Catalog No.	Secondary Volts	Secondary Amps.	Insulation Volts RMS	Net Price
F-25	2.5 CT	5.25	3500	$ 6.33
F-210*	2.5 CT	10.	5000	9.09
F-210H*	2.5 CT	10.	9000	10.65
F-215H*	2.5 CT	15.	9000	12.75
F-54	5.0 CT	4.0	2500	6.51
F-58	5.0 CT	10	2500	8.70
F-510H*	5.0 CT	10.	10000	13.50
F-516	5.0 CT	20.	2500	13.11
F-520HB*	5.0 CT	20.	10000	16.95
F-530	5.0 CT	30	2500	16.95
F-615	6.3 CT	1.5	2500	3.75
F-63	6.3 CT	3.	2500	5.25
F-65	6.3 CT	5.5	2500	7.65
F-610	6.3 CT	10.	2500	9.78
F-75	7.5 CT	5.0	2500	7.35
F-712	7.5 CT	12.	2500	12.75
F-725	7.5 CT	25	2500	16.95
F-104	10. CT	4.0	2500	7.80
F-106	10. CT	6.5	2500	10.77
F-1010	10. CT	10	2500	12.75
F-1110	11. CT	10.	2500	12.30

PLATE TRANSFORMERS

Primary: 115/230 volts, 50/60 cycles.

Chicago plate transformers, as well as the filter reactors, are conservatively designed. They are amply insulated, and built to withstand the most rugged service. Temperature rise is very moderate under both CCS and ICAS conditions.

Catalog No.	Max. Pri. VA	Secondary A-C Load Volts	D-C Volts after filter	D-C Ma. CCS	D-C Ma. ICAS	Mtg. Type	Net Price
P-45	185	675-0-675 575-0-575	400 500	250	325	S	$ 14.10
P-67	250	900-0-900 735-0-735	750 600	250	325	S	16.80
P-107	310	1150-0-1150 870-0-870	1000 750	250	350	FS	41.70
P-1240	360	1425-0-1425* 600-0-600	1250 400	150 200	200 260	S	23.70
P-1512	550	1710-0-1710 1430-0-1430	1500 1250	300	425	FS	47.70
P-2520	915	2820-0-2820 2260-0-2260	2500 2000	300	425	FS	71.70
P-2126	1600	2900-0-2900 2320-0-2320	2600 2100	500	700	FS	89.70
P-3025	1850	3450-0-3450 2850-0-2850	3000 2500	500	700	FS	119.70
P-4353	3050	4600-0-4600 4050-0-4050 3400-0-3400	4000 3500 3000	600	800	FS	161.70

*Both secondaries may be rectified simultaneously

FILTER REACTORS

Catalog No.	Inductance in Henries	Max. D-C Ma.	D-C Resistance, Ohms	Insulation Volts RMS	Mtg. Type	Net Price
R-67	6	700	35	10,000	FS	$33.90
R-105	10	500	40	9,000	FS	34.50
R-65	6	500	35	9,000	FS	27.48
R-103	10	300	40	7,500	SX	16.95
R-63	6	300	35	7,500	SX	14.34

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POCKETSCOPE®

The Pocket Oscilloscope by Waterman

...light... compact... accurate... portable

The HIGH, WIDE and TWIN POCKETSCOPES have become the "triple threat" of the oscilloscope industry. Their small size, light weight and incredible performance, has skyrocketed this team of truly portable instruments into unparalleled prominence. Each oscilloscope features DC coupled amplifiers in both its vertical and horizontal channels. The HIGH GAIN, S-14-A POCKETSCOPE, has a vertical sensitivity of 10 millivolts rms/inch, and a frequency response within −2 db from DC to 200 KC, while the WIDE BAND S-14-B POCKETSCOPE is characterized by frequency response within −2 db from DC to 700 KC and a sensitivity of 50 millivolts rms/inch.

The TWIN POCKETSCOPE is essentially two HIGH GAIN POCKETSCOPES with individual cathode ray tubes, amplifiers, controls, but a common sweep generator. All these are endowed with many identical characteristics. Their sweep generators can be operated as triggered or repetitive over a frequency range from 0.5 cycles to 50 KC, with synchronization polarity optional. Return traces are blanked and provisions are made for modulating the intensity in each cathode ray tube.

Laboratory quality has not been sacrificed in order to accomplish portability and ruggedness. Investigate the many advantages of Waterman POCKETSCOPES.

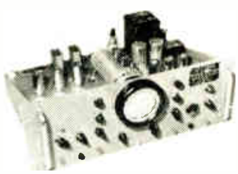
The INDUSTRIAL POCKETSCOPE, model S-11-A, has become America's most popular DC coupled oscilloscope because of its small size, light weight, and unique flexibility. This compact instrument has identical vertical and horizontal amplifiers which permit the observation of low frequency repetitive phenomena, while simultaneously eliminating undesirable trace bounce. Each amplifier sensitivity is 0.1 Volt rms/inch. The frequency responses are likewise identical, within −2 db from DC to 200 KC.

Discover for yourself the amazing utility of this tiny work-horse of industrial electronics.

POCKETSCOPE

S-11-A

RAKSCOPE®



S-12-B

The S-12-B RAKSCOPE is a rack mounted, JANized version of the famous Waterman S-11-A POCKETSCOPE, with the addition of a triggered sweep and a special calibrating circuit for rapid frequency comparisons. The entire oscilloscope is built to occupy but seven inches when mounted in a standard relay rack.

Because provisions are made for applying input signals from the rear, as well as the front, the S-12-B is the ideal combination, systems monitor and trouble-shooting oscilloscope. Investigate the multiple applications of this instrument as an integral part of your own rack mounted apparatus.

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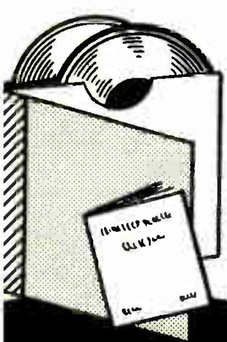

78 RPM RECORDS

"RADIO AMATEUR QUESTION & ANSWER LICENSE GUIDE"

A "must" for anyone preparing for the Novice, Technician, Conditional or General Class exams. This power-packed book contains approximately 200 questions and answers — mostly of the multiple choice type — similar to the ones given on FCC exams.

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PAR-METAL RACKS

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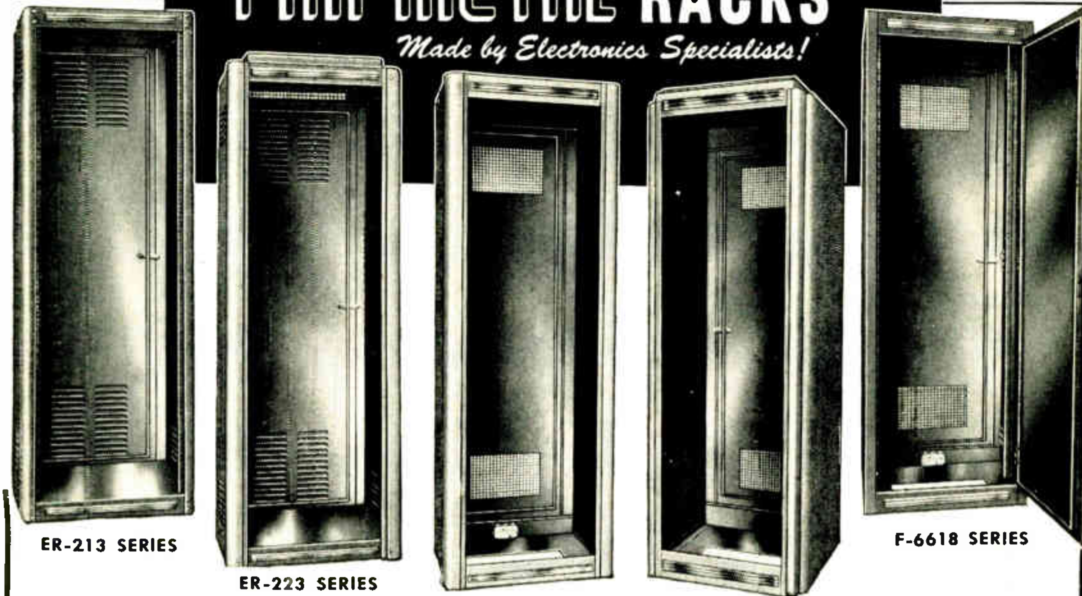

ER-213 SERIES ER-223 SERIES P-3618 SERIES P-6918 SERIES F-6618 SERIES

TYPE "A" KNOCKDOWN TYPE STANDARD RACKS

All racks are constructed from 1/16" furniture steel. Panel mounting angles are 7/64" thick steel with holes tapped 10/32 thread on universal centers. Removable doors are a new formed design, with smooth edges, equipped with sturdy automobile type handles.

Series ER-223 racks have quick fingertip detachable corner trims. These racks are available for assembly in "rows" or "gangs". When so ordered, intermediate sides and connecting front trim strips are supplied at same unit cost.

Finish: Slate grey ripple enamel is standard. Black ripple enamel is optional if specified in your order.

Series	*Cat. No.*	*Overall Dimensions L. W. D.*	*Panel Space*	*Net Price*
ER-213	ER-213	42 x22x16½"	36¼x19"	$32.70
ER-213	ER-215	66½x22x16½"	61¼x19"	48.60
ER-213	ER-217	82¼x22x16½"	77 x19"	58.50
ER-223	ER-223	43¼x22x18"	36¼x19"	46.50
ER-223	ER-225	67¼x22x18"	61¼x19"	57.00
ER-223	ER-227	83½x22x18"	77 x19"	69.00

TYPE "C" COMPLETELY ASSEMBLED STANDARD RACKS 18" and 24" Deep

Entire cabinet is welded together into one integral unit; substantially constructed from furniture steel with panel mounting angles of 3/16" thick steel. Mounting holes drilled and tapped 12/24 thread on multiple 1¼"/½" spacings.

Doors are of a new design stamped from one piece of steel into a rigid unit, with smooth double thick edges and sturdy automobile type handles.

Duplex receptacle, outlet box and opening is provided at bottom for leads, and conduit on all racks. Quick detachable corner trims fastened with finger type "captive screws."

Series P-3618 may be set up in "gangs" or "rows"; suitable front connecting strips will be supplied if so specified on your order.

Series F-6618 are equipped with front and rear doors; panel mounting angles are adjustable and may be positioned at any distance from front or rear doors. No corner trims are required.

Finish: "P" denotes black ripple, and "PG" denotes slate grey ripple for all Type "C" Racks.

Series	*Cat. No.*	*Overall Dimensions L. W. D.*	*Panel Space*	*Net Price*
P-3618	P,PG-3618	42⅞x22 x18"	36¼x19"	$ 60.00
P-3618	P,PG-6618	67⅜x22 x18"	61¼x19"	78.00
P-3618	P,PG-8318	83⅛x22 x18"	77 x19"	99.00
P-6918	P,PG-6918	69⅝x23⅝x18½"	61¼x19"	94.50
P-6918	P,PG-7818	78¾x23⅝x18½"	70 x19"	103.50
P-6918	P,PG-8518	85¾x23⅝x18½"	77 x19"	117.00
P-6918	P,PG-6924	69⅝x23⅝x24"	61¼x19"	111.00
P-6918	P,PG-7824	78¾x23⅝x24"	70 x19"	120.00
P-6918	P,PG-8524	85¾x23⅝x24"	77 x19"	135.00
F-6618	F-6618	67⅜x22 x18"	61¼x19"	105.00
F-6618	F-8318	83⅛x22 x18"	77 x19"	129.00

ROLLER TRUCKS FOR RACKS

These trucks are designed for use with the racks shown above. Overall size is about 3" wider than racks for better distribution of weight. Has rubber composition wheels. **Finished** in slate grey ripple with chrome trim.

Cat. No.	*Use with Rack Numbers*	*Inside Clearance*	*Net Price*
RT-411	ER-213, ER-215, ER-217	22⅝x17¼"	$10.20
RT-412	ER-223, ER-225, ER-227, F-6618, F-8318, P & PG-3618, P & PG-6618, P & PG-8318	22⅝x18¾"	11.40
RT-418	P & PG-6918, P & PG-7818, P & PG-8518	23⅞x19"	12.75
RT-424	P & PG-6924, P & PG-7824, P & PG-8524	23⅞x25"	13.80

CHANNEL RELAY RACKS

(For Standard 19" Rack Panels)

These racks are ideal for use on all types of Public Address Systems, and all transmitters of Commercial or Amateur types. Made from 7/64" pressed steel with tapped 10/32 holes for standard panels. Two vertical members and top cross-brace are rigidly welded together into one unit; bolts supplied to assemble bottom and braces. **Finished** in black ripple enamel.

Cat. No.	*Overall Dimensions L. W. D.*	*Panel Space*	*Net Price*
RR-193	38¼x20x18⅜"	36¾x19"	$17.40
RR-195	73¼x20x20⅞"	71¾x19"	20.70

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PAR-METAL CABINETS • CHASSIS PANELS

DESK PANEL CABINET RACKS

(For Standard 19" Rack Panels)
21½" Long x 15" Deep

Streamlined design. Vertical front corners are rounded and trimmed with mouldings. Panels fit into recess, and mounting holes are drilled for all our panels. Holes tapped 10/32, and can be used with **any chassis** up to 13" x 17". **Finished** in black ripple enamel.

Cat. No.	Overall Height	Panel Space	Net Price
	With Door in Top Only		
DL-128	10½"	8¾"	$12.00
DL-1210	12¼"	10½"	13.50
DL-1225	14"	12¼"	14.55
DL-1413	15¾"	14"	16.20
	With Door in Top and Door on Rear Panel		
DL-1713	19¼"	17½"	19.20
DL-2613	28"	26¼"	22.50
DL-3513	36¾"	35"	24.90

HINGED STEEL CABINETS

(DeLuxe Type)

Has rounded front corners. New type formed top door hinged at back with large opening for full access to interior. Chrome finished mouldings, modern handle at top and bottom opening at rear for leads. Finished in grey ripple enamel. **Prices do not include chassis.**

Cat. No.	Overall Dimensions H. L. D.	For Chassis	Net Price
CA-300	9x12⅛x 8"	7x 9x2"	$ 5.40
CA-301	9x16⅛x 8"	7x13x2"	6.30
CA-302	10x17⅛x11"	10x14x3"	9.30
CA-303	10x20⅛x 9"	8x17x3"	9.30
CA-304	13x20⅛x12"	10x17x3"	10.20

HINGED STEEL CABINETS

(Rounded Corner Type)

Excellent for monitors, oscillators, etc. These cabinets are similar in design to DeLuxe type with hinges at back for full access to interior. Rear opening allows for necessary leads, etc. Slate grey ripple finish. **Prices do not include chassis.**

Cat. No.	Overall Dimensions H. L. D.	For Chassis	Net Price
CA-200	8½x10⅛x 8"	7x 7x2"	$3.60
CA-201	8½x12⅛x 8"	7x 9x2"	3.90
CA-202	8½x16⅛x 8"	7x13x2"	5.07
CA-203	9½x17⅛x11"	10x14x3"	7.89
CA-204	12½x20⅛x12"	10x17x3"	9.30

SLOPING FRONT CABINETS

Adaptable as instrument cases for studios, laboratories, etc. Top corner rounded and trimmed with chrome moulding. Slate grey ripple finish. A chassis may be mounted to front panel, and removed as a unit. Rear of case ventilated, with openings for connections. **Prices do not include chassis.**

Cat. No.	Overall Dimensions H. L. D.	For Chassis	Net Price
SF-500	8x 8x 8"	7x 7x2"	$3.84
SF-501	8x10x 8"	7x 9x2"	4.26
SF-502	8x14x 8"	7x13x2"	4.59
SF-503	9x18x 8"	7x17x3"	6.60
SF-504	12x18x12"	10x17x3"	8.40

TABLE TYPE RELAY RACKS

These racks are designed for table mounting, in place of heavy duty floor units. Constructed in one piece with mounting holes drilled on universal centers. Finished in black ripple and shipped "knocked-down," to fit 19" wide rack panels.

Cat. No.	H.	W.	D.	Panel Space	Net Price
TR-2520	25"	21"	12"	21x19"	$6.60
TR-3220	32"	21"	12"	28x19"	8.25

19" WIDE BLANK RACK PANELS

(⅛" Steel or Aluminum)

These panels have the standard 1¼" x ½" spaced slotted holes. Aluminum panels can also be supplied 3/16" thick at an additional cost of 60%. **Finish** is black ripple or slate grey ripple enamel.

Steel Cat. No. Black	Steel Cat. No. Grey	Net Price	Heights Steel or Alum.	Aluminum Cat. No. Black	Aluminum Cat. No. Grey	Net Price
6600	G-6600	$ .66	1¾'	6675	G-6675	$ .75
6601	G-6601	.75	3½"	6676	G-6676	1.08
6602	G-6602	.93	5¼"	6677	G-6677	1.59
6603	G-6603	1.08	7"	6678	G-6678	1.92
6604	G-6604	1.32	8¾"	6679	G-6679	2.31
6605	G-6605	1.59	10½"	6680	G-6680	2.91
6606	G-6606	1.89	12¼"	6681	G-6681	3.39
6607	G-6607	2.16	14"	6682	G-6682	3.90
6608	G-6608	2.46	15¾"	6683	G-6683	4.35
6609	G-6609	2.70	17½"	6684	G-6684	4.80
6610	G-6610	3.00	19¼"	6685	G-6685	5.25
6611	G-6611	3.30	21"	6686	G-6686	5.70

STANDARD DESK PANELS

Tables are rigidly made of 1/16" thick steel, and are securely mounted to regular ⅛" steel panels, which are 10½" x 19". All tables are 22" wide.

Black Enamel Cat. No.	Depth	Net Price	Grey Enamel Cat. No.	Depth	Net Price
BT-2220	20"	$13.80	GT-2220	20"	$15.90
BT-2216	16"	13.50	GT-2216	16"	14.40

DOOR PANELS

(Grille or Solid)
Grille Type Illustrated

These panels have flush hinged doors which are equipped with piano hinges, knob and concealed catch. All doors start 1" from top to allow space for chassis at bottom. Regular chassis brackets may be used. **Finish:** P — Black Ripple or G — Grey Ripple.

SOLID Cat. No.	SOLID Net Price	SOLID or GRILLE Panel Size	SOLID or GRILLE Door Size	GRILLE Cat. No.	GRILLE Net Price
P,G-670	$4.20	8¾"	4½x15⅞"	P,G-680	$5.40
P,G-671	4.50	10½"	6 x15⅞"	P,G-681	5.85
P,G-672	4.95	12¼"	7½x15⅞"	P,G-682	6.45

BLANK STEEL CHASSIS BASES

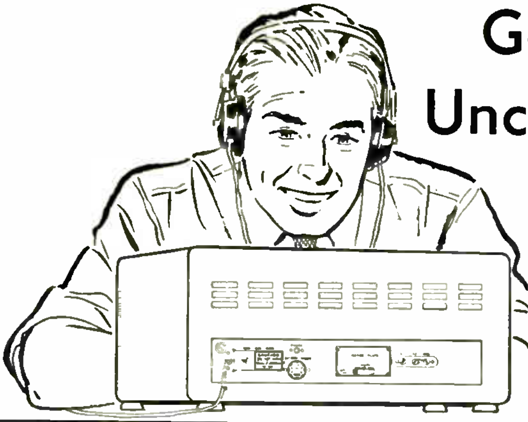
These bases are stamped from one piece of cold rolled steel. Has 4 solid sides with welded corners. Bottom edges are flanged in on 4 sides to provide additional reinforcement, and are drilled for bottom plates.

Chassis are made from No. 20 gauge steel except those marked (*) which are stamped from 1/16" steel.

Cat. No. Black Ripple	Cat. No. Zinc Plated	Size	Net Price	Cat. No. Black Ripple	Cat. No. Zinc Plated	Size	Net Price
B-4500	C-4500	5½x9½x1½"	$0.78	B-4525	C-4525	10x12x3"	$1.50
B-4507	C-4507	5x 7x2"	.78	B-4524	C-4524	10x14x3"	1.59
B-4508	C-4508	5x10x3"	1.08	B-4528	C-4528	10x17x2"	1.59
B-4509	C-4509	6x14x3"	1.26	B-4529	C-4529	10x17x4"	2.04
B-4510	C-4510	7x 7x2"	.90	B-4526	C-4526	10x17x3"	1.65
B-4511	C-4511	7x 9x2"	1.08	B-4527	C-4527	10x23x3"	2.04
B-4512	C-4512	7x11x2"	1.11	B-4533*	C-4533*	11x17x2"	2.25
B-4513	C-4513	7x13x2"	1.20	B-4534*	C-4534*	11x17x3"	2.46
B-4514	C-4514	7x15x3"	1.44	B-4516	C-4516	12x17x2"	1.71
B-4518	C-4518	4x17x3"	1.29	B-4517	C-4517	12x17x3"	2.04
B-4515	C-4515	7x17x3"	1.56	B-4530	C-4530	12x17x4"	2.25
B-4502	C-4502	8x12x3"	1.11	B-4535*	C-4535*	13x17x2"	2.46
B-4531	C-4531	8x17x2"	1.44	B-4536*	C-4536*	13x17x3"	2.85
B-4532	C-4532	8x17x3"	1.56	B-4537*	C-4537*	13x17x4"	3.25

PAR-METAL PRODUCTS CORPORATION
32-62—49th STREET . . . LONG ISLAND CITY 3, N. Y.

Export Dept.:
13 East 40 Street
New York 16



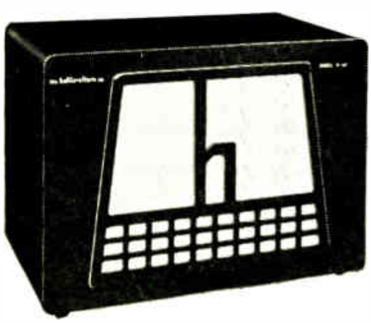
Get in touch with Uncle Dave, W2APF*

for

hallicrafters

Model S-40B—Covers Broadcast Band 540-1700 kc and three short-wave bands. 1.7-43 Mc. One r-f, two i-f stages. Electrical bandspread tuning. Switches for automatic noise limiter. code reception and three-position tone control. CW pitch control. Built-in PM speaker. 115 V, A.C. Only **$129.95**

TVI Suppressed 100 Watter—Model HT20
Here's the transmitter you've been waiting for! Continuous coverage from 1.7 Mc to 30 Mc. Full band switching, no more plug-in coils; choice of 10 crystals. Shielded, filtered r-f compartments plus low-pass 52 ohm co-axial line output filter assures at least 90 db suppression of all harmonics above 40 Mc. Only **$449.50**



Matched Speaker, Model R46—The perfect speaker for SW. Includes transformer of 500/600-ohm input. Voice coil impedance 3.2 ohms. 10" cone. Black finish. 15" x 10⅞" x 10⅞" deep. Only **$24.95**

He can help you pick the right gear. On the air for over a quarter of a century is your guarantee that Uncle Dave knows the Ham and his needs from a practical standpoint.

Fort Orange
RADIO DISTRIBUTING COMPANY
904 BROADWAY, ALBANY, N. Y.
TELEPHONE ALBANY 5-1594

Sooner or later you'll want COLLINS amateur radio equipment

KW-1 TRANSMITTER—This kilowatt rig operates on phone or CW in the 160, 80, 40, 20, 15, 11, and 10 meter bands but tunes as easily as a 32V-3. Just pick your band, set the built-in VFO to the desired spot, and adjust the final amplifier for 1000 watts input. That's all there is to it! The KW-1 design reduces spurious radiation to a very low value, particularly at television frequencies.

KW-1 Transmitter. Net Domestic Price **$3,850.00**

35C-2 LOW-PASS RF FILTER—Although designed as an accessory for the 32V-3 and as standard equipment in the KW-1, coaxial fittings make the 35C-2 readily adaptable to any amateur transmitter having 52 ohm output. It has an insertion loss of only 0.25 db at frequencies below 30 mc, but its three individually shielded filter sections provide about 75 db attenuation at television frequencies.

35C-2 Low-Pass RF Filter. Net Domestic Price . . **$40.00**

75A-3 RECEIVER—The 75A-3 is a double conversion superheterodyne designed for top performance in the 160, 80, 40, 20, 15, 11, and 10 meter amateur bands. A crystal controlled high frequency oscillator and highly stable low frequency VFO permit accurate calibration of two kilocycles per dial division on 10 and 11 meters and one kilocycle per dial division on all other bands. The Collins mechanical filter gives the 75A-3 the most nearly optimum selectivity ever achieved in a communications receiver.

75A-3 Receiver (includes 3 kc Mechanical Filter). Net Domestic Price **$530.00**

Fort Orange RADIO DISTRIBUTING COMPANY

904 BROADWAY, ALBANY, N. Y.

TELEPHONE ALBANY 5-1594

WHY THE EXCITEMENT ABOUT SINGLE SIDEBAND?

by Bill Cummings, W1RMG

When I heard the first stories about Single Sideband transmission, I found it hard to believe the "miracle" promises. Imagine 10 watts on SSB performing better than 1 KW on regular AM!

The stories seemed to be backed by good authority, so we decided to go into it all the way here at Dale. During years of working the Ham circuit, I've tried all the new gimmicks, so I don't startle easily . . . but SSB gave me the same kind of excitement I experienced when I first became a Ham. Here's an example of what happened:

On one occasion, I caught a signal from a KW transmitter in New Mexico on AM. He said he would switch right over to SSB on his 10-watter using the same antenna. There was a sharp improvement in readability, and I copied him with the greatest ease. I've worked the whole route from the W1's through the W0's with my own 10-watter.

Amazing? No, it's just SSB. The only amazing thing about it is you can buy that kind of performance for a small fraction of the price of AM gear.

Actually, no AM set can "match" the job SSB does in cutting through QRM in the noisy 75 meter band. SSB takes less space on the spectrum, knifes through sharply, and permits higher selectivity in your receiver. The Dale ad adjoining this column tells you more about specific advantages and the best kind of equipment for SSB operation.

If you'd like to check those big claims for SSB, how about a simple test? Write me and we'll arrange a schedule so you can hear for yourself. I'll send you tuning instructions too. Send along any questions — I'll be glad to answer them. Just drop me a note c/o Dale Electronic Distributors Inc. in New Haven.

Write to Bill Cummings at Dale for trade-in allowances and easy terms.

DALE takes the

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MODEL SX-88 RECEIVER

$499.95

Hallicrafters' finest receiver for amateur use, with highest degree of usable selectivity. Full coverage in six bands, from 535 kc to 33.3 Mc. Double conversion superhet over entire frequency range. Two tuned r-f stages on bands 2-6, one on broadcast. IF's 2075 kc and 50 kc (bands 1-6), 1550 kc and 50 kc (band 2). AM, CW, SSSC signals. Sensitivity on bands 1-6: 1 microvolt for ½ w output. 1 microvolt for 10 db signal-to-noise ratio. Band 6: 10 microvolts for ½ w output. Image rejection not less than 80 db on frequencies lower than 20 Mc and not less than 60 db from 20-30 Mc. Seventeen tubes plus voltage regulator, ballast tube, and rectifier.

SELECTIVITY

10 kc	6 db — 10 kc	60 db — 30 kc
5 kc	6 db — 5 kc	60 db — 15 kc
2.5 kc	6 db — 2.5 kc	60 db — 7.5 kc
1.25 kc	6 db — 1.25 kc	60 db — 3.75 kc
.500 kc	6 db — 500 cps	60 db — 1.50 kc
.250 kc	6 db — 250 cps	60 db — 850 cps

hallicrafters

$249.95

MODEL SX-71 RECEIVER

560-1600 kc broadcast plus four bands covering 1650 kc — 34 Mc and 46 — 56 Mc. Intermediate bands 1 and 2: 455 kc; bands 3, 4, 5: 2075 Mc and 455 kc. Double conversion superhet circuit. Eleven tubes plus rectifier and regulator.

MODEL S-76 RECEIVER

$199.95

Standard broadcast plus three bands covering 1720 kc — 34 Mc. Double conversion superhet over entire range. Intermediate frequencies: 1650 kc and 50 kc. Nine tubes plus rectifier and voltage regulator.

SELECTIVITY

1: (Broad)	6 db-5 kc	60 db-15 kc
2:	6 db-3 kc	60 db-12 kc
3:	6 db-2 kc	60 db-10 kc
4:	6 db-1.3 kc	60 db- 7 kc
5: (Sharp)	6 db- .5 kc	60 db- 5 kc

DALE ELECTRONIC DISTRIBUTORS INC. • 150 JAMES STREET

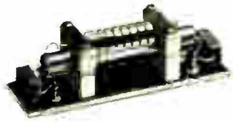
lead with Single Sideband!

Advantages of Single Sideband Transmission

Here are some reasons why SSB offers an exciting new kind of performance at a fraction of AM cost:

1. With Single Sideband, your final amplifier tubes can provide a 9 db gain, or a power increase of 8 times over conventional double sideband AM.
2. Your final amplifier delivers TALK POWER, not wasteful carrier.
3. No high power modulator and modulator power supply required.
4. The newer selective receivers and receiving adapters will pass only one sideband — in this case half of your sideband power is wasted with double sideband.
5. SSB eliminates the heterodyning carriers that plague the overcrowded phone bands.
6. Less spectrum space occupied by phone signals using SSB.
7. Take full advantage of SSB technique by switching sidebands to avoid QRM.
8. Round-table operation of two independent QSOs on the same suppressed carrier frequency using opposite sidebands.
9. Distortion due to selective fading eliminated.
10. Harmonic TVI virtually eliminated through the use of linear amplifiers.

COLLINS



MODEL 75A-3 RECEIVER

with built-in mechanical filter

The nearly flat top and sharp cutoff at the sides of the selectivity curve of the 3 kc Mechanical Filter permit all AM signals to be tuned so as to accept the carrier and either one of the sidebands at will, while other sideband is rejected.

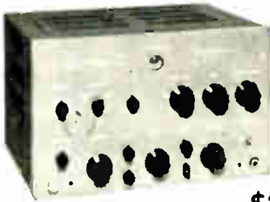
Alternatively, both AM and SSSC signals may be received with carrier supplied by the BFO; and the ideal selectivity curve of the Mechanical Filter permits full advantage to be taken of the benefits of local carrier reinsertion.

75A-3 receiver	$ 530.00
Matching speaker	20.00
32V3 transmitter	775.00
KW-1 transmitter	3,850.00
70E8A-VFO	97.50
35C2 lo-pass filter	40.00

COLLINS MECHANICAL FILTERS

F-455B-31	3.1 KC plug in	55.00
F-455B-08	800 cycle plug in	55.00
F-455A-31	3.1 KC solder term	55.00
F-455A-08	800 cycle solder term	55.00

CENTRAL ELECTRONICS NEW!



$249.50

MODEL 20A MULTIPHASE SSB EXCITER

Also rack-mounted in grey or black $7.50 additional

20 peak watts output — SSB, AM, PM, CW. Bandswitched 160-10 meters. Magic eye carrier null and modulation peak indicator. *New! Carrier level control*: Separate knob inserts any amount of carrier without disturbing carrier suppression adjustments. *New! Calibrate circuit*: Simply talk yourself exactly on frequency as you set VFO. *New! Calibrate level control*: Adjusts signal strength to suit band conditions. *New! Phone patch input jack*: Inserting phone plug disconnects speaker. PLUS all the time-proven features of popular Model 10A.

***Available soon!* Model V-10 all-band VFO for model 10A, 20A, or your present rig.**

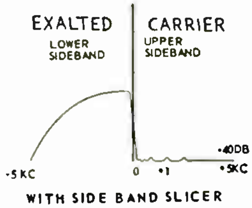
MODEL 10A MULTIPHASE EXCITER

$159.50

10 watts peak output — SSB, AM, PM, CW. Multiband operation 10-160 meters. Reduced harmonic TVI. Voice operated break-in. CW break-in. Switchable single sideband, with or without carrier. Double sideband AM. Narrow band phase modulation.

SIDEBAND SLICER

$74.50



WITH SIDE BAND SLICER

An adapter that will improve *any* receiver. Utilizes the phase shift principle to provide selectable single sideband reception of SSB, AM, PM and CW signals. Reduces the unwanted sideband an average of 40 db, a power ratio of 10,000 -- 1, or approximately 6 to 8 S-units on the average receiver. Improves SSB and CW reception, since the noise bandwidth is cut in half. May be used to check the sideband rejection of SSB signals. *Simple connection to any receiver.*

DALE

ELECTRONIC DISTRIBUTORS

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everybody agrees . . .

by every test ATR is BEST!

AUTO RADIO VIBRATORS

Have Ceramic Stack Spacers

A COMPLETE LINE OF VIBRATORS Designed for Use in Standard Vibrator-Operated Auto Radio Receivers. Built with Precision Construction, featuring Ceramic Stack Spacers for Longer Lasting Life. Backed by more than 22 years of experience in Vibrator Design, Development, and Manufacturing.

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DICTATE IN YOUR CAR with ATR INVERTERS

For changing your battery current to A. C. *Household* ELECTRICITY *Anywhere* IN YOUR OWN CAR!!

ATR INVERTERS . . . especially designed for operating standard 110 volt A.C. Tape Recorders, Wire Recorders, Dictating Machines and Electric Razors IN YOUR CAR.

There is an ATR INVERTER for most applications for inverting D.C. Voltages (ranging from 6 volts D.C. to 220 volts D.C.) to 110 volts A.C. . . . Specially Designed for operating A.C. Radios, Tape Recorders, Wire Recorders, Dictating Machines, Electric Razors, Record-Changers, Television Sets, Amplifiers, Address Systems, Radio Test Equipment and most small electrical and electronic devices FROM D.C. VOLTAGES IN AUTOMOBILES, BUSES, TRUCKS, SHIPS, TRAINS, PLANES AND IN D.C. DISTRICTS. Additional information available upon request.

Note Carefully: Inverters not recommended for operating refrigerators, washing machines or other similar motor-driven appliances, nor for operating toasters, electric irons, sun lamps, or similar appliances of high wattage or low power factor.

A VALUABLE TIME SAVER FOR:

- TRAVELING SALESMEN—Dictate reports in your own car. Send your dictated reports in daily to your home office or sales manager.
- EXECUTIVES—Dictate in your car all business matters while on trips for pleasure or business.
- FISHERMEN & HUNTERS—Use your electric razor on camping trips, operating in your car. Also small home radios and other electrical or electronic items.
- CAMPERS—Make your camping and outing trips more exciting using mixmasters, tape recorders, or wire recorders operating from your car battery.

ATR products are TOPS!

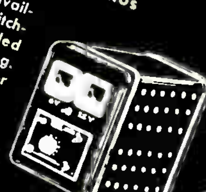
ATR VIBRATOR PACKS

for changing low voltage D.C. to high voltage D.C.

ATR "A" BATTERY ELIMINATORS

for demonstrating & testing Auto Radios

6 VOLT OR 12 VOLT! 6 or 12 volt D. C. combination output available by means of simple toggle switching arrangement, properly safeguarded to prevent accidental switching. NEW MODELS . . . Designed for testing D. C. Electrical Apparatus on Regular A. C. lines. Equipped with Full-Wave Dry Disc Type Rectifier, Assuring Noise-less, Interference-Free Operation and Extreme Long Life and Reliability.



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C.U. on 10-20-40 & 75 Meters

Now it is easier than ever to buy from WRL. Pay as little as 10% down and the balance in 18 months. No red tape — we handle our own financing. We will give you a generous trade-in allowance on your present equipment. We have over 600 reconditioned used items — with new factory guarantee. Send for free list. You will profit most when you deal with WRL — THE WORLD'S MOST PERSONALIZED RADIO SUPPLY HOUSE and THE WORLD'S LARGEST DISTRIBUTOR OF AMATEUR RADIO TRANSMITTING EQUIPMENT.

The NEW MODEL 40A

Complete Band Switching

GLOBE SCOUT

(50 Watts CW—40 Watts Phone)

$10. cash down
12 monthly payments of $8.95

KIT FORM **$89.95** WIRED **$99.95**

LATEST TRIUMPH OF THE WRL ENGINEERING STAFF
Complete bandswitching 160 thru 10M transmitter with combination pi-network antenna tuner which will work into any antenna. Three-stage modulator allows 100% modulation of final. Has complete power supply. XMTR housed in new special grey TVI screened cabinet. Ideal XMTR for the novice or experienced ham.

"It Is A Pleasure To Do Business With WRL"

Dear Leo:

I placed my Globe King in operation several months ago and am MORE than pleased with the results. For rugged dependability, an outstanding signal, and watt-per-dollar value, I do not believe it can be beat. Although I live, literally, in a forest of TV antennas, I have never received a complaint.

GEORGE FAIRBANKS, W1FWX
Seymour, Connecticut

NEW TVI MODEL 400 C

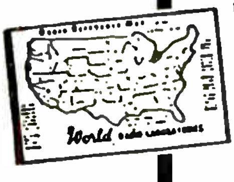
GLOBE KING XMTR

TVI SCREENED CABINET NOW STANDARD EQUIPMENT

$51.50 Cash Down

$495.00 KIT FORM
$515.00 WIRED

12 Monthly Payments of $40.95
18 Monthly Payments of $28.07



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25¢
HANDY WALL SIZE

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"EL" XENON GAS-FILLED TUBES

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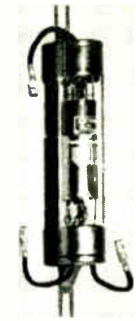


FULL WAVE RECTIFIER
EL 1C

D.C. Output (Amps.) ..	1.0
Peak Anode Current ..	4.0
Peak Inverse Volts	725
Filament Volts	2.5
Filament Amperes	6.0
Overall Length	5½″

FULL WAVE RECTIFIER
EL 3C

D.C. Output (Amps.) ..	2.5
Peak Anode Current ..	10.0
Peak Inverse Volts	725
Filament Volts	2.5
Filament Amperes	11.5
Overall Length	7″

FULL WAVE RECTIFIER
EL 6C

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	25.6
Peak Inverse Volts	725
Filament Volts	2.5
Filament Amperes	17.0
Overall Length	7½″

HALF WAVE RECTIFIER
EL 6B & EL 6F

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	40.0
Peak Inverse Volts	920
Filament Volts	2.5
Filament Amperes	21
Overall Length (6B) ..	9″
Overall Length (6F) ..	8¼″

(Panel Mounting)

HALF WAVE RECTIFIER
EL 16F

D.C. Output (Amps.) ..	16.0
Peak Anode Current ..	96.0
Peak Inverse Volts	620
Filament Volts	2.5
Filament Amperes	36
Overall Length	15⅝″

(Panel Mounting)

GRID CONTROL RECTIFIERS (THYRATRONS)





EL C1K

D.C. Output (Amps.) ..	1.0
Peak Anode Current ..	8.0
Peak Forward Volts	1000
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	6.3
Overall Length	4¼″

EL C3J

D.C. Output (Amps.) ..	2.5
Peak Anode Current ..	30.0
Peak Forward Volts	750
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	9.0
Overall Length	6⅛″

EL C6J

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	77.0
Peak Forward Volts	750
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	21.0
Overall Length	9″

EL C16J

D.C. Output (Amps.) ..	16.0
Peak Anode Current	160.0
Peak Forward Volts	1000
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	31.0
Overall Length	10″

(Panel Mounting)

EL C6C

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	77.0
Peak Forward Volts	2000
Peak Inverse Volts	4000
Filament Volts	2.5
Filament Amperes	24.0
Overall Length	11″

EL C1B/A

D.C. Output (Amps.) ..	1.0
Peak Anode Current ..	8.0
Peak Forward Volts	750
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	6.3
Overall Length	4½″

EL C3J/A

D.C. Output (Amps.) ..	2.5
Peak Anode Current ..	30.0
Peak Forward Volts	1000
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	9.0
Overall Length	6⅛″

EL C6J/A

D.C. Output (Amps.) ..	6.4
Peak Anode Current ..	77.0
Peak Forward Volts	1000
Peak Inverse Volts	1250
Filament Volts	2.5
Filament Amperes	21.0
Overall Length	9″

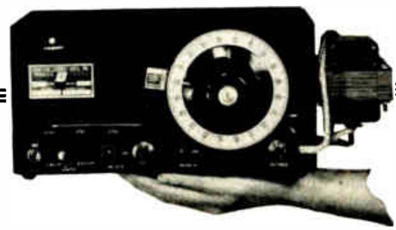
ENGINEERING MANUAL & CATALOG WILL BE SENT AT YOUR REQUEST

ELECTRONS, INCORPORATED
127 SUSSEX AVENUE
NEWARK 3, N. J.

Why LAMPKIN METERS

are the PERFECT PAIR for Mobile-Service Measurements

- LOWEST COST . . . for multiple channels. No additional crystals needed for added frequencies.
- VERSATILE . . . measure unlimited number of channels.
- PORTABLE . . . weigh less than 14 pounds apiece.
- LAMPKIN REPUTATION . . . over 60% of sales are repeat orders . . . satisfaction guaranteed or your money refunded.



LAMPKIN 205 FM MODULATION METER

FREQUENCY RANGE—continuous 25 mc. to 200 mc. No coils to change. Rough and vernier tuning control. MEETS FCC SPECS—reads peak voice deviation directly, 0–25 kc., positive or negative. No charts or tables. ACCURATE—within 10% at full scale. FIELD STRENGTH METER—reads relative transmitter output. PROTECTED—Panel components recessed behind edges of the case. PORTABLE—just a 2-finger load.

LAMPKIN 105-B MICROMETER FREQUENCY METER

FREQUENCY RANGE—continuous 100 kc. to 175 mc. No extra crystals needed. MEETS FCC SPECS—Accuracy guaranteed better than 0.0025%. CALIBRATION—Calibration Table for each meter. Charts show percentage transmitter off frequency from FCC assignment. DIAL—4″ circular dial with 8000 dial divisions spread over 42 feet. TEMPERATURE COMPENSATION—crystal thermometer on panel, automatically indicates dial checkpoint for any temperature. SIGNAL GENERATOR—highly accurate CW source for mobile-receiver final alignment.

Lampkin meters are used by many municipalities . . . including New York, Chicago, and Los Angeles—by agencies of 37 states—by the service organizations of most two-way radio manufacturers—and by hundreds of independent mobile-service engineers. They are guaranteed to please you, too, or your money will be refunded.

For price and complete technical information, mail coupon today!

LAMPKIN LABORATORIES INC.

Bradenton, Florida

Measurements Division
LAMPKIN LABORATORIES, INC.
Bradenton, Florida

With no obligation to me, please send information and prices on Lampkin meters.

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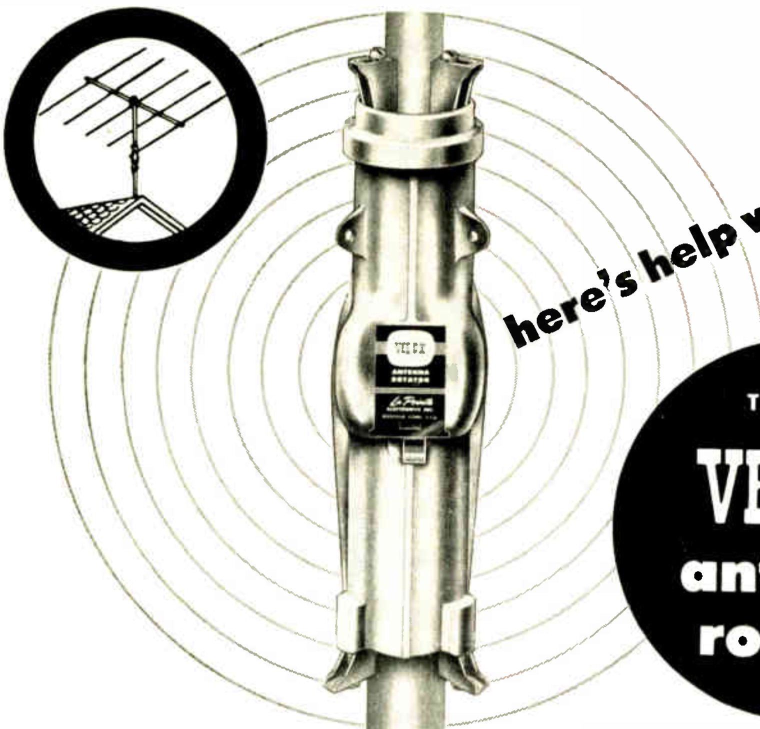

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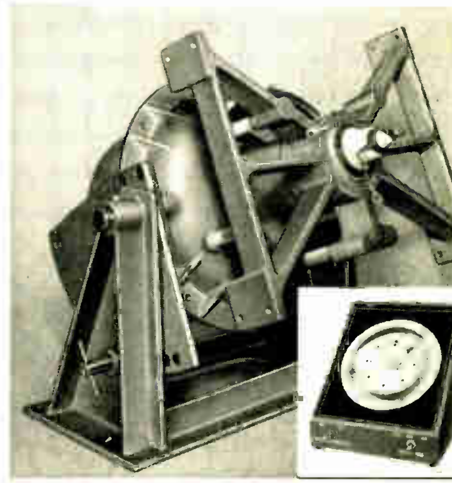

TRYLON 2-BAND ROTARY BEAM: This highly directional, high-gain antenna covers *any two* bands between 12 and 50 mc. Each array is available with either 3 or 4 elements consisting of adjustable telescoping aluminum tubes, plus adjustable stubs and "T" matches. This allows precise adjustment for maximum radiation efficiency in each band and insures an impedance match for all transmission lines.

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Typical Towers Illustrated at Right

Fig. 1	*Fig. 2*	*Fig. 3*
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Face width—6.5 in.	Face width—24⅝ in.	Face width—60 in.
Wt. per 10-ft. sec.—22 lbs.	Wt. per 10-ft. sec.—112 lbs.	Wt. per 10-ft. sec.—653 lbs.
Use—Mast for TV, Amateur, portable, and wire-type antennas.	*Use*—Tower for Trylon Rotary Beam, AM Broadcast, and Microwave antennas.	*Use*—TV Broadcasting and Curtain antennas for International Broadcasting.

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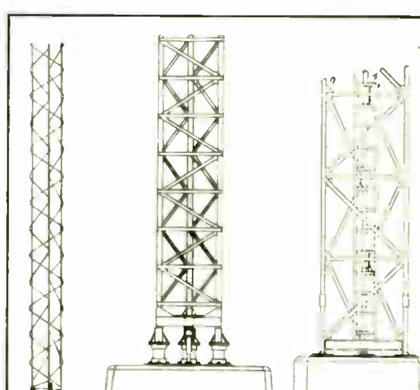

Fig. 1 Fig. 2 Fig. 3

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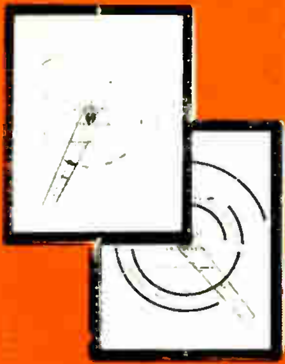

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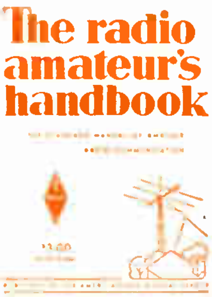

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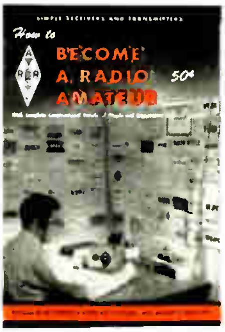

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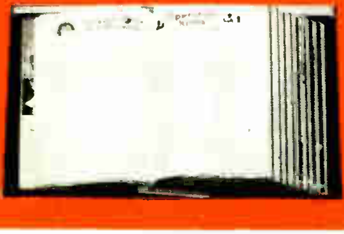

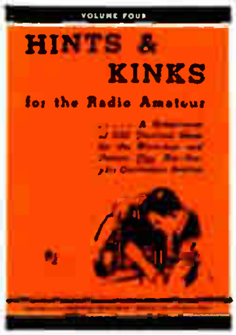

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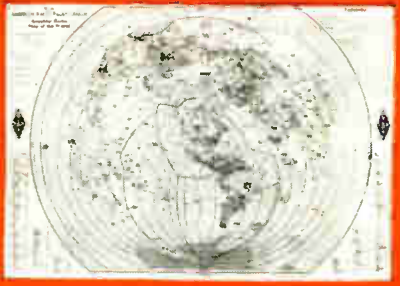

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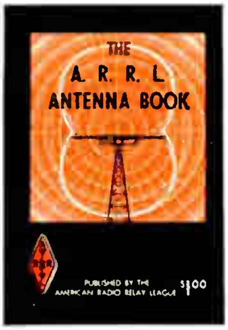

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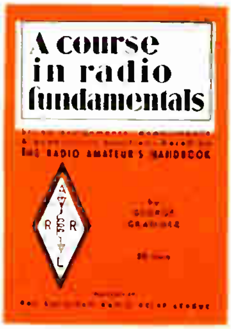

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15-25 Mc.; 195-225 Mc. 15-25 Mc.; 90-125 Mc. Other ranges on order	1 to 100,000 microvolts	AM 8200 400 cycles 625—400 cycles Fixed at approximately 30%

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86 Mc.—108 Mc.	1 to 100,000 microvolts	Deviation 0-300 Kc. 2 ranges FM 400-8200 cycles External mod. to 15 Kc.

STANDARD SIGNAL GENERATOR MODEL 80

FREQUENCY RANGE	OUTPUT RANGE	MODULATION
2 Mc.—400 Mc.	0.1 to 100,000 microvolts	AM. 0 to 30% 400 cycles or 1000 cycles External mod., 50-10,000 cycles

STANDARD SIGNAL GENERATOR MODEL 82

FREQUENCY RANGE	OUTPUT RANGE	MODULATION
20 cycles to 200 Kc. 80 Kc. to 50 Mc.	0-50 volts 0.1 microvolt to 1 volt	Continuously variable 0-50% from 20 cycles to 20 Kc.

STANDARD SIGNAL GENERATOR MODEL 84

FREQUENCY RANGE	OUTPUT RANGE	MODULATION
300 Mc.—1000 Mc.	0.1 to 100,000 microvolts	AM. 0 to 30%, 400, 1000, or 2500 cycles. Internal pulse modulator External mod., 50-30,000 cycles.

STANDARD SIGNAL GENERATOR MODEL 84-TV

FREQUENCY RANGE	OUTPUT RANGE	MODULATION
300 Mc. to 1000 Mc.	Continuously variable from 0.1 microvolt to 1.0 volt	Continuously variable 0 to 30% External modulation 20 to 20,000 cycles

STANDARD SIGNAL GENERATOR MODEL 90

FREQUENCY RANGE	OUTPUT RANGE	MODULATION
20 Mc.—250 Mc.	0.3 microvolt to 0.1 volt	Continuously variable, 0 to 100% Sinusoidal modulation 30 cycles 5 mc. Composite TV modulation.

PULSE GENERATOR MODEL 79-B

FREQUENCY RANGE	PULSE WIDTH	OUTPUT
60 to 100,000 cycles	Continuously variable from 0.5 to 40 microseconds	Approx. 150 v. positive with respect to ground. "Sync Output" 75 v. positive with respect to ground.

SQUARE WAVE GENERATOR MODEL 71

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Continuously variable 6 to 100,000 cycles	Rise time less than 0.2 microseconds with negligible overshoot	Step attenuator 75, 50, 25, 15, 10, 5 peak volts fixed and 0 to 2.5 volts continuously variable.

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MODEL 62

VOLTAGE RANGE	FREQUENCY RANGE	INPUT IMPEDANCE
0-1, 0-3, 0-30 and 0-100 volts AC or DC	30 cycles to over 150 Mc.	Approximately 7 mmfd.

MODEL 67

VOLTAGE RANGE	FREQUENCY RANGE	INPUT IMPEDANCE
.0005 to 300 volts peak-to-peak	5 to 100,000 sine-wave cycles per second	1 megohm shunted by 30 mmfd.

CRYSTAL CALIBRATOR MODEL 111

STANDARD SIGNAL GENERATOR MODEL 80

MEGACYCLE METER MODEL 59

U.H.F. FIELD STRENGTH METER MODEL 58

FREQUENCY RANGE	INPUT VOLTAGE RANGE
15 Mc. to 150 Mc.	1 to 100,000 microvolts in antenna. 1 to 100 microvolts on semi-logarithmic output meter, balanced resistance attenuator with ratios of 10, 100 and 1000 ahead of all tubes.

INTERMODULATION METER MODEL 31

INTERMODULATION RANGE	FREQUENCIES (CYCLES)	ANALYZER INPUT VOLTAGES
0.5% to 30%	LF: 60 cps HF: 3000 cps	Full scale ranges of 3, 10, 30 volts RMS

MEGACYCLE METER MODEL 59

FREQUENCY RANGE	FREQUENCY ACCURACY	MODULATION
2.2 Mc. to 400 Mc.	Within ±2%	CW or 120 cycles fixed at approximately 30%. Provision for external modulation

CRYSTAL CALIBRATORS

MODEL 111

FREQUENCY RANGE	FREQUENCY ACCURACY	HARMONIC RANGE
250 Kc.—1000 Mc.	0.001%	.25 Mc. Oscillator: .25-450 Mc. 1 Mc. Oscillator: 1-600 Mc. 10 Mc. Oscillator: 10-1000 Mc.

MODEL 111-B

FREQUENCY RANGE	FREQUENCY ACCURACY	HARMONIC RANGE
100 Kc.—1000 Mc.	0.001%	.1 Mc. Oscillator: .1 — 450 Mc. 1 Mc. Oscillator: 1 — 600 Mc. 10 Mc. Oscillator: 10— 1000 Mc.

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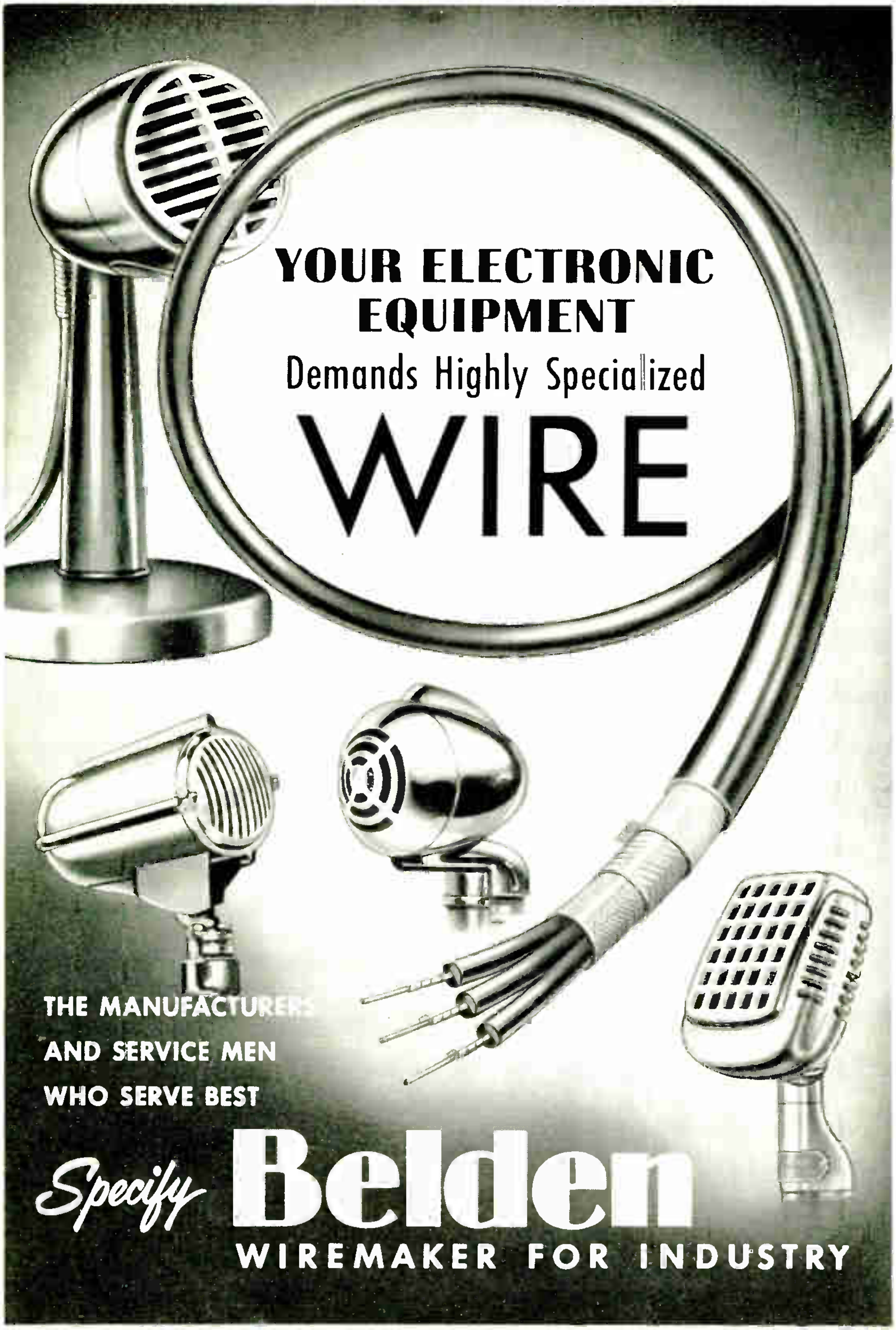
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Small size, high output, single-button carbon microphone for maximum intelligibility. Close-talking, noise-cancelling Differential* type. High articulation. Blast proof, water proof, shock resistant. Comfortably hand-held. Press-to-talk switch. Panel mounting bracket. List, **$16.50**

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Model 428 "Break-in" Touch-to-Talk Stand with locking feature. Fits any microphone with standard ⅝"-27 thread. Lever-type switch gives finger-tip relay operation or microphone "On-Off." Single-pole double-throw. List, **$16.00**

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TV ACCESSORIES • PA PROJECTORS

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VOLT-OHM-MIL-AMMETERS

for your every need

★ Note . . . These Sensational Improvements

TRIPLETT Model 630

- ★ Large 5½ Inch Meter in special molded case under panel.
- ★ Resistance Scale Markings from .2 Ohms to 100 Megohms. Zero Ohms control flush with panel.
- ★ Only one switch. Has extra large knob 2½" long, easy to turn, flush with panel surface.
- ★ New molded selector switch, contacts are fully enclosed.
- ★ All resistors are precision film or wire wound types, all sealed in molded compartment.
- ★ Unit construction—Resistors, Shunts, Rectifier, Batteries all are housed in a molded base built right over the switch. Provides direct connections without cabling. No chance for shorts.
- ★ Batteries easily replaced—New Double Suspended Contacts.
- ★ Ranges: DC Volts 0 to 6000, 20,000 Ohm Volt; AC Volts 0 to 6000, 5000 Ohm Volt; DB: — 30 to +70; Direct Current from 0–60 Microamp. to 0–12 Amps; Resistance: 0–1000–10,000 Ohms, 0–1–100 Megs.

ABSORPTION FREQUENCY METER

Model 3256

A band-switching, tuned absorption type frequency meter that covers five amateur bands. Has Germanium crystal and a DC Milliammeter indicator for greater sensitivity. Direct calibration on panel — no coils to change. Switching permits instantaneous band change. Audio jack provides for monitoring of phone signals — another new feature. Calibration is in Megacycles in following bands: 3.5–4 MC; 7–7.3 MC; 14–14.4 MC; 20–21.5 MC; 28–30 MC. Coil is removable and other coils may be substituted for special bands. Useful for checking: Fundamental frequency of oscillating circuits; Presence, order and relative amplitude of harmonics; Parasitic oscillations, etc. Size: 7½" x 2½" x 2¼". Metal case with gray enamel finish black trim.

POCKET SIZE Model 666-R

For your AC and DC Voltage, Direct Current and Resistance analyses to 3 Megohms. Enclosed selector switch of molded construction keeps dirt out. Retains contact alignment permanently. Unit Construction — All Resistors, shunts, rectifier and batteries housed in a molded base integral with the switch. Eliminates chance for shorts. Direct connections. No cabling. All precision film or wire-wound resistors, mounted in their own compartment—assures greater accuracy. 3" RED•DOT Lifetime Guaranteed instrument. Red and black markings on a white background. Easy to read scale. Precalibrated rectifier unit. Self-contained batteries.

RANGES: AC-DC Volts: 0–10–50–250–1000–5000, 1000 Ohms Volt; Direct Current: 0–10–100 Ma., 0–1 Amp.; Resistance: 0–3000–300,000 Ohms, 3 Meg. Black molded case, completely insulated, 3 ⁷⁄₁₆" x 5⅞" x 2 ⁵⁄₁₆". White panel markings.

Model 666-HH VOLT-OHM-MILLIAMMETER

A complete miniature laboratory for DC-AC Voltage, Direct Current and Resistance analyses. The answer to V-O-Ma. requirements of radio servicemen, amateurs, industrial engineers, etc. Greater scale readability on the 3" RED•DOT Lifetime guaranteed instrument with red and black scale markings. Simplified switching.

RANGES: AC-DC Volts: 0–10–50–250–1000–5000, 1000 Ohm Volt; DC Ma: 0–10–100–500; OHMS: 0–2000–400,000. Self-contained plug-in batteries. Black molded case, completely insulated, 3 ⁷⁄₁₆" x 5⅞" x 2 ⅛". Panel with white markings.

A COMPLETE LINE OF METERS

Triplett panel and portable meters are available in more than 26 case styles — round, square and fan — 2" to 7" sizes. Included are voltmeters, ammeters, milliammeters, millivoltmeters, microammeters, thermo-ammeters, DB meters, VU meters and electrodynamometer type instruments

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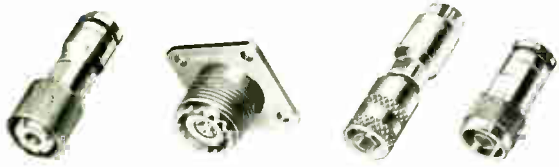


RF CONNECTORS

AMPHENOL RF connectors are unsurpassed for mechanical design and electrical efficiency. They provide low-loss continuity in critical RF circuits with little or no impedance change or increase in standing wave ratio.

AMPHENOL RF connectors are available in Types BNC, BN, HN, LC, N, C and the popular 83 series, including plugs, jacks, receptacles and adapters. All AMPHENOL RF connectors meet or surpass rigid government specifications.

"S" SOCKETS and "CP" PLUGS

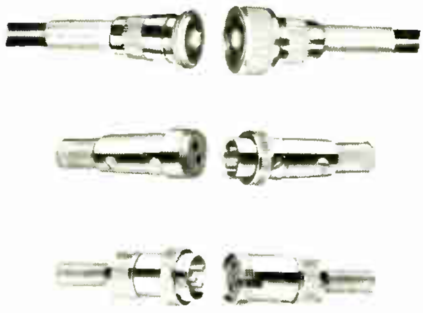
"S" Sockets and "CP" Plugs mate with each other and feature the AMPHENOL retainer ring design – mount without screws or rivets on panel or chassis. They are compact in size and are ruggedly built for trouble-free service. Available in black bakelite or mica-filled bakelite. "S" type sockets are also available in AMPHENOL Steatite and in a variety of sizes with the number of contacts ranging from 4 to 11. Supplied with retainer rings for chassis mounting. Plugs and sockets are also available with plates and caps.

STEATITE TRANSMITTING TYPE TUBE SOCKETS

Designed for use where other, less rugged, sockets cannot do the job. Made of low-loss AMPHENOL Steatite, they feature the "clover leaf" contacts that provide four full lines of contacts. Barriers provide long creepage paths that prevent arcing and flash-over. Available in various sizes with 4, 5, 6, 7 and 8 contacts.

MINIATURE 7 and 9 PIN SOCKETS

AMPHENOL has a complete line of miniature 7 and 9 pin sockets for every application. Materials include black bakelite, mica-filled bakelite, Steatite or AMPHENOL'S own Ethylon-A. Zip-in sockets are molded of Ethylon-A, a resilient dielectric, and need no mounting plate or retainer ring.



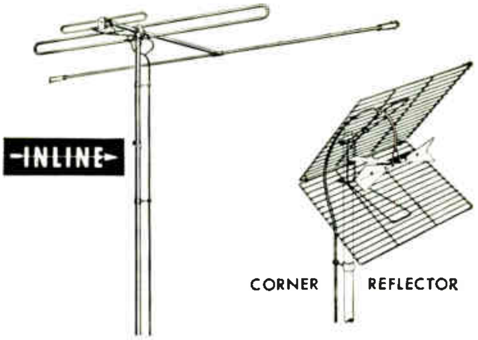
MICROPHONE CONNECTORS

AMPHENOL manufactures an extensive line of connectors to fit practically all makes of microphones. The 75-MC1F Microphone connectors, illustrated above, function as either male or female fitting, include jacks, plugs, receptacles, adapters and switches. The 80 Series, single and double contact, connectors are designed for shielded cables. Obtainable in any combination of male or female cable connectors or as chassis units. The 91 Series includes both three and four contact connectors, polarized to prevent incorrect insertion. They are procurable as plugs, cable jacks and chassis receptacles in any combination of male or female types.

"MIP" SOCKETS

The world's strongest socket! The plated steel mounting plate is molded right into the solid bakelite body, cannot come loose or vibrate. Two holes in each contact provide wiring and anchoring points for resistors, condensers, chokes, etc. Available in black bakelite or mica-filled bakelite in a wide variety of contact arrangements. "MIP" sockets are also available for 8 pin Octal and Loktal tubes. The "MIP" sockets are unequalled in either strength or versatility.

Quality AMPHENOL



VHF and UHF TELEVISION ANTENNAS

AMPHENOL manufactures a complete line of quality television antennas. For VHF, the famous INLINE (Reissue Pat. No. 23,273) design is available in either single bay or stacked array. AMPHENOL UHF antennas include BO-TY, Corner Reflector, Rhombic and Yagi types. The Stacked-V model will receive VHF, UHF or both.

Antennas are available for extended distances reception of either VHF or UHF. Stacking for greater db gain may be done with the INLINE, BO-TY, Corner Reflector and Yagi.

The materials used in all AMPHENOL antennas are the very finest available. Sturdy aluminum and steel with non-rusting finishes insure like-new appearance on the rooftop. All AMPHENOL antennas feature *stay-up* construction—and perform well for many years.

COAX and TWINAX CABLE

AMPHENOL cables are produced in strict conformity to the rigid military specifications. Constant checks and inspections are made to assure the best in mechanical and electrical construction.

Most of the RF cables in the AMPHENOL line have top grade polyethylene dielectric for low-loss, flexibility and mechanical stability. For high temperature applications, cables are also available with other types of dielectric.

FLAT and TUBULAR TWIN-LEAD

AMPHENOL twin-lead, flat or tubular, is made of the finest materials available, manufactured with constant and rigid inspection. The brown pigmented virgin polyethylene assures a minimum of signal loss and constant impedance.

AMPHENOL flat twin-lead is available in a variety of types and sizes. AIR-CORE Tubular twin-lead (U.S. Pat. No. 2,543,696) is a must for UHF television lead-in purposes.

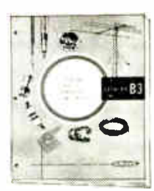
AMATEUR ANTENNA KIT

The AMPHENOL Amateur Antenna has been designed to meet your need for a simple, effective folded dipole antenna system. The efficiency of the AMPHENOL Antenna for both transmitting and receiving has been demonstrated by years of satisfied amateur use. The Amateur Antenna is available in an economical, easy-to-assemble kit form. All the kits are pre-cut to band length and are ready for final assembly and installation. Complete assembly instructions are included.

YOUR AMPHENOL DISTRIBUTOR
has what you need in radio and electronic components. Save time by seeing him for the part you want. All the components listed on these pages, and many others in addition, are carried in his stock and are immediately available.

NEW! CATALOG B-3

The new, revised AMPHENOL *general catalog B-3 will be sent upon request. It contains illustrations and specifications on the over 9,000 items now manufactured by the* AMERICAN PHENOLIC CORPORATION.



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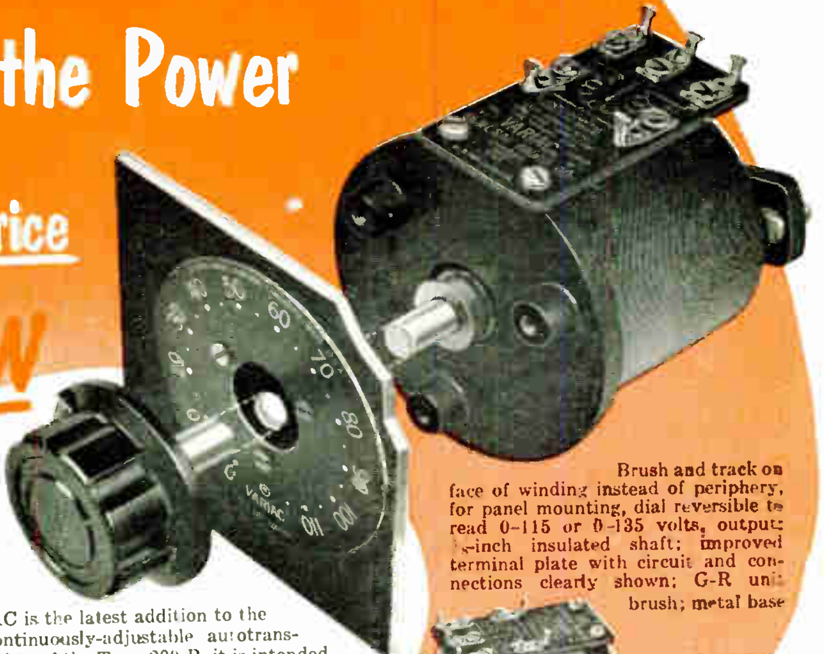
Twice the Power for the Same Price

with the NEW type V-2 Variac



Brush and track on face of winding instead of periphery, for panel mounting, dial reversible to read 0-115 or 0-135 volts, output; 3/8-inch insulated shaft; improved terminal plate with circuit and connections clearly shown; G-R unit brush; metal base

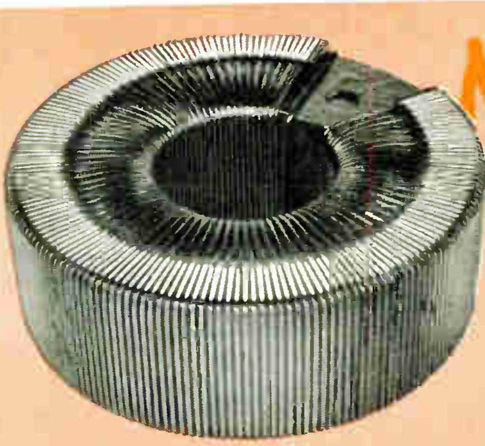
The *new* Type V-2 VARIAC is the latest addition to the popular "V" series of VARIAC continuously-adjustable autotransformers. With a power rating of *twice* that of the Type 200-B, it is intended eventually to supplant the latter.

The Type V-2 represents a considerable improvement over the 200-B, making it even more adaptable to a wide variety of applications where a relatively low-power, moderately-priced, a-c voltage control with all of the latest exclusive VARIAC features is needed. Its electrical and mechanical features include:

- ★ New DURATRAK brush-track coating
- ★ Load rating of 0.345 kva (instead of 0.17 kva)
- ★ Output voltages continuously adjustable from 0 to 135 v, or 0 to 115 v.
- ★ Line frequency 50 to 60 cycles; no load loss at 60 c.: less than 3.5 watts
- ★ Standard mounting holes (interchangeable with Type 200-B)
- ★ Considerably less panel space
- ★ Metal Base for improved cooling and added strength
- ★ More rugged stops
- ★ G-R Unit Brush construction
- ★ Terminal board with circuit and connections clearly shown
- ★ Molded knob-pointer and reversible dial plate for panel installation
- ★ Meets applicable requirements of Military Specification for Transformers: MIL-T-27
- ★ **Moderate Price: $12.50**

NOW ALL "V" TYPE VARIACS HAVE NEW Duratrak CONSTRUCTION

for Increased Performance Even Longer Life



The new DURATRAK coating is one of the most important improvements in VARIACS; another step in the continuing process of development which makes the VARIAC the best voltage control available

The new DURATRAK brush-track coating eliminates all brush track oxidization problems which formerly might become critical under severe conditions of operation.

DURATRAK construction adds these features to increase the utility of VARIACS:

Long Life — *insures a life as long as that of fixed-ratio power transformers operated at rated loads*

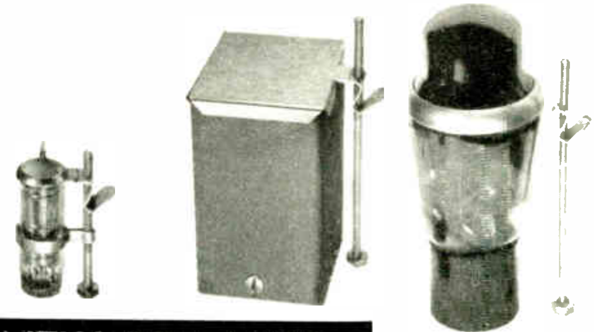
Unaffected By Surges — *VARIACS with DURATRAK will withstand initial surges as high as ten times rated current*

Minimum Maintenance — *the brush track will not deteriorate under normal operation — only maintenance needed is occasional wiping of track with alcohol-moistened rag*

Write for the NEW VARIAC BULLETIN for Complete Specifications

TOP HAT RETAINERS

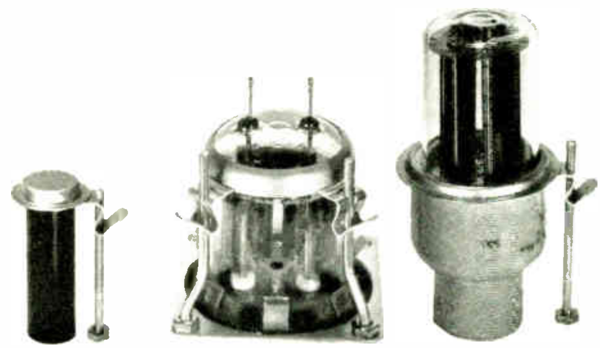
An improved clamp that holds electron tubes, relays, capacitors and pluggable components in position.



EASY TO APPLY...INSTANTLY RELEASED

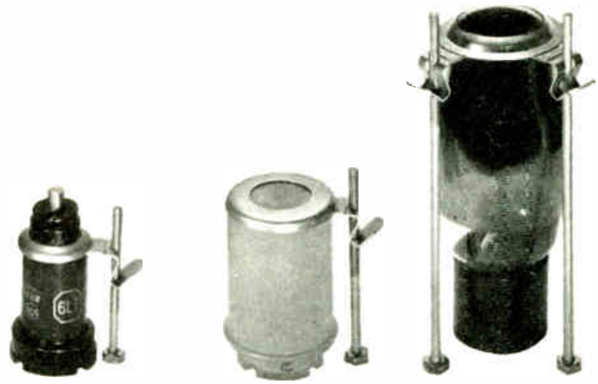
Made in a variety of sizes to fit almost any type tube.

Recommended for use in Army, Navy and Air Force electronic equipment.



POSITIVE LOCKING ACTION

POSITIVE—Component even when inverted will not loosen under the most severe shock or vibration. **RESILIENT**—Positive retention is achieved with minimum rigidity or strain on component. **ACCESSIBLE**—Clamp fits on top of component and can be fastened or released easily without tools. **CORROSION RESISTANT**—Hat and Posts are stainless steel. All materials and finishes comply with Armed Forces Specifications. **VERSATILE**—Top Hat Retainers may be applied to practically any tube or plug-in component.



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CANNON PLUGS

• **Highest Quality** • **Greatest Variety**— meet every need in radio, sound, TV and related fields • Standard on leading microphones

O SERIES

For microphone and related uses. 3 30a contacts only. Plug shown O3-12. Available in 6 basic shapes. Insert 1" wide. Zinc alloy shell, satin chrome finish. Molded phenolic insert. Latch lock coupling method.

P SERIES

For audio, TV and instrument uses. 2 to 8 contacts 30a max. Plug shown P3-CG-11. Available in 16 basic shapes. Insert Dia. 1". Zinc or steel shells, satin chrome finish. Molded phenolic insert. Latch lock coupling method.

X SERIES

For audio, instrument and related uses. 1 to 4 contacts 15a max. Receptacle shown X-3-14. Available in 5 basic shapes. Insert Dia. .625". Zinc alloy shell, bright nickel finish. Molded phenolic insert. Coupling held by contact friction.

XL SERIES

For audio, instrument and related uses. 3-15a contacts only. Plug shown XL-3-11. Available in 14 basic shapes. Insert dia. .625". Zinc or steel shells, bright nickel and satin chrome finish. Molded phenolic insert. Latch lock coupling method.

XK SERIES

For audio, instrument and related uses. 1 to 4 contacts. 15a max. Plug shown XK-3-11. Available in 4 basic shapes. Insert Dia. .625". Zinc or steel shells, bright nickel finish. Molded phenolic insert. Acme thread coupling nut.

UA SERIES

The ultimate audio quick disconnect to R.T.M.A. Specs. 3-15a contacts only. Plug shown UA-3-11. Available in 7 basic shapes. Insert Dia. .750". Zinc or steel shells, satin chrome finish. Insulation, phenolic with rubber seal. Latch lock coupling method.

XKW SERIES

Similar to XK Series, but weatherproofed by addition of a rubber bushing, special packing ring within the coupling nut and rubber sealing washers on the retaining screw. Plug shown XKW-3-12. Available in 4 basic shapes.

GB SERIES

For storage batteries, engine starters and other high current uses. 2—600a contacts. 1 contact to accommodate No. 8, 10 or 12 wire for signal or starter relay circuits. Molded rubber shell. Coupling maintained by contact friction. Plug shown is GB-3-21CFS.

XKW-B1 SERIES

For coaxial cable applications. One Standard Cannon type "R" coaxial contact only. The coax carries one 10a contact for No. 16 B & S wire. Shells, coupling ring and mounting flanges are same as XKW Series. Plug shown is XKW-B1-11.

D SERIES

For miniature circuitry, radio or audio. 15 to 50 contacts 5a Plug shown DA-15S. Insert is 1.049" long. Cold rolled steel shell Cadmium Iridite bleached. Inserts are molded Nylon. Coupling is held by contact friction.

U SERIES

For hermetically sealed instruments, indicators miniature switches, etc. 3 to 12 contacts 5a in three shell sizes. Shell is cold rolled steel. Inserts (min. dia. .294") are vitreous material and Silcan silicone rubber. Bayonet lock.

M1 SERIES

For heavy duty power, lighting and sound uses. 4—30a contacts only. Plug shown is M1-4-22. Available in 17 basic shapes. Insert Dia. 2.250". Aluminum alloy shells sand blasted. Molded phenolic inserts. Coupling held by contact friction.

K SERIES — TELEPHONE RECORDER

The receptacle SK-M7-32S (shown) and mating plug SK-M7-21C-1/2" are standard equipment for the recorder connectors used by Telephone Companies as subscribers' voice recorder.

K SERIES — TELEVISION

For TV cameras and cable. Coax contacts available. Insert Dia. 2.250". Plug shown LKT-R24C-22-7/8". Straight and 90° shells, ribbed coupling nut, gland nut, friction washer, bushing, gland washer and packing ring to support cable are features.

All connectors shown are stock items for Cannon Franchised Distributors. All other Cannon Plugs may be purchased from these same distributors by arrangement between the distributor and the customer. For further detail, request RJC Bulletin from factory.

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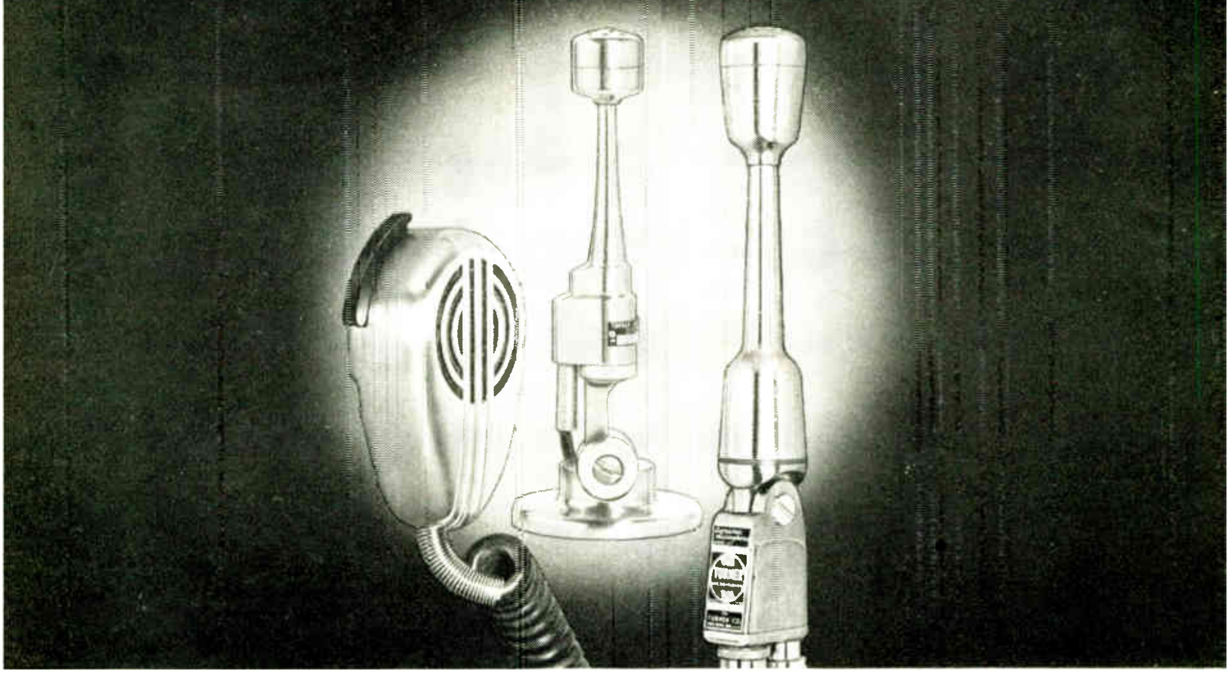
TURNER MICROPHONES



IN THE SPOTLIGHT FOR HAMS

TURNER Model 80 Crystal (center) A compact convenient microphone for increased rig efficiency. Excellent response to voice, 80 to 7000 cps. High quality Bimorph moisture sealed crystal, mechanical and shock proofed. Solidly built of die-cast zinc overlaid with beautiful satin chrome plate. 7-ft. attached, single conductor, shielded cable. Shown with Turner C-4 Stand. Standard 5/8"-27 thread.

Model 80 .... $15.95
C-4 Matching Stand 5.75

TURNER Model ADA 95D Dynamic (right) Newest Turner microphone. Superior sound characteristics make it ideal for amateur work. Maximum sensitivity to voice assured with Alnico V Magnets and moving coils. And this graceful beauty sells for less. Frequency response, 70 to 10,000 cps. Level, -58 db; standard 5/8"-27 coupler, 20 ft. removable cable set; high impedance wired single ended (single conductor shielded cable); 50, 200 or 500 ohms wired for balanced line (two conductor shielded cable).

ADA 95D Dynamic $35.00
ADA S95D, with slide switch 38.50

TURNER 90 Carbon Mobile Microphone (left) The answer to mobile amateur operations. Adds real eye appeal—gives superior sound performance — solidly built for lasting use. Choice of heavy duty push-to-talk switch, on-off relay circuit switch, or on-off relay switch. Equipped with 2 or 4 conductor cable, regular or Koiled Kord (5-ft. extended). Supplied with hook for hanging and bracket for wall or dashboard mounting. Response, 200 to 4000 cps; level, -42 db; 200 ohms nominal DC resistance.

Standard Model SR90R with 4 conductor unshielded Koiled Kord $26.50

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This engineering and manufacturing experience has enabled ILLINOIS CONDENSER COMPANY to meet every demand for new capacitor types, whether for peace or defense applications, with components of exceptional value and dependability.

Illustrated are but a few of the many capacitor types now manufactured by ILLINOIS CONDENSER. There is a guaranteed ILLINOIS Condenser for every electronic application. Whatever your requirement— specify ILLINOIS CONDENSER. Benefit by the years of engineering and manufacturing experience behind these condensers of "Time Tested Quality".

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IHC [illegible]

LN [illegible]

UMP [illegible]

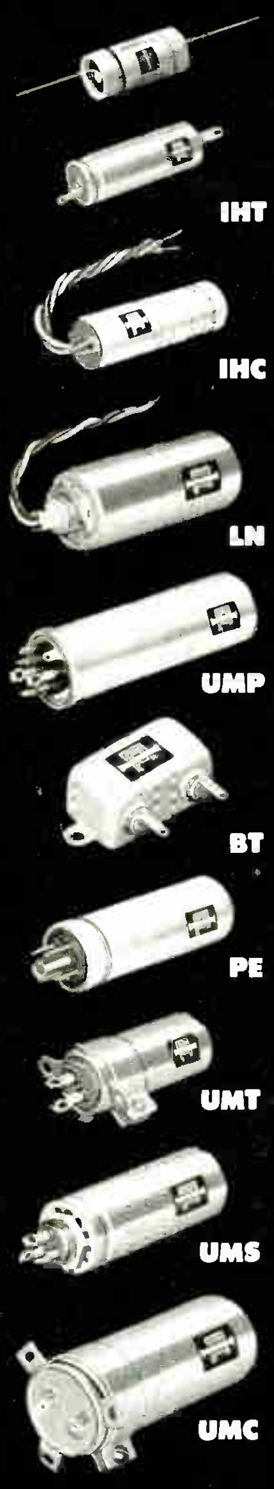
BT [illegible]

PE [illegible]

UMT [illegible]

UMS [illegible]

UMC [illegible]



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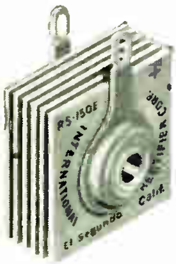


RADIO & TV STACKS

Half wave, full wave and voltage doubler units with input ratings of 25 to 165 volts AC. DC output current from 65 milliamperes to 1200 milliamperes. High quality selenium universal replacement Rectifiers for radio and TV sets. Also useful for amateur and experimental applications. Can be used in voltage multiplier circuits directly from line supply to provide a DC voltage output of any multiple of the input voltage. Universal mounting.

Technical information on request

(on your letterhead)

BOOSTER AND CONVERTER TYPES

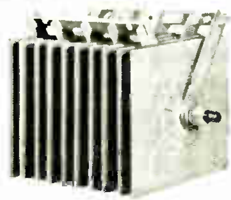
These miniature Selenium Rectifier units are ideal for replacement in TV Boosters, UHF Converters and Record Player Amplifiers. Small, compact. Provided with pigtail leads for easy wiring into crowded spaces. Available in half wave units with AC inputs of from 130 to 165 volts and DC current output of 20 to 50 milliamperes. These units are very useful for experimental devices, especially if the load current is 50 milliamperes or less.

HIGH VOLTAGE CARTRIDGES

Designed with ferrule terminals for quick insertion into standard 30-ampere fuse clips. Recommended for use in high voltage power supplies where long life and extreme reliability are of prime importance. DC output from 1,000 volts at 5 milliamperes to 3500 volts at 5 milliamperes.

INDUSTRIAL AND POWER TYPES

Typical uses include heavy duty plating, general purpose power supplies, cathodic protection, battery chargers and laboratory equipment. DC output from 10 volts at 0.4 amperes to 120 volts at 13 amperes. Available in half wave, center tap and single phase bridge units.



SELENIUM DIODES

Designed for stable operation in an ambient temperature range of minus 60° C to plus 100° C. Rectifying elements are potted in a thermosetting compound for protection against corrosion and fungus. DC output voltage from 20 to 160 volts. DC output current average from 200 microamperes to 11 milliamperes. Assortments available in kit form. Diam. from 1/8" to 13/32". Length: from 1/4" to 1/2".

GERMANIUM DIODES

Available in all types in demand by service men, laboratories, experimenters and hobbyists. Case diam: 1/4". Case length: 1/2". Axial pigtail leads for easy soldering and clipping into breadboard set-ups. Assortments available in kit form.

WIDEST RANGE IN THE INDUSTRY

SELENIUM PHOTOCELLS

When connected directly to the actuating device or the amplifier, current of the order of 600 microamperes per lumen will flow at an external resistance of 100 ohms. Available in unmounted and mounted units. 8 different cell sizes with average output at 100 foot-candles illumination of from 75 to 750 microamperes.

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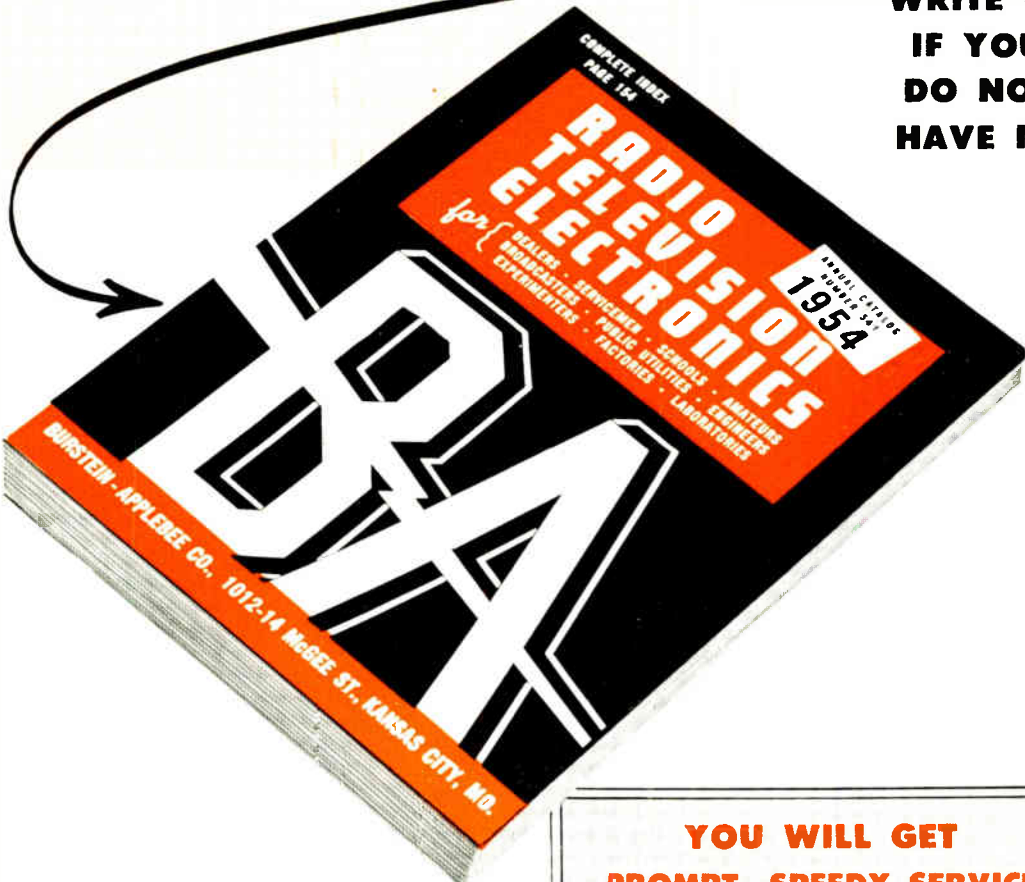
1954 B-A CATALOG NO. 541

156 BIG PAGES

WRITE US IF YOU DO NOT HAVE IT!



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BURSTEIN-APPLEBEE CO.
1012-14 McGEE ST., KANSAS CITY 6, MISSOURI

VALPEY *Quartz Crystals*

are famous for accuracy!

Since 1931, Valpey Crystals have earned and maintained a foremost reputation in aiding amateur and professional electronic engineers in experimentation and development. A portion of the line of crystals is shown here.

COMPACT—HERMETICALLY SEALED

Type VR6, hermetically sealed in compact metal case, has 1500 to 75000 Kc. frequency range. For mobile or fixed stations, VHF and experimental work. Vertical mount for .486" spacing special crystal socket. Jan type HC6.

FOR FREQUENCY STANDARDS

Type XL100, with a frequency of 100 Kc., is used extensively in frequency standards. Mounts in ¾" spacing crystal socket or the standard 5-prong socket. Compact and dependable.

FOR ALL TYPES OF APPLICATIONS

Type CM1, ideal for marine, police, aircraft and general applications has frequency range from 850 to 20000 Kc. Fixed air-gap; vertical mount; available in standard ¾" spaced pins and GR ¾", ⅝", ⅞" or .850" spaced pins.

FAVORITE OF AMATEURS

Type CM5. Frequency range 1500 to 30000 Kc.; used extensively for marine, police and other mobile or fixed stations. Fits Valpey Xtalector for instant, accurate frequency shift; fixed air gap, vertical mount, standard .486" spacing socket or standard octal socket.

AUTHORIZED FOR BROADCAST USE

Type CBC-O Used widely by broadcast, fixed stations and frequency standards. Frequency range 60 to 10000 Kc. Available with 6, 8, or 10 volt over plus-minus ½ degree C. temperature stability. Micrometer-*adjustable air-gap*. Mounts in standard 5-prong socket.

HIGH ACCURACY

Type DFS Features separate 100 and 1000 Kc. crystals in one compact mounting, with accuracy plus or minus .005% over range of minus 10°C to plus 60°C when used in recommended circuits. This is a Valpey development for secondary standards and receiver calibration.

SINGLE OR DUAL CRYSTALS

Type VD5 Frequency range 1000 to 6000 Kc. Single or dual crystals, popular for marine, aircraft and police applications. Mounts in special 3-prong socket.

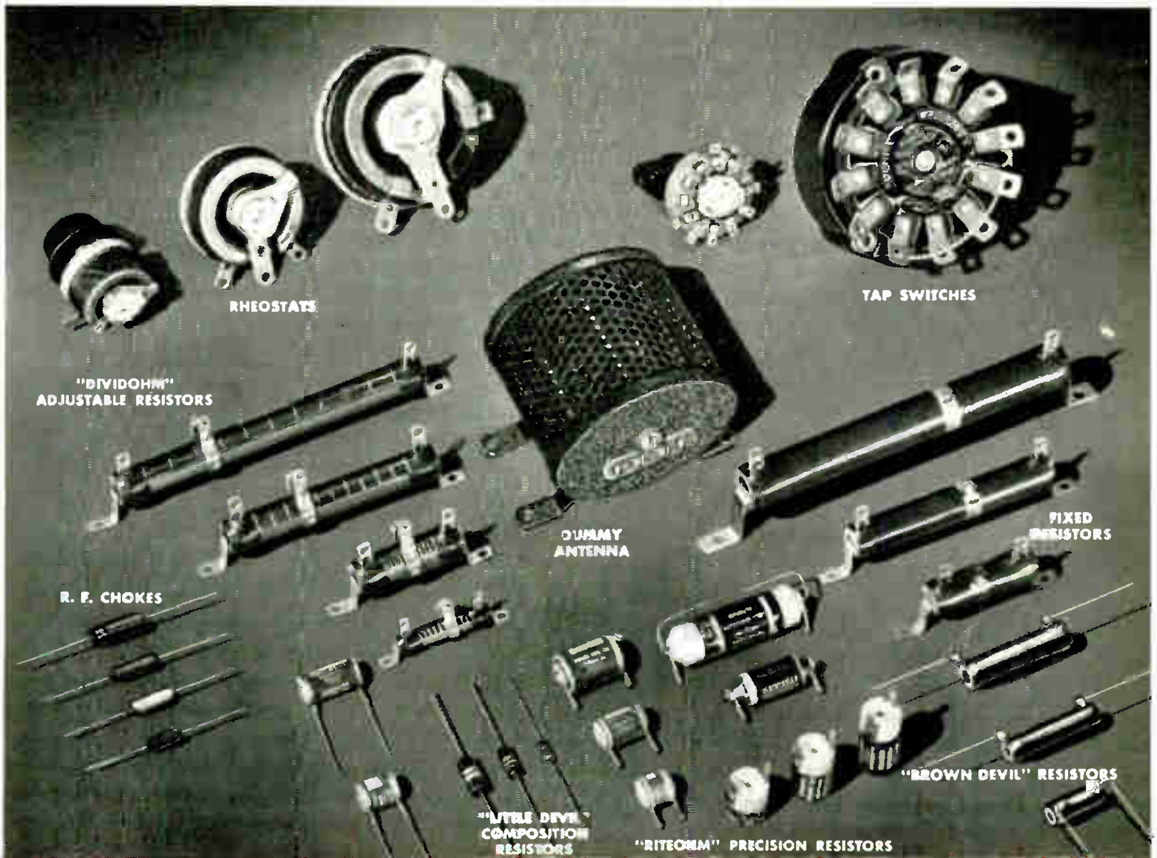
FOR FIXED OR MOBILE TRANSMITTER-RECEIVERS

Type VDO Frequency range 1000 to 10000 Kc. Used for railroad and bus communications, fixed or mobile units. Single or dual crystals with 6 volt over plus-minus ½ degree temperature stability. Mounts in standard 5-prong socket.

For additional details on these or other types write:

VALPEY *Crystal* CORPORATION
1200 HIGHLAND ST., HOLLISTON, MASS.

Where Dependability counts— Use OHMITE!



RHEOSTATS
Insure permanently smooth, close control. All-ceramic vitreous enameled: 25, 50, 75, 100, 150, 225, 300, 500, 750, and 1000-watt sizes.

"BROWN DEVIL" RESISTORS
Sturdy, wire-wound, vitreous-enameled resistors for voltage dropping, bias units, bleeders, etc. In 5, 10, and 20-watt sizes; values from 0.4 to 100,000 ohms.

"LITTLE DEVIL" RESISTORS
Tiny, molded, composition resistors—each marked with resistance and wattage—½, 1, and 2-watt sizes, ±10% or ±5% tol. 10 Ohms to 22 megohms.

FIXED RESISTORS
Resistance wire is locked in place and protected by vitreous enamel. Stock sizes—25, 50, 100, 160, and 200 watts; values 1 to 250,000 ohms.

ADJUSTABLE RESISTORS
Vitreous enameled. Quickly adjustable to the value needed. Adjustable lugs can be attached for multi-tap resistors and voltage dividers. Sizes 10 to 200 watts, to 100,000 ohms.

R. F. CHOKES
Single-layer-wound on low power factor cores, with moisture-proof coating. Seven stock sizes, 3 to 520 mc. Two units rated 600 ma, others 1000 ma.

TAP SWITCHES
Compact, high-current rotary selectors for a-c use. All ceramic. Self-cleaning silver-to-silver contacts. Rated at 10, 15, 25, 50, and 100 amperes.

PRECISION RESISTORS
Three types available: vitreous-enameled, vacuum-impregnated, or glass-sealed. Tolerance ±1%, in ½ and 1-watt sizes, from 0.1 to 2,000,000 ohms.

DUMMY ANTENNA
These rugged, vitreous-enameled units are practically nonreactive within their recommended frequency range. In 100 and 250-watt sizes, 52 to 600 ohms, ±5%.

Write for Stock Catalog No. 24

OHMITE MANUFACTURING COMPANY
3608 W. Howard St., Skokie, Ill., (Suburb of Chicago)

Be Right with **OHMITE®**

RHEOSTATS • RESISTORS • TAP SWITCHES • CHOKES

SHURE MICROPHONES

Field-Proved from SHACK to CAR!

Model 520SL
List Price $38.50

In the Shack . . .

This sturdy Controlled Reluctance unit is designed to handle the most severe requirements of amateur broadcasting, paging, and dispatching systems. It provides high speech intelligibility, makes your messages instantly understood. The "Dispatcher" has a 2-conductor shielded cable, and is wired to operate both microphone and relay circuits. Firm downward pressure on the grip-bar locks the switch. The "Dispatcher" is immune to severe conditions of heat and humidity. Output is 52.5 db below one volt per microbar. High impedance. Furnished with 7-foot cable.

In the Car . . .

A high-quality carbon microphone specially designed for mobile equipment. Used throughout the world for Ham, Police, Fire, and Transportation Services—more *than all other makes combined!* Rugged, dependable unit with clear, crisp voice response and high output. Fits snugly into palm of hand. Heavy duty switch for push-to-talk performance. Furnished with bracket for wall mounting, plus coiled-cord cable. Output level: 5 db below 1 volt for 100 microbar speech signal. 70 to 80 ohms impedance.

"100 Series"

Model	Cable	Switch Arrangement	List Price
101C	Standard Coiled Cord 11" retracted; 5' extended	Two Wire Relay Switch normally open. (No microphone switch)	$27.50
101E	Tinsel Coiled Cord 11" retracted, 5' extended with Amphenol MC4M Connector		$32.50
102C	Standard Coiled Cord 11" retracted, 5' extended	Relay normally open. Microphone switch normally open.	$27.50
102E	Tinsel Coiled Cord 11" retracted; 5' extended with spade lugs		$30.00
103	Standard Coiled Cord 11" retracted; 5' extended with Amphenol MC4M Connector	Two Wire Relay Switch normally open. (No microphone switch)	$29.00

Microbar = one dyne per sq. cm.

SHURE

SHURE BROTHERS, Inc. ★ Microphones and Acoustic Devices

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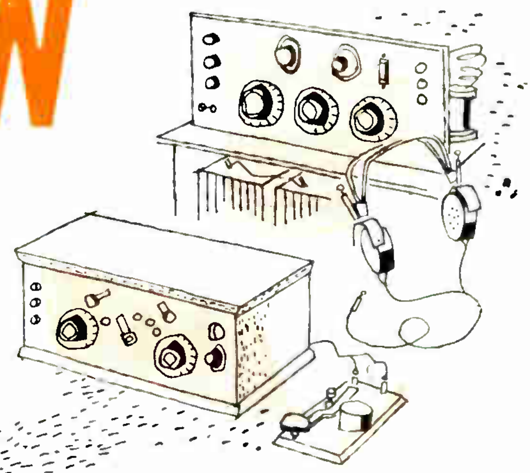
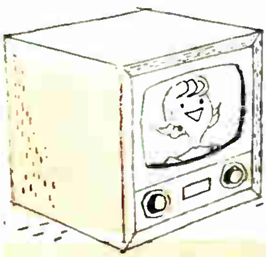
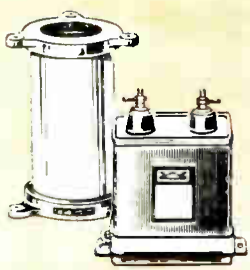
you know

from your own
experience
that Cornell-Dubilier products
are time-tested

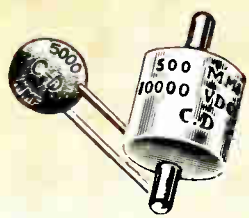




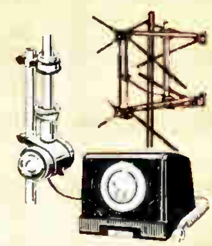
Consistently Dependable



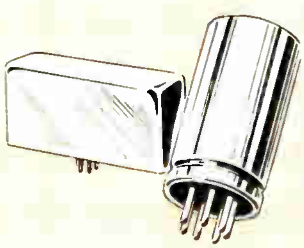
High and Medium Voltage Transmitter Capacitors in all popular capacities.



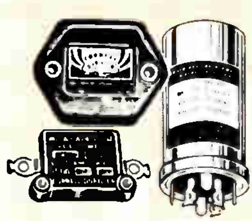
Disc and hi-voltage ceramic capacitors.



CDR Rotators, TV-AM-FM and Auto Antennas.



Auto radio, heavy duty, railroad converter vibrators.



Receiver and transmitter capacitors.

Vibrator-powered converters and inverters.

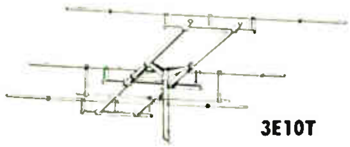
Faradon

CORNELL-DUBILIER ELECTRIC CORPORATION

PLANTS IN SOUTH PLAINFIELD, N. J.; NEW BEDFORD, WORCESTER AND CAMBRIDGE, MASS.; PROVIDENCE, R. I.; INDIANAPOLIS, IND.; FUQUAY SPRINGS, N. C.; AND SUBSIDIARY, THE RADIART CORP., CLEVELAND OHIO

HY-LITE AMATEUR ANTENNAS

3E10F



3E10T

THE FAMOUS INSULATED STANDARD

6, 10-11, 15, 20 METERS • 2, 3, 4 ELEMENTS

Quickly and easily assembled, the HY-LITE ROTARY BEAMS are ideal for any transmission location. They are built ruggedly and will stand up under stress and strain far beyond that encountered under actual use.

The elements on the 6, 10-11 mtr. beams are made of ⅝" and ¾" telescopic aluminum tubing of the highest grade. (The elements of the 15 & 20 mtr. beam are of 1" and ⅞" tubing.) with locking clamps for securing after tuning. The elements are supported on heavy Steatite insulators. Strong aluminum castings support the reflectors and directors and the antenna elements.

The HY-LITE ROTARY BEAMS are rigid and sturdy and mounted on a standard size 6' mast.

Any BEAM can be purchased and later a BEAM of a different frequency added to the original. Stacked array can be had in any combination on order.

6, 10 & 15 meter beams supplied with .1 and .15 spacing. 3 element 20 meter units supplied with .075 & .1 spacing.

FULLY ADJUSTABLE • ALL ALUMINUM • PRICED RIGHT

PD 3E10

THE ALL-GROUNDED SKY-LITE

6, 10-11 METERS • 2, 3 ELEMENTS

Lightest Beam ever built by HY-LITE. Beam complete includes Mast. Can be assembled and adjusted in less than 30 minutes by one man with only a screw driver. Element lengths and spacing fully adjustable. Will withstand high winds and gale, built of all dural tubing and rigid aluminum castings. Elements are ¾" and ⅝" diameter, boom is heavy wall 1". "T" match is adjustable by movable castings to match 300 ohm line. So light it can be rotated with almost any TV antennae rotator. Completely grounded. Good lightning protection. No wood used. Good forward gain and front to back ratio. Very little sag in either boom or elements. Using .1 and .15 spacing. 6 meter or 10 meter can be added separately at later date for stacking. Full instructions are given for length of elements and spacing at different frequencies.

WATCH OUR ADS IN QST FOR CURRENT PRICES

2 meter arrays

4E2

8E2

16E2

Can be mounted for horizontal polarization, built of all dural tubing. All elements are grounded. Excellent forward gain and front to back ratio. Boom is 1" diameter hard tubing. Using .2, .2, and .25 spacing. Elements are ⅜" and 1" outside diameter. Unit is supplied with clamp to fit any size mast up to 1⅜ inch outside diameter (no mast is included with unit.) So light it can be rotated with almost any TV Antennae Rotator. Each four element unit weighs under 3 lbs. — 16 Element unit weighs approximately 12 lbs. Made to match 300 ohm line. Can be stub matched for use with low impedance coax.

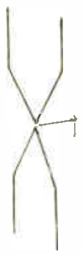
A FULL LINE OF UHF-VHF FM-TV ANTENNAS

YAGIS • CONICALS • SINGLE and DOUBLE V's • CORNER REFLECTORS • BOW TIES • SPECIALS •

Free CATALOG HF-12 ON REQUEST

300-375-450 OHM OPEN WIRE LINE in 100 FOOT SPOOLS

QUADRAPOLE *or the NEW* ROCKETQUAD



HIGH GAIN • OMNIDIRECTIONAL • CD • POLICE • FIRE • HAM

These famous revolutionarily designed antennas are the design of 'Doc' Ross, W2NGA. Real gain is offered over a ground plane and conventional type all-direction antennas. Featuring the same all aluminum construction for which HY-LITE is famous, the QUADS are ideal for service other than, as well as, amateur; CD Headquarters, Police Stations, Fire Stations, Aircraft towers or any application where omnidirectional signals with added gain is desirable. Impedance at antenna is approximately 300 ohms but instructions for balun matching for 52 or 72 ohm coax is included. QUADRAPOLE owners can obtain an additional QUAD and matching stub to assemble the ROCKET QUAD at reasonable cost. Complete with all hardware less mast.

HY-LITE BEAMS

HY-LITE BEAMS HY-LITE BEAMS

HY-LITE *Antennae* inc., 243 EAST 137th STREET, NEW YORK 51, N. Y.

Just Right FOR YOUR RIG!

SPRAGUE CAPACITORS

HYPASS®—feed-through capacitors for bypassing harmonic currents in transmitters and for eliminating v-h-f interference from a-c mains and control circuits. Ideal for eliminating TVI . . . developed to meet transmitter needs outlined by ARRL Headquarters.

FILTEROL®—line filters for suppressing "man-made" radio noises and television "scrambles" on practically any application. Small, self-contained, and easily installed. Filterols provide maximum noise suppression.

CERA-MITE★—the first complete line of miniature disc ceramics. Tough, dependable, and inexpensive. A type for every use—GA (General Application), high-K, and temperature-compensating.

METAL-ENCASED PAPER CAPACITORS—long a favorite with amateurs, these metal encased capacitors are impregnated and filled with KVO, the efficient dielectric material. Long-life and reliability are outstanding.

TELECAP★—phenolic molded paper tubulars assure maximum dependability under extreme heat, humidity, and shock. These Black Beauties are oil-impregnated in ratings from 600 to 12,500 volts.

ATOM®—metal-encased, dry electrolytics are engineered especially for tough television replacement applications. Small enough to fit anywhere. Atoms withstand punishing 85°C. (185 F.) temperatures.

TWIST-LOK★—dry electrolytics, sealed in aluminum cans with twist-prong tabs for washer or direct-to-chassis mounting. Top dependability in small size at 85 C. (185 F.). Sprague makes more ratings than any other manufacturer. Ideal for TV servicing.

MICAS—of maximum quality for R-F applications where high "Q" and high insulation resistance are required. There is a Sprague mica capacitor for every amateur requirement, from tiny half "postage stamp" types to giant ceramic-jacketed types.

KOOLOHM® RESISTORS

The only resistors wound with ceramic coated wire and doubly protected by a glazed ceramic housing and unique end seal to guard against moisture. Wires cannot short. Koolohms can be mounted anywhere, even flat against a chassis or grounded parts. They can be used safely at full wattage ratings.

WRITE FOR CATALOG C-609 OR
SEE YOUR RADIO PARTS SUPPLIER.

★TRADEMARK

SPRAGUE®

PIONEERS IN ELECTRIC AND ELECTRONIC DEVELOPMENT

SPRAGUE PRODUCTS COMPANY
(Distributors' Division of Sprague Electric Co.)
North Adams, Massachusetts

SUPREME

Since 1927

"Supreme by Comparison"

MULTI-METERS

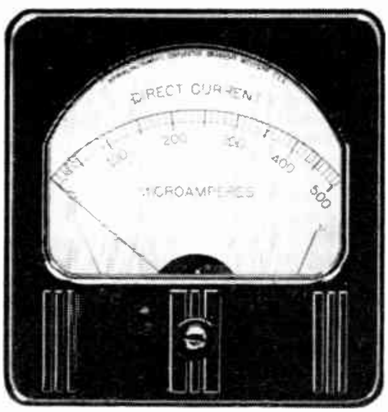
The Supreme Model 543 (left) and Model 542 (right) volt-ohm-milliammeters have become real companions to maintenance men everywhere. Both instruments are the same size (approx. 6 x 3 x 2 inches) and are available in either the moulded "B" style case (illustrated) or the metallic type case with leather handle and meter protector shield (Style "M"). Model 543 has 4 functions and 12 ranges and the Model 542 has 6 functions and 24 ranges.

Model 543B	**$24.60**	**Model 542B**	**$27.70**
Model 543M	**$28.30**	**Model 542M**	**$32.10**
Test Leads	**$ 1.40**	**Test Leads**	**$ 1.40**

MINIMETERS

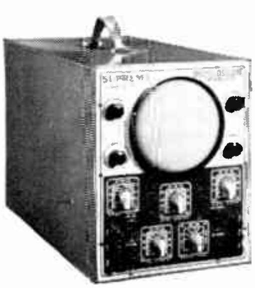
About the size of a cigarette pack, these fine little multi-range, single function instruments have found their way into a lot of shirt and vest pockets of electrical troubleshooters. Convenient size and durable construction for portability. Built with highest quality materials and workmanship for dependability. Quantity production for a budget price. Model 440 Ohmmeter only $12.90 complete with test leads. Other functions available from $10.65 to $14.50.

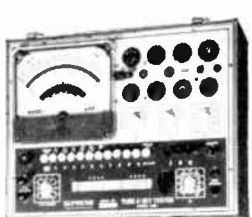
PANEL METERS



Leading manufacturers of electrical and electronic devices select Supreme meters for use in their products because they are quality built and dependable. Supreme meters are supplied for end equipment uses in a variety of sizes and designs "customed" to meet the need for special scales, dials and case appearance. A sketch or drawing of your meter needs will bring you a prompt quotation and delivery schedule from Supreme.

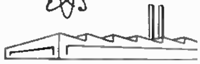
WRITE FOR ADDITIONAL INFORMATION: For prompt action, address your request for additional information to **SUPREME, INC., Attn: HB-4, 1754 CARROLTON AVENUE, GREENWOOD, MISSISSIPPI.** In addition to the items illustrated above, Supreme also makes Oscilloscopes, Tube Testers, Signal Generators, and special purpose instruments for both government and industry. Special instrument problems are welcome.





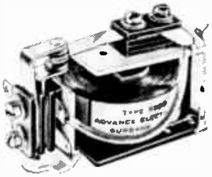
ADVANCE RELAYS

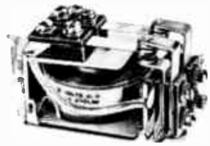
FOR EVERY APPLICATION

COMMUNICATIONS 
AIRCRAFT 
ELECTRONICS 
INDUSTRIAL 

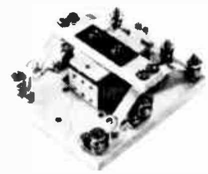
Whatever your relay requirements may be, and whatever you need in the way of specialized facilities, ADVANCE is equipped to help you.
Here are capable personnel and modern machines of a specialized organization with a 37-year record of successful operation.
Here are unexcelled facilities for the development and production of precision relays for general industry.

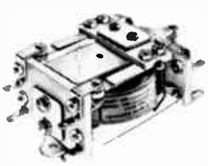

DM2C
(Formerly 304BF)

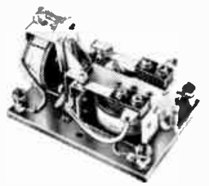

PC1C
(Formerly 953B)

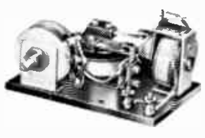

PC2C
(Formerly 964B)


PV1A
(Formerly 961C)


AT2C
(Formerly 400)


AH2C
(Formerly 1000)

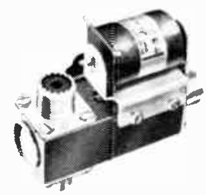

LE2C
(Formerly 604B)


OE2B
(Formerly 750A)


SV1C
(Formerly 1200)


AM2C
(Formerly K1504RF)


MG2C
(Formerly K1504)


CB1C
(Formerly 7200)

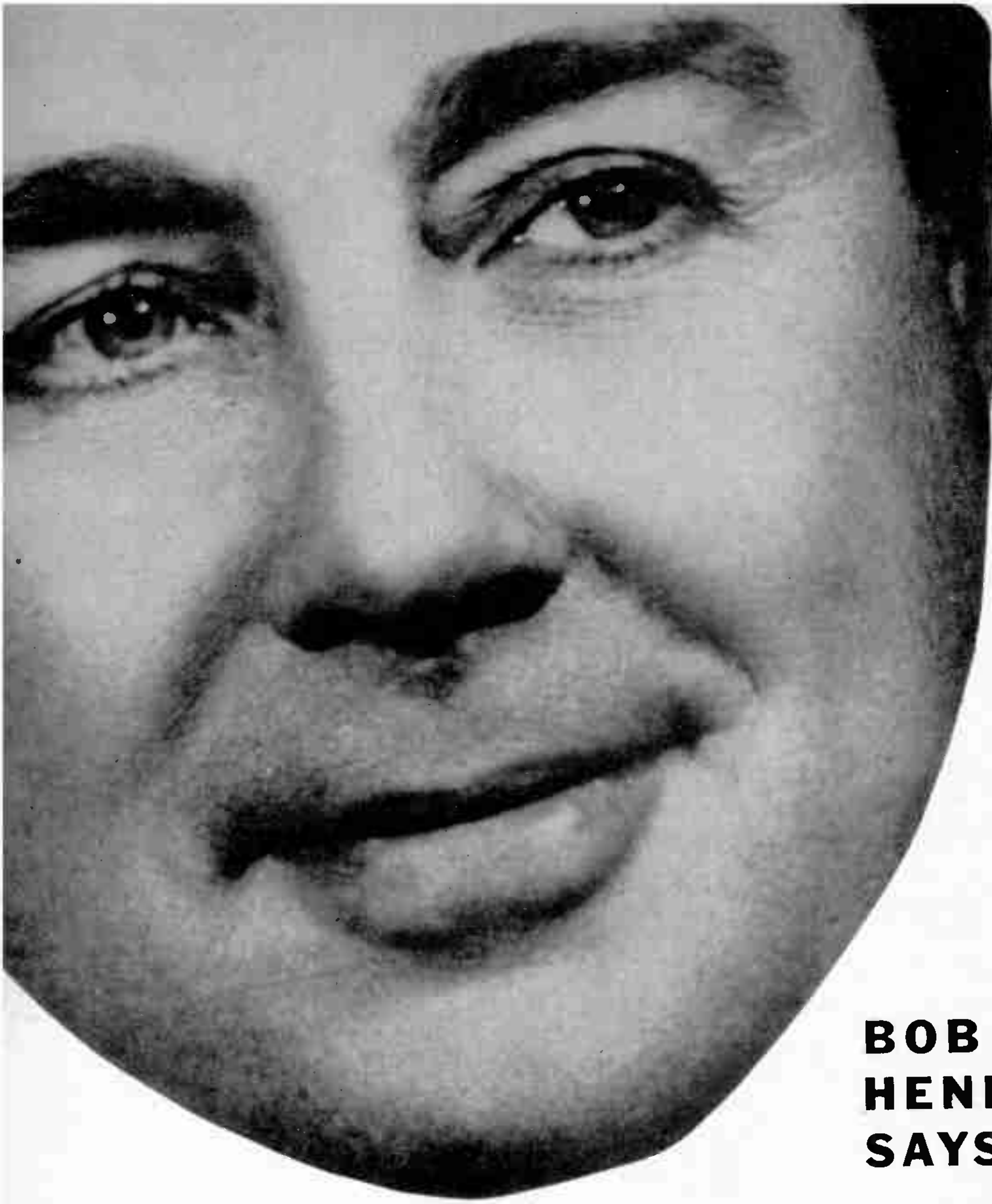
TYPE	USE	CONTACTS		COILS	DIMENSIONS		
DM	Time Delay	10 Amps	DPDT	6-12-24-110-220 A.C. 6-12-24-48-110-220 D.C.	1 5/8	2 5/8	3 3/4
PV	Power Transfer	30 Amps	SPDT	1 to 440 VAC 1 to 220 VDC	1 5/8	2	3
AH	Antenna	1/2 KW	3PDT	1 to 440 VAC 1 to 220 VDC	1 7/16	1 5/8	2 3/4
OE	Overload	10 Amps	DPDT	1 to 440 VAC 1 to 220 VDC	2 3/8	2 3/4	4 1/4
AM	Antenna	1/4 KW	DPDT	1 to 220 VAC 1 to 150 VDC	1 1/4	15/16	1 7/8
PC	Power Transfer	10 Amps	4PDT	1 to 440 VAC 1 to 220 VDC	1 9/16	1 5/8	2 5/8
AT	Antenna	1 KW	DPDT	1 to 440 VAC 1 to 220 VDC	1 11/16	2 3/4	3 15/16
LE	Latching	10 Amps	DPDT	Adjustable 250 MA to 500 MA	2 3/8	2 1/2	3 5/8
SV	Sensitive	1 Amp	SPDT	1 to 40,000 OHMS	1 1/2	2	2 9/16
MG	Miniature	1 1/2 Amps	3PDT	1 to 220 VAC 1 to 150 VDC	1 5/8	1 1/4	1 13/16
PC	Power Transfer	10 Amps	4PDT	1 to 440 VAC 1 to 220 VDC	1 9/16	1 5/8	2 5/8
CB	Coaxial	8/10 KW	SPDT	1 to 440 VAC 1 to 220 VDC	3 1/16	1 7/16	3 9/16

Illustrated are a few typical examples of the ADVANCE relay line. The complete line includes sensitive, midget, midget telephone, keying, instrument, time delay, overload, transmission line, hermetically sealed and ceramic insulated types and variations.
For accurate circuit behavior . . . lower unit cost . . . increased efficiency . . . and uniformly high quality relays — *specify* ADVANCE.

ADVANCE
RELAYS

Write for new, descriptive Catalog containing detailed information about ADVANCE Relays and facilities.

ADVANCE ELECTRIC AND RELAY COMPANY
2435 NORTH NAOMI STREET, BURBANK, CALIFORNIA
Sales Representatives in Principal Cities of U. S. and Canada



BOB HENRY SAYS—

"When you deal with Bob Henry I want you to be entirely satisfied, get the lowest wholesale price, the quickest service, the best trade-in and the best time sale plan financed by me with the least red tape and the lowest carrying charges. I want everything to be entirely satisfactory. It will pay to contact me by mail, phone, or telegraph about any equipment.

"I want you to say, '*You can bet on Bob Henry for the best.*'"

Bob Henry

Butler 1, Missouri
Phone: 395

HENRY RADIO STORES

11240 Olympic Blvd.
LOS ANGELES 64
BRadshaw 2-2917

"LARGEST DISTRIBUTORS OF SHORT WAVE RECEIVERS"

YOU CAN BET ON *Bob Henry* FOR THE BEST—HE HAS IT!

31 REASONS WHY hallicrafters SX88

IS HAM NEWS AT BOB HENRY'S:

1. Heavy gauge steel welded chassis for mechanical stability.
2. Full precision gear drive for main and band spread tuning.
3. Six position Band Width Control (selectivity) from 250 cycles to 10 kc.
4. 10 watt inverse feed back and push-pull audio output.
5. Exhalted B.F.O. for tops in single side band reception.
6. Buffer amplifier in B.F.O. circuit.
7. Antenna trimmer.
8. Amplified and delayed A.V.C.
9. Built-in 100 kc calibration crystal.
10. Second conversion oscillators crystal controlled.
11. Inertia tuning (fly wheels both dials).
12. Full frequency coverage from 535 kc to 33 mc.
13. Calibrated electrical band spread 160, 80, 40, 20, 15, 11, and 10 meters.
14. Logging scales on each tuning shaft.
15. Dial locks on each tuning shaft.
16. Tuning dial indicators resettable from front panel for maximum calibration accuracy.
17. Auxiliary A.C. socket on rear of chassis.
18. Illuminated band-in-use indicator.
19. Illuminated S meter.
20. Dual S meter calibration S units and microvolts.
21. Auxiliary power socket plus .6 amps at 6.3 volts and 10 ma at 150 volts for accessories.
22. Standard 8¾" by 19" panel for rack mounting if desired.
23. 50 kc i.f. output jack via cathode follower for teletype converter, etc.
24. Five position response control (tone control).
25. Two r.f. stages (Bands II to VI).
26. 17 tubes plus voltage regulator, ballast tube and rectifier.
27. Automatic noise limiter circuit.
28. Phono Jack.
29. Audio output transformer for 3.2, 8, 500/600 ohm loads.
30. Fuse for overload protection.
31. Auxiliary sensitivity control permits monitoring of local transmissions in standby position.

PRICE $499.50 SUBJECT TO CHANGE WITHOUT NOTICE.

BOB HENRY SAYS—"I offer you the lowest prices, the easiest terms, top trade-ins, my personal attention to your satisfaction. Get my offer before you buy."

Bob Henry

Butler 1, Missouri
Phone: 395

HENRY RADIO STORES

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LOS ANGELES 64
BRadshaw 2-2917

"LARGEST DISTRIBUTORS OF SHORT WAVE RECEIVERS"

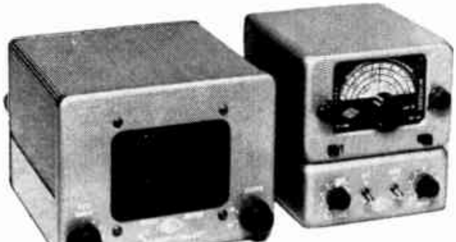
YOU CAN BET ON *Bob Henry* FOR THE BEST—HE HAS IT!

THE NEW GONSET "COMMUNICATOR"

Ideal for Amateur • Civil Defense • C.A.P. Commercial Activities • Fixed-Mobile • Airport Unicom.

Superheterodyne receiver is complete with 6BQ7 Cascode rf stage, 3 stages if, and employs the famous GONSET NOISE CLIPPER circuit. Tuning eye metering on all stages including transmitter antenna circuit (indicates max. transmitter output). No external meter required. Transmitter employs 8 mc. crystals for stability. Modulator can be used for P.A. system, for emergency and CD work. **$209.50** Including Federal Excise Tax (less microphone and crystal). For deluxe model with Squelch add $20.00.

THE NEW GONSET "SUPER-CEIVER" MOBILE RECEIVER



The Super-ceiver combination consists of three elements: HF tuning head, which may be a Super-Six or other standard, good quality converter, a control box and the all-important Model 3041 unit, the heart of the combination. The latter is actually a crystal controlled, superheterodyne receiver with input circuits fixed-tuned to the output frequency used for the average converter. (1430 kc for Super Six) Six band operation, (10-11-15-20-40-75) when used with "Super Six."

"Finger-tip control" with remote control head 2" high to match Super Six converter.

COMPACT! Model 3041 unit is 6⅞" wide, 6¾" deep and 5¼" high. Control head is 5" wide, 3½" deep, 2" high.

Convertible (dual) 6-12 volt pack. **$119.50** Including Federal Excise Tax. (Price does not include Super Six converter.)

THE GONSET "SUPER 6"

A Six Band Amateur Converter covers the amateur 10, 11, 15, 20, 40 and 75 meter phone bands with lots of bandspread. Also covers 6 mc. (49 meter) nighttime and 15 mc. (19 meter) daytime Short Wave broadcast bands, therefore ideal for BC reception in remote areas where standard broadcast reception is poor. Very high sensitivity on an 8 ft. whip antenna. Low drift, noise factor, and image response. Net **$52.50.**

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"I offer you the lowest prices, the easiest terms, top trade-ins, my personal attention to your satisfaction. Get my offer before you buy."

Bob Henry

THE GONSET "COMMANDER"

35-50 watt MULTI-BAND TRANSMITTER

FREQUENCY RANGE: 1.7 to 54 Mc. continuous.

TUBES: R.f. — 6AG7, 6146. A.f. — 12AT7, 2-6AG5's

POWER REQUIREMENTS: 300 volts d.c. at 200-225 ma. (phone) and 6.3 volts a.c. or d.c. at 3.15 amp. 35 watts input on phone, 50 watts on c. w.

MODULATOR: Class AB_2 beam tetrodes and integral high level speech clipping (requires no adjustment) permit heavy modulation without splatter.

MICROPHONE INPUT: Any standard carbon or p.a. type high impedance dynamic or crystal. No optional model or preamp required for crystal mike.

ANTENNA FEED: All conventional feed lines including COAX, RIBBON, and OPEN LINE, or DIRECT TO MARCONI ANTENNA.

SIZE: Very compact: 5⅜" high by 8⅛" wide by 7⅛" deep. Suitable for under dash mounting.

(Completely wired and tested, with all tubes, and including two high-Q final tank coils which cover 10-11, 15, 20, 40, 75, and 80 meters. Final coils for other frequencies are available separately.) **Model #3016 — Net $124.50**

Butler 1, Missouri
Phone: 395

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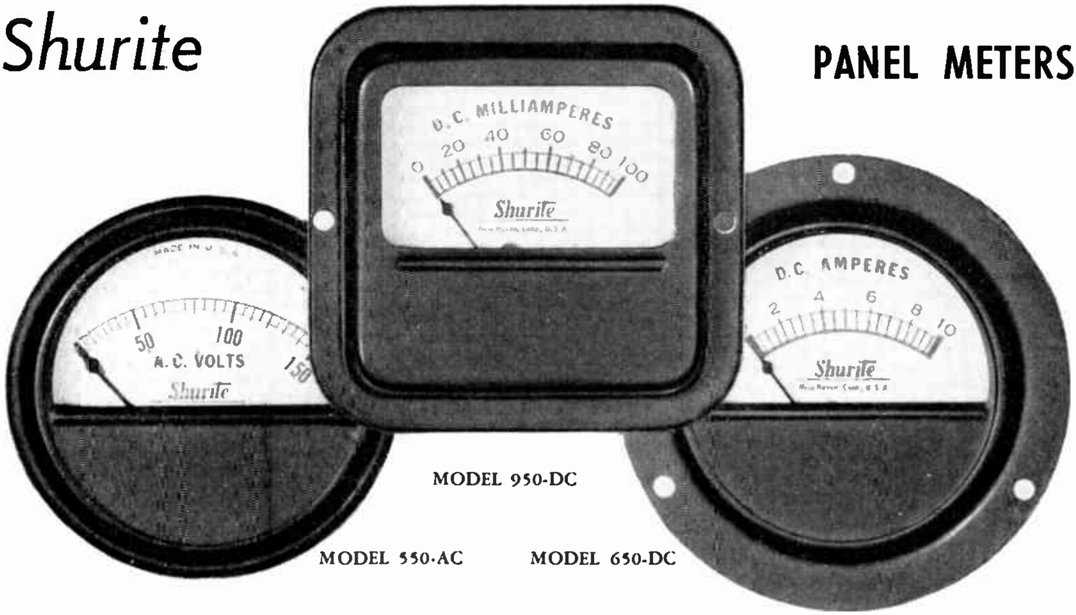
"LARGEST DISTRIBUTORS OF SHORT WAVE RECEIVERS"

Shurite PANEL METERS



MODEL 950-DC

MODEL 550-AC

MODEL 650-DC

POPULAR WITH HAMS BECAUSE THEY'RE . . .

Rugged. . . . . Exceptionally high ratio, torque to weight, for fast pointer response. Sturdy construction throughout. Molded inner unit with internal and external locking nuts for maximum rigidity.

Good Looking. . . . . Dials are metal so they stay attractive in spite of age and moisture. Metal cases throughout with rich telephone black finish. Concealed coils and good, readable scales.

Well Designed. . . . . AC meters are double-vane repulsion type using hair-spring and jeweled bearing. Most DC meters are polarized-vane solenoid type. High internal resistance voltmeters available in popular ranges. Accuracy well within 5%.

Guaranteed. . . . . For one year from date of purchase against defective workmanship and material, and will be repaired or replaced if sent to the factory postpaid with 40¢ handling charge.

Priced Right. . . . . For instance, the meters pictured above are priced at

Model 950, 0–100 DC Ma. $1.60
Model 650, 0–10 DC Amps. 1.55
Model 550, 0–150 AC Volts. 3.30

Other meters are correspondingly reasonable in price. You get the benefit of low costs made possible by large quantity production.

A Complete Line. . . . . All of these features are available in 323 ranges and types; AC, DC, Voltmeters, Ammeters, Milliammeters, Resistance Meters. For instance, DC Milliammeters are made in 65 types and ranges. The newest meters are a 0–3 DC Milliammeter with 500 ohms internal resistance and built-in zero adjustor with ten times the sensitivity of previous 0–3 DC Milliammeters, also a 0–1 DC Milliammeter with 1000 ohms internal resistance and zero adjustor.

Available. Stocked by leading electronic parts distributors in a wide variety of types and ranges. For prompt shipment, contact the authorized Shurite distributor nearest you.

THEY'RE TODAY'S BEST VALUE

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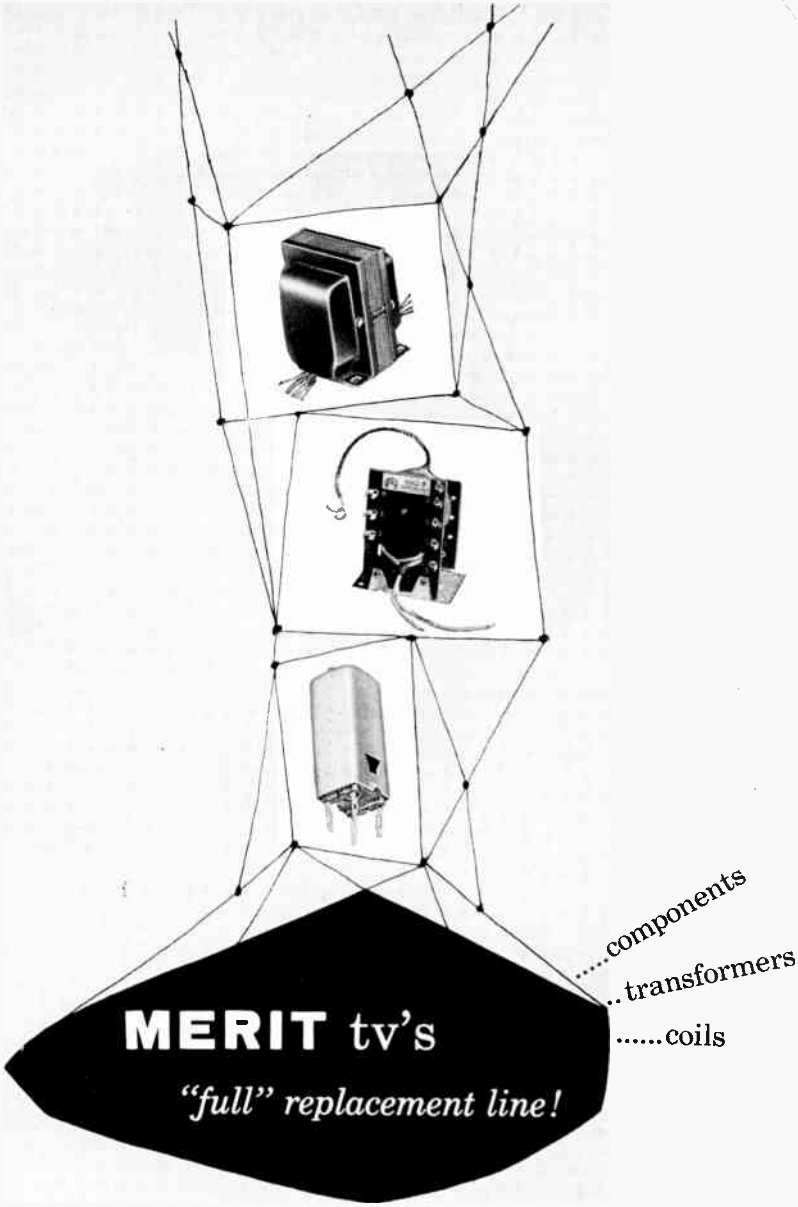
Shurite TRADE MARK

SHURITE METERS • 89 Hamilton Street, New Haven 8, Conn.



Flybacks: Universal replacement with complete installation data. Exact replacement for the "big 10" *where necessary.
**RCA—Zenith—Motorola—Admiral—etc.*

Transformers: Complete coverage of the replacement market.

IF-RF Coils: A complete line constantly expanding to meet current and future needs.

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- Listed in John Rider's Tek-File and Howard Sams' Counter Facts and Photo Facts—Tape Marked* to help you.

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Complete Stocks—

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Meets all MIL-T-27 specs

C-TYPE
With Leads

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Lug terminals

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THE WORLD'S TOUGHEST
Hermetically-Sealed Units

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We can supply the *complete* line of CHICAGO "Sealed-in-Steel" Transformers — units designed to fit today's amateur circuits . . . units to meet all MIL-T-27 specifications for military requirements . . . units for a wide range of industrial applications. Whatever your transformer needs, there's nothing tougher, nothing better than CHICAGO "Sealed-in-Steel" units — and BRILL has them in stock for quick shipment. Order from your BRILL Catalog for prompt delivery.

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Write for your FREE copy of the latest W. D. BRILL Catalog — the complete Buying Guide for all Amateur and Industrial Electronic Equipment. Features the widest selection and largest stocks of equipment — includes the CHICAGO Transformers you need. Get and use your 200 page W. D. BRILL Catalog now.

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THE NEW MODEL 770-A

The FIRST Pocket-Sized

VOLT-OHM MILLIAMETER

USING THE NEW "FULL-VIEW" METER

71% MORE SCALE AREA!

Yes, although our new FULL-VIEW D'Arsonval type meter occupies exactly the same space used by the older standard 2½" Meters, it provides 71% more scale area. As a result, *all* calibrations are printed in large easy-to-read type and for the first time it is now possible to obtain measurements instead of approximations on a popular priced pocket-sized V.O.M.

Features

★ Compact—measures 3⅛" x 5⅞" x 2¼"

★ Uses "Full View" 2% accurate, 850 Micro-ampere D'Arsonval type meter

★ Housed in round-cornered, molded case

★ Beautiful black etched panel. Depressed letters filled with permanent white, insures long-life even with constant use.

SPECIFICATIONS

6 A.C. VOLTAGE RANGES:
0-15/30/150/300/1500/3000 Volts

6 D.C. VOLTAGE RANGES:
0-7.5/15/75/150/750/1500 Volts

2 RESISTANT RANGES:
0-10,000 Ohms 0-1 Megohm

3 D.C. CURRENT RANGES:
0-15/150 Ma. 0-1.5 Amps.

3 DECIBEL RANGES

-6 db to +18 db
+34 db to +58 db
+14 db to +38 db

The Model 770-A comes complete with self-contained batteries, test leads and all operating instructions.

$15.85 NET

SUPERIOR INSTRUMENTS CO.

Dept. HB-54 ★ 2435-41 WHITE PLAINS ROAD ★ NEW YORK 67, N.Y.

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NYLON Lacing Cords and Flat Braided Tapes

meet Army, Navy and Civilian "Specs"

Lacing Cord or Tying Twine

Intensive research in the laboratories of The Heminway & Bartlett Mfg. Co. has resulted in the development of fungus-proof Nylon Lacing Cords (also known as Tying Twines) and Flat Braided Tapes. These new products are especially important to you because of their special synthetic resin coating which resists the growth of mold and micro-organisms — factors most often responsible for the deterioration of old-type linen and cotton lacing cords and tapes and the subsequent corrosion and failure of electronic equipment.

Heminway & Bartlett's new, special-finish Nylon Lacing Cords and Flat Braided Tapes are much faster and easier to lace. They have high abrasion resistance . . . low moisture absorption. They retain the desirable malleability of wax and yet have a melting point of over 190°F. In addition, they are non-toxic to humans.

Heminway & Bartlett's Nylon Lacing Cords and Flat Braided Tapes are sold through leading distributors throughout the U. S. If your regular distributor does not have them in stock, write us for the name of the nearest distributor who carries them.

EST. 1888

HEMINWAY & BARTLETT

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942 KINGS HIGHWAY **BROOKLYN 23, NEW YORK** **Tel. ES 5-9435**

ONE HUNDRED THOUSAND VOLTS!

At Fifty Thousand Ohms/Volt
KILOVOLTER Model 4000

- **Shielded Polysterene Probe**
- **Choice of Ranges**
- **Simple Foolproof Circuits**

Measures television and X-ray voltages with extra high input impedances. All voltage is dissipated in the shielded polysterene probe. "Normal-Reverse" key is provided so that probe may be used regardless of polarity of voltage under test. Indicator has large clear scale for easy reading. Additional low leakage path, special insulated probe is supplied on all 100 KV models.

Model	Range Kilovolts	Price
4000	0-25/50	**$67.50**
4000-A	0-100	**80.00**
4000-B	0-50/100	**85.00**
4000-C	0-10/50/100	**95.00**

HIGH VOLTAGE POWER SUPPLIES

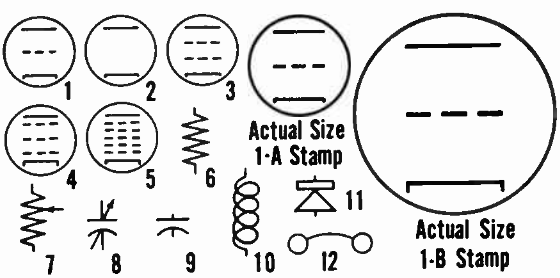
A precision, well constructed high voltage supply for television, meter testing and calibration, electrostatic painting, breakdown tests, nuclear physics and wherever high potentials at low currents are needed. Output is well filtered direct current. Adjustable by means of a control on the front panel. Available with or without meter. Input voltage is 115 v., 60 cycles.

Model		Maximum Voltage	Price
6000		2,500	**$60.00**
6000-A	With Meter	2,500	**80.00**
6005		5,000	**75.00**
6005-A	With Meter	5,000	**95.00**
6010		10,000	**85.00**
6010-A	With Meter	10,000	**105.00**
6015		15,000	**95.00**
6015-A	With Meter	15,000	**115.00**
6025		25,000	**115.00**
6025-A	With Meter	25,000	**155.00**

Also available with two meters (as illustrated) thus current and voltage can be read simultaneously. Perfect for breakdown and leakage measurements. Current limiter circuits protect meter on shorts. Price $35 additional. Specify 50 µamps, 100 µamps or 1 mil meter. Same delivery time.

Regulated Power Supplies. Any model can be had with built-in regulation circuits. Output remains within 5% over most of the range. Same delivery time. Closer regulation on special order.

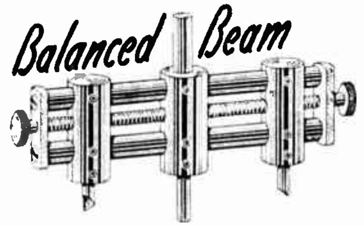
TRANSPARENT CIRCUIT STAMPS



Provide clear sharp impressions of all the most used radio and electrical circuit symbols. Saves considerable drawing and drafting time and provides a neater looking appearance. Used by leading manufacturers and laboratories. Stamps are crystal clear and easily lined up by looking right through them. Available in two popular sizes. May be purchased separately or in complete set. When ordering, specify stamp number and size.

Size A	CIRCUIT STAMP SET	(12 stamps)	**$8.50**
Size A	Individual Circuit Stamps	each	**.85**
Size B	CIRCUIT STAMP SET	(12 stamps)	**10.00**
Size B	Individual Circuit Stamps	each	**.95**

Balanced Beam MICRO CIRCLE CUTTER



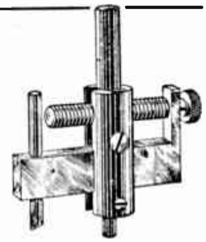
Cuts Metals • Woods • Plastics

For quickly cutting those extra large hole sizes. Perfect for making washers.

Model	Shank	Size	Price
9	Round	10"	**$12.50**
10	Round	15"	**15.00**

Extra cutting bits 50¢ each.

MICRO CIRCLE CUTTER



Cuts holes in all types of metals from stainless steel to magnesium. Perfect for plastics and wood. Especially recommended for cutting meter holes in panels.

Built-in micrometer type size control for precise settings. Extra heavy construction of the main beam and body make it useful for production jobs as well as experimental work. All are equipped with a ¼" high speed steel cutting bit.

Model	Type	Size	Price
1	Round Shank (for drill press or hand drill)	4"	**$5.00**
1-A	Square Tapered (for hand brace)	4"	**5.00**
5	Round Shank	6"	**7.50**
6	Morse No. 2 Taper Shank	6"	**8.50**

Extra cutting bits 60¢

SCALE PRINTING MACHINE

Prints lettering, numbers and divisions on meter scales, metal, paper, plastics. For making experimental name plates, decals, etc. Perfect lettering is assured by use of standard printers type. Many leading instrument manufacturers and laboratories are using our machines. The Model 1500 Scale Printer is illustrated. Send for Catalog A-4.

HIGH VOLTAGE TRANSFORMER

Small physical size is the result of engineering studies into this problem. Used for Oscilloscopes, Photoflash Units, and Miniature High Voltage Power Supplies. Primary has tap for rectifier tube. Several diagrams furnished. Provides 4500 Volts D.C. Price **$15.00.**

Part No	Primary	Secondary	Size
3 KVT	115 Volts 60 Cycle with 1¼ Volt Tap.	3200 Volts AC at 1 Ma	2" high x 2¼" wide. Base 2⅜" Mtg. Centers 2⅞"

MINIATURE HIGH VOLTAGE CONVERTER

Connect primary to batteries and obtain high voltage output. Single compact unit contains both an efficient vibrator and special step-up transformer. Use for photoflash, isolated power supplies and where high potentials and small size are required. Easily rectified with any of these tubes: 1X2A, 1B3GT, 5642 or 1654. Diagrams for complete power supplies furnished with each unit.

Part No.	Primary	Secondary	Price
5 MVT	2 to 6 Volts DC or 60 Cycle AC	2,000 to 7,000 Volts	**$8.50**

ELECTRO-GRAPHIC RECORDER

As a result of a number of field tests, we are in the process of adding a number of new features to the ever popular Model 50 Electro-Graphic Recorder. This direct writing instrument will continue to be the same compact size and rugged construction.

INSTRUMENTS BUILT TO SPECIFICATIONS. Precise Measurements Company equipment is used by leading industrial concerns, research laboratories, government and serious amateurs everywhere. **Your Local Distributor will be glad to show you our products.**

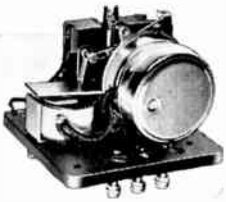
FOR ECONOMY and DEPENDABILITY... ORDER HUSKY RELAYS!

For quality, efficiency and reliability — select the relay you need from Price Electric's complete line. A few of the many types available appear on this page. A more complete listing and description is available in our Relay Reference Folder, which will be sent free upon request.

HUSKY relays consistently meet the most exacting specifications and perform their required functions under the most adverse conditions.

Series 8200, High Current

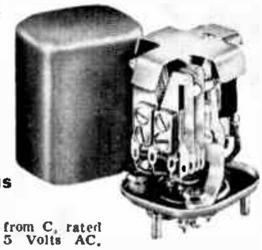
CONTACTS: 4 PNO. Either one or two contacts rated up to 150 amp at 28 V.D.C. Balance rated at 25 amp 115 V.A.C.
COIL VOLTAGE: 6 to 48 V.D.C.
SHOCK: 70 G
VIBRATION: 30 G at 10 to 500 CPS
SIZE: 3⅛" L x 2-1/16" W x 2-9/16" H



SYNCHRONOUS TIME DELAY Series 4050

CONTACTS: S.P.N.O. to D.P.D.T. rated 10 Amps at 115 Volts AC, non-inductive.
COIL VOLTAGE: 115 Volts, 60 cycle AC or 230 Volts, 60 cycle AC.
TIME INTERVALS (FIXED): 15 sec. to 15 min.

AVAILABLE HERMETICALLY SEALED



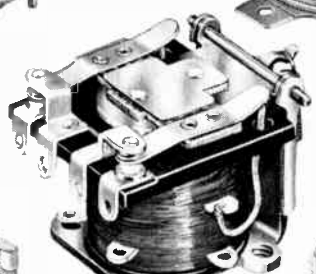
Series 5724-1HS

(AN 3304-1)
(MIL-R-6106)
CONTACTS: 4 from C, rated 3 Amps at 115 Volts AC, non-inductive.
COIL VOLTAGE: 26.5 Volts DC.

ALSO AVAILABLE AS MIDGET OPEN STYLE RELAY

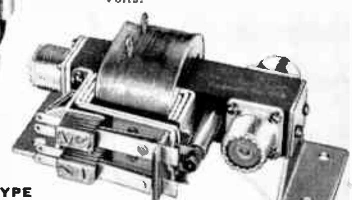
Series 8000 ROTARY RELAY

CONTACTS: 11 P.D.T., make-break or break-make
TORQUE: 7½ inch ounces
COIL VOLTAGE: 6 to 115 V.D.C.



TYPE 6351 CO-AXIAL

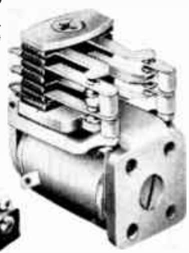
CONTACTS: SPDT rated at 750 watts (RF load). Auxiliary SPNO-SPNC rated at 2 amps at 115 V.A.C.
COIL VOLTAGE: AC — 115 or 230 Volts. DC — 6 to 110 Volts.

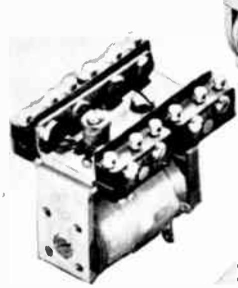


MIDGET, Series 1050

CONTACTS: D.P.D.T., rated 2 Amps at 115 Volts AC, non-inductive.
COIL VOLTAGE: 3 to 24 Volts DC.
MAXIMUM COIL RESISTANCE: 8500 Ohms.

AVAILABLE HERMETICALLY SEALED





MINIATURE TELEPHONE TYPE Series 5500

OPERATING VOLTAGES: 6 to 48 volts DC.
MAXIMUM COIL RESISTANCE: 5000 ohms.
CONTACT CAPACITY: Standard contacts are rated at 3 amps 115 volts AC or 24 volts DC, non-inductive.
CONTACT COMBINATIONS: Up to 6 Form A, or 4 Form C (total of 12 springs).

AVAILABLE HERMETICALLY SEALED

HIGH SHOCK RELAY, Series 4650

CONTACTS: 4 PDT rated 5 amperes at 125 V.A.C. non-inductive
COIL VOLTAGE: 6 to 115 V.D.C.
SIZE: 1⅝" L x 1-5/16" W x 1-9/16" H

RELAY REFERENCE FOLDER

A handy guide to Husky relays, containing photographs and specifications of 42 types. Helps you find the type relay you need in just a few seconds. Address request to Dept. AR54.

PRICE ELECTRIC CORP., FREDERICK, MD.
HUSKY RELAYS, CONTROLS

Price Electric CORPORATION
FREDERICK, MARYLAND

GEIGER COUNTERS

75N	120C	200C
75NB3	120CB	200CB
90NB	120N	200N
100C	120NB	200NB
100CB	150N	230N
100N	150NB	240C
100NB	153C	240N

HIGH VOLTAGE RECTIFIERS

AVT-13 AVT-14

IGNITRONS

5551/652 5554/679
5552/651 5555/653B
5553/655

MAGNETRONS

2J48	4J47	4J58
2J55	4J52	4J59
2J56	4J57	725A

MERCURY RECTIFIERS

249B	575A	866AX
249C	673	869B
255B	857B	872A
266B	866A	8008
315A		

PENTODES

828 6083/AX-9909

Secondary Emission Pentode

EFP60

Subminiature Pentodes

6007/5913 6008/5911

TETRODES

AX4-125A/6155 5895/AX-9905
4X150A 6075/AX-9907
AX4-250A/6156 6076/AX-9907R
807 6097/AX-9908
813

TWIN TETRODES

5894/AX-9903 6360
6252/AX-9910 832A

THYRATRONS
Hydrogen

4C35 6268/AX-9911
5C22 6279/AX-9912

Inert Gas

2D21

Mercury

FG17/1701/967/ 5559
5557 5869/AGR-9950
FG105/AX105 5870/AGR-9951
678

Xenon

3B28 4B32 5685/C6J

TRIODES

HF125	473	889A
HF200	498	889RA
207	501R	891
211	502	891R
212E	502R	892
212F	504R	892R
232C	508/6246	893A
242C	750TL	893AR
HF250	805	HF-3000
250TH	810	ZB3200
250TL	833A	5604*
251A	834	5619
279A	838	5658
284D	845	5666
HF300	849	5667
311CH	849A	5736
342A	849H	5771
343A	851	5920
343AA	858	8002
450TH	859	8002R
450TL	880	

3X2500A3 5868/AX-9902
3X2500F3 5923/AX-9904
EC55/5861 5924/AX-9904R
HF201A/468 6077/AX-9906
5866/AX-9900 6078/AX-9906R
5867/AX-9901

VACUUM CONDENSERS

VC6/20	VC50/32
VC6/32	VC100/20
VC12/20	VC100A/20
VC12/32	VC100/32
VC25/20	VC100A/32
VC25/32	VC250/20
VC50/20	

VOLTAGE REGULATORS

[illegible]

AMPEREX® TUBES

FOR ALL INDUSTRIAL and COMMUNICATIONS REQUIREMENTS

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MATTHEW MANDL is author of the "Television Servicing Clinic" in *Radio-Electronics* and of regular articles on the technical and servicing problems of radio and television in other leading trade magazines; technical and engineering consultant; former operator of a radio station and owner of his own servicing business.

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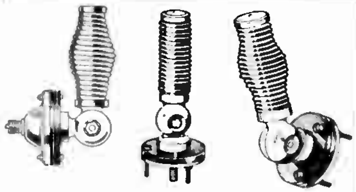

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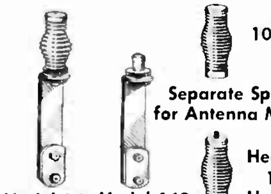

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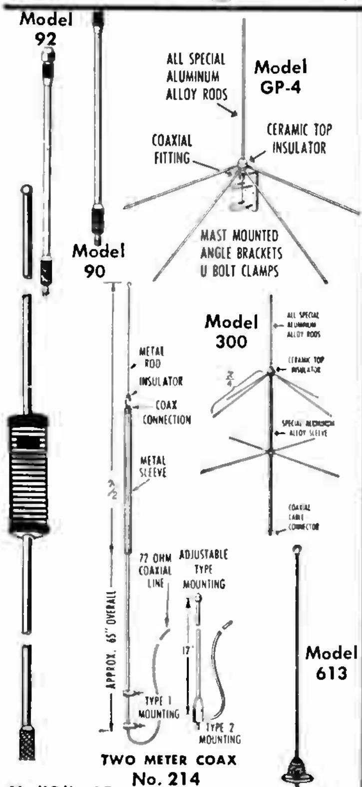

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ANTENNAS

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106-78S	78"	Plain End 3/16" Dia. (Fits Model 92 Ext.)	4.20
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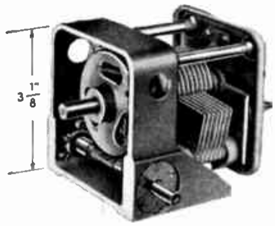

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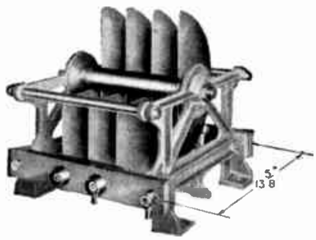

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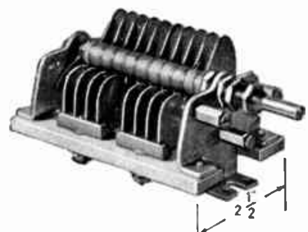

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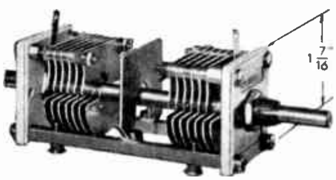

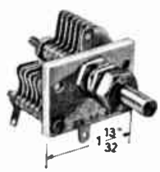

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