The radio amateur's handbook

THE STANDARD MANUAL OF AMATEUR RADIO COMMUNICATION

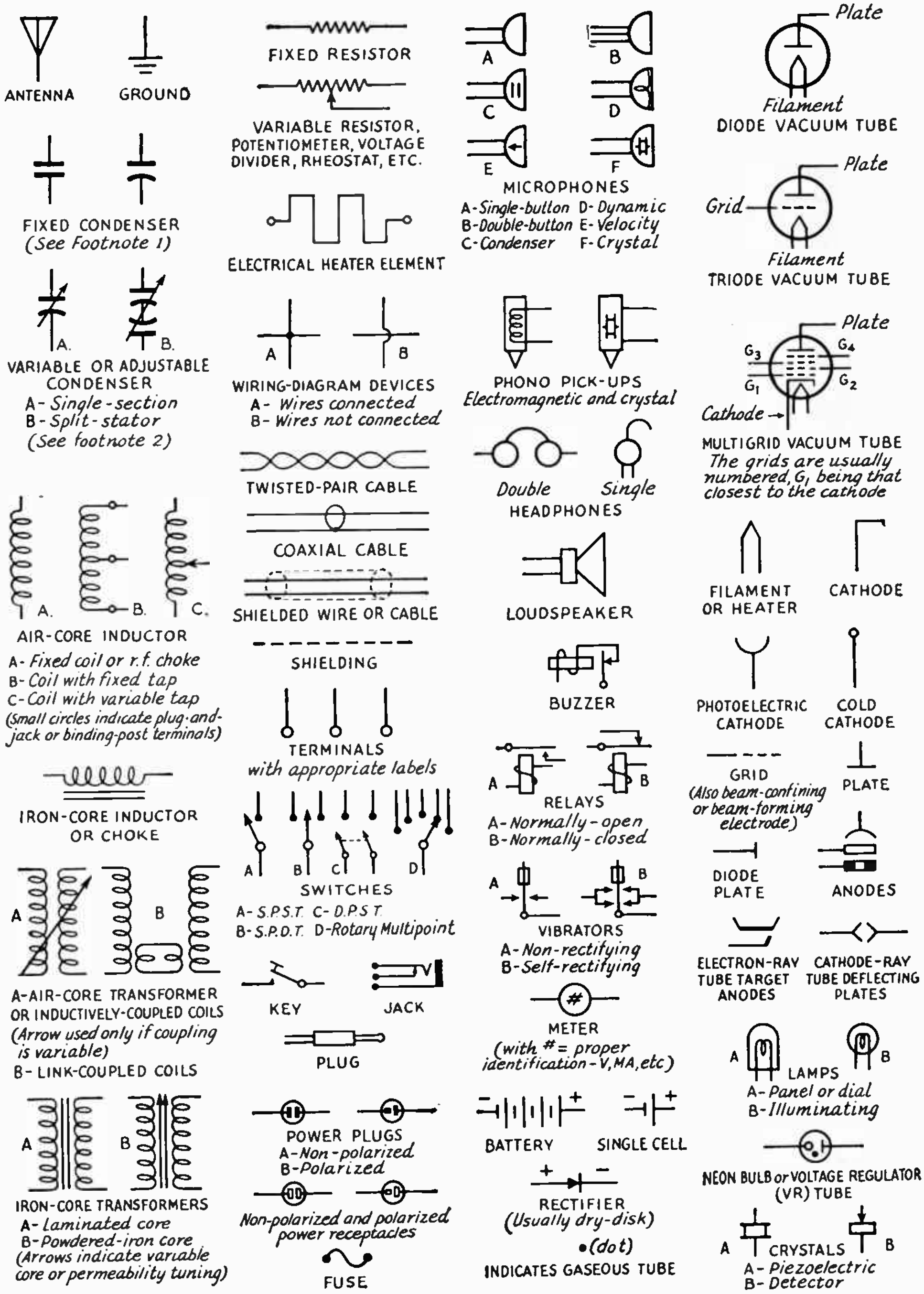
1947 EDITION

$1.25 IN U.S.A. PROPER

PUBLISHED BY THE AMERICAN RADIO RELAY LEAGUE

The Radio Amateur's Handbook

SCHEMATIC SYMBOLS USED IN CIRCUIT DIAGRAMS



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

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

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

TWENTY-FOURTH EDITION

1947

The Radio Amateur's Handbook

by the

HEADQUARTERS STAFF
OF THE AMERICAN
RADIO RELAY LEAGUE

published by

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Foreword

Twenty-one years ago — in 1926 — the first edition of *The Radio Amateur's Handbook* was presented to the amateur world. Produced by the amateur's own organization, the American Radio Relay League, and written with the needs of the practical amateur constantly in mind, its publication was eagerly greeted by the radio enthusiasts of that day. Subsequent editions have earned ever-increasing acceptance not only by amateurs but by all segments of the radio world, from students to engineers, servicemen to operators.

This wide dependence on the *Handbook*, evidenced by a total printing of over a million and a half copies, primarily is founded on its practical utility, its treatment of radio communication problems in terms of how-to-do-it rather than by abstract discussion and abstruse formulas.

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

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

When war came to this nation it was discovered by the military and other agencies that the *Handbook* was precisely what was needed to help make practical radiomen for the Army and Navy and to help those who were training themselves for wartime radio work. Not only was the *Handbook* used as a text or reference in many training programs, but it also provided source data for many service-written special courses. During the war years the training aspects have been given increasing emphasis — not, however, to the detriment of other long-established features, but rather by increasing the size and scope of the book.

With the constant editorial problem before us of gearing each year's edition to the needs of amateur radio of that year, as we perceive them, it has seemed best to leave intact in this edition the entire section on principles and design factors, large as that portion of the book grew during the war years. During this early postwar period there are many new people coming into amateur radio who need sound guidance, and it is a commonplace among practising amateurs that we all grew so rusty during the war that we have forgotten many of even the simple and fundamental things in radio. The preservation of this material in a connected and related manner seems to our staff to be the best possible way of presenting it during this transition period. The section of the book dealing with the construction of equipment, on the other hand, has been thoroughly revised in terms of postwar practices and postwar components. Many new pieces of apparatus, employing the best known amateur technique, have been designed and built for this year's edition, and proved by thorough testing, so that we are confident that other amateurs will find them reliable guides in their constructional projects.

A word about the reference system: It will be noted that each chapter is divided into sections and that these are numbered serially within each chapter. The number takes the form of two digits or groups separated by a hyphen. The first figure is the chapter number, the second the section number within the chapter. Cross-references in the text take such a form as (§ 4-7), for example, which means that the subject referred to will be found discussed in Chapter Four, Section 7. Throughout the book, illustrations are serially numbered within each chapter. Thus Fig. 1107 can be readily identified as the seventh illustration in Chapter Eleven. There is a carefully-prepared index at the rear of the book.

To a long-established reputation of indispensability in the amateur station of prewar days the *Handbook* now has added a proud record of participation in the national war effort. With the opening of the new postwar era in amateur communication, we earnestly hope that the present edition will succeed in bringing as much assistance and inspiration to amateurs and would-be amateurs as have its predecessors.

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

West Hartford, Conn.
December, 1946

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THE AMATEUR'S CODE

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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.

Chapter One

Amateur Radio

COUNTLESS thousands of persons all over the world have enjoyed the thrills and pleasures of amateur radio. This is a brief account of how it grew into the magnificently-useful institution it is today.

Amateur radio is as old as the art itself. There were amateurs before the present century. Shortly after the late Marconi astounded the world with his experiments proving that wireless telegraph messages actually could be sent, "amateurs" were attempting to duplicate his results. But amateur radio actually began when private citizens discovered this means for personal communication with others, and set about learning enough about "wireless" to build home-made stations. Its subsequent development may be divided into two phases, the period before 1917 and the years between that war and December 7, 1941. Plus, of course, the new phase now opening.

Amateur radio of pre-World War I bore little resemblance to radio as we know it today, except in principle. Transmitting and receiving equipment was of a type now long obsolete. No U. S. amateur had ever heard a foreign one nor had any foreigner ever reported an American signal. The oceans were an impenetrable wall. Cross-country communication could be accomplished only by relays. "Short waves" meant 200 meters; the entire spectrum below that was a vast silence undisturbed by any signals. By 1912, however, there were numerous Government and commercial stations and hundreds of amateurs; regulation was needed; and laws, licenses and wavelength specifications for the various services appeared.

"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 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 6,000 amateurs. Over 4,000 of them served in the armed forces during that war.

Today, few amateurs realize that World War I not only marked the close of the first phase of amateur development but came very 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 to get back on the air.

From the start, amateur radio took on new aspects. 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.

As DX became 1,000, then 1,500 and then 2,000 miles, amateurs began to dream of trans-Atlantic work. Could they get across? In December, 1921, in what has been called the greatest sporting event of all time, 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 W3IBZ, 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 conferences, in 1924, wisely obtained amateur bands not only at 80 meters but at 40, 20, 10 and even 5 meters.

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

From then until "Pearl Harbor," when U. S. amateurs were again closed down "for the duration," amateur radio thrilled with a series of unparalleled accomplishments. Countries all over the world came on the air, and the world total of amateurs passed the 100,000 mark. . . . ARRL representatives deliberated with the representatives of twenty-two other nations in Paris in 1925 where, on April 17th, the International Amateur Radio Union was formed — a federation of national amateur radio societies. . . . The League began issuing certificates to those who could prove they had worked all six continents. By 1941 over five thousand WAC certificates had been issued!

Amateur radio is a grand and glorious hobby but this fact alone would hardly merit such wholehearted support as was 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 4,000 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 co-operative activities which resulted, in 1925, in the establishment of the Naval Communications Reserve and the Army-Amateur Radio System. In World War II thousands of amateurs in the Naval Reserve were called to active duty, where they served with distinction, while many other thousands served in the Army, Air Forces, Coast Guard and Marine Corps. Altogether, more than 25,000 radio amateurs served in the armed forces of the United States. Other thousands were engaged in vital civilian electronic research, development and manufacturing.

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 co-operation with expeditions began in '23 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 United States 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, and the Southern California flood and Long Island-New England hurricane disaster in '38 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.

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 5-meter DX had been accomplished! 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. Many amateur developments have come to represent valuable contributions to the art. The complete record would fill a book! 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 for superheterodynes. During the war, thousands of skilled amateurs contributed their knowledge to the development of secret radio devices, both in Government and private laboratories. Equally as important, the prewar technical progress by amateurs provided the keystone for the development of modern military communications equipment.

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

The American Radio Relay League

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

The League is organized to represent the amateur in legislative matters. It is pledged to promote interest in two-way amateur communication and experimentation. It is interested in the relaying of messages by amateur radio. It is concerned with the advancement of the radio art. It stands for the maintenance of fraternalism and a high standard of conduct. One of its principal purposes is to keep amateur activities so well conducted that the amateur will continue to justify his existence.

The operating territory of ARRL is divided into fourteen U. S. and six Canadian divisions. The affairs of the League are managed by a Board of Directors. One director is elected every two years by the membership of each U. S. division, and a Canadian General Manager is elected every two years by the Canadian membership. These directors then choose the president and vice-president, who are also members of the Board. The managing secretary, treasurer and communications manager are appointed by the Board.

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 ARRL 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.

Among its other activities the League maintains, at its headquarters offices in West Hartford, Conn., a Communications Department concerned with the operating activities of League members. A large field organization is headed by a Section Communications Manager in each of the country's seventy-one sections. There are appointments for qualified members as Official Relay Station or Official 'Phone Station for traffic-handling; as Official Observer for monitoring frequencies and the quality of signals; as Route Manager and

'Phone Activities Manager for the establishment of trunk lines and networks; as Emergency Coördinator for the promotion of amateur preparedness to cope with natural disasters. Mimeographed bulletins keep appointees informed of the latest developments. Special activities and contests promote operating skill and thereby add to the ability of amateur radio to function "in the public interest, convenience and necessity." A special section is reserved each month in *QST* for amateur news from every section of the country.

Amateur Licensing in the United States

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

A radio amateur is a duly authorized person interested in radio technique solely with a personal aim and without pecuniary interest. Amateur operator licenses are given to U. S. citizens who pass an examination on operation and apparatus and on the provisions of law and regulations affecting amateurs, and who demonstrate ability to send and receive code at 13 words per minute. Station licenses are granted only to licensed operators and permit communication between such stations for amateur purposes, i.e., for personal noncommercial aims flowing from an interest in radio technique. An amateur station may not be used for material compensation of any sort nor for broadcasting. Narrow bands of frequencies are allocated exclusively for use by amateur stations. Transmissions may be on any frequency within the assigned bands. All the frequencies may be used for c.w. telegraphy and some are available for radio-telephony by any amateur, while others are reserved for radiotelephone use by persons having at least a year's experience and who pass the examination for a Class A license. The input to the final stage of amateur stations is limited to 1,000 watts and on frequencies below 60 Mc. must be adequately-filtered direct current. Emissions must be free from spurious radiations. The licensee must provide for measurement of the transmitter frequency and establish a procedure for checking it regularly. A complete log of station operation must be maintained, with specified data. The station license also authorizes the holder to operate portable and portable-mobile stations on certain frequencies, subject to further regulations. An amateur station may be operated only by an amateur operator licensee, but any licensed amateur operator may operate any amateur station. All radio licensees are subject to penalties for violation of regulations.

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

The Amateur Bands

During 1946, FCC announced its final determination of postwar frequency allocations above 25 Mc., with certain alterations and additions to prewar amateur frequencies. Similarly the Commission announced proposed allocations below 25 Mc., these still being under consideration as this is written in the late summer of 1946. The final recommendations for the region below 25 Mc. will then be subject to further consideration at the next international conference.

Meanwhile, as of our press date, the following are the postwar amateur bands:

3,500– 4,000 kc.	50– 54 Mc.	2,300– 2,450 Mc.
7,000– 7,300 "	144– 148 "	3,300– 3,500 "
14,000–14,400 "	235– 240 "	5,650– 5,850 "
27,185–27,455 "	420– 450 "	10,000–10,500 "
28,000–29,700 "	1,215–1,295 "	21,000–22,000 "

The future of the prewar amateur band at 1.75 Mc. has not been determined as of this date but, at the least, it is expected that the amateur, along with other services, will be given nonexclusive rights in the frequencies 1750–1800 kc. for the maintenance of emergency networks and necessary tests and drills incident thereto. There is also a pending proposal for a new amateur band at 21 Mc. but this will not likely be made available until after the agreement of the next world conference.

It should be carefully noted that as of this writing the 420 Mc. band has not yet been opened in its entirety to amateur use, being still partly in use by other services as a result of the war. Moreover, the portion of each band available for 'phone operation is customarily varied from time to time in accordance with changes in amateur operational habits. In such respects each amateur should keep himself currently informed by consulting *QST* or by writing ARRL for latest information.

Chapter Two

Fundamentals

2-1 Fundamentals of a Radio System

THE BASIS of radio communication is the transmission of electromagnetic waves through space. The production of suitable waves constitutes radio transmission, and their detection, or conversion at a distant point into the intelligence put into them at the originating point, is radio reception. There are several distinct processes involved in the complete chain. At the transmitting point, it is necessary first to generate power in such form that when it is applied to an appropriate radiator, called the *antenna*, it will be sent off into space in electromagnetic waves. The message to be conveyed must be superimposed on that power by suitable means, a process called *modulation*.

As the waves spread outward from the transmitter they rapidly become weaker, so at the receiving point an antenna is again used to abstract as much energy as possible from them as they pass. The wave energy is transformed into an electric current which is then amplified, or increased in amplitude, to a suitable value. Then the modulation is changed back into the form it originally had at the transmitter. Thus the message becomes intelligible.

Since all these processes are performed by electrical means, a knowledge of the basic principles of electricity is necessary to understand them. These essential principles are the subject of the present chapter.

2-2 The Nature of Electricity

Electrons — All matter — solids, liquids and gases — is made up of fundamental units called *molecules*. The molecule, the smallest subdivision of a substance retaining all its characteristic properties, is constructed of *atoms* of the elements comprising the substance.

Atoms in turn are made up of particles, or charges, of electricity, and atoms differ from each other chiefly in the number and arrangement of these charges. The atom has a nucleus containing both "positive" and "negative" charges, with the positive predominating so that the nature of the nucleus is positive. The charges in the nucleus are closely bound together. Exterior to the nucleus are negative charges — *electrons* — some of which are not so closely bound and can be made to leave the vicinity of the nucleus without too much urging. These electrons whirl around the nucleus like the planets around the sun, and their orbits are not random paths but geometrically-regular ones determined by the charges on the nucleus and the number of electrons. Ordinarily the atom is electrically neutral, the outer negative electrons balancing the positive nucleus, but when something disturbs this balance electrical activity becomes evident, and it is the study of what happens in this unbalanced condition that makes up electrical theory.

Electrons are exceedingly small particles — so small that many billions of them must act together before measurable electrical effects are observed.

Insulators and conductors — Materials which will readily give up an electron are called *conductors*, while those in which all the electrons are firmly bound in the atom are called *insulators*. Most metals are good conductors, as are also acid or salt solutions. Among the insulators are such substances as wood, hard rubber, bakelite, quartz, glass, porcelain, textiles, and many other non-metallic materials.

Resistance — No substance is a perfect conductor — a "perfect" conductor would be one in which an electron could be detached from the atom without the expenditure of energy — and there is also no such thing as a perfect insulator. The measure of the difficulty in moving an electron by electrical means is called *resistance*. Good conductors have low resistance, good insulators very high resistance. Between the two are materials which are neither good conductors nor good insulators, but they are nonetheless useful since there is often need for intermediate values of resistance in electrical circuits.

Conduction — Under the influence of a suitable force — that is, an electric *field* — electrons tend to move. If the substance is one in which electrons can be detached from atoms as explained above, these electrons will move through the substance. This is the process of *conduction*, and the moving electrons constitute an electric *current*. The intensity of the current depends upon the amount of force exerted on the electrons, and also upon the resistance of the material through which they are moving.

Strictly speaking, this description applies only to conduction through solid substances. However, conduction in liquids and gases, although different in detail, is similar in principle. These cases are treated later in chapter.

Circuits — A circuit is simply a complete path along which electrons can transmit their charges. There will normally be a source of energy (a battery, for instance) and a *load* or portion of the circuit where the current is made to do work. There must be an unbroken path through which the electrons can move, with the source of energy acting as an electron pump and sending them around the circuit. The circuit is said to be *open* when no charges can move, because of a break in the path. It is *closed* when no break exists — when switches are closed and all connections are made.

2-3 Static Electricity

The electric charge — Many materials that have a high resistance can be made to acquire a charge (surplus or deficiency of electrons) by mechanical means, such as friction. The familiar crackling when a hard-rubber comb is run through hair on a dry winter day is an example of an electric charge generated by friction. Objects can have either a surplus or a deficiency of electrons — a surplus of electrons is called a *negative* charge: a lack of them is called a *positive* charge. The kind of charge is called its *polarity*. A negatively charged object is frequently called a negative *pole*, while a positively charged object similarly is called a positive pole.

Attraction and repulsion — Unlike charges (one positive, one negative) exert an attraction on each other. This can be demonstrated by giving charges of opposite polarity to two very light, well-insulated conductors, such as bits of metal foil suspended from dry thread (Fig. 201). Pith balls covered with foil frequently are used in this experiment.

When the two charged objects are brought close together, it will be observed that they will be attracted to each other. If the charges are equal and the charged bodies are permitted to touch, the surplus electrons on the negatively charged object will transfer to the positively charged object (i.e., the one deficient in electrons) and the two charges will neutralize,

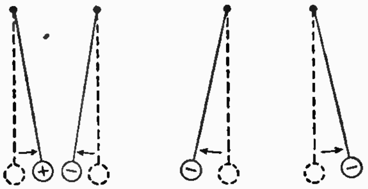


Fig. 201 — Attraction and repulsion of charged objects, as demonstrated by the familiar pith-ball experiment.

leaving both bodies uncharged. If the charges are not equal, the weaker charge neutralizes an equal amount of the stronger when the two bodies touch, upon which the excess of the stronger charge distributes itself over both. Both bodies then have charges of the same polarity, and a force of repulsion is exercised between them. Consequently, the bits of foil tend to spring away from each other. *Unlike charges attract, like charges repel.*

Electrostatic field — From the foregoing it is evident that an electric charge can exert a force through the space surrounding the charged object. The region in which this force is exerted is considered to be pervaded by an *electrostatic field*, this concept of a field being adopted to explain the "action at a distance" of the charge. The field is pictured as consisting of *lines of force* originating on the charge and

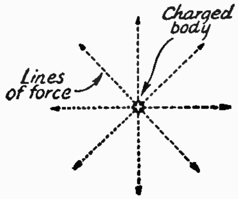


Fig. 202 — Lines of force from a charged object extend outward radially. Although only two dimensions are shown, the field extends in all directions from the charge, and should be visualized in three dimensions.

spreading in all directions, finally terminating on other charges of opposite polarity. These other charges may be a very large distance away. The number of lines of force per unit area is, however, a measure of the intensity of the field.

The general picture of a charged object in isolated space is shown in Fig. 202. This is an idealized situation, since in practice the charged object could not be completely isolated. The presence of other charges, or simply of insulators or conductors, in the vicinity will greatly change the configuration of the field. The direction of the field, as indicated by the arrowheads, is away from a positively charged object; if the charge were negative, the direction would be toward the charge.

It should be understood that the field picture as represented above is merely a convenient method of explaining observed effects, and is not to be taken too literally. The electric force does not consist of separate lines like strings or rods; instead, it completely pervades the medium through which the force is exerted. With this understanding in mind, it is *convenient* to talk of lines of force and to measure the field intensity in terms of number of lines per unit area.

The intensity of the field dies away with distance from the charged object in a manner determined by its shape and the circumstances of its surroundings. In the case of an isolated charge at a point (an infinitesimally small object), the field strength is inversely proportional to the square of the distance. However, this relationship is not true in many other cases; in some important practical applications the field intensity is inversely proportional to the distance involved, and not to its square.

Electrostatic induction — If a piece of conducting material is brought near a charged object, the field will exert a force on the electrons of the metal so that those free to move will do so. If the object is positively charged, as indicated in Fig. 203, the free electrons will move toward the end of the conductor nearest the charged body, leaving a deficiency of electrons at the other end. Hence, one end of the conductor becomes negatively charged while the other end has an equal positive charge. The lines of force from the charged body terminate on the conductor, where sufficient electrons accumulate to provide an electric intensity equal and opposite to that of the field at that point. Because of this effect, the electrostatic field inside the conductor is completely neutralized by the induced charge; in other words, the field does not penetrate the conductor. In radio work this principle provides the means by which electrostatic fields may be excluded from regions where they are not wanted.

Charges induced in a conductor as shown in Fig. 203-A are held in existence by the field from the charged object. On taking the conductor out of the field the electrons will redistribute themselves so that the charges disappear. However, if the conductor is connected to the earth through a wire while under the influence of the field, as shown in Fig. 203-B, the induced positive charge will tend to move as far as possible from the source of the field (that is, electrons will flow from the earth to the conductor). If the grounding wire is then removed, the conductor will be left with an excess of electrons and will have acquired a "permanent" charge — permanent, that is, so long as the conductor is well enough insulated to prevent the charge from escaping to earth or to other objects. The polarity of the induced charge always is opposite to the polarity of the charge which set up the original field.

Energy in the electrostatic field — The expenditure of energy is necessary to place an electrical charge upon an object and thus establish an electrostatic field. Once the field is established and is constant, no further expenditure of energy is required. The energy supplied to establish the field is stored in the field; thus the field represents *potential* energy (that is, energy available for use). The potential energy is acquired in the same way that potential energy is given any object (a 10-pound weight, for instance) when it is lifted against the gravitational pull of the earth. If

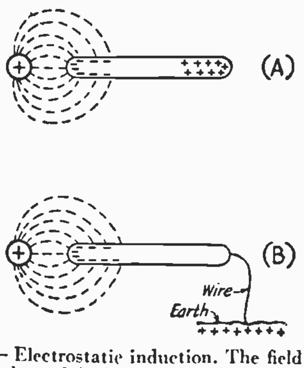


Fig. 203 — Electrostatic induction. The field from the positively charged body attracts electrons, which accumulate to form a negative charge. The opposite end of the conductor consequently acquires a positive charge. This charge may be "drained off" to earth as shown at B.

the weight is allowed to drop, its potential energy is changed into the energy of motion. Similarly, if the electrostatic field is made to disappear its potential energy is transformed into a movement of electrons; in other words, into an electric current.

The potential energy of the lifted weight is measured by its weight and the distance it is lifted; that is, by the work done in lifting it. Similarly, the potential energy (called simply *potential*) of the electrostatic field at any point is measured by the work done in moving a charge of specified value to that point, against the repulsion of the field. In practice, absolute potential is of less interest than the *difference of potential* between two points in the field.

Potential difference — If two objects are charged differently, a potential difference exists between them. Potential difference is measured by an electrical unit called the *volt*. The greater the potential difference, the higher (numerically) the voltage. This voltage exerts an electrical pressure or *force* as explained above, and is often called *electromotive force* or, simply, *e.m.f.* It is not necessary to have unlike charges in order to have a difference of potential; both, for instance, may be negative, so long as one charge is more intense than the other. From the viewpoint of the stronger charge, the weaker one appears to be positive in such a case, since it has a smaller number of excess electrons; in other words, its *relative polarity* is positive. The greater the potential difference, the more intense is the electrostatic field between the two charged objects.

Capacity — More work must be done in moving a given charge against the repulsion of a strong field than against a weak one; hence, potential is proportional to the strength of the field. In turn, field strength is proportional to the charge or quantity of electricity on the charged object, so that potential also is proportional to charge. By inserting a suitable constant, the proportionality can be changed to an equality:

$$Q = CE$$

where Q is the quantity of charge, E is the potential, and C is a constant depending upon the charged object (usually a conductor) and its surroundings and is called the *capacity* of the object. Capacity is the ratio of quantity of charge to the potential resulting from it, or

$$C = \frac{Q}{E}$$

When Q is in coulombs and E in volts, C is measured in *farads*. A conductor has a capacity of one farad when the addition of one coulomb to its charge raises its potential by one volt.

The farad is much too large a unit for practical purposes. In radio work, the *microfarad* (one millionth of a farad) and the *micro-microfarad* (one millionth of a microfarad) are the units most frequently used. They are abbreviated $\mu fd.$ and $\mu\mu fd.$, respectively.

The capacity of a conductor in air depends upon its size and shape. A given charge on a small conductor results in a more intense electrostatic field in its vicinity than the same charge on a larger conductor. This is because the charge distributes itself over the surface, hence its density (the quantity of electricity per unit area) is smaller on the larger conductor. Consequently, the potential of the larger conductor is smaller, for the same amount of charge. In other words, its capacity is greater because a greater charge is required to raise its potential by the same amount.

Condensers — If a grounded conductor, *A* (Fig. 204), is brought near a second conductor, *B*, which is charged, the former will acquire a charge by electrostatic induction. Since the charge on *A* is opposite in polarity to that on *B*, the field set up by the induced charge on *A* will oppose the original field set up by the charge on *B*, hence the potential of *B* will be lowered. Because of this, more charge must be placed on *B* to raise its potential to its original value; in other words, its *capacity has been increased* by the presence of the second conductor. The combination of the two conductors separated by a dielectric is called a *condenser.*

The capacity of a condenser depends upon the areas of the conductors, as before, and also becomes greater as the distance between the conductors is decreased, since, with a fixed amount of charge, the potential difference between them decreases as they are moved closer together.

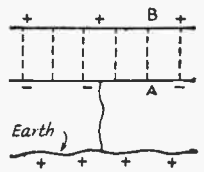


Fig. 204 — The principle of the condenser.

If insulating or dielectric material other than air is inserted between the conductors, it is found that the potential difference is lowered still more — that is, there is a further increase in capacity. This lowering of the potential difference is considered to be the result of *polarization* of the dielectric. By this it is meant that the molecules of the substance tend to be distorted under the influence of the electrostatic field in such a way that the negative charges within the molecule are drawn toward the positively charged conductor, leaving the other end of the molecule with a positive charge facing the negatively charged conductor. Since the electrons are firmly bound in the atoms of the dielectric, there is no flow of current and the total charge on each atom is still zero, but there is a *tendency* toward separation which causes a reaction on the electrostatic field. The dielectric of a charged condenser thus is under mechanical stress, and if the potential difference between the plates of the condenser is great enough the dielectric may break down mechanically and electrically.

The ratio of the capacity of a condenser with a given dielectric material between its plates to the capacity of the same condenser with air as a dielectric is called the *specific inductive capacity* of the dielectric, or, probably more commonly, the *dielectric constant.* Strictly speaking, the comparison should be made to empty space (i.e., a vacuum) rather than to air, but the dielectric constant of air is so nearly that of a vacuum that the practical difference is negligible. A table of dielectric constants is given in Chapter Twenty.

Condensers have many uses in electrical and radio circuits, all based on their ability to store energy in the electric field when a potential difference or voltage is caused to exist between the plates — energy which later can be released to perform useful functions.

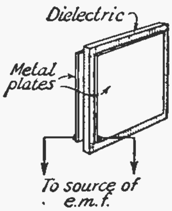


Fig. 205 — A simple condenser, consisting of two metal plates separated by dielectric material.

¶ 2-4 The Electric Current

Conduction in metals — When a difference of potential is maintained between the ends of a metallic conductor, there is a continuous drift of electrons through the conductor toward the end having a positive potential (relative polarity positive). This electron drift constitutes an electric current through the metal (§ 2-2). The speed with which the electron movement is established is very nearly the speed of light (300,000,000 meters, or approximately 186,000 miles, per second), so that the current is said to travel at nearly the speed of light. By this it is meant that the time interval between the application of the electromotive force and the flow of current in all parts of a circuit, even one extending over hundreds of miles, is negligible. However, the individual electrons do not move at anything approaching such a speed. The situation is similar to that existing when a mechanical force is transmitted by means of a rigid rod. A force applied to one end of the rod is transmitted practically instantaneously to the other end, even though the rod itself moves relatively slowly or not at all.

The magnitude of the electric current is the rate at which electricity is moved past a point in the circuit. If the rate is constant, then the current is equal to the quantity of electricity moved past a given point in some selected time interval. That is,

$$I = \frac{Q}{t}$$

where I is the intensity or magnitude of the current, Q is the quantity of electricity, and t is the time. If Q is in coulombs and t in seconds, the unit for I is called the *ampere*. One ampere of current is equal to one coulomb of electricity moving or "flowing" past a given point in a circuit in one second.

The currents used by different electrical devices vary greatly in magnitude. The current which flows in an ordinary 60-watt lamp, for instance, is about one-half ampere. the current in an electric iron is about 5 amperes, and that in a radio tube may be as low as 0.001 ampere.

When a current flows through a metallic conductor there is no visible or chemical effect on the conductor. The only physical effect is the heat developed (§ 2-2) as the result of energy loss in the conductor. Under normal conditions the rate at which heat is generated and that at which it is radiated by the conductor will quickly reach equilibrium. However, if the heat is developed at a more rapid rate than it can be radiated, the temperature will continue to rise until the conductor burns or melts.

Experimental measurements have shown that the current which flows in a given metallic conductor is directly proportional to the applied e.m.f., so long as the temperature of the conductor is held constant. There is no e.m.f. so small but that some current will flow as a result of its application to a metallic conductor.

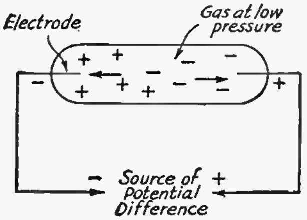


Fig. 206 — Illustrating conduction through a gas at low pressure. Positive ions are attracted to the negative electrode, while electrons are attracted to the positive electrode. This takes place only after the gas is ionized.

Gaseous conduction — In any gas or mixture of gases (such as air, for example) there are always some free electrons — that is, electrons not attached to an atom — and also some atoms lacking an electron. Thus there are both positively and negatively charged particles in the gas, as well as many neutral atoms. An atom lacking an electron is called a *positive ion,* while the free electron is called a *negative ion.* The term *ion* is, in fact, applied to any elemental particle which has an electric charge.

If the gas is in an electric field, the free electrons will be attracted toward the source of positive potential and the positive ions will be attracted toward the source of negative potential. If the gas is at atmospheric pressure neither particle can travel very far before meeting an ion of the opposite kind, when the two combine to form a neutral atom. Since a neutral atom is not affected by the electric field, there is no flow of current through the gas.

However, if the gas is enclosed in a glass container in which two separate metal pieces called *electrodes* are sealed, and the gas pressure is then reduced by pumping out most of the gas, a different set of conditions results. At low pressure there is a comparatively large distance between each atom, and when an electric field is established by applying a difference of potential to the electrodes the ions can travel a considerable distance before meeting another ion or atom. The farther the ion travels the greater the velocity it acquires, since the effect of the field is to accelerate its motion. If the field is strong enough the ions will acquire such velocity that when one happens to collide with a neutral atom the force of the collision will knock an electron out of the atom, so that this atom also becomes *ionized.* The process is cumulative, and the freed electrons are attracted to the positive electrode while the positive ions are attracted to the negative electrode. This movement of charged particles constitutes an electric current through the gas.

Since an ion must acquire a certain velocity before it can knock an electron out of a neutral atom, a definite field strength is required before conduction can take place in a gas. That is, a certain value of potential difference, called the *ionizing potential,* must be applied to the electrodes. If less voltage is applied, the gas does not ionize and the current is negligible. On the other hand, once the gas is ionized an increase in potential does not have much effect on the current, since the ions already have sufficient velocity to maintain the ionization. The ionizing potential required depends upon the kind of gas and the pressure. Ionization is usually accompanied by a colored glow, different gases having different characteristic colors.

Current flow in liquids — A very large number of chemical compounds have the peculiar characteristic that, when they are put into solution, the component parts become ionized. For example, common table salt (sodium chloride), each molecule of which is made up of one atom of sodium and one of chlorine, will, when put into water, break down into a sodium ion (positive, with one electron deficient) and a chlorine ion (negative, with one excess electron). This can only occur so long as the salt is in solution — take away the

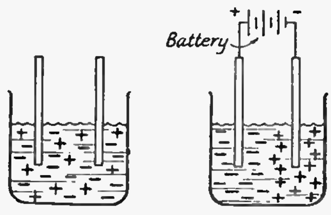


Fig. 207 — Electrolytic conduction. When an e.m.f. is applied to the electrodes, negative ions are attracted to the positively charged plate and positive ions to the negatively charged plate. The battery, which is the source of the e.m.f., is indicated by its customary symbol.

water and the ions are recombined into the neutral sodium chloride. This spontaneous *dissociation* in solution is another form of ionization. If two wires with a difference of potential between them are placed in the solution, the negative wire will attract the positive sodium ions while the positive wire will attract the negative chlorine ions and an electric current will flow through the solution. When the ions reach the wires the electron surplus or deficiency will be remedied, and a neutral atom will be formed.

In this process, the water is decomposed into its gaseous constituents, hydrogen and oxygen. The energy used up in decomposing the water and in moving the ions is supplied by the source of potential difference. The energy used in decomposing the water is equivalent to an opposing e.m.f., of the order of a volt or two. If this constant "back voltage" is subtracted from the applied voltage, it is found that the current flowing through a given solution, or *electrolyte*, is proportional to the difference between the two voltages.

Current flow in vacuum — If a suitable metallic conductor is heated to a high temperature in a vacuum, electrons will be emitted from the surface. The electrons are freed from this *filament* or *cathode* because it has been

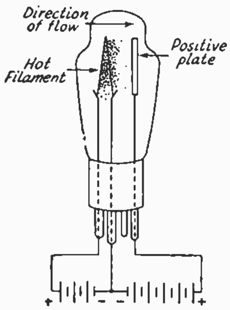


Fig. 208 — Conduction by thermionic emission in a vacuum tube. One battery is used *only* to heat the filament to a temperature where it will emit electrons. The other battery places a potential on the plate which is positive with respect to the filament, and as a result the electrons are attracted to the plate. The electron flow from filament to plate completes the circuit.

heated to a temperature that gives them sufficient energy of motion to allow them to break away from the surface. The process is called *thermionic electron emission*. Now, if a metal plate is placed in the vacuum and given a positive charge with respect to the cathode, this plate or *anode* will attract a number of the electrons that surround the cathode. The passage of the electrons from cathode to anode constitutes an electric current. All thermionic vacuum tubes depend for their operation on the emission of electrons from a hot cathode.

Since the electrons emitted from the hot cathode are negatively charged, it is evident that they will be attracted to the plate only when the latter is at a positive potential with respect to the cathode. If the plate is negatively charged with respect to the cathode the electrons will be repelled back to the cathode, hence no current will flow through the vacuum. Consequently, a thermionic vacuum tube conducts current *in one direction only*. When the plate is positive, it is found that (if the potential is not too large) the current increases with an increase in potential difference between the plate and cathode. However, the relationship between current and applied voltage is not a simple one. If the voltage is made large enough all the electrons emitted by the cathode will be drawn to the plate, and a further increase in voltage therefore cannot cause a further increase in current. The number of electrons emitted by the cathode depends upon the temperature of the cathode and the material of which it is constructed.

Direction of current flow — Use was being made of electricity for a long time before its electronic nature was understood. While it is now clear that current flow is a drift of negative electrical charges or electrons toward a source of positive potential, in the era preceding the electron theory it was *assumed* that the current flowed from the point of higher positive potential to a point of lower (i.e., less positive or more negative) potential. While this assumption turned out to be wholly wrong, it is still customary to speak of current as flowing "from positive to negative" in many applications. The practice often causes confusion, but this distinction between "current" flow and "electron" flow often must be taken into account. If electron flow is specifically mentioned there can be, of course, no doubt as to the meaning; but when the direction of *current* flow is specified, it may be taken, by convention, as being opposite to the direction of electron movement.

Primary cells — If two electrodes of dissimilar metals are immersed in an electrolyte, it is found that a small difference of potential exists between the electrodes. Such a combination is called a *cell*. If the two electrodes are connected together by a conductor external to the cell, an electric current will flow between them. In such a cell, chemical energy is converted into electrical energy. The difference of potential arises as a result of the fact that material from one or both of the electrodes goes into solution in the electrolyte, and in the process ions are formed in the vicinity of the electrodes. The electrodes acquire charges because of the electric field associated with the charged ions. The difference of potential between the electrodes is principally a function of the metals used, and is more or less independent of the kind of electrolyte or the size of the cell.

When current is supplied to an external circuit, two principal effects occur within the cell. The negative electrode (negative as viewed from outside the cell) loses weight as its material is used up in furnishing energy, and hydrogen bubbles form on the positive electrode. Since the gas bubbles are non-conducting, their accumulation tends to reduce the effective area of the positive electrode, and consequently reduces the current. The effect is cumulative, and eventually the electrode will be completely covered and no further current can flow. This effect is called *polarization*. If the bubbles are removed, or prevented from forming by chemical means, polarization is reduced and current can flow as long as there is material in the negative electrode to furnish the energy. A chemical which prevents the formation of hydrogen bubbles in a cell is called a *depolarizer*.

In addition to polarization effects, a cell has a certain amount of *internal resistance* because of the resistance of the electrodes and the electrolyte and the contact resistance between the electrodes and electrolyte. The internal resistance depends upon the materials used and the size and electrode spacing of the cell. Large cells with the electrodes close together will have smaller internal resistance than small cells made of the same materials.

A collection of cells connected together is called a *battery*. The term battery also is applied (although incorrectly) to a single cell.

Dry cells — The most familiar form of primary cell is the *dry cell*. Like the elementary type of cell just described, it has a liquid electrolyte, but the liquid is mixed with other materials to form a paste. The cell therefore can be used in any position and handled as though it actually were dry.

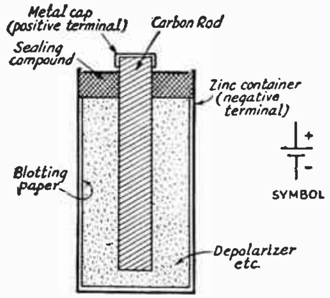


Fig. 209 — Construction of a dry cell.

The construction of an ordinary dry cell is shown in Fig. 209. The container is the negative electrode and is made of zinc. Next to it is a section of blotting material saturated with the electrolyte, a solution of sal ammoniac. The positive electrode is a carbon rod, and the space between it and the blotting paper is filled with a mixture of carbon, manganese dioxide (the depolarizer) and the electrolyte. The top is filled with sealing compound to prevent evaporation, since the cell will not work when the electrolyte drys out. The e.m.f. of a dry cell is about 1.5 volts.

Dry cells are made in various sizes, depending upon the current which they will be called upon to furnish. The construction frequently varies from that shown in Fig. 209, although in general the basic materials are the same in all dry cells. Batteries of small cells are assembled together as a unit for furnishing plate current for the vacuum tubes used in portable receiving sets; such "B" batteries, as they are called, can supply a current of a few hundredths of an ampere continuously. Larger cells, such as the common "No. 6" cell, can deliver currents of a fraction of an ampere continuously, or currents of several amperes for very short periods of time. The total amount of energy delivered by a dry cell is larger when the cell is used only intermittently, as compared with continuous use. The cell will deteriorate even without use, and should be put into service within a year or so from the time it is manufactured. The period during which it is usable (without having been put in service) is known as the "shelf life" of the cell or battery.

Secondary cells — The types of cells just described are known as *primary cells*, because the electrical energy is obtained directly from chemical energy. In some types of cells the chemical actions are reversible; that is, forcing a current through the cell, in the opposite direction to the current flow when the cell is delivering electrical energy, causes just the reverse chemical action. This tends to restore the cell to its original condition, and electrical energy is transformed into chemical energy. The process is called *charging* the cell. A cell which must first be charged before it can deliver electrical energy is called a *secondary cell*.

A simple form of secondary cell can be made by immersing two lead electrodes in a dilute solution of sulphuric acid. If a current is forced through the cell, the surface of the electrode which is connected to the positive terminal of the charging e.m.f. will be changed to lead peroxide and the surface of the electrode connected to the negative terminal will be changed to spongy lead. After a period of charging the charging source can be disconnected, and the cell will be found to have an e.m.f. of about 2.1 volts. It will furnish a small current to an external circuit for a period of time. This *discharge* of electrical energy is accompanied by chemical action which forms lead sulphate on both electrodes. When the lead peroxide and spongy lead are converted to lead sulphate there is no longer a difference of potential, since both electrodes are now the same material, and the cell is completely discharged.

The lead storage battery — The most common form of secondary cell is the lead storage cell. The common storage battery for automobile starting consists of three such cells connected together electrically and assembled in a single container. The principle of operation is similar to that just described, but the construction of the cell is considerably more complicated. To obtain large currents it is necessary to use electrodes having a great deal of surface area and to put them as close together as possible. The electrodes are made in the form of rectangular flat plates, consisting of a latticework or grid of lead or an alloy of lead. The interstices of the latticework are filled with a paste of lead oxide. The electrolyte is a solution of sulphuric acid in water. When the cell is charged, the lead oxide in the positive plate is converted to lead peroxide and that in the negative plate to spongy lead. To obtain high current capacity, a cell consists of a number of positive plates, all connected together, and a number of negative plates likewise connected together. They are arranged as shown in Fig. 210, with alternate negative and positive plates kept from touching by means of thin *separators* of insulating material, generally treated wood or perforated hard rubber. The separators preferably should be porous, so that the electrolyte can pass through them freely; thus they do not impede the passage of current from one plate to the next. There is always one extra negative plate in such an assembly, because the active material in the positive plate expands when the cell is being charged and if all the expansion took place on one side the plate would be distorted out of shape.

The e.m.f. of a fully charged storage cell is about 2.1 volts. When the e.m.f. drops to about 1.75 volts on discharge, the cell is considered to be completely discharged. Discharge beyond this limit may result in the formation of so much lead sulphate on the plates that the cell cannot be recharged, since lead sulphate is an insulator. During the charging process water in the electrolyte is used up, with the result that the sulphuric acid solution becomes more concentrated. The higher concentration increases the specific gravity of the solution, so that the specific gravity may be used to indicate the state of the battery with respect to charge. In the ordinary lead storage cell the solution is such that a specific gravity of 1.285 to 1.300 indicates a fully charged cell, while a discharged cell is indicated by a specific gravity of 1.150 to 1.175. The specific gravity can be measured by means of a *hydrometer*, shown in Fig. 211. For use with portable batteries, the hydrometer usually consists of a glass tube fitted with a syringe so that some of the electrolyte can be drawn from the cell into the tube. The hydrometer float is a smaller glass tube, air-tight and partly filled with shot to make it sink into the solution. The lower the specific gravity of the solution, the farther the float sinks into it. A graduated scale on the float shows the specific gravity directly, being read at the level of the solution.

Storage cells are rated in *ampere-hour capacity*, based on the number of amperes which can be furnished continuously for a stated period of time. For example, the cell may have a rating of 100 ampere-hours at an 8-hour discharge rate. This means that the cell will deliver 100/8 or 12.5 amperes continuously for 8 hours after having been fully charged. The ampere-hour capacity of a cell will vary with the discharge rate, becoming smaller as the rated time of discharge is made shorter. It also depends upon the size of the plates and their number. In automobile-type batteries the dimensions of the plates are fairly well standardized, so that the ampere-hour capacity is chiefly determined by the number of plates in a cell. It is, therefore, common practice to speak of "11-plate," "15-plate," etc., batteries as an indication of the battery capacity.

Lead storage batteries must be kept fully charged if they are to stay in good condition. If a discharged battery is left standing idle, lead sulphate will form on the plates and eventually the battery will be useless. When the battery is being charged, hydrogen bubbles are given off by the electrolyte which, in bursting at the surface, throw out fine drops of the electrolyte. This is called "gassing." The sulphuric-acid solution spray from gassing will attack many materials, and consequently care must be used to see that it is not permitted to fall on near-by objects. It should also be wiped off the battery itself.

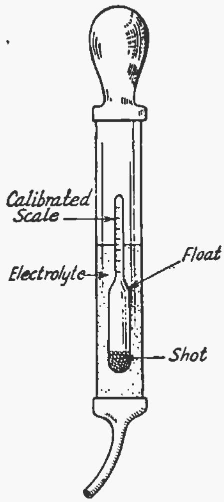


Fig. 211—The hydrometer, a device with a calibrated scale for measuring the specific gravity of the electrolyte, used to determine the state of charge of a lead storage battery.

A lead battery may be charged at its nominal discharge rate; i.e., a 100-ampere-hour battery, 8-hour rating, can be charged at 100/8, or 12.5 amperes. The charging voltage required is slightly more than the output voltage of the cell. The preferred method is to charge at the full rate until the cells start to "gas" freely, after which the charging rate should be dropped to about half its initial value until the battery is fully charged, as indicated by the hydrometer reading. Alternatively, the battery may be charged from a constant-potential source (about 2.3 volts per cell), when the rise of terminal voltage of the battery as it accumulates a charge will automatically "taper" the charging rate.

The solution in a lead storage battery will freeze at a temperature of about zero degrees Fahrenheit when the battery is discharged, but a fully charged battery will not freeze until the temperature reaches about 90 degrees below zero. Keeping the battery

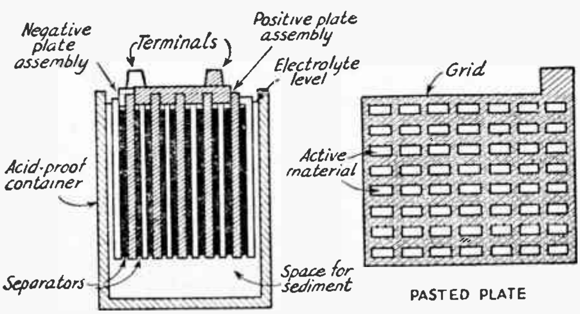


Fig. 210—Details of typical lead storage-battery construction.

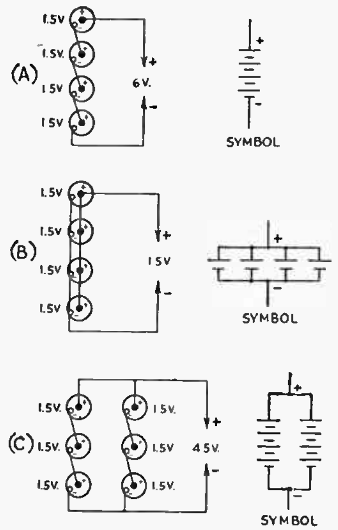


Fig. 212 — Series, parallel, and series-parallel connection of cells. Series connection increases the total voltage without changing current capacity; parallel connection increases current capacity without increasing voltage.

charged therefore is the best way to insure against damage by freezing.

Cells in series and parallel — For proper operation, many electrical devices require higher voltage or current than can be obtained from a single cell. If greater voltage is needed, cells may be connected in *series*, as shown in Fig. 212-A. The negative terminal of one cell is connected to the positive terminal of the next, so that the total e.m.f. of the battery is equal to the sum of the e.m.f.s of the individual cells. For radio purposes, batteries of 45 and 90 volts or more are built up in this way from 1.5-volt dry cells. An automobile storage battery consists of three lead storage cells in series, totalling 6.3 volts — or, in round figures, 6 volts. The current which may be taken safely from a battery composed of cells in series is the same as that which may be taken safely from one cell alone; since the same current flows through all cells, the current capacity is unchanged.

When the device or load to which the battery is to be connected requires more current than can be taken safely from a single cell, the cells may be connected in parallel, as shown in Fig. 212-B. In this case the total current is the sum of the currents contributed by the individual cells, each contributing the same amount if the cells are all alike. When cells are connected in parallel it is essential that the e.m.f.s all be the same, since if one cell generated a larger voltage than the others it would force current through the other cells in the reverse direction and thus would take most, if not all, of the load. Also, if one cell has a lower terminal voltage than the others it will take current from the others rather than carrying its fair share.

Cells may be connected in series-parallel, as in Fig. 212-C, to increase both the voltage and the current-carrying capacity of the battery.

2-5 Electromagnetism

The magnetic field — Everyone is familiar with the fact that a bar or horseshoe magnet will attract small pieces of iron. Just as in the case of electrostatic attraction (§ 2-3) the concept of a *field*, in this case a field of *magnetic* force, is adopted to explain the magnetic action. The field is visualized as being made up of *lines* of magnetic force, the number of which per unit area determines the field strength. As in the case of the electrostatic field, the lines of force do not have physical existence but simply represent a convenient way of describing the properties of the force.

Magnetic attraction and repulsion — The forces exerted by the magnetic field are analogous to electrostatic forces. Corresponding to positive and negative electric charges, it is found that there are two kinds of magnetic *poles*. Instead of being called "positive" and "negative," however, the magnetic poles are called "north" (*N*) and "south" (*S*) poles. These names arise from the fact that, when a magnetized steel rod is freely suspended, it will turn into such a position that one end points toward the north. The end which points north is called the "north-seeking," or simply the "north," pole.

Unlike electric lines of force, which terminate on charges of opposite polarity (§ 2-3) magnetic lines of force are *closed upon themselves*. This is illustrated by the field about a bar magnet, as shown in Fig. 213-A. The lines extend through the magnet, the direction being taken from *S* to *N* inside the magnet and from *N* to *S* outside the magnet. If similar poles of two magnets are brought near each other, there is a force of repulsion between them, while dissimilar poles are attracted when brought close together. As in the case of electric charges, like poles repel, unlike poles attract.

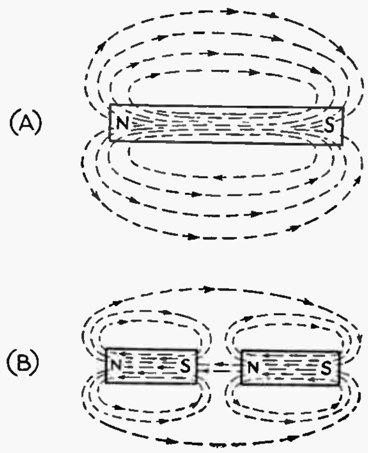


Fig. 213 — (A) The field about a bar magnet. The magnetic lines of force are continuous, part of the path being inside the magnet and part outside. (B) Cutting a magnet produces two magnets, each complete with N and S poles. With the magnets in the positions shown, some of the lines of force are common to both magnets.

If a bar magnet is cut in half, as in Fig. 213-B, it is found that the cut ends also are poles, of opposite kind to the original poles on the same piece. Such cutting can be continued indefinitely, and, no matter how small the pieces are made, there are always two opposite poles associated with each piece. In other words, a single magnetic pole cannot exist alone; it must always be associated with a pole of the opposite kind.

To explain this property of a magnet, it is considered that each molecule of a magnetic substance is itself a miniature magnet. If the material is not magnetized, the molecules are in random positions and the total magnetic effect is zero since there are just as many molecules tending to set up a magnetic field in one direction as there are others tending to set up a field in the opposite direction. When the substance becomes magnetized, however, the molecules are aligned so that most or all of the *N* poles of the molecular magnets are turned toward one end of the material while the *S* poles point toward the other end.

Magnetic induction — When an unmagnetized piece of iron is brought into the field of a magnet, its molecules tend to align themselves as described in the preceding paragraph. If one end of the iron is near the *N* pole of the magnet, the *S* poles of the molecules will turn toward that end and an *S* pole is said to be *induced* in the iron. An *N* pole will appear at the opposite end. Because of the attraction between opposite poles, the iron will be drawn toward the magnet. Since the iron has become a magnet under the influence of the field, it also possesses the property of attracting other pieces of iron.

When the magnetic field is removed, the molecules may or may not resume their random positions. If the material is soft iron the magnetism disappears quite rapidly when the field is removed, but in some types of steel the molecules are slow to resume their random positions and such materials will retain magnetism for a long time. A magnet which loses its magnetism quickly when there is no external magnetizing force is called a *temporary magnet*, while one which retains its magnetism for a long time is called a *permanent magnet*. The tendency to retain magnetism is called *retentivity*. The process of destroying magnetism can be hastened by heating, which increases the motion of the molecules within the substance, as well as by mechanical shock, which also tends to disturb the molecular alignment.

Electric current and the magnetic field — Experiment shows that a moving electron generates a magnetic field of exactly the same nature as that existing about a permanent magnet. Since a moving electron, or group of electrons moving together, constitutes an electric current, it follows that the flow of current is accompanied by the creation of a magnetic field. When the conductor is a wire the magnetic lines of force are in the form of concentric circles around it and lie in planes at right angles to it, as shown in Fig. 214. The direction of this field is controlled by the direction of current flow.

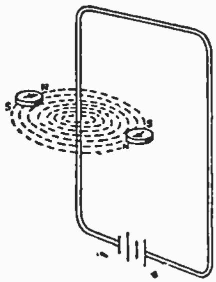


Fig. 214 — **Whenever electric current passes through a wire, magnetic lines of force are set up, in the form of concentric circles, at right angles to the wire, and a magnetic field is said to exist around the wire. The direction of this field is controlled by the direction of current flow, and can be traced by means of a small compass.**

There is an easily remembered method for finding the relative directions of the current and of the magnetic field it sets up. Imagine the fingers of the right hand curled about the wire, with the thumb extended along the wire in the direction of current flow (the conventional direction, from positive to negative, not the direction of electron movement). Then the fingers will be found to point in the direction of the magnetic field; that is, from *N* to *S*.

Magnetomotive force — The force which causes the magnetic field is called *magnetomotive force*, abbreviated *m.m.f.* It corresponds to electromotive force or e.m.f. in the electric circuit. The greater the magnetomotive force, the stronger the magnetic field; that is, the larger the number of magnetic lines per unit area. Magnetomotive force is proportional to the current flowing. When the wire carrying the current is formed into a coil so that the magnetic flux will be concentrated instead of being spread over a large area, the m.m.f. also is proportional to the number of turns in the coil. Consequently magnetomotive force can be expressed in terms of the product of current and turns, and the *ampere-turn*, as this product is called, is in fact the common unit of magnetomotive force. The same magnetizing effect can be secured with a great many turns and a weak current or with a few turns and a strong current. For example, if 10 amperes flow in one turn of wire, the magnetizing effect is 10 ampere-turns. If there is one ampere flowing in 10 turns of wire, the magnetomotive force also is 10 ampereturns.

The magnetic circuit — Since magnetic lines of force are always closed upon themselves, it is possible to draw an analogy between the magnetic *circuit* and the ordinary electrical circuit. The electrical circuit also must be closed so that a complete path is provided around which the electrons or current can flow. However, there is no insulator for the magnetic field, so that the magnetic circuit is always complete even though no magnetic material (such as iron) may be present.

The number of lines of magnetic force, or *flux*, is equivalent in the magnetic circuit to current in the electric circuit. However, it is usual practice to express the strength of the field in terms of the number of lines per unit area, or *flux density*. The unit of flux density is the *gauss*, which is equal to one line per square centimeter, but the terms "lines per square centimeter" or "lines per square inch" are commonly used instead.

Corresponding to resistance in the electric circuit is the tendency to obstruct the passage of magnetic flux, which is called *reluctance*. The reluctance of good magnetic materials, such as iron and steel, is quite low.

The *permeability* of a material is the ratio of the flux which would be set up in a closed magnetic path or circuit of the material to the flux that would exist in a path of the same dimensions in air, the same m.m.f. being used in both cases. The permeability of air is assigned the value 1. The permeability of steels of various types varies from about 50 to several thousand, depending upon the materials alloyed with the steel. Very high permeabilities are attained in certain special magnetic materials, such as "permalloy," which is an alloy of iron and nickel.

The permeability of magnetic materials depends upon the density of magnetic flux in the material. At very high flux densities the permeability is less than its value at low or moderate flux densities. This is because the flux in magnetic materials is proportional to the applied m.m.f. only over a limited range. As the m.m.f. increases more and more of the molecular magnets within the material become aligned, until eventually a point is reached where a very great increase in m.m.f. is required to cause a relatively small increase in flux. This is called *magnetic saturation*. In this region of saturation the permeability decreases, since the *ratio* between the number of lines in the material and the number in air, for the same m.m.f., is smaller than when the flux density is below the saturation point.

Energy in the magnetic field — Like the electrostatic field (§ 2-3), the magnetic field represents potential energy. Consequently the expenditure of energy is necessary to set up a magnetic field, but once the field has been established and remains constant no further energy is consumed in maintaining it. If by some means the field is caused to disappear, the stored-up magnetic energy is converted to energy in some other form. In other words the energy undergoes a transformation when the magnetic field is *changing*, being stored in the field when the field strength is increasing and being released from the field when the field strength is decreasing.

When a magnetic field is set up by a current flowing in a wire or coil, a certain amount of energy is used initially in bringing the field into existence. Thereafter the current must continue to flow, if the field is to be maintained at steady strength, but no expenditure of energy is required for this purpose. (There will be a steady energy loss in the circuit, but only because of the resistance of the wire.) If the current stops the energy of the field is transformed back into electrical energy, tending to keep the current flowing. The amount of energy stored and subsequently released depends upon the strength of the field, which in turn depends upon the intensity of the current and the circuit conditions; i.e., it depends upon the relationship between field strength and current in the circuit.

Induced voltage — Since a magnetic field is set up by an electric current, it is not surprising to find that, in turn, a magnetic field can cause a current to flow in a closed electrical circuit. That is, an e.m.f. can be *induced* in a wire in a magnetic field. However, since a *change* in the field is required for energy transformation, an e.m.f. will be induced only when there is a change in the field with respect to the wire.

This change may be an actual change in the field strength or may be caused by relative motion of the field and wire; e.g., a moving field and a stationary wire, or a moving wire and a stationary field. It is convenient to consider this induced e.m.f. as resulting from the wire's "cutting through" the lines of force of the field. The strength of the e.m.f. so induced is proportional to the *rate* of cutting of the lines of force.

If the conductor is moving parallel with the lines of force in a field, no voltage is induced since no lines are cut. Maximum cutting results when the conductor moves through the field in such a way that both its longer dimension and direction of motion are perpendicular to the lines of force, as shown in Fig. 215. When the conductor is stationary and the field strength varies, the induced voltage results from the alternate increase and decrease in the number of lines of force cutting the wire as the m.m.f. varies in intensity.

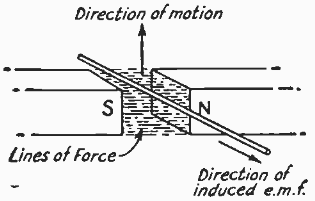


Fig. 215 — Showing how e.m.f. is induced in a conductor moving through a stationary magnetic field, cutting the lines of force. Conversely, a current sent through the conductor in the same direction by means of an external e.m.f. will cause the conductor to move downward.

Lenz's Law — When a voltage is induced and current flows in a conductor moving in a magnetic field, energy of motion is transformed into electrical energy. That is, mechanical *work* is done in moving the conductor when an induced current flows in it. If this were not so the induced voltage would be creating electrical energy, in violation of the fundamental principle of physics that energy can neither be created nor destroyed but only transformed.

It is found, therefore, that the flow of current creates an opposing magnetic force tending to stop the movement of the wire. The statement of this principle is known as Lenz's Law: "In all cases of electromagnetic induction, the induced currents have such a direction that their reaction tends to stop the motion which produces them."

Motor principle — The fact that current flowing in a conductor moving through a magnetic field tends to oppose the motion indicates that current sent through a *stationary* conductor in a magnetic field would tend to *set the conductor in motion*. Such is the case. If moving the conductor through the field in the direction indicated in Fig. 215 causes a current to flow as shown, then, if the conductor is stationary and an e.m.f. is applied to send a current through the conductor in the same direction, the conductor will tend to move across the field in the *opposite* direction.

This principle is used in the electric motor. The same rotating machine frequently may be used either as a generator or motor; as a generator it is turned mechanically to cause an induced e.m.f., and as a motor electric current through it causes mechanical motion.

Self-induction — When an e.m.f. is applied to a wire or coil, current begins to flow and a magnetic field is created. Just before closing the circuit there was no field; just after closing it the field exists. Consequently, at the instant of closing the circuit the *rate of change* of the field is very rapid. Since the wire or coil carrying the current is a conductor in a changing field, an e.m.f. will be induced in the wire. This induced voltage is the *e.m.f. of self-induction*, so called because it results from the current flowing in the wire itself.

By the principle of conservation of energy (and Lenz's Law), the polarity of the induced voltage must be such as to oppose the applied voltage; that is, the induced voltage must tend to send current through the circuit in the direction opposite to that of the current caused by the applied voltage. At the instant of closing the circuit the field changes at such a rate that the induced voltage equals the applied voltage (it cannot exceed the applied voltage, because then it would be supplying energy to the source of applied e.m.f.). but after a short interval the rate of change of the field no longer is so rapid and the induced voltage decreases. Thus the current flowing is very small at first when the applied and induced e.m.f.s are about equal, but rises as the induced voltage becomes smaller. The process is cumulative, the current eventually reaching a final value determined only by the resistance in the circuit.

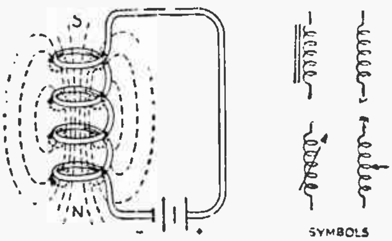


Fig. 216 — When the conducting wire is coiled, the individual magnetic fields of each turn are in such a direction as to produce a field similar to that of a bar magnet. The schematic symbols for inductance are shown at the right. The symbol at the left in the top row indicates an iron-core inductance; at the right, air core. Variable inductances are shown in the bottom row.

In forcing current through the circuit against the pressure of the induced or "back" voltage, work is done. The total amount of work done during the time that the current is rising to its final value is equal to the amount of energy stored in the magnetic field, neglecting heat losses in the wire itself. As explained before, no further energy is put into the field once the current becomes steady. However, if the circuit is opened and current flow caused by the applied e.m.f. ceases, the field collapses. The rate of change of field strength is very great in this case, and a voltage is again induced in the coil or wire. This voltage causes a current flow in the *same* direction as that of the applied e.m.f., since energy is now being restored to the circuit. The energy usually is dissipated in the spark which occurs when such a circuit is opened. Since the field collapses very rapidly when the switch is opened, the induced e.m.f. at such a time can be extremely high.

Inductance — As explained above, the strength of the self-induced voltage is proportional to the rate of change of the field. However, it is also apparent from the foregoing that the voltage also depends upon the properties of the circuit, since, if a number of similar conductors are in the same varying field, the same voltage will be induced in each. By combining the conductors properly, the total induced voltage in such a case will be the sum of the voltages induced in each wire. Also, the rate of change of field strength depends upon the strength of the field set up by a given amount of current flowing in the wire or coil, and this in turn depends upon the ampere-turns, permeability, length and cross-section of the magnetic path, etc.

For a given circuit, however, the field strength will be determined by the current, and the rate of change of the field consequently will be determined by the rate of change of current. Hence, it is possible to group all of these other factors into one quantity, a property of the circuit. This property is called *inductance*. When this is done, the equation giving the value of the induced voltage becomes:

$$\text{Induced voltage} = L \times \text{rate of change of current}$$

where L is the value of inductance in the circuit.

Inductance is a property associated with all circuits, although in many cases it may be so small in comparison to other circuit properties (such as resistance) that no error results from neglecting it. The inductance of a straight wire increases with the length of the wire and decreases with increasing wire diameter. The inductance of such a wire is small, however. For a given length of wire, much greater inductance can be secured by winding the wire into a coil so that the total flux from the wire is concentrated into a small space and the flux density correspondingly increased. The unit of inductance is the *henry*. A circuit or coil has an inductance of one henry if an e.m.f. of one volt is induced when the current changes at the rate of one ampere per second. In radio work it is frequently convenient to use smaller units; those commonly used are the *millihenry* (one thousandth of a henry) and the *microhenry* (one millionth of a henry).

It will be recognized that the relationship between inductance and the magnetic field is similar to that between capacity and the electrostatic field. The greater the inductance, the greater the amount of energy stored in the magnetic field for a given amount of current; the greater the capacity, the greater the amount of energy stored in the electrostatic field for a given voltage.

The inductance of a coil of wire depends upon the number of turns, the cross-sectional dimensions of the coil, and the length of the winding. It also depends upon the permeability of the material on which the coil is wound, or *core*. Formulas for computing the inductance of air-core coils of the type commonly used in radio work, are given in Chapter Twenty.

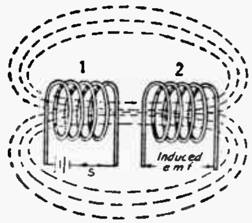
Mutual inductance — If two coils are arranged with their axes coinciding, as shown in Fig. 217, a current sent through Coil 1 will cause a magnetic field which cuts Coil 2. Consequently, an e.m.f. will be induced in Coil 2 whenever the field strength is changing. This induced e.m.f. is similar to the e.m.f. of self-induction; that is,

Induced e.m.f.
$= M \times$ rate of change of current

where M is a quantity called the *mutual inductance* of the two coils. The mutual inductance may be large or small, depending upon the self-inductances of the coils and the proportion of the total flux set up by one coil which cuts the turns of the other coil. If all the flux set up by one coil cuts all the turns of the other coil the mutual inductance has its maximum possible value, while if only a small part of the flux set up by one coil cuts the turns of the other the mutual inductance may be relatively small. Two coils having mutual inductance are said to be *coupled*.

The *degree of coupling* expresses the ratio of actual mutual inductance to the maximum possible value. Coils which have nearly the maximum possible mutual inductance are said to be *closely*, or *tightly*, coupled, while 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

Fig. 217 — Mutual inductance. When the switch, *S*, is closed current flows through coil No. 1, setting up a magnetic field which induces an e.m.f. in the turns of coil No. 2.



respect to each other. Maximum coupling exists when they have a common axis, as shown in Fig. 217, and are as close together as possible.

If two coils having mutual inductance are connected in the same circuit, the directions of the respective magnetic fields may be such as to add or oppose. In the former case the mutual inductance is said to be "positive"; in the latter case, "negative" Positive mutual inductance in such a circuit means that the total inductance is greater than the sum of the two individual inductances, while negative inductance means that the total inductance is less than the sum of the two individual inductances. The mutual inductance may be made either positive or negative simply by reversing the connections to *one* of the coils.

2-6 Fundamental Relations

Direct current — A current which always flows in the same direction through a circuit is called a *direct current*, frequently abbreviated *d.c.* Current flow caused by batteries, for example, is direct current One terminal of each cell is always positive and the other always negative, hence electrons are attracted only in the one direction around the circuit. To make the current change direction, the connections to the battery terminals must be reversed.

Work, energy and power — When a quantity of electricity is moved from a point of one potential to a point at a second potential, work is done. The work done is the product of the quantity of electricity and the difference of potential through which it is moved; that is,

$$W = QE$$

In the practical system of units, with Q in coulombs and E in volts, the unit of work is called the *joule*. Energy, which is the capacity for doing work, is measured in the same units.

Since $I = Q/t$ when the current is constant (§ 2-1), $Q = It$. Substituting for Q in the equation above gives

$$W = EIt$$

where E is in volts, I in amperes, and t in seconds. One ampere flowing through a difference of potential of one volt for one second does one joule of work. *Power* is the time rate at which work is done, so that, if the work is done at a uniform rate, dividing the equation by t will give the electrical power:

$$P = EI$$

The unit of electrical power is the *watt*.

In practical work, the term "joule" is seldom used for the unit of work or energy. The more common name is *watt-second* (one joule is equal to one watt applied for one second). The watt-second is a relatively small unit; a larger one, the *watt-hour* (one watt of power applied for one hour) is more frequently used. Again, for some purposes the watt is too small a unit, and the *kilowatt* (1000 watts) is used instead. A still larger energy unit is the *kilowatt-hour*, the meaning of which is easily interpreted.

Fractional and multiple units — As illustrated by the examples in the preceding paragraph, it is frequently convenient to change the value of a unit so that it will not be necessary to use very large or very small numbers. As applied to electrical units, the practice is to add a prefix to the name of the fundamental unit to indicate whether the modified unit is larger or smaller. The common prefixes are *micro* (one millionth), *milli* (one thousandth), *kilo* (one thousand) and *mega* (one million). Thus, a microvolt is one millionth of a volt, a milliampere is one thousandth of an ampere, a kilovolt is one thousand volts, and so on.

Unless there is some indication to the contrary, it should be assumed that, whenever a formula is given in terms of unprefixed letters (E, I, P, R, etc.), the fundamental units are meant. If the quantities to be substituted in the equation are given in fractional or multiple units, conversion to the fundamental units is necessary before the equation can be used.

Ohm's Law — In any metallic conductor, the current which flows is directly proportional to the applied electromotive force. This relationship, known as *Ohm's Law*, can be written

$$E = RI$$

where E is the e.m.f., I is the current, and R is a constant, depending on the conductor, called the *resistance* of the conductor. By definition, a conductor has one unit of resistance when an applied e.m.f. of one volt causes a current of one ampere to flow. The unit of resistance is called the *ohm.*

Ohm's Law does not apply to all types of conduction, particularly to conduction through gases and in a vacuum. The law is of very great importance, however, because practically all electrical circuits use metallic conduction.

By transposing the equation, the following equally useful forms are obtained:

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

The three equations state that, in a circuit to which Ohm's Law applies, the voltage across the circuit is equal to the current multiplied by the resistance; the resistance of the circuit is equal to the voltage divided by the current; and the current in the circuit is equal to the voltage divided by the resistance.

Resistance and resistivity — The resistance of a conductor is determined by the material of which it is made and its temperature, and is directly proportional to the length of the conductor (that is, the length of the path of the current through the conductor) and inversely proportional to the area through which the current flows. If the temperature is constant,

$$R = k\frac{L}{A}$$

where R is the resistance, k is a constant depending upon the material of which the conductor is made, L is the length and A the area. For the purpose of giving a specific value to k, L is taken as one centimeter and A as one square centimeter (a cube of the material measuring one centimeter on a side); k is then the resistance in ohms of such a cube at a specified temperature. It is called the *specific resistance* or *resistivity* of the material. If the resistivity is known, the resistance of any conductor of known length and uniform cross-section readily can be determined by the formula above. The length must be in centimeters and the area in square centimeters.

The relationships given above are true only for unidirectional (direct) currents and low-frequency alternating currents. Modifications must be made when the current reverses its direction many times each second (§ 2-8).

Conductance and conductivity — The reciprocal of resistance is called *conductance*, and has the opposite properties to resistance. The lower the resistance of a circuit, the higher is the conductance, and vice versa. The symbol of conductance is G, and the relationship to resistance is

$$G = \frac{1}{R} \qquad R = \frac{1}{G}$$

The unit of conductance is called the *mho.* A circuit or conductor which has a resistance of one ohm has a conductance of one mho. By substituting $1/G$ for R in Ohm's Law,

$$G = \frac{I}{E} \qquad I = EG \qquad E = \frac{I}{G}$$

The reciprocal of resistivity is called the *specific conductance* or *conductivity* of a material, and is measured in mhos per centimeter cube. It is frequently useful to know the *relative* conductivity of different materials. This is usually expressed in *per cent conductivity*, the conductivity of annealed copper being taken as 100 per cent. A table of per cent conductivities is given in Chapter Twenty.

Power used in resistance — If two conductors of different resistances have the same current flowing through them, then by Ohm's Law the conductor with the larger resistance will have a greater difference of potential across its terminals. Consequently, more energy is supplied to the larger resistance, since in a given period of time the same amount of electricity is moved through a greater potential difference. The energy appears in the form of heat in the conductor. With a steady current, the heat will raise the temperature of the con-

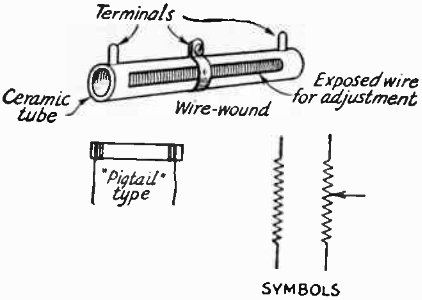


Fig. 218 — Two common types of fixed resistors. The wire-wound type is used for dissipating power of the order of 5 watts or more. "Pigtail" resistors, usually made of carbon or other resistance material in the form of a molded rod or as a thin coating on an insulating tube, rather than being wound with wire, are small in size but do not safely dissipate much power. Schematic symbols for fixed and variable resistors are shown at lower right.

ductor until a balance is reached between the heat generated and that radiated to the surrounding air or otherwise carried away.

Since $P = EI$, substituting for E the appropriate form of Ohm's Law ($E = IR$) gives

$$P = I^2R$$

and making a similar substitution for I gives

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

That is, the power used in heating a resistance (or *dissipated* in the resistance) is proportional to the square of the voltage applied or to the square of the current flowing. In these formulas P is in watts, E in volts and I in amperes.

Further transposition of the equations gives the following forms, useful when the resistance and power are known:

$$E = \sqrt{PR} \qquad I = \sqrt{\frac{P}{R}}$$

Unless the circuit containing the resistance is being used for the specific purpose of generating heat, the power used in heating a resistance is generally considered as a loss. However, there are very many applications in radio circuits where, despite the loss of power, a useful purpose is served by introducing resistance deliberately. Resistances made to specified values and provided with connecting terminals are called *resistors*. They are frequently wound on ceramic or other heat-resisting tubing with wire having high resistivity.

Temperature coefficient of resistance — The resistance of most pure metals increases with an increase in temperature. The resistance of a wire at any temperature is given by

$$R = R_0\,(1 + at)$$

where R is the required resistance, R_0 the resistance at 0°C. (temperature of melting ice), t is the temperature (Centigrade), and a is the *temperature coefficient of resistance*. For copper, a is about 0.004; that is, starting at 0°C., the resistance increases 0.4 per cent per degree above zero.

Temperature coefficient of resistance becomes of importance when conductors operate at high temperatures. In the case of resistors used in electrical and radio circuits, the heat developed by current flow may raise the temperature of the resistance wire to several hundred degrees F. Thus the resistance at operating temperatures can be very much higher than the resistance at room temperature. Consequently such resistors are wound with wire which has a low temperature coefficient of resistance, so that the resistance will be more nearly constant under all conditions.

Resistances in series — When two or more resistances are connected so that the same current flows through each in turn, as shown in Fig. 219, they are said to be connected *in series*. Then, by Ohm's Law,

$$E_1 = IR_1$$
$$E_2 = IR_2$$
$$E_3 = IR_3$$

etc., where the subscripts 1, 2, 3 indicate the first, second and third resistor, and the voltages E_1, E_2 and E_3 are the voltages appearing across the terminals of the respective resistors. Adding the three voltages gives the total voltage across the three resistors:

$$E = E_1 + E_2 + E_3 = IR_1 + IR_2 + IR_3 = I\,(R_1 + R_2 + R_3) = IR$$

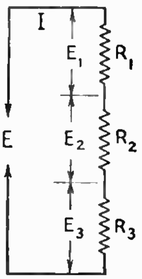


Fig. 219 — Resistances in series.

That is, the voltage across the resistors in series is equal to the current multiplied by the sum of the individual resistances. In the above equation, R, which denotes this sum, may be called the *equivalent* resistance or *total* resistance. The equivalent resistance of a number of resistors connected in series is, therefore, equal to the sum of the values of the individual resistors.

Resistances in parallel — When a number of resistances are connected so that the same voltage is applied to all, as shown in Fig. 220,

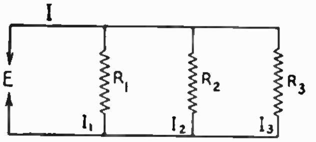


Fig. 220 — Resistances in parallel.

they are said to be connected *in parallel*. By Ohm's Law,

$$I_1 = \frac{E}{R_1} \qquad I_2 = \frac{E}{R_2} \qquad I_3 = \frac{E}{R_3}$$

so that the total current, I, which is the sum of the currents in the individual resistors, is

$$I = I_1 + I_2 + I_3 = \frac{E}{R_1} + \frac{E}{R_2} + \frac{E}{R_3} =$$

$$E\left(\frac{1}{R_1} + \frac{1}{R_2} + \frac{1}{R_3}\right) = E\frac{1}{R}$$

where R is the equivalent resistance — i.e., the resistance through which the same total current would flow if such a resistance were substituted for the three shown. Therefore,

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

That is, the reciprocal of the equivalent resistance of a number of resistances in parallel is equal to the sum of the reciprocals of the individual resistances. Since the reciprocal of resistance is conductance,

$$G = G_1 + G_2 + G_3$$

where G is the total conductance and G_1, G_2, G_3, etc., are the individual conductances in parallel.

To obtain R instead of its reciprocal the equation above may be inverted, so that

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

The number of terms in the denominator of this equation will, of course, be equal to the actual number of resistors in parallel.

For the special case of only two resistances in parallel, the equation reduces to

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

Series-parallel connection of resistors is shown in Fig. 221. When circuits of this type are encountered the equivalent or total resistance can be found by first adding the series resistances in each group, then treating each group as a single resistor so that the formula for resistors in parallel can be used.

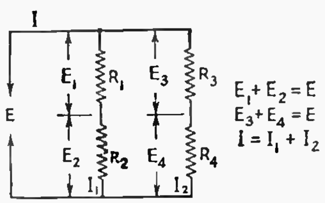


Fig. 221 — Series-parallel connection of resistances. Voltage and current relationships are given at the right.

Voltage dividers and potentiometers — Since the same current flows through resistors connected in series, it follows from Ohm's Law that the voltage (termed *voltage drop*) across each resistor of a series-connected group is proportional to its resistance. Thus, in Fig. 222-A, the voltage E_1 across R_1 is equal to the applied voltage, E, multiplied by the ratio of R_1 to the total resistance, or

$$E_1 = \frac{R_1}{R_1 + R_2 + R_3} \cdot E$$

Similarly, the voltage, E_2, is equal to

$$\frac{R_1 + R_2}{R_1 + R_2 + R_3} \cdot E$$

Such an arrangement is called a *voltage divider*, since it provides a means for obtaining smaller voltages from a source of fixed voltage. When current is drawn from the divider at the various tap points the above relations are no longer strictly true, for then the same current does not flow in all parts of the divider. Design data for such cases are given in § 8-10.

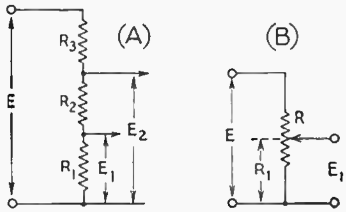


Fig. 222 — Voltage divider (A) and potentiometer (B).

A similar arrangement is shown in Fig. 222-B, where the resistor, R, is equipped with a sliding tap for fine adjustment. Such a variable resistor is frequently called a *potentiometer*.

Inductances in series and parallel — As explained in § 2-5, inductance determines the voltage induced when the current changes at a given rate. That is, $E = L \times$ rate of change of current. This resembles Ohm's Law, if L corresponds to R and the rate of change of current to I. Thus, by reasoning similar to that used in the case of resistors, it can be shown that, for inductances in series,

$$L = L_1 + L_2 + L_3$$

and for inductances in parallel,

$$L = \frac{1}{\frac{1}{L_1} + \frac{1}{L_2} + \frac{1}{L_3}}$$

where the number of terms in either equation is determined by the actual number of inductances connected in series or parallel.

These equations do not hold if there is mutual inductance (§ 2-5) between the coils.

Condensers in series and parallel — When a number of condensers are in parallel, as in Fig. 223-A, the same e.m.f. is applied to all. Consequently, the quantity of electricity stored in each is in proportion to its capacity. The total quantity stored is the sum of the quantities in the individual condensers:

$$Q = Q_1 + Q_2 + Q_3 = C_1E + C_2E + C_3E = (C_1 + C_2 + C_3)E = CE$$

where C is the equivalent capacity. The equivalent capacity of condensers in parallel is equal to the sum of the individual capacities.

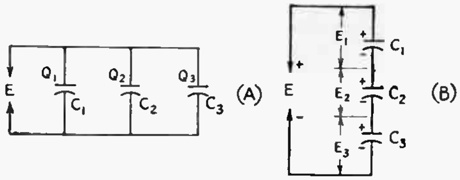


Fig. 223 — **Condensers in parallel (A) and in series (B).**

When condensers are connected in series, as in Fig. 223-B, the application of an e.m.f. to the circuit causes a certain quantity of electricity to accumulate on the top plate of C_1. By electrostatic induction, an equal charge of opposite polarity (negative in the illustration) appears on the bottom plate of C_1, and, since the lower plate of C_1 and the upper plate of C_2 are connected together, this must leave an equal positive charge on the upper plate of C_2. This, in turn, causes the lower plate of C_2 to assume an equal negative charge, and so on down to the plate connected to the negative terminal of the source of e.m.f. In other words the same quantity of electricity is placed on each condenser, and this is equal to the total quantity stored. The voltage across each condenser will depend upon its capacity, and the sum of these voltages must equal the applied voltage. Thus,

$$E = E_1 + E_2 + E_3 = \frac{Q}{C_1} + \frac{Q}{C_2} + \frac{Q}{C_3} =$$

$$Q\left(\frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}\right) = \frac{Q}{C}$$

where C is the equivalent capacity. This leads to an expression similar to that for resistances in parallel:

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

where the number of terms in the denominator should be the same as the actual number of condensers in series.

Time constant — When a condenser and resistor are connected in series with a source of e.m.f., such as a battery, the initial flow of current into the condenser is limited by the resistance, so that a longer period of time is required to complete the charging of the condenser than would be the case without the resistor. Likewise, when the condenser is discharged through a resistor a measurable period of time is taken for the current flow to reach a negligible value. In the case of either charge or discharge the time required is proportional to the capacity and resistance, the product of which is called the *time constant* of the circuit. If C is in farads and R in ohms, or C in microfarads and R in megohms, the product gives the time in seconds required for the voltage across a discharging condenser to drop to $1/e$, or approximately 37 per cent of its original value. (The constant e is the base of the natural series of logarithms.)

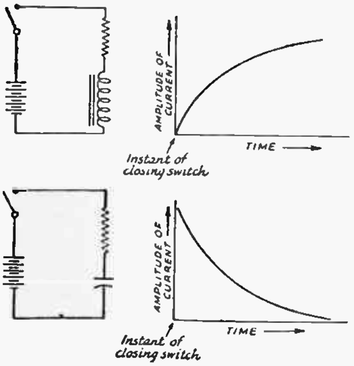


Fig. 224 — The *RL* and *RC* circuits at the left, together with the curves of current amplitude vs. time, show how the current in a circuit combining resistance with inductance or capacity takes a finite period of time to reach a steady-state value.

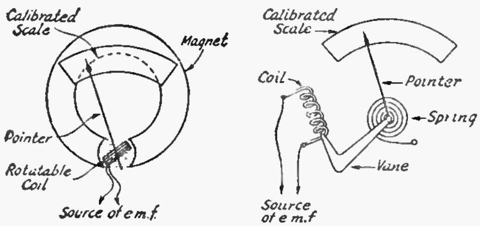


Fig. 225 — *Left* — The d'Arsonval or moving-coil meter for d.c. current measurement. Current flowing through the rotatable coil in the field of the permanent magnet causes a force to act on the coil, tending to turn it. The turning tendency is counteracted by springs (not shown) so that the amount of movement is proportional to the value of the current in the coil. *Right* — In the simpler moving-iron-vane type, a light-weight soft-iron plunger is attracted by current flowing in a fixed coil. As the plunger moves the pointer to which it is linked also moves, until the magnetic force in the coil is balanced by the spiral spring restraining the plunger movement.

In a circuit containing inductance and resistance in series, the effect of the resistance is to *shorten* the period required for the current to reach its final value (§ 2-5) after an e.m.f. is applied to the circuit. The time constant of such a circuit is equal to L/R, where L is in henrys and R in ohms. It gives the time in seconds required for the current to reach $1-1/e$, or approximately 63 per cent of its final steady value when a constant voltage is applied.

By proper application to associated circuits and devices such as vacuum tubes, it is possible by suitable selection of time constant to create almost any desired wave or pulse shape. This is of practical importance in many circuit applications in amateur transmission and reception, as in electronic keyers, automatic volume control, resistance-capacity filters and remote control. Apart from these applications, many of the techniques employed in television and specialized electronic devices are based on this principle.

Measuring instruments — Instruments for measuring d.c. current and voltage make use of the force acting on a coil carrying current in a magnetic field (§ 2-5), produced by a permanent magnet, to move a pointer along a calibrated scale. The magnetic field may be produced by a permanent magnet acting upon a moving coil, or by a fixed coil acting upon a moving iron vane or plunger.

The first type of instrument, based on what is known as the d'Arsonval moving-coil movement, is shown at the left in Fig. 225. The moving-iron vane instrument shown at the right is less accurate and requires higher energizing current, making it relatively insensitive as compared to the moving-coil type. Only the cheaper measuring instruments available to amateurs are based on this principle.

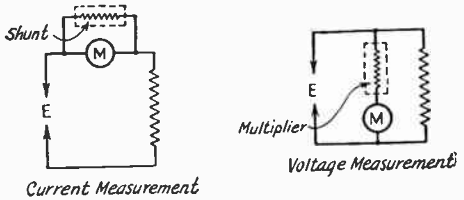


Fig. 226 — Circuit connections for measuring current and voltage. The shunt resistor is used for increasing the value of the current which the instrument can measure, by providing an alternate path through which some of the current can flow. The series multiplier limits the current when the instrument is used to measure voltage.

In such instruments the current required for full-scale deflection of the pointer varies from several milliamperes to a few microamperes, according to the sensitivity required. If the instrument is to read high currents, it is *shunted* (paralleled) by a low resistance through which most of the current flows, leaving only enough flowing through the instrument to give a full-scale deflection corresponding to the total current flowing through both meter and shunt. An instrument which reads microamperes is called a *microammeter* or *galvanometer;* one calibrated in milliamperes is called a *milliammeter;* one calibrated in amperes is an *ammeter.* A *voltmeter* is simply a milliammeter with a high resistance in series so that the current will be limited to a suitable value when the instrument is connected across a voltage source; it is calibrated in terms of the voltage which must appear across the terminals to cause a given value of current to flow. The series resistance is called a *multiplier.* A *wattmeter* is a combination voltmeter and ammeter in which the pointer deflection is proportional to the power in the circuit.

An ammeter or milliammeter is connected in series with the circuit in which current is being measured, so that the current flows through the instrument. A voltmeter is connected in parallel with the circuit.

ℂ 2-7 Alternating Current

Description — An alternating current is one which periodically reverses its direction of flow. In addition to this alternate change in direction, usually the amount or *amplitude* of the current also varies continually during the period when the current is flowing in one direction. These variations are accompanied by corresponding variations in the magnetic field set up by the current, and it is this feature which makes the alternating current so useful. By means of the varying field, energy may be continually transferred (by induction) from one circuit to another without direct connection, and the voltage may be changed in the process. Neither of these is possible with direct current because, except for brief periods when the circuit is closed or opened, the field accompanying a steady direct current is unchanging, and hence there is no way of inducing an e.m.f. except by moving a conductor through the field (§ 2-5).

Alternating currents may be generated in several ways. Rotating electrical machines (*a.c. generators* or *alternators*) are used for developing large amounts of power when the rate of reversal is relatively slow. However, such machines are not suitable for producing currents which reverse direction thousands or millions of times each second. The thermionic vacuum tube is used for this purpose, as described in Chapter Three.

The simplest form of alternating current (or voltage) is shown graphically in Fig. 227. This chart shows that the current starts at zero value, builds up to a maximum in one direction, comes back down to zero, builds up to a maximum in the opposite direction and comes back to zero. The curve follows the sine law and is known as a *sine wave,* because of the wavelike nature of the curve which results when sine values are plotted on rectangular coördinates as a function of angle or time.

Frequency — The complete wave shown in Fig. 227 is called a *cycle,* and the length of time required to complete one cycle is called the *period.* Each half of the cycle, during which the current is flowing in one direction, although its strength is varying, is known as an *alteration.* The number of cycles the wave goes through each second of time is called the *frequency.* In radio work, where frequencies are extremely large, it is convenient to use two other units, *kilocycles* per second (cycles per second ÷ 1000) and megacycles per second (cycles per second ÷ 1,000,000). These are usually abbreviated kc. and Mc., respectively. Occasionally these abbreviations are written kcs. and Mcs. to indicate "kilocycles per second" and "megacycles per second" rather than simply "kilocycles" and "megacycles," but it is understood that "per second" is meant when the shorter forms are used.

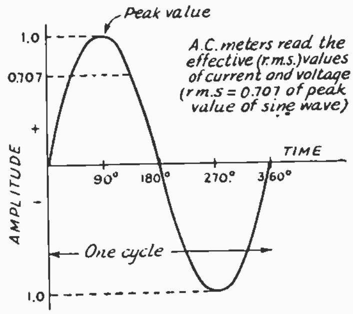


Fig. 227 — Sine wave of alternating current or voltage.

Electrical degrees — If we take a fixed point on the periphery of a revolving wheel, we find that at the end of each revolution, or cycle, the point has come back to its original starting place. Its position at any instant can be expressed in terms of the angle between two lines, one drawn from the center of the wheel to the point at the instant of time considered, the other drawn from the wheel center to the starting point. In making one complete revolution the point has travelled through 360 degrees, a half revolution 180 degrees, a quarter revolution 90 degrees, and so on. The periodic wave of alternating current may be treated similarly, one complete cycle equalling one revolution or 360 degrees, one alternation (half cycle) 180 degrees, and so on. With the cycle divided up in this way, the sine curve simply means that the value of current at any instant is proportional to the sine of the angle which corresponds to the particular fraction of the cycle considered.

The concept of angle is universally used in alternating currents. Generally, it is expressed in the fundamental form, using the radian rather than the degree as a unit, whence a cycle is equal to 2π radians, or a half cycle to π radians. The expression $2\pi f$, for which the symbol ω is often used, simply means electrical degrees per cycle times frequency, and is called the *angular velocity*. It gives the total number of electrical radians passed through by a current of given frequency in one second.

Peak, instantaneous, effective and average values — The highest value of current or voltage during the time when the current is flowing in one direction is called the *maximum* or *peak* value. For the sine wave, the peak has the same absolute value on both the positive and negative halves of the cycle. This is not necessarily true of waves having shapes other than the true sine form.

The value of current or voltage existing at any particular point of time in the cycle is called the *instantaneous* value. The instant for which a particular value is to be found can be specified in terms of time (fraction of the period) or of angle.

Since both the voltage and current are swinging continuously between their positive maximum and negative maximum values, it might be wondered how one can speak of so many amperes of alternating current when the value is changing continuously. The problem is simplified in practical work by considering that an alternating current has an effective value of one ampere when it produces heat, in flowing through a given resistance, at the same average rate as one ampere of continuous direct current flowing through the same resistance. This effective value is the square root of the mean of all of the instantaneous current values squared. In the case of the sine-wave form,

$$E_{eff} = \sqrt{\tfrac{1}{2}E^2_{max}}$$

For this reason, the effective value of an alternating current or voltage is also known as the *root-mean-square*, or *r.m.s.*, value. Hence, the effective value is the square root of ½, or 0.707, times the maximum value.

In a purely a.c. circuit the average current over a whole cycle must be zero, because if the average current on, say, the positive half of the cycle were greater than the average on the negative half, there would be a net current flow in the positive direction. This would correspond to a direct (although intermittent) current, and hence must be excluded because a purely alternating current was assumed. The "average" value of an alternating current is defined as the average current during the part of the cycle when the current is flowing in one direction only. It is of particular importance when alternating current is changed to direct current by the methods considered in later chapters. For a sine wave, the average value is equal to 0.636 of the peak value.

In the sine wave the three voltage values, peak, effective and average, are related to each other as follows:

$$E_{max} = E_{eff} \times 1.414 = E_{ave} \times 1.57$$
$$E_{eff} = E_{max} \times 0.707 = E_{ave} \times 1.11$$
$$E_{ave} = E_{max} \times 0.636 = E_{eff} \times 0.9$$

The relationships for current are equivalent to those given above for voltage.

Phase — As the next few paragraphs will show, the current and voltage in an alternating-current circuit may not pass through their maximum and minimum values at the same time, even though both are sine waves of the same frequency. The time at which a particular part of the cycle (such as the positive peak) occurs is called the *phase* of the wave. If two waves are not exactly in step there is a *phase difference* between them. The phase difference can be expressed in terms of the actual difference in time between the two instants at which the two waves reach corresponding parts of their cycles, but it is generally more convenient to measure it in angular units. A phase difference of 90 degrees, for example, means that one wave reaches its maximum value one-quarter cycle before the other wave reaches its maximum value in the same direction.

The phase relationships between two currents (or two voltages) of the same frequency are defined in the same way. When two such currents are combined the resultant is a single current of the same frequency, but having an instantaneous amplitude equal to the algebraic sum of the amplitudes of the two components at the same instant. The amplitude of the resultant current hence is determined by the phase relationship between the two currents before combination. Thus if the two currents are exactly in phase, the maximum value of the resultant will be the numerical sum of the maximum values of the individual currents; if they are 180 degrees out of phase, one reaches its positive maximum at the instant the other reaches its negative maximum, hence the resultant current is the difference between the two. In the latter case, if the two currents have the same amplitude the resultant current is zero.

Current, voltage and power in an inductance — When alternating current flows through an inductance, the continually varying magnetic field causes the continuous generation of an e.m.f. of self-induction (§ 2-5). The induced voltage at any instant is proportional to the rate at which the current is changing at that instant. If the current is a sine wave, it can be shown that the rate of change is greatest when the current is passing through zero and least when the current is maximum. For this reason, the induced voltage is maximum when the current is zero and zero when the current is maximum. The direction or polarity of the induced voltage is such as to tend to sustain the current flow when the current is decreasing and to prevent it from flowing when the current is increasing (§ 2-5). As a result, the induced voltage in an inductance lags 90 degrees behind the current.

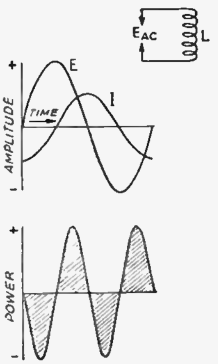


Fig. 228 — **Voltage, current and power relations in an alternating-current circuit consisting of inductance only.**

By Lenz's Law, the induced voltage must always oppose the applied voltage; that is, the induced and applied voltages must be in phase opposition, or 180 degrees out of phase. Consequently, the *applied* voltage leads the current by 90 degrees. Or, using the voltage as a reference, the current in an inductance lags 90 degrees, or one-quarter cycle, behind the voltage. These relationships are shown in Fig. 228.

When the current is increasing in either direction, energy is being stored in the magnetic field. At such times the voltage has the same polarity as the current, so that the product of the two, which gives the instantaneous power fed to the inductance, is positive. When the current is decreasing energy is being restored to the circuit and the applied voltage has the opposite polarity, so that the product of current and voltage is negative. This is also shown in Fig. 228. Positive power means power taken from the source (i.e., the source of the applied e.m.f.), while negative power means power returned to the source. Power is alternately taken and given back in each quarter cycle, and, since the amount given back is the same as that taken, the *average* power in an inductance is zero when considering a whole cycle. In a practical inductance the wire will have some resistance, so that some of the power supplied will be consumed in heating the wire, but if the resistance of the circuit is small compared to the inductance the power consumption is very small compared to the power which is alternately stored and returned.

Current, voltage and power in a condenser — When an alternating voltage is applied to a condenser, the condenser acquires a charge while the voltage is rising and loses its charge while the voltage is decreasing. The quantity of electricity stored in the condenser at any instant is proportional to the voltage across its terminals at that instant ($Q = CE$). Since current is the *rate* of transfer of quantity of electricity, the current flowing into the condenser (when it is being charged) or out of it (when it is discharging) consequently will be proportional to the rate of change of the applied voltage. If the voltage is a sine wave, its rate of change will be greatest when passing through zero and least when the voltage is maximum. As a result, the current flowing into or out of the condenser is greatest when the voltage is passing through zero and least when the voltage reaches its peak value.

This relationship is shown in Fig. 229. Whenever the voltage is rising (in either direction) the current flow is in the same direction as the applied voltage. When the voltage is decreasing and the condenser is discharging, the current flows in the opposite direction. The energy stored in the condenser on the charging part of the cycle is restored to the circuit on the discharge part, and the total energy consumed in a whole cycle therefore is zero. A condenser operating on a.c. takes no average power from the source, except for such actual energy losses as may occur as the result of heating of the dielectric (§ 2-3). The energy loss in air condensers used in radio circuits is negligibly small except at extremely high frequencies.

As shown by Fig. 229, the phase relationship between current flow and applied voltage is such that the current leads the voltage by 90 degrees. This is just the opposite to the inductance case.

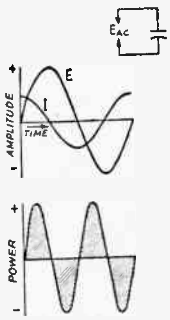


Fig. 229 — **Voltage, current and power relations in an alternating-current circuit consisting of capacity only.**

Current, voltage and power in resistance — In a circuit containing resistance only there are no energy storage effects, and consequently the current and voltage are in phase. The current therefore always flows in the same direction as the applied voltage, and, since the power is always positive, there is continual power *dissipation* in the resistance. The relationships are shown in Fig. 230.

Strictly speaking, no circuit can have resistance only, because the flow of current always is accompanied by the creation of a magnetic field and every conductor also has a certain amount of capacity. Whether or not such residual inductance and capacity are large enough to require consideration is determined by the frequency at which the circuit is to operate.

The a.c. spectrum — Alternating currents of different frequencies have different properties and are useful in a variety of ways. For the transmission of power to light homes, run motors and perform familiar everyday tasks by electrical means, low frequencies are most suitable. Frequencies of 25, 50 and 60 cycles are in common use, the latter being most widely used in this country. The range of frequencies between about 15 and 15,000 cycles is known as the *audio-frequency* range, because when frequencies of this order are converted from a.c. into air vibrations, as by a loudspeaker or telephone receiver, they are distinguishable as sounds having a tone pitch proportional to the frequency.

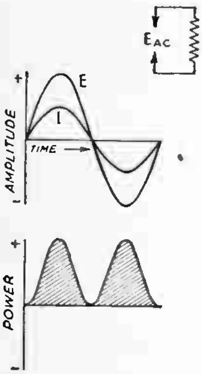


Fig. 230 — Voltage, current and power relations in an alternating-current circuit consisting of resistance only.

Frequencies above 15,000 cycles (15 kilocycles) are used for radio communication, because at frequencies of this order it is possible to convert electrical energy into radio waves which can be radiated over long distances.

For convenience in reference, the following classifications for radio frequencies have been recommended by an international technical conference and are now increasingly in use:

10 to 30 kilocycles	Very-low frequencies
30 to 300 kilocycles	Low frequencies
300 to 3000 kilocycles	Medium frequencies
3 to 30 megacycles	High frequencies
30 to 300 megacycles	Very-high frequencies
300 to 3000 megacycles	Ultrahigh frequencies
3000 to 30,000 megacycles	Superhigh frequencies

Until recently, other terminology was used; for example, the region above 30 megacycles formerly was considered the "ultrahigh" frequencies.

Waveform, harmonics — The sine wave is not only the simplest but for many purposes is the most desirable waveform. Many other waveforms are met in practice, however, and they may differ considerably from the simple sine case. It is possible to show by analysis that any such waveform can be resolved into a number of components of differing frequencies and amplitudes, but related in frequency in such a way that all are integer multiples of the lowest frequency present. The lowest frequency is called the *fundamental*, and the multiple frequencies are called *harmonics*. Thus a wave may consist of fundamental, 3rd, 5th, and 7th harmonics, meaning, if the fundamental frequency is say 100 cycles, that frequencies of 300, 500 and 700 cycles also are present in the wave.

Fig. 231 shows how a fundamental and a second harmonic might combine to form a non-sinusoidal wave. An infinite number of waveforms could be obtained from the combination of two such waves, since the shape of the combined wave will depend upon the amplitude and phase of the two component waves.

The square wave, also shown in Fig. 231, consists of a fundamental and an infinite number of harmonics. This type of wave is useful in a variety of applications.

ℂ 2-8 Ohm's Law for Alternating Currents

Resistance — Since current and voltage are always in phase through a resistance, the instantaneous relations for a.c. are equivalent to those in d.c. circuits. By definition, the effective units of current and voltage for a.c. are made equal to those for d.c. in resistive circuits (§ 2-7). Therefore the various formulas expressing Ohm's Law for d.c. circuits apply without any change to a.c. circuits containing resistance only, or for purely resistive parts of complex a.c. circuits. See § 2-6.

In applying the formulas, it must be remembered that consistent units must be used. For example, if the instantaneous value of current is used in finding voltage or power, the voltage found will be the instantaneous voltage and the power will be the instantaneous power. Likewise, if the effective value is used for one quantity in the formula, the unknown will be expressed in effective value. Unless otherwise indicated, the effective value of current or voltage is always understood to be meant when reference is made to "current" or "voltage."

Reactance — In the preceding section it was shown that energy-storage effects in inductance and capacitance cause a phase difference to exist between the applied voltage and the cur-

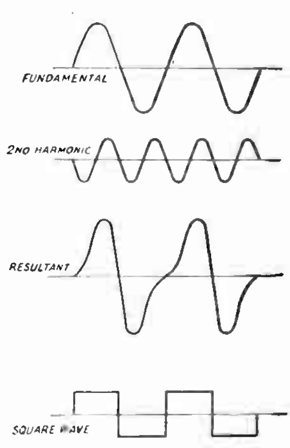


Fig. 231 — Combination of a fundamental and second harmonic with the amplitude and phase relationships shown gives the non-sinusoidal resultant. The square wave, below, contains an infinite number of harmonics.

rent that flows as a result. Because of this, Ohm's Law cannot be applied in its entirety to a.c. circuits containing inductance and/or capacitance, particularly for the calculation of power consumed. However, the amplitude of the current that flows in such circuits is directly proportional to the voltage applied, just as it is in purely resistive circuits. In other words, both inductance and capacity offer opposition to current flow, and this opposition can be measured in ohms just as it is in the case of resistance. But the opposition is called *reactance* to indicate that it does not consume power and thereby distinguish it from resistance.

Ohm's Law formulas extended to include reactance are quite similar to the formulas for resistive circuits:

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

where X is the symbol for reactance.

Reactance differs from resistance in another respect — its value, for a given amount of inductance or capacity, varies with the frequency of the current flowing, whereas resistance is not inherently affected by frequency. However, the reactance of a given inductance or capacity is constant for all values of applied voltage so long as the frequency is constant.

Inductive reactance — When alternating current flows through an inductance it must take just the right value to make the induced voltage equal the applied voltage (§ 2-7). Since the induced voltage is equal to the inductance multiplied by the rate of change of the current, it is evident that the larger the value of inductance considered, the smaller the rate of current change required to induce a given voltage. If the frequency is fixed, the rate at which the alternating current changes is simply proportional to the amplitude of the current. Hence a small current will suffice if the inductance is large, while a large current will be required if the inductance is small, assuming that the applied voltage is the same in both cases. In other words, the reactance of an inductance is directly proportional to the value of the inductance, at a fixed frequency.

However, the rate of change of current is proportional to *frequency* as well as to amplitude, because the greater the number of cycles per second the more rapidly the current goes through its regular variations. Consequently, increasing the frequency will have the same effect as increasing the amplitude of the current insofar as the induced voltage is concerned; or, to put it another way, if the frequency is increased the amplitude may be decreased in the same proportion to maintain the same induced voltage in a given inductance. Smaller current amplitude through a fixed value of inductance means that the reactance is higher, so it is apparent that the reactance of an inductance increases with increasing frequency.

Thus three factors, inductance, current amplitude, and frequency (angular velocity) determine the induced voltage. Combining them, we have, for sine-wave current,

$$E = 2\pi fLI, \text{ or } \frac{E}{I} = 2\pi fL$$

Since $X = E/I$, then

$$X_L = 2\pi fL$$

where the subscript L indicates that the reactance is inductive.

The fundamental units (ohms, cycles, henrys) must be used in the above equation, or appropriate factors inserted if other units are employed. If inductance is in millihenrys, the frequency should be stated in kilocycles; if inductance is in microhenrys, the frequency should be given in megacycles, to bring the answer in ohms.

Capacitive reactance — The quantity of electricity stored in a condenser depends upon the capacity and the applied voltage ($Q = CE$), and if losses are negligible the same quantity of electricity is taken out of the condenser on discharge. Current must flow into the condenser to charge it, and must flow out of it to discharge it; the value of the current is the rate at which the quantity of electricity is put into the condenser or taken out (§ 2-4). When an a.c. voltage is applied to a condenser the alternate movement of a quantity of electricity to charge and discharge it as the applied voltage rises and falls and reverses polarity, constitutes current flow "through" the condenser.

The amplitude of the current at any instant is proportional to the rate of change of the voltage at that instant; the greater the rate of change the faster the given quantity of electricity is moved. The amplitude is also proportional to the capacitance of the condenser, since a larger capacitance will take a larger quantity of electricity at a given voltage. Since the rate of change of voltage is proportional to the amplitude of the voltage and its frequency, then for a sine-wave voltage

$$I = 2\pi fCE, \text{ or } \frac{E}{I} = \frac{1}{2\pi fC}$$

Since $X = E/I$, then

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

where the subscript C indicates that the reactance is capacitive. Capacitive reactance is *inversely* proportional to capacity and to the applied frequency. For a given value of capacity, the reactance decreases as the frequency increases.

Fundamental units (farads, cycles per second) must be used in the right-hand side of the equation to obtain the reactance in ohms. Conversion factors must be used if the frequency and capacity are in units other than cycles and farads. If C is in microfarads and f in megacycles, the conversion factors cancel.

Impedance — In any series circuit the same current flows through all parts of the circuit. If a resistance and inductance are connected in series to form an a.c. circuit they both carry the same current, but the voltage across the resistance is in phase with the current while the voltage across the inductance leads the current by 90 degrees. In a d.c. circuit with resistances in series, the applied voltage is equal to the sum of the voltages across the individual resistances (§ 2-6). This is also true of the a.c. circuit with resistance and inductance in series if the *instantaneous* voltages are added algebraically to find the instantaneous value of applied voltage. But, because of the phase difference between the two voltages, the maximum value of the applied voltage will not be the sum of the maximum values of the two voltages, so that the effective values cannot be added directly. The same considerations hold in the case of resistance and capacity in series.

In either case the total voltage is given by the following expressions:

$$E^2 = E^2_X + E^2_R, \text{ or } E = \sqrt{E^2_R + E^2_X}$$

where E_X indicates the voltage across the reactance, which may be either inductive or capacitive, and E_R is the voltage across the resistance.

Since $E_R = IR$ and $E_X = IX$, substitution gives

$$E = I\sqrt{R^2 + X^2}, \text{ or } \frac{E}{I} = \sqrt{R^2 + X^2}$$

E/I is called the *impedance* of the circuit and is designated by the letter Z. Hence,

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

The impedance determines the voltage which must be applied to the circuit to cause a given current to flow. The unit of impedance is, therefore, the ohm, just as in the case of resistance and reactance, which also determine the ratio of voltage to current. Ohm's Law for alternating current circuits then becomes

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

It should be noted that the equivalent Ohm's Law relationship for *power* in a d.c. circuit does not apply directly in the case of an a.c. circuit where Z replaces R. As will be explained, the power factor of the circuit must be taken into consideration.

In summary, impedance is a generalized quantity applying to a.c. or d.c. circuits, simple or complex. In a d.c. circuit or in an a.c. circuit containing resistance only, the phase angle is zero (current and voltage are in phase) and the impedance is equal to the resistance.

In an a.c. circuit containing reactance only the phase angle is 90 degrees, with current lagging the voltage if the reactance is inductive and current leading the voltage if the reactance is capacitive. In either case, the impedance is equal to the reactance.

In an a.c. circuit containing both resistance and reactance the phase angle may have any value between zero and 90 degrees, with the current lagging the voltage if the reactance is inductive and leading the voltage if the reactance is capacitive. The value of impedance, in ohms, may be found from the equation given above.

Power is consumed in a circuit only when the current flow produced by the applied voltage is less than 90 degrees out of phase with that voltage. Power consumption decreases from maximum with in-phase conditions to zero at a 90-degree phase difference.

Series circuits with L, C and R — When inductance, capacity and resistance all are in series in an a.c. circuit, the voltage relations are a combination of the separate cases just considered. The voltage across each element will be proportional to the resistance or reactance of that element, since the current is the same through all. The voltages across the inductance and capacity are 180 degrees out of phase, since one leads the current by 90 degrees and the other lags the current by 90 degrees. This means that the two voltages tend to cancel; in fact, if the voltage across only the inductance and capacity in series is considered (leaving out the resistance), the total voltage is the *difference* between the two voltages.

The *total* reactance in a series circuit is, therefore, the difference between the individual inductive and capacitive reactances; or

$$X = X_L - X_C$$

If more than one inductance element is present in the circuit, the total inductive reactance is the sum of the individual reactances; similarly, the same is true for capacitive reactances. Inductive reactance is conventionally taken as "positive" (+) in sign and capacitive reactance as "negative" (−). With this convention, algebraic addition of all the reactances in a series circuit gives the total reactance of the circuit.

Parallel circuits with L, C and R — The equivalent resistance of a number of resistances in parallel in an a.c. circuit is found by the same rules as in the case of d.c. circuits (§ 2-6). Parallel reactances *of the same kind* have an equivalent reactance given by a similar rule:

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

This formula applies to reactances of the same sign; it cannot be used if both inductive and capacitive reactance are in parallel.

When both resistance and reactance are in parallel the same voltage is applied to both, but the current in the resistance branch will not be in phase with the current in the reactive branch. The phase difference will be 90 degrees if each branch contains *only* resistance or *only* reactance, so that the total current may be found by a rule similar to that used for finding the total voltage in a series circuit. That is,

$$I = \sqrt{I_R^2 + I_X^2}$$

The impedance of the circuit is equal to E/I, so

$$Z = \frac{E}{\sqrt{I_R^2 + I_X^2}}$$

By assuming some convenient value for the applied voltage and then solving for the currents in the resistance and reactance, the values so found may be substituted in this equation to find the impedance of the circuit.

The formulas above may be used for either inductive or capacitive reactance. When inductive reactance and capacitive reactance are in parallel, the current through the inductance is 180 degrees out of phase with the current through the condenser, hence the total current is the *difference* between the two currents. This difference may be substituted for I_X in the above equations.

It is of interest to note that, since the total current flowing in a circuit containing inductive and capacitive reactance in parallel is the difference between the currents in the two branches, the impedance of such a parallel combination always is larger than the reactance of either branch alone. Any resistance which also may be in parallel is unaffected, since the current taken by the resistance is determined solely by the applied voltage.

With series-parallel circuits the solution becomes considerably more complicated, since the phase relationships in any parallel branch may not be either 90 degrees or zero. However, the majority of parallel circuits used in radio work can be solved by the rather simple approximate methods described in § 2-10.

Power factor — The power dissipated in an a.c. circuit containing both resistance and reactance is consumed entirely in the resistance, hence is equal to I^2R. However, the reactance is also effective in determining the current or voltage in the circuit, even though it consumes no energy. Hence the product of volts times amperes (which gives the power consumed in d.c. circuits) for the whole circuit may be several times the actual power used up. The ratio of power dissipated (watts) to the *volt-ampere* product is called the power factor of the circuit, or

$$\text{Power factor} = \frac{\text{Watts}}{\text{Volt-amperes}}$$

Distributed capacity and inductance — It should not be thought that the reactance of coils becomes infinitely high as the frequency is increased to a high value and, likewise, that the reactance of condensers becomes infinitely low at high frequencies. All coils have some capacity between turns, and the reactance of this capacity can become low enough at some high frequencies to tend to cancel the high reactance of the coil. Likewise, the leads and plates of condensers will have considerable inductance at very high frequencies, which will tend to offset the capacitive reactance of the condenser itself. For these reasons, coils constructed for high-frequency use must be designed to have low "distributed" capacity. Similarly, condensers must be made with short, heavy leads so that they will have low self-inductance.

Units and instruments — The units used in a.c. circuits may be divided or multiplied to give convenient numerical values to different orders of magnitude, just as in d.c. circuits (§ 2-6). Because the rapidly reversing current is accompanied by similar reversals in the magnetic field, instruments used for measurement of d.c. (§ 2-6) will not operate on a.c.

At low frequencies suitable instruments can be constructed by making the current produce both magnetic fields, one by means of a fixed coil and the other by the moving coil. Instruments having movements of this kind are variously known as *dynamometer*, *electrodynamometer* and *electrodynamic* types.

Another type of instrument suitable for measuring alternating current is less expensive in construction and therefore more widely used. This is the *repulsion-type* moving-iron a.c. ammeter shown in Fig. 232. Fundamentally, the movement is based on the same principle as the inexpensive moving-iron-vane meter for d.c. shown in Fig. 225. In the repulsion-type instrument current flowing through the stationary coil magnetizes two iron vanes, one fixed and the other attached to the movable pointer shaft. Inasmuch as the two vanes are in the same plane and magnetized by the same source, the magnetic effect upon them by the current through the coil will be identical regardless of its polarity. When the two vanes are magnetized they repel each other (§ 2-2) and the movable vane moves away from the fixed vane, causing the pointer to travel along the scale. The degree of travel is controlled by a spring which brings the pointer to rest at a point where the electrical and mechanical forces balance, and returns the pointer to zero on the scale when current flow ceases.

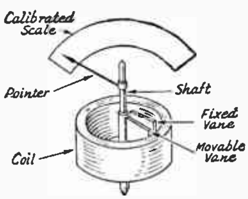


Fig. 232 — **Ammeter based on a repulsion-type moving-iron movement used for a.c. measurements.**

Such instruments are used for measurement of either current or voltage. However, when employed for voltage measurement by the use of high-resistance series multipliers, the minimum current drain required by such instruments because of their inherent insensitivity is so great that excessive load is placed upon the measurement source. For this reason, in radio work it is more common practice to convert the a.c. voltage to d.c. by means of a copper-oxide or vacuum-tube rectifier and then measure the resulting indication on a d.c. instrument, as described in § 2-6.

At radio frequencies instruments of the type described above are inaccurate because of distributed capacity and other effects, and the only reliable type of direct-reading instrument is the *thermocouple* ammeter or milliammeter. This is a power-operated device consisting of a resistance wire heated by the flow of r.f. current through it, to which is attached a thermocouple or pair of wires of dissimilar metals joined together and possessing the property of developing a small d.c. voltage between the terminals when heated. This voltage, which is proportional to the heat applied to the couple, is used to operate a d.c. instrument of ordinary design.

2-9 The Transformer

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

Two coils operating in this way are said to be *coupled*, and the pair of coils constitutes a *transformer*. The coil connected to the source of energy is called the *primary* coil, and the other is called the *secondary* coil. Energy may be taken from the secondary, being transferred from the primary through the medium of the varying magnetic field.

Types of transformers — The usefulness of the transformer lies in the fact that 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, 120 volts and only a 440-volt source is available, a transformer can be used to change the source voltage to that required. The transformer, of course, can be used only on a.c., since no voltage will be induced in the secondary if the magnetic field is not changing. If d.c. is applied to the primary of a transformer, a voltage will be induced in the secondary only at the instant of closing or opening the primary circuit, since it is only at these times that the field is changing.

As shown in Fig. 233, 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. 233 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*, an effect which occurs because the iron tends to retain its magnetism, and hence requires the expenditure of energy to overcome this *residual* magnetism every time the alternating current reverses in direction, and because of *eddy currents*, or currents induced in the core by the varying magnetic field.

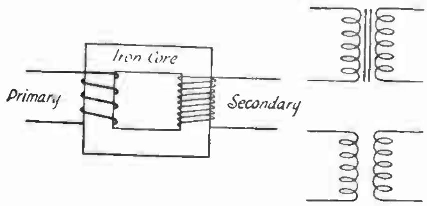


Fig. 233 — The transformer. Power is transferred from the primary coil to the secondary by means of the magnetic field. The upper symbol at right indicates an iron-core transformer, the lower one an air-core transformer.

Core losses increase with frequency to such an extent that they become excessive at radio frequencies if a transformer is wound on the type of core used for power and audio frequencies. Transformers for use at radio frequencies either are wound on non-magnetic material ("air core") or on special cores made of powdered iron particles held in an insulating binder. In the latter case the core is not used as a means of carrying the magnetic field from the primary to the secondary, but simply to give a larger inductance with a fixed number of turns. In radio-frequency transformers relatively little of the magnetic flux set up by the primary cuts the turns of the secondary. The discussion in this section is confined to low-frequency iron-core transformers, where practically all of the primary flux cuts the secondary. Radio-frequency transformers are considered in § 2-10.

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, it follows that the induced voltages will be proportional to the number of turns on each coil. In the case of the primary, or coil connected to the source of power, the induced voltage is practically equal to, and opposes, the applied voltage. Hence, for all practical purposes,

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

where E_s is the secondary voltage, E_p is the primary voltage, and n_s and n_p are the number of turns on the secondary and primary, respectively. The ratio n_s/n_p is called the *turns ratio* of the transformer.

This relationship is true *only* when all the flux set up by the primary current cuts all the turns of the secondary. If some of the magnetic flux follows a path which does not make it cut the secondary turns then the secondary voltage is less than given by this formula, since this reduces the number of lines of force (and thus reduces the effective strength of the magnetic field affecting the secondary) by causing the rate of change of flux to be less in the secondary than in the primary. In general, the equation can be used only when both coils are wound on a closed core of high permeability, so that practically all of the flux can be confined to definite paths.

Effect of secondary current — The primary current which has been discussed above is usually called the *magnetizing current* of the transformer. Like the current in any inductance, it lags the applied voltage by 90 degrees, neglecting the small energy losses in the resistance of the primary coil and in the iron core.

When current is drawn from the secondary winding, the secondary current sets up a magnetic field of its own in the core. The phase relationship between this field and that caused by the magnetizing current will depend upon the phase relationship between current and voltage in the secondary circuit. In every case there will be an effect upon the original field. To maintain the induced primary voltage equal to the applied voltage, however, the original field *must* be maintained. Consequently, the primary current must change in such a way that the effect of the field set up by the secondary current is completely canceled. This is accomplished when the primary draws additional current that sets up a field exactly equal to the field set up by the secondary current, but which opposes the secondary field. The additional primary current is thus 180 degrees out of phase with the secondary current.

In rough calculations on transformers it is convenient to neglect the magnetizing current and to assume that the primary current is caused entirely by the secondary load. This is justifiable, because in any well-designed transformer the magnetizing current is quite small in comparison to the load current when the latter is near the rated value.

For the fields set up by the primary and secondary load currents to be equal, the number of ampere turns in the primary must equal the number of ampere turns in the secondary. That is,

$$n_s I_s = n_p I_p$$

Hence,

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

The load current in the primary for a given load current in the secondary is proportional to the turns ratio, secondary to primary. This is the opposite of the voltage relationships.

If the magnetizing current is neglected, the phase relationship between current and voltage in the primary circuit will be identical with that existing between the secondary current and voltage. This is because the applied voltage and induced voltage are 180 degrees out of phase, and the primary current and secondary current likewise are 180 degrees out of phase.

Energy relationships; efficiency — A transformer cannot create energy; 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. Since there is always some power loss in the resistance of the coils and in the iron core, the power taken from the source always will exceed that taken from the secondary. Thus,

$$P_o = n P_i$$

where P_o is the power taken from the secondary, P_i is the power input to the primary, and n is a factor which always is less than 1. It is called the *efficiency* of the transformer and is usually expressed as a percentage. The efficiency of small power transformers such as are used in radio receivers and transmitters may vary 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 which cuts one coil and not the other is only a small percentage of the total flux. This *leakage flux* acts in the same way as flux about any coil which is not coupled to another coil; that is, it gives rise to self-induction. Consequently, there is a small amount of *leakage inductance* associated with both windings of the transformer, but not common to them. Leakage inductance acts in exactly the same way as an equivalent amount of ordinary inductance inserted in series with the circuit. It has, therefore, a certain reactance, depending upon the amount of inductance and the frequency. This reactance is called *leakage reactance.*

In the primary the practical effect of leakage reactance is equivalent to a reduction in applied voltage, since the primary current flowing through the leakage reactance causes a voltage drop. This voltage drop increases with increasing primary current, hence it increases as more current is drawn from the secondary. The induced voltage consequently decreases, since the applied voltage (which the induced voltage must equal in the primary) has been effectively reduced. The secondary induced voltage also decreases proportionately. When current flows in the secondary circuit the secondary leakage reactance causes an additional voltage drop, which results in a further reduction in the voltage available from the secondary terminals. Thus, the greater the secondary current, the smaller the secondary terminal voltage becomes. The resistance of the primary and secondary windings of the transformer also causes voltage drops when current is flowing, and, 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.

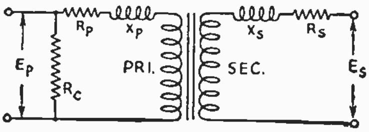


Fig. 234 — The equivalent circuit of a transformer includes the effects of leakage inductance and resistance of both primary and secondary windings. The resistance R_c is an equivalent resistance representing the constant core losses. Since these are comparatively small, their effect may be neglected in many approximate calculations.

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 under load. The drop in voltage may be considerably more than this in a transformer operating at audio frequencies, however, since the leakage reactance in a transformer increases directly with the frequency.

Impedance ratio — In an ideal transformer having no losses or leakage reactance, the primary and secondary volt-amperes are equal; that is,

$$E_p I_p = E_s I_s$$

On this assumption, and by making use of the relationships between voltage, current and turns ratio previously given, it can be shown that

$$\frac{E_p}{I_p} = \frac{E_s}{I_s}\left(\frac{n_p}{n_s}\right)^2$$

Since $Z = E/I$, E_s/I_s is the impedance of the load on the secondary circuit, and E_p/I_p is the impedance of the loaded transformer as viewed from the line. The equation states that the impedance presented by the primary of the transformer to the line, or source of power, is equal to the secondary load impedance multiplied by the square of the primary-to-secondary turns ratio. This primary impedance is called the *reflected impedance* or *reflected load.* The reflected impedance will have the same phase angle as the secondary load impedance, as previously explained. If the secondary load is resistive only, then the input terminals of the transformer primary will appear to the source of e.m.f. as a pure resistance.

In practice there is always some leakage reactance and power loss in the transformer, so that the relationship above does not hold exactly. However, it gives results which are adequate for many practical cases. The *impedance ratio* of the transformer consequently is considered to be equal to the square of the turns ratio, both ratios being taken from the same winding to the other.

Impedance matching — Many devices require a specific value of load resistance (or impedance) for optimum operation. The resistance of the actual load which is to dissipate the power may differ widely from this value, hence the transformer, with its impedance-transforming properties, is frequently called upon to change the actual load to the desired value. This is called *impedance matching*. From the preceding paragraph,

$$\frac{n_s}{n_p} = \sqrt{\frac{Z_s}{Z_p}}$$

where n_s/n_p is the required secondary-to-primary turns ratio, Z_s is the impedance of the actual load, and Z_p is the impedance required for optimum operation of the device delivering the power.

Transformer construction — Transformers are generally built so that flux leakage is minimized insofar as possible. The magnetic path is laid out so that it is as short as possible, since this reduces its reluctance and hence the number of ampere-turns required for a given flux density, and also tends to minimize flux leakage. Two core shapes are in common use, as shown in Fig. 235. 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 there is a large difference of potential between primary and secondary.

Core material for small transformers is usually silicon steel, called "transformer iron." The core is built up of thin sheets, called *laminations*, insulated from each other (by a thin coating of shellac, for example) to prevent the flow of *eddy currents* which are induced in the iron at right angles to the direction of the field. If allowed to flow, these eddy currents would cause considerable loss of energy in overcoming the resistance of the core material. The separate laminations are overlapped, to make the magnetic path as continuous as possible and thus reduce leakage.

The number of turns required on the primary for a given applied e.m.f. is determined by the maximum permissible flux density in the

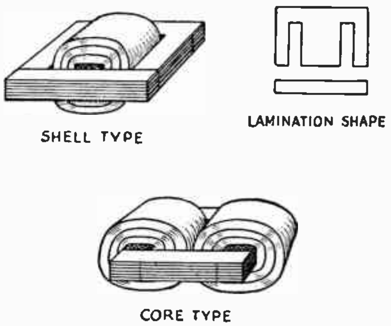


Fig. 235 — Two common types of transformer construction. Core pieces are interleaved to provide a continuous magnetic path with as low reluctance as possible.

type of core material used, the frequency, and the magnetomotive force required to force the flux through the iron. As a rough indication, windings of small power transformers frequently have about two turns per volt for a core of 1 square inch cross-section and a magnetic path 10 or 12 inches in length. A longer path or smaller cross section would require 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 separate coils and between the coils and the core.

In power transformers distributed capacity in the windings is of little consequence, but in audio-frequency transformers it may cause undesired resonance effects (see § 2-10 for a discussion of resonance). High-grade audio transformers often have special types of windings designed to minimize distributed capacity.

The autotransformer — The transformer principle can be utilized with only one winding instead of two, as shown in Fig. 236; the principles just discussed apply equally well. The *autotransformer* has the advantage that, since

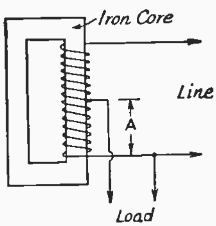


Fig. 236 — The auto-transformer is based on the transformer principle, but uses only one winding. The line and load currents in the common winding (A) flow in opposite directions, so that the resultant current is the difference between them. The voltage across A is proportional to the turns ratio.

the line and load currents are out of phase, the section of the winding common to both circuits carries less current than the remainder of the coil. This advantage is not very marked unless the primary and secondary voltages do not differ very greatly, while it is frequently disadvantageous to have a direct connection between primary and secondary circuits. For these reasons, application of the autotransformer is usually limited to boosting or reducing the line voltage by a relatively small amount for purposes of voltage correction.

ℂ 2-10 Resonant Circuits

Principle of resonance — It has been shown (§ 2-8) that the inductive reactance of a coil and the capacitive reactance of a condenser are oppositely affected by frequency. In any series combination of inductance and capacitance, therefore, there is one particular frequency for which the inductive and capacitive reactances are equal. Since these two reactances cancel each other, the net reactance in the circuit becomes zero, leaving only the resistance to impede the flow of current. The frequency at which this occurs is known as the *resonant frequency* of the circuit and the circuit is said to be *in resonance* at that frequency, or *tuned* to that frequency.

Series circuits — 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 (§ 2-8) gives

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

Solving this equation for frequency gives

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

This equation is in the fundamental units — cycles per second, henrys and farads — and so, if fractional or multiple units are used, the appropriate factors must be inserted to change them to the fundamental units. A formula in units commonly used in radio circuits is

$$f = \frac{1}{2\pi\sqrt{LC}} \times 10^6$$

where f is the frequency in kilocycles per second, 2π is 6.28, L is the inductance in microhenrys (μh.), and C is the capacitance in micro-microfarads ($\mu\mu$fd.).

The resistance that may be present does not enter into the formula for resonant frequency.

When a constant a.c. voltage of variable frequency is applied, as shown in Fig. 237-A, the current flowing through such a circuit will be maximum at the resonant frequency. The magnitude of the current at resonance will be determined by the resistance in the circuit. The curves of Fig. 237 illustrate this, curve *a* being for low resistance and curves *b* and *c* being for increasingly greater resistances.

In the circuits used at radio frequencies the reactance of either the coil or condenser at resonance is usually several times as large as the resistance of the circuit, although the *net* reactance is zero. As the applied frequency departs from resonance, say on the low-frequency side, the reactance of the condenser increases and that of the inductance decreases, so that the net reactance (which is the difference between the two) increases rather rapidly. When it becomes several times as high as the resistance, it becomes the chief factor in determining the amount of current flowing. Hence, for circuits having the same values of inductance and capacity but varying amounts of resistance, the resonance curves tend to coincide at fre-

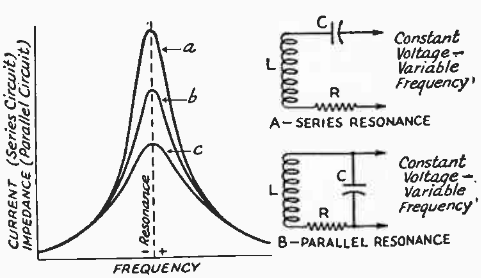


Fig. 237 — Characteristics of series-resonant and parallel-resonant circuits with variations in resistance, *R*.

quencies somewhat removed from resonance. The three curves in the figure show this tendency.

Parallel circuits — The parallel-resonant circuit is illustrated in Fig. 237-B. This circuit also contains inductance, capacitance and resistance in series, but the voltage is applied in parallel with the combination instead of in series with it as in A. As explained in connection with parallel inductance and capacity (§ 2-8), the total current through such a combination is less than the current flowing in the branch having the smaller reactance. If the currents through the inductive and capacitive branches are equal in amplitude and exactly 180 degrees out of phase, the total current, called the *line* current, will be zero no matter how large the individual branch currents may be. The impedance ($Z = E/I$) of such a circuit, viewed from its parallel terminals, would be infinite. In practice the two currents will not be exactly 180 degrees out of phase, because there is always some resistance in one or both branches. This resistance makes the phase relationship between current and voltage less than 90 degrees in the branch containing it, hence the phase difference between the currents in the two branches is less than 180 degrees and the two currents will not cancel completely. However, the line current may be very small if the resistance is small compared to the reactance, and thus the parallel impedance at resonance may be very high.

As the applied frequency is increased or decreased from the resonant frequency, the reactance of one branch decreases and that of the other branch increases. The branch with the smaller reactance takes a larger current, if the applied voltage is constant, and that with the larger reactance takes a smaller current. As a result, the difference between the two currents becomes larger as the frequency is moved farther from resonance. Since the line current is the difference between the two currents, the current increases when the frequency moves away from resonance; in other words, the parallel impedance of the circuit decreases.

The variation of parallel impedance of a parallel-resonant circuit with frequency is illustrated by the same curves of Fig. 237 that show the variation in current with frequency for the series-resonant circuit. The parallel impedance at resonance increases as the series resistance is made smaller.

In the case of parallel circuits, resonance may be defined in three ways: the condition which gives maximum impedance, that which gives a power factor of 1 (impedance purely resistive), or (as in series circuits) when the inductive and capacitive reactances are equal. If the resistance is low, the resonant frequencies obtained on the three bases are practically identical. This condition usually is satisfied in radio work, so that the resonant frequency of a parallel circuit is generally computed by the series-resonance formula given above.

Resistance at high frequencies — When current flows in a conductor a magnetic field is set up inside the conductor as well as externally. When the current is alternating, the internal magnetic field induces a voltage inside the conductor which opposes the applied voltage and becomes larger as the center of the conductor is approached. As a result, the current is forced to distribute itself so that the greater proportion flows near the surface and less near the center. This is known as *skin effect.*

Skin effect is negligible at low frequencies, but increases with increasing frequency to such an extent that at radio frequencies the major portion of the current flows near the surface. In the u.h.f. range, all the current may be concentrated within one or two thousandths of an inch of the surface, so that for all practical purposes the current flows entirely on the surface.

Since little current flows in the interior of a conductor at radio frequencies, the effect is the same as though the current were flowing in a thin conducting tube. This is the same as reducing the cross-sectional area of the conductor, which increases its resistance. Consequently skin effect increases the resistance of a solid conductor as compared to its value for d.c. and low-frequency a.c.

Low resistance at radio frequencies can be achieved by using conductors with large surface area. Since the inner part of the conductor does not carry current, thin-walled tubing may be used for coils equally as well as solid wire of the same diameter.

In the case of inductance coils, the magnetic field close to the wire causes the current to tend to concentrate in the part of the conductor where the field is weakest, again causing an effective decrease in the conductor size and raising the resistance. These effects, plus the effects of stray currents flowing through the distributed capacity (§ 2-8) between turns, raise the effective resistance of a coil at radio frequencies to many times the d.c. resistance of the wire.

Sharpness of resonance — As the internal series resistance is increased the resonance curves become "flatter" for frequencies near the resonance frequency, as shown in Fig. 237. The relative sharpness of the resonance curve near resonance frequency is a measure of the *sharpness of tuning* or *selectivity* (ability to discriminate between voltages of different frequencies) in such circuits. This is an important consideration in tuned circuits for radio work.

Flywheel effect; Q — A resonant circuit may be compared to a flywheel in its behavior. Just as such a wheel will continue to revolve after it is no longer driven, so also will oscillations of electrical energy continue in a resonant circuit after the source of power is removed. The flywheel continues to revolve because of its stored mechanical energy; current flow continues in a resonant circuit by virtue of the energy stored in the magnetic field of the coil and the electric field of the condenser. When the applied power is shut off the energy surges back and forth between the coil and condenser, being first stored in the field of one, then released in the form of current flow, and then restored in the field of the other. Since there is always resistance present some of the energy is lost as heat in the resistance during each of these *oscillations* of energy, and eventually all the energy is so dissipated. The length of time the oscillations will continue is proportional to the ratio of the energy stored to that dissipated in each cycle of the oscillation. This ratio is called the Q (quality factor) of the circuit.

Since energy is stored by either the inductance or capacity and may be dissipated in either the inductive or capacitive branch of the circuit, a Q can be established for either the inductance or capacity alone as well as for the entire circuit. It can be shown that the energy stored is proportional to the reactance and that the energy dissipated is proportional to the resistance, so that, for either inductance or capacity associated with resistance,

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

This relationship is useful in a variety of circuit problems.

In resonant circuits at frequencies below about 28 Mc. the internal resistance is almost wholly in the coil; the condenser resistance may be neglected. Consequently, the Q of the circuit as a whole is determined by the Q of the coil. Coils for use at frequencies below the very-high-frequency region may have Qs ranging from 100 to several hundred, depending upon their size and construction.

The sharpness of resonance of a tuned circuit is directly proportional to the Q of the circuit. As an indication of the effect of Q, the current in a series circuit drops to a little less than half its resonance value when the applied frequency is changed by an amount equal to $1/Q$ times the resonant frequency. The parallel impedance of a parallel circuit similarly decreases with change in frequency. For example, in a circuit having a Q of 100, changing the applied frequency by 1/100th of the resonant frequency will decrease the parallel impedance to less than half its value at resonance.

Damping, decrement — The rate at which current dies down in amplitude in a resonant circuit after the source of power has been removed is called the *decrement* or *damping* of the circuit. A circuit with high decrement (low Q) is said to be highly damped; one with low decrement (high Q) is lightly damped.

Voltage rise — When a voltage of the resonant frequency is inserted in series in a resonant circuit, the voltage which appears across either the coil or condenser is considerably higher than the applied voltage. This is because the current in the circuit is limited only by the actual resistance of the coil-condenser combination in the circuit, and hence may have a relatively high value; however, the same current flows through the high reactances of the coil and condenser, and consequently causes large voltage drops (§ 2-8). As explained above, the reactances are of opposite types and hence the voltages are opposite in phase, so that the net voltage around the circuit is only that which is applied. The ratio of the reactive voltage to the applied voltage is proportional to the ratio of reactance to resistance, which is the Q of the circuit. Hence, the voltage across either the coil or condenser is equal to Q times the voltage inserted in series with the circuit.

If, for 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 the 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 × 10 = 2000 volts.

The ratio of reactive voltage to applied voltage is equal to the ratio of the reactance of the coil or the condenser to the resistance. Since the latter ratio equals the Q of the circuit, the reactive voltage equals the applied voltage times the Q (200/5 or 40 × 50 = 2000 volts).

Parallel-resonant circuit impedance — The parallel-resonant circuit offers pure resistance (its *resonant impedance*) between its terminals because the line current is practically in phase with the applied voltage. At frequencies off resonance the current increases through the branch having the lower reactance (and vice versa) so that the circuit becomes reactive, and the resistive component of the impedance decreases as shown in Fig. 238.

If the circuit Q is 10 or more, the parallel impedance at resonance is given by the formula

$$Z_r = X^2/R = XQ$$

where X is the reactance of either the coil or the condenser and R is the internal resistance.

Q of loaded circuits — In many applications, particularly in receiving, the only power dissipated is that lost in the resistance of the resonant circuit itself. Hence the coil should be designed to have as high Q as possible. Since, within limits, increasing the number of turns raises the reactance faster than it raises the

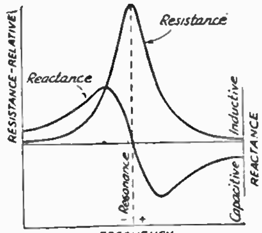


Fig. 238 — The impedance of a parallel-resonant resistance circuit is shown here separated into its reactance and resistance components. The parallel resistance of the circuit is equal to the parallel impedance at resonance.

resistance, coils for such purposes are made with relatively large inductance for the frequency under consideration.

On the other hand, 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 can be represented as shown in Fig. 239-A, where the parallel resistor represents the load to which power is delivered. If the power dissipated in the load is greater by 10 times or more than the power lost in the coil and condenser, the parallel impedance of the resonant circuit alone will be so high compared to the resistance of the load that the latter may be considered to determine the impedance of the combined circuit. (The parallel impedance of the tuned circuit alone is resistive at resonance, so that the impedance of the combined circuit may be calculated from

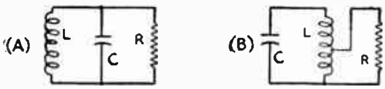


Fig. 239 — The equivalent circuit of a resonant circuit delivering power to a load. The resistor *R* represents the load resistance. At (B) the load is tapped across part of *L*, which by transformer action is equivalent to using a higher load resistance across the whole circuit.

the formula for resistances in parallel. If one of two resistances in parallel has 10 times the resistance of the other, the resultant resistance is practically equal to the smaller resistance.) The error will be small, therefore, if the losses in the tuned circuit alone are neglected. Then, since $Z = XQ$, the Q of a circuit loaded with a resistive impedance is

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

where Z is the load resistance connected across the circuit and X is the reactance of either the coil or condenser. Hence, for a given parallel impedance, the effective Q of the circuit including the load is inversely proportional to the reactance of either the coil or the condenser. A circuit loaded with a relatively low resistance (a few thousand ohms) must therefore have a large capacity and relatively small inductance to have reasonably high Q.

From the above it is evident that connecting a resistance in parallel with a resonant circuit decreases the impedance of the circuit. However, the reactances in the circuit are unchanged, hence the reduction in impedance is equivalent to a reduction in the Q of the circuit. The same reduction in impedance also could be brought about by increasing the *series* resistance of the circuit. The *equivalent series resistance* introduced in a resonant circuit by an actual resistance connected in parallel is that value of resistance which, if added in series with the coil and condenser, would decrease the circuit Q to the same value it has when the parallel resistance is connected. When the resistance of the resonant circuit alone can be neglected, the equivalent resistance is

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

the symbols having the same meaning as in the formula above.

The effect of a load of given resistance on the Q of the circuit can be changed by connecting the load across only part of the circuit. The most common method of accomplishing this is by tapping the load across part of the coil, as shown in Fig. 239-B. The smaller the portion of the coil across which the load is tapped, the less the loading on the circuit; in other words, tapping the load "down" is equivalent to connecting a higher value of load resistance across the whole circuit. This is similar in principle to impedance transformation with an iron-core transformer (§ 2-9). However, in the high-frequency resonant circuit the impedance ratio does not vary exactly as the square of the turn ratio, because all the magnetic flux lines do not cut every turn of the coil. A desired reflected impedance usually must be obtained by experimental adjustment.

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

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

$$LC = \frac{25330}{f^2}$$

where L is in microhenrys, C in micromicrofarads, and f is in megacycles.

2-11 Coupled Circuits

Energy transfer; loading — Two circuits are said to be *coupled* when energy can be transferred from one to the other. The circuit delivering energy is called the primary circuit; that receiving energy is called the secondary circuit. The energy may be practically all dissipated in the secondary circuit itself, as in receiver circuits, or the secondary may simply act as a medium through which the energy is transferred to a load resistance where it does work. In the latter case, the coupled circuits may act as a radio-frequency impedance-matching device (§ 2-9) where the matching can be accomplished by adjusting the loading on the secondary (§ 2-10) and by varying the coupling between the primary and secondary.

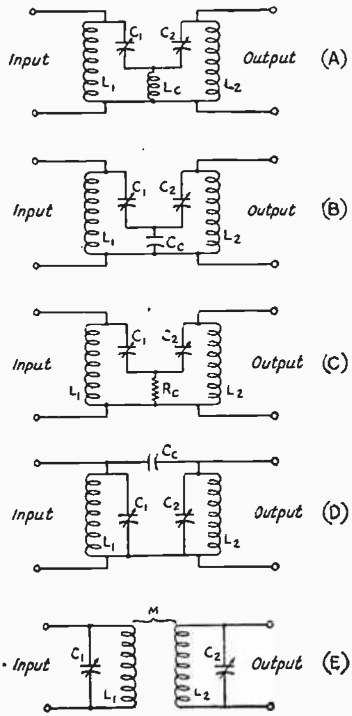


Fig. 240 — Basic methods of circuit coupling.

Coupling by a common circuit element — One method of coupling between two resonant circuits is to have some type of circuit element common to both circuits. The three variations of this type of coupling (often called *direct coupling*) shown at A, B and C of Fig. 240, utilize a common inductance, capacity 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. The degree of coupling between the two circuits becomes greater as the reactance (or resistance) of the common element is increased in comparison to the remaining reactances in the two branches.

If both circuits are resonant to the same frequency, as is usually the case, the common impedance — reactance or resistance — required for maximum energy transfer is generally quite small compared to the other reactances in the circuits.

Capacity coupling — The circuit at D shows electrostatic coupling between two resonant circuits. The coupling increases as the capacity of C_c is made greater (reactance of C_c is decreased). When two resonant circuits are coupled by this means, the capacity 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, the reactance of the coupling condenser need not be lower than 10,000 ohms or so for ample coupling. The corresponding capacity required is only a few micromicrofarads at high frequencies.

Inductive coupling — Fig. 240-E illustrates inductive coupling, or coupling by means of the magnetic field. A circuit of this type resembles the iron-core transformer (§ 2-9) but, because only a small percentage of the 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. To determine the operation of such circuits, it is necessary to take account of the mutual inductance (§ 2-5) between the coils.

Link coupling — A variation of inductive coupling, called *link coupling*, is shown in Fig. 241. This gives the effect of inductive coupling between two coils which may be so separated that they have no mutual inductance; the link may be considered simply as a means of providing the mutual inductance. Because mutual inductance between coil and link is involved at each end of the link, the total mutual inductance between two link-coupled circuits cannot be made as great as when normal inductive coupling is used. In practice, however, this ordinarily is not disadvantageous. Link coupling frequently is convenient in the design of equipment where inductive coupling would be impracticable for constructional reasons.

The link coils generally have few turns compared to the resonant-circuit coils, since the coefficient of coupling is relatively independent of the number of turns on either coil.

Coefficient of coupling — The degree of coupling between two coils is a function of their mutual inductance and self-inductances:

$$k = \frac{M}{\sqrt{L_1 L_2}}$$

where k is called the *coefficient of coupling*. It is often expressed as a percentage. The coefficient of coupling cannot be greater than 1, and generally is much smaller in resonant circuits.

Inductively coupled circuits — Three types of circuits with inductive coupling are in general use. As shown in Fig. 242, one type has a tuned-secondary circuit with an untuned-primary coil, the second a tuned-primary circuit and untuned-secondary coil, and the third uses tuned circuits in both the primary and

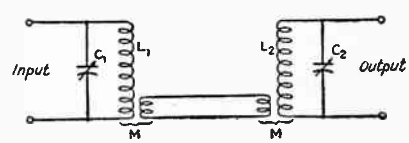


Fig. 241 — Link coupling. The mutual inductances at both ends of the link are equivalent to mutual inductance between the tuned circuits, and serve the same purpose.

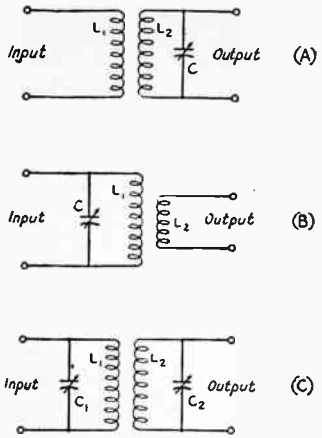


Fig. 242 — Types of inductively coupled circuits. In A and B, one circuit is tuned, the other untuned. C shows the method of coupling between two tuned circuits.

secondary. The circuit at A is frequently used in receivers for coupling between amplifier tubes when the tuning of the circuit must be varied to respond to signals of different frequencies. Circuit B is used principally in transmitters, for coupling a radio-frequency amplifier to a resistive load. Circuit C is used for fixed-frequency amplification in receivers. The same circuit also is used in transmitters for transferring power to a load which has both reactance and resistance.

If the coupling between the primary and secondary is "tight" (coefficient of coupling large), the effect of inductive coupling in circuits A and B, Fig. 242, is much the same as though the circuit having the untuned coil were tapped on the tuned circuit (§ 2-10). Thus any resistance in the circuit to which the untuned coil is connected is coupled into the tuned circuit in proportion to the mutual inductance. This is equivalent to an increase in the series resistance of the tuned circuit, and its Q and selectivity are reduced (§ 2-10). The higher the coefficient of coupling, the lower the Q for a given value of resistance in the coupled circuit. These circuits may be used for impedance matching by adjustment of the coupling and of the number of turns in the untuned coil.

If the circuit to which the untuned coil is connected has reactance, a certain amount of reactance will be "coupled in" to the tuned circuit depending upon the amount of reactance present and the degree of coupling. The chief effect of this coupled reactance is to require readjustment of the tuning when the coupling is increased, if the tuned circuit has first been adjusted to resonance under conditions of very loose coupling.

Coupled resonant circuits — The effect of a tuned-secondary circuit on a tuned primary is somewhat more complicated than in the simpler circuits just described. When the secondary is tuned to resonance with the applied frequency, its impedance is resistive only. If the primary also is tuned to resonance, the current flowing in the secondary circuit (caused by the induced voltage) will, in turn, induce a voltage in the primary which is opposite in phase to the voltage acting in series in the primary circuit. This opposing voltage reduces the effective primary voltage, and thus causes a reduction in primary current. Since the actual voltage *applied* in the primary circuit has not changed, the reduction in current can be looked upon as being caused by an increase in the resistance of the primary circuit. That is, the effect of coupling a resonant secondary to the primary is to increase the primary resistance. The resistance under consideration is the *series* resistance of the primary circuit, not the parallel impedance or resistance. The parallel resistance decreases, since the increase in series resistance reduces the Q of the primary circuit.

If the secondary circuit is not tuned to resonance, the voltage induced back in the primary by the secondary current will not be exactly out of phase with the voltage acting in the primary; in effect, reactance is coupled into the primary circuit. If the applied frequency is fixed and the secondary circuit tuning is being varied, this means that the primary circuit will have to be retuned to resonance each time the secondary tuning is changed.

If the two circuits are initially tuned to resonance at a given frequency and then the applied frequency is varied, both circuits become reactive at all frequencies off resonance. Under these conditions, the reactance coupled into the primary by the secondary retunes the primary circuit to a new resonant frequency. Thus, at some frequency off resonance, the primary current will be maximum, while at the actual resonant frequency the current will be smaller because of the resistance coupled in from the secondary at resonance. There is a point of maximum primary current both above and below the true resonant frequency.

These effects are almost negligible with very "loose" coupling (coefficient of coupling very small), but increase rapidly as the coupling increases. Because of them, the selectivity of a pair of coupled resonant circuits can be varied over a considerable range simply by changing the coupling between them. Typical curves showing the variation of selectivity are shown in Fig. 243, lettered in order of increasing co-

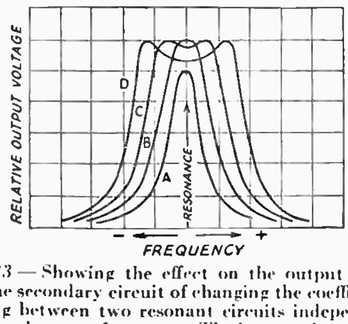


Fig. 243 — **Showing the effect on the output voltage from the secondary circuit of changing the coefficient of coupling between two resonant circuits independently tuned to the same frequency. The input voltage is held constant in amplitude while the frequency is varied.**

efficient of coupling. At loose coupling, A, the voltage across the secondary circuit (induced voltage multiplied by the Q of the secondary circuit) is less than the maximum possible because the induced voltage is small with loose coupling. As the coupling increases the secondary voltage also increases, until critical coupling, B, is reached. At still closer coupling the effect of the primary current "humps" causes the secondary voltage to show somewhat similar humps, while when the coupling is further increased the frequency separation of the humps becomes greater. Resonance curves such as those at C and D are called "flat-topped," because the output voltage is substantially constant over an appreciable frequency range.

Critical coupling — It will be observed that maximum secondary voltage is obtained in the curve at B in Fig. 243. With tighter coupling the resonance curve tends to be double-peaked, but in no case is such a peak higher than that shown for curve B. The coupling at which the secondary voltage is maximum is known as *critical coupling*. With this coupling the resistance coupled into the primary circuit is equal to the resistance of the primary itself, corresponding to the condition of matched impedances. Hence, the energy transfer is maximum at critical coupling. The over-all selectivity of the coupled circuits at critical coupling is intermediate between that obtainable with loose coupling and tight coupling. At very loose coupling, the selectivity of the system is very nearly equal to the product of the selectivities of the two circuits taken separately; that is, the effective Q of the circuit is equal to the product of the Qs of the primary and secondary.

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

With loaded circuits it is not impossible for the Q to reach such low values that critical coupling cannot be obtained even with the highest practicable coefficient of coupling (coils as close physically as possible). In such case the only way to secure sufficient coupling is to increase the Q of one or both of the coupled circuits. This can be done either by decreasing the L/C ratio or by tapping the load down on the secondary coil (§ 2-10). One or the other of these methods often must be used with link coupling, because the maximum coefficient of coupling between two coils seldom runs higher than 50 or 60 per cent and the net coefficient is approximately equal to the products of the coefficients at each end of the link. If the load resistance is known beforehand, the circuits may be designed for a Q in the vicinity of 10 or so with assurance that sufficient coupling will be available; if unknown, the proper Qs can be determined by experiment.

Shielding — Frequently it is necessary to *prevent* coupling between two circuits which, for constructional reasons, must be physically near each other. Capacitive coupling may readily be prevented by enclosing one or both of the circuits in grounded low-resistance metallic containers, called *shields*. The electrostatic field from the circuit components does not penetrate the shield, because the lines of force are short-circuited (§ 2-3). A metallic plate called a *baffle shield*, inserted between two components, may suffice to prevent electrostatic coupling between them, since very little of the field tends to bend around such a shield if it is large enough to make the components invisible to each other.

Similar metallic shielding is used at radio frequencies to prevent magnetic coupling. In this case the magnetic field induces a current (*eddy current*) in the shield, which in turn sets up its own magnetic field opposing the original field (§ 2-5). The induced current is proportional to the frequency and also to the conductivity of the shield, hence the shielding effect increases with frequency and with the conductivity and thickness of the shielding material. A closed shield is required for good magnetic shielding; in some cases separate shields, one about each coil, may be required. The baffle shield is rather ineffective for magnetic shielding, although it will give partial shielding if placed at right angles to the axes of, as well as between, the two coils to be shielded from each other.

Cancellation of part of the field of the coil reduces its inductance, and, since some energy is dissipated in the shield, the effective resistance of the coil is raised as well. Hence the Q of the coil is reduced. The effect of shielding on coil Q and inductance becomes less as the distance between the coil and shield is increased. The losses also decrease with an increase in the conductivity of the shield material. Copper and aluminum are satisfactory materials. The Q and inductance will not be greatly reduced if the spacing between the sides of the coil and the shield is at least half the coil diameter, and is not less than the coil diameter at the ends of the coil.

At audio frequencies the shielding container should be made of magnetic material, preferably of high permeability (§ 2-5), to provide a low-reluctance path for the external flux about the coil to be shielded. A nonmagnetic shield is quite ineffectual at these low frequencies since the induced current is small.

Filters — By suitable choice of circuit elements a coupling system may be designed to pass, without undue attenuation, all frequencies below and reject all frequencies above a certain value, called the *cut-off frequency*. Such a coupling system is called a *filter*, and in the above case is known as a *low-pass filter*.

If frequencies above the cut-off frequency are passed and those below attenuated, the filter is a *high-pass filter*. Simple filter circuits of both

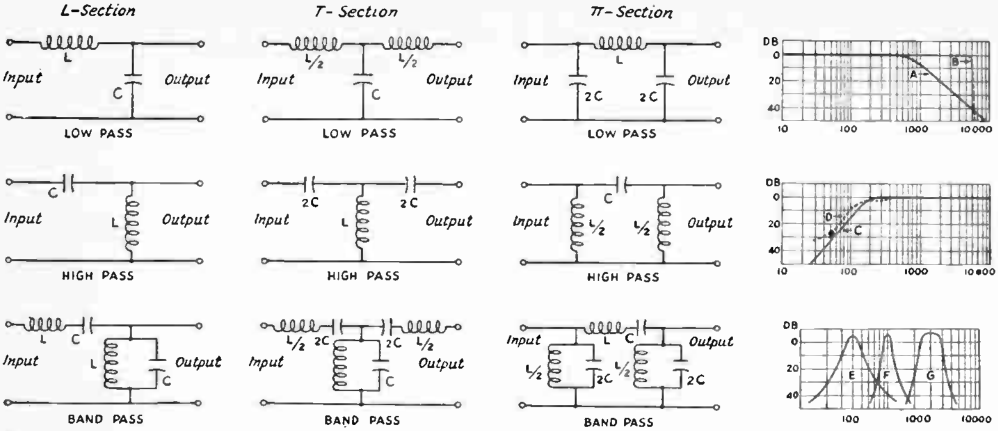


Fig. 244 — Basic forms of filter networks. Typical frequency response curves for each type are shown at the right.

types are shown in Fig. 244, along with typical frequency-response curves. The fundamental circuit, from which more complex filters are constructed, is the *L-section*. Fig. 244 also shows *π-section* and *T-section* filters, both constructed from the basic L-section.

A *band-pass* filter; also shown in Fig. 244, is a combination of high- and low-pass filter elements designed to pass without attenuation all frequencies between two selected cut-off frequencies, and to attenuate all frequencies outside these limits. The group of frequencies which is passed by the filter is called the *pass-band*. Two resonant circuits with greater than critical coupling represent a common form of band-pass filter.

In curves of Fig. 244, A shows the attenuation at high frequencies of a single-section low-pass filter with high-Q components; B illustrates the extremely sharp cut-off obtainable with a more elaborate three-section filter. Curve C is that of a high-pass section having high Q, comparable to A. D shows the attenuation by a less-efficient section having some resistance in the inductance branch. Curves E, F and G illustrate various band-pass characteristics, E being a low-Q narrow-band filter, F a high-Q narrow-band, and G a wide-band high-Q two-section filter.

Filter circuits are frequently encountered both in low-frequency and r.f. applications. The proportions of L and C for proper operation depend upon the load resistance connected across the output terminals, L being larger and C smaller as the load resistance is increased. The type of section does not affect the attenuation curve, provided the input and output resistances are correct. In a symmetrical filter the input and output impedances must be equal to the impedance for which the filter is designed. Assuming these relationships, the following design equations apply to the sections illustrated in Fig. 244.

Low-pass filter:

$$L = \frac{R}{\pi f_c} \qquad C = \frac{1}{\pi f_c R}$$

$$R = \sqrt{\frac{L_1}{C_2}} \qquad f_c = \frac{1}{\pi \sqrt{L_1 C_2}}$$

High-pass filter:

$$L = \frac{R}{4\pi f_c} \qquad C = \frac{1}{4\pi f_c R}$$

$$R = \sqrt{\frac{L_2}{C_1}} \qquad f_c = \frac{1}{4\pi \sqrt{L_2 C_1}}$$

Band-pass filter:

$$L_1 = \frac{R}{\pi(f_2 - f_1)} \qquad C_1 = \frac{f_2 - f_1}{4\pi f_1 f_2 R}$$

$$L_2 = \frac{(f_2 - f_1)R}{4\pi f_1 f_2} \qquad C_2 = \frac{1}{\pi(f_2 - f_1)R}$$

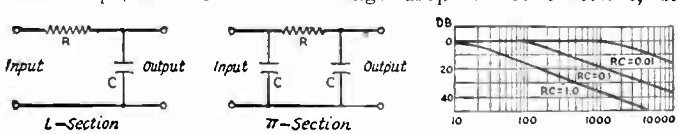
$$R = \sqrt{\frac{L_1}{C_2}} = \sqrt{\frac{L_2}{C_1}} \qquad f_M = \sqrt{f_1 f_2}$$

$$f_M = \frac{1}{2\pi\sqrt{L_1 C_1}} = \frac{1}{2\pi\sqrt{L_2 C_2}}$$

In these formulas, R is the terminal impedance and f_c the design cut-off frequency for low-pass and high-pass filters. For band-pass filters, f_1 and f_2 are the pass-band limits and f_M the middle frequency. L_2 C_2 the parallel shunt elements.

The *resistance-capacity filter*, shown in Fig. 245, is used where both d.c. and a.c. are flowing through a circuit and greater attenuation is desired for the a.c. than for d.c. It is usually employed where the direct current is small so that d.c. voltage drop is not excessive, or

Fig. 245 — L-section and π-section resistance-capacity filter circuits (left) and curves showing the attenuation in db. for three different RC products at various frequencies in the audio-frequency range.



when a voltage drop actually is required. The time constant, RC, (§ 2-6) must be large compared to the time of one cycle of the lowest frequency to be attenuated. In determining the time constant, the resistance of the load must be included as well as that in the filter itself.

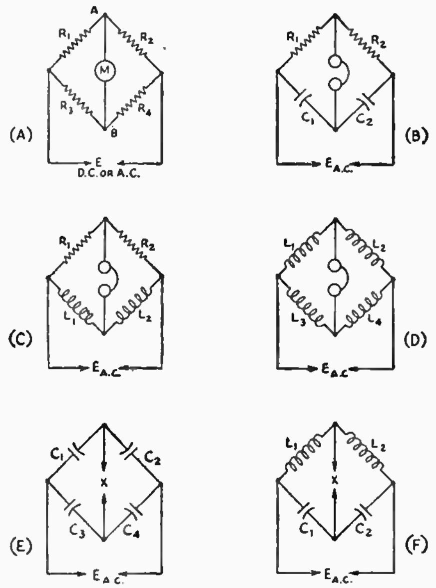


Fig. 246 — Bridge circuits utilizing resistance, inductance and capacity arms, both alone and in combination.

Bridge circuits — A *bridge circuit* is a device primarily used in making measurements of resistance, reactance or impedance (§ 2-8), and frequency, although bridges also have other applications in radio circuits.

The fundamental form is shown in Fig. 246-A. It consists of four resistances (called *arms*) connected in series-parallel to a source of voltage, E, with a sensitive galvanometer, M, connected between the junctions of the series-connected pairs. When the equation

$$\frac{R_1}{R_2} = \frac{R_3}{R_4}$$

is satisfied there is no potential difference between points A and B, since the drop across R_2 equals that across R_4 and the drop across R_1 equals that across R_3. Under these conditions the bridge is said to be *balanced*, and no current flows through M. If R_3 is an unknown resistance and R_4 is a variable known resistance, R_3 can be found from the following equation after R_4 has been adjusted to balance the bridge (*null* indication on M):

$$R_3 = \frac{R_1}{R_2} R_4$$

R_1 and R_2 are known as the *ratio arms* of the bridge; the ratio of their resistances is usually adjustable (frequently in steps of 1, 10, 100, etc.), so that a single variable resistor, R_4, can serve as a standard for measuring widely different values of unknown resistance.

Bridges similarly can be formed with arms containing capacity or inductance, and with combinations of either with resistance. Typical simple arrangements are shown in Fig. 246. For measurements involving alternating current the bridge must not introduce phase shifts which will destroy the balance, hence similar impedances should be used in each branch, as shown in Fig. 246, and the Qs of the coils and condensers should be the same. When bridges are used at audio frequencies, a telephone headset is a suitable null indicator. The bridges at E and F are commonly used in r.f. neutralizing circuits (§ 4-7); the voltage from the source, E_{ac}, is balanced out at X.

2-12-A Linear Circuits

Standing waves — If an electrical impulse is started along a wire, it will travel at approximately the speed of light until it reaches the end. If the end of the wire is open circuited, the impulse will be *reflected* at this point and will travel back again. When a high-frequency alternating voltage is applied to the wire a current will flow toward the open end, and reflection will occur continuously. If the wire is long enough so that time comparable to a half cycle or more is required for current to travel to the open end, the phase relations between the reflected current and outgoing current will vary along the wire. At one point the two currents will be 180° out of phase and at another in phase, with intermediate values between. Assuming negligible losses, the resultant current along the wire, as measured by a current-indicating instrument such as a thermo-couple ammeter, will vary in amplitude from zero to a maximum value. Such a variation is called a *standing wave*. The voltage along the wire also goes through standing waves, reaching its maximum value where the current is minimum and vice versa.

When the wire is cut to such a length that the current traverses it in one direction in exactly the time of one-half cycle, a single standing wave will occur along the wire and the wire is said to be resonant to the applied frequency. Although the inductance and capacity are distributed along the wire rather than being concentrated in a coil and condenser, such a wire is in many ways equivalent to an ordinary resonant circuit.

Frequency and wavelength — It is possible to describe the constants of such line circuits in terms of inductance and capacitance, but it is more convenient to give them simply in terms of fundamental resonant frequency or of length. Since the velocity at which the current travels is 300,000 kilometers (186,000 miles) per second, the *wavelength*, or distance the current will travel in the time of one cycle, is

$$\lambda = \frac{300{,}000}{f_{kc.}}$$

where λ is the wavelength in meters and $f_{kc.}$ is the frequency in kilocycles.

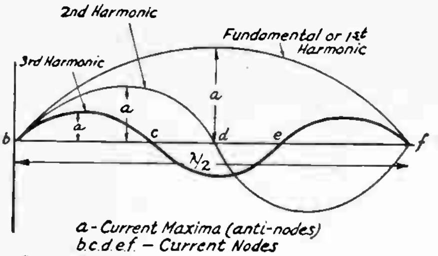


Fig. 247 — Standing-wave current distribution on a wire operating as an oscillatory circuit, at the fundamental, second harmonic and third harmonic frequencies.

Harmonic resonance — Although a coil-condenser combination having lumped constants (capacitance and inductance) resonates only at one frequency, circuits such as antennas which contain distributed constants resonate readily at frequencies which are very nearly integral multiples of the fundamental frequency. These frequencies are, therefore, in *harmonic* relationship to the fundamental frequency, and hence are referred to as *harmonics* (§ 2-7). In radio practice the fundamental itself is called the *first harmonic*, the frequency twice the fundamental is called the *second harmonic*, and so on.

Fig. 247 illustrates the distribution of current on a wire for fundamental, second and third harmonic excitation. There is one point of maximum current with fundamental operation, two when operation is at the second harmonic, and three at the third harmonic; the number of current maxima corresponds to the order of the harmonic and the number of standing waves on the wire. As noted in the figure, the points of maximum current are called *anti-nodes* (also known as "loops") and the points of zero current are called *nodes*.

In the case of the harmonic current curves, the half-wave curves are drawn alternately above and below the reference line to indicate that the phase of the current reverses in each half wavelength. In other words, if current in one half-wave section is flowing to the right, for example, the current in the adjacent half-wave section will be flowing to the left. However, when the current is measured with an r.f. ammeter there will simply be a maximum indication at the center of each half-wave section, since the ammeter cannot indicate phase.

Radiation resistance — Since a line circuit has distributed inductance and capacity, current flow causes storage of energy in magnetic and electrostatic fields (§ 2-3, 2-5). As the fields travel outward from the wire at the speed of light, some of the energy escapes from the circuit in the form of electromagnetic waves; that is, energy is *radiated* from the wire. Such a wire is, in fact, an antenna. Since the energy radiated by the line or antenna represents a loss, insofar as the line is concerned, the loss of energy can be considered to take place in an *equivalent resistance*. The value of the equivalent resistance is found from the ordinary Ohm's Law formula. $R = P/I^2$, where P is the power radiated and I is the current in the wire. R, the equivalent resistance, is called *radiation resistance*.

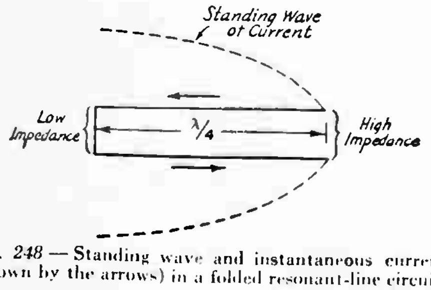


Fig. 248 — Standing wave and instantaneous current (shown by the arrows) in a folded resonant-line circuit.

Two-conductor lines — The effective resistance of a resonant straight wire is fairly high, because a large proportion of the power supplied to such a wire is radiated. In many cases it is necessary to transfer power from one point to another with the least possible loss — for example, from a transmitter to a radiating antenna which may be located some distance away. If the line is folded so that there are two conductors instead of one, as shown in Fig. 248, the currents in adjacent sections of the two wires are flowing in opposite directions, consequently the fields set up by the two oppose each other and there is very little radiation.

The quarter-wave folded line in Fig. 248 has a *total* length of one-half wavelength, hence is resonant to the frequency corresponding to its length. Since the current is large and the voltage is low at the closed end, the impedance at this point is quite low. On the other hand, the voltage is high and the current is very low at the open end, so at this point the impedance is high. These properties of a quarter-wave two-conductor line have applications to be described later.

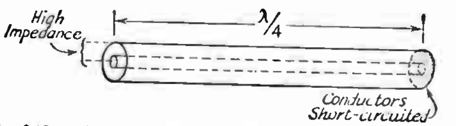


Fig. 249 — A quarter-wave coaxial-line resonant circuit.

A folded line also may be constructed in the form of two coaxial or concentric conductors, as shown in Fig. 249. In effect, this line is directly comparable with the parallel conductor line, except that one conductor may be said to have been rotated around the other in a complete circle. The *coaxial line* has even lower radiation resistance than the folded-wire line, since the outer conductor acts as a shield. Standing waves exist but are confined to the outside of the inner conductor and the inside of the outer conductor, since skin effect prevents the currents from penetrating to the other sides. Thus such a line will have no radio-frequency potentials on its exposed surfaces, and no radiation can occur. Because of the low radiation resistance and the relatively large conducting surfaces, such self-enclosed resonant lines can be made to have much higher Qs than are attainable with coils and condensers. They are most applicable at very high frequencies (very short wavelengths) (§ 2-7), where the dimensions are small.

A modified form of construction for coaxial lines is the "trough" line in which a tubular inner conductor is enclosed within a rectangular sheet-metal box or trough, usually left open on one side to facilitate tapping or other adjustments. The absence of shielding on one side does not affect the performance materially, and the simplicity of construction is an advantage.

The term *transmission line* is generally applied to all lines whether they are actually used as a means for transferring radio-frequency power between two points or whether they are used as replacements for coil-and-condenser resonant circuits. The lines shown in Figs. 248 and 249 are "short" lines of the type frequently used for the latter purpose. For transferring power the line may be many wavelengths long, depending upon the distance over which the power is to be transmitted. Furthermore, a line used for this purpose is not necessarily resonant; in fact, it may be desirable to avoid resonance effects entirely.

If a transmission line could be made infinitely long, power would simply travel along it until it was entirely dissipated in the resistance of the line; there would be nothing to reflect it and standing waves would not exist. Such a line would present a constant impedance in the form of a pure resistance to an input at *any* frequency, and hence would show no resonance effects. Practically, the characteristics of an infinitely-long line can be simulated by terminating a line of finite length in a load resistance equal to the *characteristic impedance* of the line. This and other general properties of transmission lines are discussed in the following paragraphs.

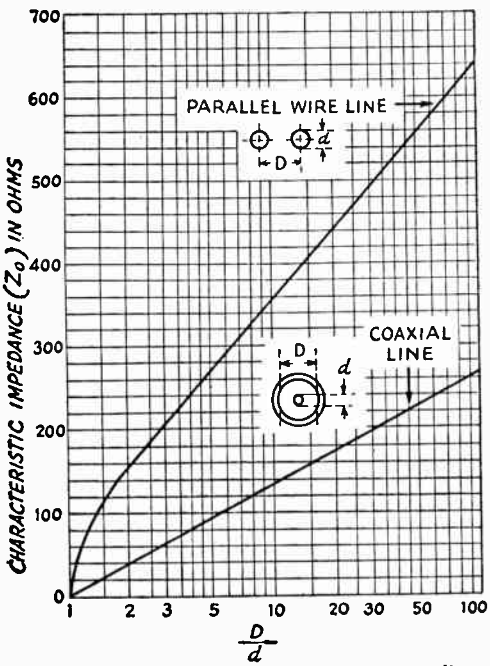


Fig. 250 — Characteristic impedance of uniform lines.

Characteristic impedance — The *characteristic impedance* of a transmission line, also known as the *surge impedance*, is defined as that impedance which a long line would present to an electrical impulse induced in the line. In an ideal line having no resistance it is equal to the square root of the ratio of inductance to capacity per unit length of the line.

The characteristic impedance of air-insulated transmission lines may be calculated from the following formulas:

Parallel-conductor line:

$$Z = 276 \log \frac{b}{a} \qquad (5)$$

where Z is the surge impedance, b the spacing, center to center, and a the radius of the conductor. The quantities b and a must be measured in the same units (inches, cm., etc.).

Coaxial or concentric line:

$$Z = 138 \log \frac{b}{a} \qquad (6)$$

where Z again is the surge impedance. In this case, b is the *inside diameter* (not radius) of the outer conductor and a is the *outside diameter* of the inner conductor. The formula is true for lines having air as the dielectric, and approximately so with ceramic insulators so spaced that the major part of the insulation is air.

The surge impedance for both parallel and coaxial lines using various sizes of conductors is given in chart form in Fig. 250.

When a solid insulating material is used between the conductors, the increase in line capacity causes the impedance to decrease by the factor $1/\sqrt{K}$, where K is the dielectric constant of the insulating material.

Although two-conductor lines have lower radiation, a single-conductor line can be used for transferring power if it is terminated in its characteristic impedance. Under such circumstances the current in the line will be small, and since radiation is proportional to current the radiation also will be small. The characteristic impedance of a single-wire transmission line varies with conductor size, height above ground, and orientation with respect to ground. An average figure is about 500 ohms.

Standing-wave ratio — The lengths of transmission lines used at radio frequencies are of the same order as the operating wavelengths, and therefore standing waves of current and voltage may appear on the line. The ratio of current (or voltage) at a loop to the value at a node (*standing-wave ratio*) depends upon the ratio of the resistance of the load connected to the output end of the line (its *termination*) to the characteristic impedance of the line itself. That is,

$$\textit{Standing-ware ratio} = \frac{Z_s}{Z_t} \text{ or } \frac{Z_t}{Z_s} \qquad (7)$$

where Z_s is the characteristic impedance of the line and Z_t is the terminating resistance. Z_t is generally called an impedance, although it must be non-reactive and therefore must correspond to a pure resistance for the line to operate as described. For example, this means that if the load or termination is an antenna, it must be resonant at the operating frequency.

The formula is given in two ways because it is customary to put the larger number in the numerator, so that the ratio will not be fractional. As an example, a 600-ohm line terminated in a resistance of 70 ohms will have a standing wave ratio of 600/70, or 8.57. The ratio on a 70-ohm line terminated in a resistance of 600 ohms would be the same. Thus, if the current as measured at a node is 0.1 ampere, the current at a loop will be 0.857 ampere.

A line terminated in a resistance equal to its characteristic impedance is equivalent to an infinitely long line; consequently there is no reflection, and no standing waves will appear. The standing wave ratio therefore is 1. The input end of such a line appears as a pure resistance of a value equal to the characteristic impedance of the line.

Electrical length — The electrical length of a line is not exactly the same as its physical length for reasons corresponding to the end effects in antennas (§ 10-2). Spacers used to separate the conductors have dielectric constants larger than that of air, so that the waves do not travel quite as fast along a line as they would in air. The lengths of electrical quarter waves of various types of lines can be calculated from the formula

$$\textit{Length (feet)} = \frac{246 \times V}{\textit{Freq. (Mc.)}}$$

where V depends upon the type of line. For lines of ordinary construction, V is as follows:

Parallel wire line	$V = 0.975$
Parallel tubing line	$V = 0.95$
Concentric line (air-insulated)	$V = 0.85$
Concentric line (rubber-insulated)	$V = 0.56$–0.65
Twisted pair	$V = 0.56$–0.65

Reactance, resistance, impedance — The input end of a line may show reactance as well as resistance, and the values of these quantities will depend upon the nature of the load at the output end, the electrical length of the line, and the line characteristic impedance. The reactance and resistance are important in determining the method of coupling to the source of power. Assuming that the load at the output end of the line is purely resistive, a line less than a quarter wavelength long electrically will show inductive reactance at its input terminals when the output termination is *less* than the characteristic impedance, and capacitive reactance when the termination is *higher* than the characteristic impedance. If the line is more than a quarter wave but less than a half wave long, the reverse conditions exist. These properties are shown in Fig. 251. With still longer lengths, the reactance characteristics reverse in each succeeding quarter wavelength. The input impedance is purely resistive if the line is an exact multiple of a quarter wave in length. The reactance at intermediate lengths is higher the greater the standing-wave ratio, being zero for a ratio of 1.

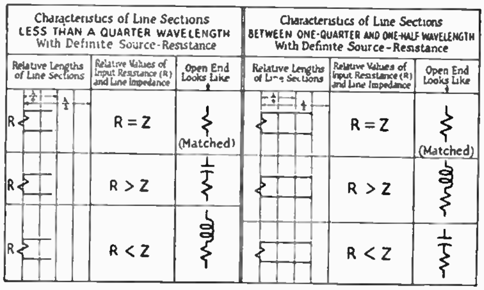


Fig. 251 — **Input reactive characteristics of resistance-terminated transmission lines as a function of line length.**

Whether lines are classified as *resonant* or *nonresonant* depends upon the standing-wave ratio. If the ratio is near 1, the line is said to be nonresonant, and reactive effects will be small even when the line length is not an exact multiple of a quarter wavelength. If the standing-wave ratio is large, the input reactance must be canceled or "tuned out" unless the line is resonant — i.e., a multiple of a quarter wavelength.

Impedance transformation — Regardless of the standing-wave ratio, the input impedance of a line a half-wave long electrically will be equal to the impedance connected at its output end; the same thing is true of a line any integral multiple of a half-wave in length. Such a line can be considered to be a one-to-one transformer. However, if the line is a quarter-wave (or an odd multiple of a quarter-wave) long, the input impedance will be equal to

$$Z_i = \frac{Z_s^{\,2}}{Z_t}$$

where Z_s is the characteristic impedance of the line and Z_t the impedance connected to the output end. That is, a quarter-wave section of line will match two impedances, Z_i and Z_t, provided its characteristic impedance, Z_s, is equal to the geometric mean of the two impedances. A quarter-wave line may, therefore, be used as an *impedance transformer*. By suitable selection of constants, a wide range of impedance-matching values can be obtained.

Since the impedance measured between the two conductors anywhere along the line will vary between the two end values, a quarter-wave line short-circuited at the output end can be used as a *linear transformer* with an adjustable impedance ratio. For best operation,

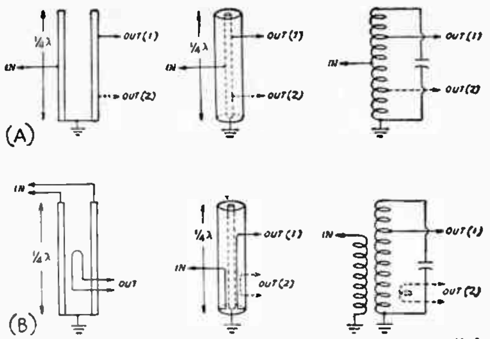


Fig. 252 — **Equivalent coupling circuits for parallel-line, coaxial-line and conventional resonant circuits.**

the two terminating impedances must be of the same order of magnitude. However, a series of quarter-wave sections can be used to obtain a step-by-step match of two terminal impedances efficiently if they are widely different.

Impedance-matching or transformation with transmission-line sections may also be effected by taps on quarter-wave resonant lines employed as coupling circuits in the same manner as conventional coil-condenser circuits. The equivalent relationships between parallel-line, coaxial-line and coil-and-condenser circuits for this purpose are shown in Fig. 252.

Other impedance-matching arrangements employ the use of matching stubs or equivalent sections so arranged so as to balance out the reactive component introduced by the coupled circuit. These are employed primarily in connection with antenna feed systems and are described in detail in § 10-8.

Transmission lines as circuit elements — Sections of transmission lines, together with combinations of such sections, can be used to simulate practically any electrical circuit property. Transmission lines can be used as resistance, inductance and capacity, as well as for resonant circuits, impedance-matching transformers, filters, and even as insulators.

When a short-circuited quarter-wavelength line is connected between a "hot" circuit and ground, the input end offers an extremely high resistive impedance. In other words, the transmission line is virtually an insulator. Insulating lines of this sort are commonly employed in ultrahigh frequency work. Such insulators can be used to provide a d.c. path between the r.f. conductor and chassis, and at the same time effectively block the flow of r.f. current.

A transmission line terminated in its characteristic impedance affords a pure resistance at high frequencies, and so may be used as a non-reactive resistor. Unterminated lines afford a variety of reactive properties. Lengths of short-circuited line less than a quarter wavelength represent pure inductive reactance, while open-circuited lines have pure capacitive reactance. Thus the former can be used in lieu of r.f. chokes, while the latter can serve as by-pass condensers.

The reactive characteristics of open- and closed-end lines are summarized in Fig. 253.

Resonant lines as tuned 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 necessary capacity. The fact that the coil has a certain amount of self-capacity of its own, as well as some resistance, while the condenser also possesses a small self-inductance, can usually be disregarded.

At the very-high and ultrahigh frequencies, however, it is no longer possible to separate these components. The connecting leads which, at lower frequencies, would serve merely to join the condenser to the coil now may have more inductance than the coil itself. The required inductance coil may be no more than a single turn of wire, yet even this single turn may have dimensions comparable to a wavelength at the operating frequency. Thus the energy in the field surrounding the "coil," may in part be radiated. At a sufficiently high frequency the loss by radiation may represent a major portion of the total energy in the circuit. Since energy which cannot be utilized as intended is wasted, regardless of whether it is consumed as heat by the resistance of the wire or simply radiated into space, the effect is as though the resistance of the tuned circuit were greatly increased and its Q greatly reduced.

For this reason, it is common practice to utilize resonant sections of transmission line as tuned circuits at frequencies above 100 Mc. A quarter-wavelength line, or any odd multiple thereof, shorted at one end and open at the other, exhibits large standing waves. When a voltage of the frequency at which such a line is resonant is applied to the open end, the response is very similar to that of a parallel resonant circuit; it will have very high input impedance at resonance and a large current flowing at the short-circuited end.

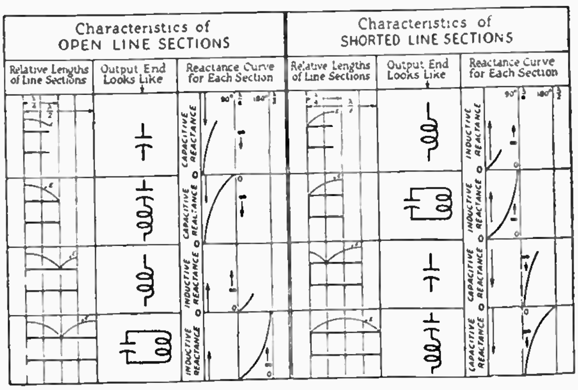


Fig. 253 — **Open and closed transmission lines as circuit elements.**

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. 253. 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 v.h.f. 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. The coaxial line is advantageous at the lower frequencies, as well, but because it is more complicated to construct and adjustments are more difficult the open type of line is generally favored at these frequencies.

Transmission-line filter networks — The same general equations can be applied to any type of electrical network whether it be an actual section of transmission line, a combination of lumped-circuit elements, or a combination of transmission-line elements. Ordinary electric filters (§ 2-11) at lower frequencies use combinations of coils and condensers, but conventional circuit elements cannot be used at extremely high frequencies. However, combinations of transmission-line sections or combinations of transmission lines and parallel-plate condensers may be used for the elements of very-high-frequency filter networks, instead.

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 the constructional 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. 254. At the left, a sliding shorting disc 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 capacity at the open end of the line has the greatest tuning effect per unit of capacity; 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 capacity "loading" of the sort illustrated will be shorter, physically, than an unloaded line resonant at the same frequency.

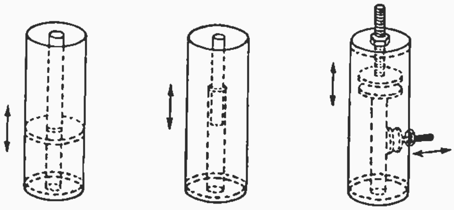


Fig. 254 — Methods of tuning coaxial resonant lines.

The short-circuiting disc at the end of the line must be designed to make perfect electrical contact. The voltage is a minimum at this end of the line; therefore, it will not break down some of the thinnest insulating films. Usually a soldered connection or a tight clamp is used to secure good contact. When the length of line must be readily adjustable, the shorting plug is provided with spring collars which make contact on the inner and outer conductors at some distance away from the shorting plug at a point where the voltage is sufficient to break down the film between the collar and conductor.

Two methods of tuning parallel-conductor lines are shown in Fig. 255. 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-capacity 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.

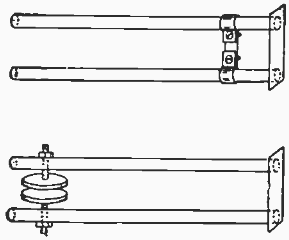


Fig. 255 — Methods of tuning parallel-type resonant lines.

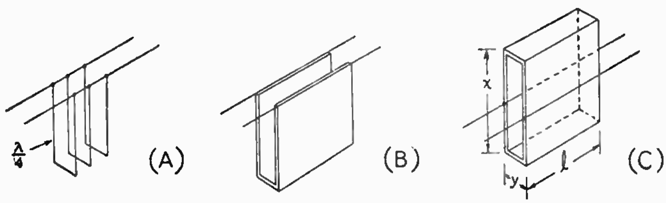


Fig. 256 — Evolution of a wave guide from a two-wire transmission line.

2-12-B Wave Guides and Cavity Resonators

Hollow wave guides — 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 v.h.f. 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. 256-A, 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. 256-C. As before, the line impedance still will not be affected. But now, instead of a two-wire transmission line, the energy is being conducted within a hollow rectangular tube.

This analogy to wave-guide operation is not exact, and therefore should not be taken too literally. In the evolution from the two-wire line to the closed tube the electric and magnetic field configurations undergo considerable change, with the result that the guide does not actually operate like a two-conductor line shunted by an infinite number of quarter-wave stubs. If it did, only waves of the proper length to correspond to the stubs would be propagated through the tube, but the fact is that such waves do *not* pass through the guide. Only waves of shorter length — that is, higher frequency — can go through. The distance x represents half the *cut-off wavelength*, or the shortest wavelength which 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 wave-length (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. 257. 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 any electric field parallel to the walls at the surface would cause an infinite current to flow in a perfect conductor. This represents an impossible situation.

Zero electric field at the end walls will result if the wave is considered to consist of two separate waves moving in zig-zag fashion down the guide, reflected back and forth from the end walls as shown in Fig. 258. Just at the walls, the positive crest of one wave meets the negative crest of the other, giving complete cancellation of the electric fields. The angle of reflection at which this cancellation occurs depends upon the width x of the guide and the length of the waves; Fig. 258-A illustrates the

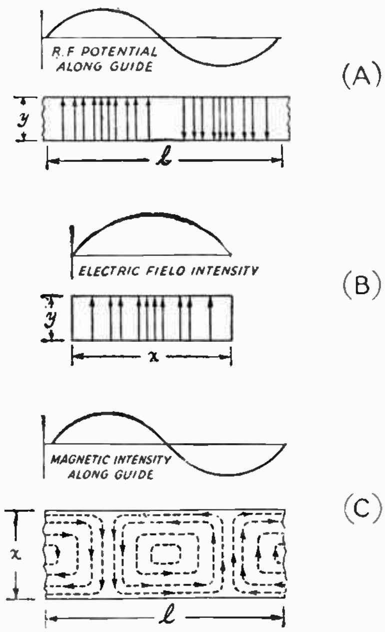


Fig. 257 — Field distribution in a rectangular wave guide. The $TE_{1,0}$ mode of propagation is depicted.

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. 258.

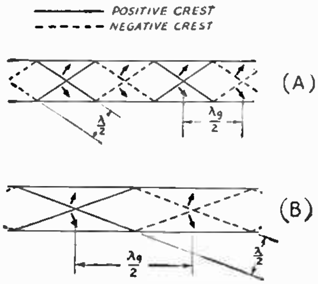


Fig. 258 — Reflection of two component waves in a rectangular guide. λ = wavelength in space. λg = wavelength in guide. Direction of wave motion is perpendicular to the wave front (crests) as shown by the arrows.

Modes of propagation — Fig. 257 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. 256; this dimension must be more than ½ 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 capacity 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 (§ 2-12-B) 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. 259. 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. 259-A 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. 259-C, the circuit may be thought of as

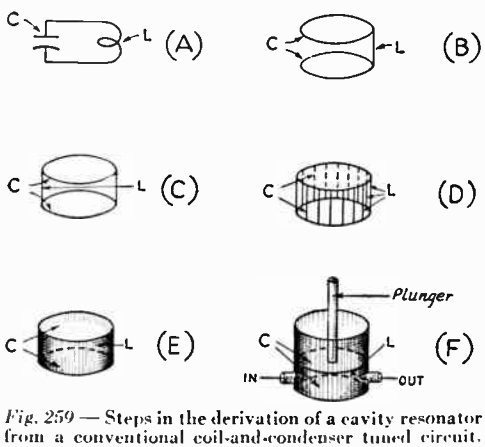


Fig. 259 — Steps in the derivation of a cavity resonator from a conventional coil-and-condenser tuned circuit.

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, as shown in Fig. 252, 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 from some v.h.f. source 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.

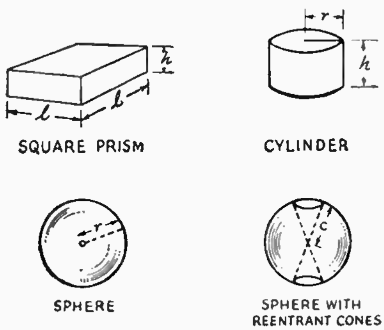


Fig. 260 — Forms of cavity resonators.

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. 260. The resonant frequency depends upon the dimensions of the cavity and the mode of oscillation of the waves (comparable to the transmission modes in a wave guide). For the lowest modes the resonant wavelengths are as follows:

Cylinder	2.61r
Square box	1.41l
Sphere	2.28r
Sphere with re-entrant cones	4r

The resonant wavelengths of the cylinder and square box are independent of the height when the height is less than a half wavelength. In other modes of oscillation the height must be a multiple of a half wavelength as measured inside the cavity. Fig. 259-F 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.

A form of cavity resonator in wide practical use is the re-entrant cylindrical type shown in Fig. 261. It is useful in connection with vac-

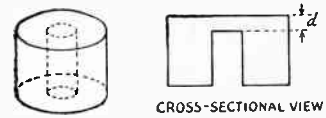


Fig. 261 — Re-entrant cylindrical cavity resonator.

uum-tube oscillators of the types described for u.h.f. use in Chapter Three. In construction it resembles a concentric line closed at both ends with capacity loading at the top, but the actual mode of oscillation may differ considerably from that occurring in coaxial lines. The resonant frequency of such a cavity depends upon the diameters of the two cylinders and the distance d between the ends of the inner and outer cylinders.

Compared to ordinary resonant circuits, cavity resonators have extremely high Q. A value of Q of the order of 1000 or more is readily obtainable, and Q values of several thousand can readily be secured with good design and construction.

Coupling to wave guides and cavity resonators— Energy may be introduced into or abstracted from a wave guide or resonator by means of either the electric or magnetic field. The energy transfer frequently is through a coaxial line, two methods for coupling to which are shown in Fig. 262. The probe shown at A is simply a short extension of the inner conductor of the coaxial line, so oriented that it is parallel to the electric lines of force. The loop shown at B is arranged so that it encloses some of the magnetic lines of force. The point at which maximum coupling will be secured depends upon the particular mode of propagation in the guide or cavity; the coupling will be maximum when the coupling device is in the most intense field.

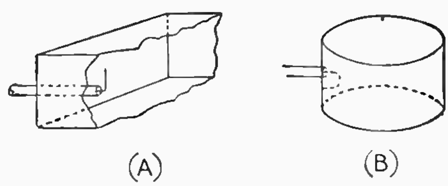


Fig. 262 — Coupling to wave guides and resonators.

Coupling can be varied by turning either the probe or loop through a 90-degree angle. When the probe is perpendicular to the electric lines the coupling will be minimum; similarly, when the plane of the loop is parallel to the magnetic lines the coupling will have its least possible value.

2-12-C Lumped-Constant Circuits

V.h.f. resonator 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.

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 discs combining half-turn inductance loops with semi-circular 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 capacity in the circuit is that between the wide shells, while the central rod comprises the inductance.

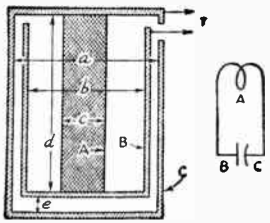


Fig. 263 — Concentric-cylinder or "pot"-type tank for v.h.f. The equivalent circuit diagram is also shown. Connections are made to the terminals marked T. For maximum Q the ratio of b to c should be between 3 and 5.

"Pot"-type tank circuits — The lumped-constant concentric-element tank in Fig. 263, commonly referred to as the "pot" circuit, is equivalent to a very short coaxial line (no linear dimension should exceed 1/20th wavelength), loaded by a large integral capacity.

The inductance is supplied by the copper rod, A. The capacity is provided by the concentric cylinders, B and C, plus the capacity between the plates at the bottoms of the cylinders.

Approximate values of capacity and inductance for tank circuits of the "pot" type can be determined by the following:

$$L = 0.0117 \log \frac{b}{c} \mu\text{h}.$$

$$C = \left(\frac{0.6225\, d}{\log \dfrac{a}{b}}\right) + \left(\frac{0.1775\, b^2}{e}\right) \mu\mu\text{fd}.$$

where the symbols are as indicated in Fig. 263, and all dimensions are in inches. The left-hand term for capacity applies to the concentric cylinders, B and C, while the second term gives the capacity between the bottom plates.

"Butterfly" circuits — The tank circuits described in the preceding section are primarily fixed-frequency devices. The "butterfly" circuits shown in Fig. 264 are capable of being tuned over an exceptionally wide range,

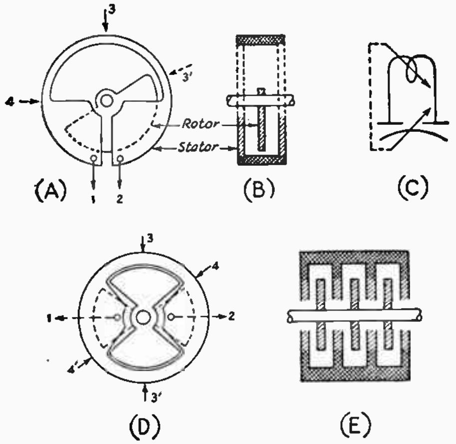


Fig. 264 — "Butterfly" tank circuits for v.h.f., showing front and cross-section views and the equivalent circuit.

while still having high Q and reasonable physical dimensions. The circuit at A is derived from a conventional balanced-type variable condenser. The inductance is in the wide circular band connecting the stator plates. At its minimum setting the rotor plate fills the opening of the loop, reducing the inductance to a minimum. Connections are made to points 1 and 2. This basic structure eliminates all connecting leads and avoids all sliding or wiping electrical contacts to a rotating member. A disadvantage is that the electrical midpoint shifts from point 3 to point 3′ as the rotor is turned. Constant magnetic coupling may be obtained by a coupling loop located at point 4, however.

In the modification shown at D, two sectoral stators are spaced 180 degrees, thereby achieving the electrical symmetry required to permit tapping for balanced operation. Connections to the circuit should be made at points 1 and 2 and it may be tapped at points 3 and 3′, which are the electrical midpoints. Where magnetic coupling is employed, points 4 and 4′ are suitable locations for coupling links.

The capacity of any butterfly circuit may be computed by the standard formula for parallel-plate condensers given in Chapter 20. 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 usual 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 may either be made in one piece or with separate sectoral stator plates and spacing rings assembled with machine screws.

2-12-D Piezoelectric Crystals

Piezoelectricity — Properly ground plates or bars of quartz and certain other crystalline materials, such as Rochelle salts, show a mechanical strain when subjected to an electric charge and, conversely, a difference in potential between two faces when subjected to mechanical stress. The relationship between mechanical force and electrical stress under such conditions is known as the *piezoelectric effect*. The charges appearing on the crystal as a result of mechanical force applied to the crystal, or of mechanical vibration of the crystal itself, are termed *piezoelectricity*.

Piezoelectric crystals may be employed as devices either for changing mechanical energy to electrical energy or for changing electrical energy to mechanical energy. In the former category are such devices as crystal microphones and phonograph pickups; in the latter, crystal headphones, crystal loud-speakers and crystal recording heads.

A properly cut crystal is a mechanical vibrator electrically equivalent to a series-resonant circuit of very high Q, and so can be also used for many of the purposes for which ordinary resonant circuits are used. The resonant frequency depends upon shape, thickness, length and cut.

Natural quartz crystals are usually in the form of a hexagonal prism terminated at one or both ends by a six-sided pyramid. Joining the vertices of these pyramidal ends, and perpendicular to the plane of the hexagonal cross section, is the optical or Z axis. The three electrical or X axes lie in a plane perpendicular to the optical axis and passing through opposite corners of the hexagon. The three mechanical or Y axes lie in the same plane but perpendicularly to the sides of the hexagon.

Active plates cut from a raw crystal at various angles to its optical, electrical and mechanical axes have differing characteristics as to thickness, frequency-temperature coefficient, power-handling capabilities, etc. The basic cuts are designated X and Y after their respective axes, but a variety of specialized cuts, such as the AT, are in more common use.

Frequency-thickness ratio — At frequencies above about 500 kc. the thickness of the crystal is the principal frequency-determining factor, the other dimensions being of relatively minor importance. Thickness and frequency are related by a constant, K, such that

$$f = \frac{K}{t}$$

where f is the frequency in megacycles and t the thickness of the crystal in mils. For the X-cut, $K = 112.6$; Y-cut, $K = 77.0$; AT-cut, $K = 66.2$, BT-cut, $K = 97.3$.

At frequencies above about 10 Mc. the crystal becomes very thin and correspondingly fragile, so that crystals seldom are manufactured for fundamental operation above this frequency. Direct crystal control on 14 and 28 Mc. is secured by use of "harmonic" crystals, which are ground to be active oscillators when excited at a harmonic (usually the third).

Temperature coefficient of frequency — The resonant frequency of a crystal varies with temperature, the variation depending upon the type of cut. The frequency change is usually expressed as a coefficient relating the number of cycles of frequency change per megacycle per °C. It may be either positive (increasing frequency with increasing temperature) or negative (decreasing frequency with increasing temperature). X-cut crystals have a negative coefficient of 15 to 25 cycles/Mc./°C. The coefficient of Y-cut crystals may vary from −20 cycles/Mc./°C. to +100 cycles/Mc./°C.

Variations in frequency caused by temperature changes can be minimized by proper cutting of the plate. By orienting the plate through various angles in relation to its optical, electrical and mechanical axes, a compensatory relationship can be derived between the dimensions of the plate, its density, and its elastic constants — the components responsible for the temperature coefficient.

The AT cut is the type perhaps most extensively used for transmitter frequency control. This plate can be ground to almost any frequency between 300 and 5000 kc. Its complement, the BT cut, is used for frequencies within the range 4500 to 10,000 kc.

For frequencies below 500 kc., CT and DT shear-type cuts have been developed which depend not upon thickness but on length and width for determining frequency. Plates of the CT and DT type vibrating at a harmonic mode are designated ET or FT cuts.

The low-drift types described above show a zero temperature coefficient through only a few degrees of change. Another type of cut, the GT, will drift less than 1 cycle/Mc./°C. over a temperature change of 100° C. In this plate a face shear vibration is changed into two longitudinal vibrations coupled together. At a certain ratio of length to width one mode

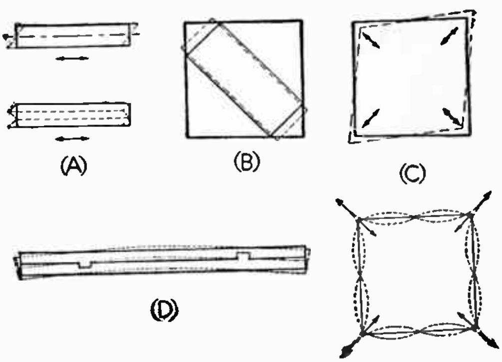


Fig. 265 — Modes of vibration for various crystal cuts. A — Fundamental (above) and harmonic (below) of the AT and BT cuts. B — The GT cut. C — CT and DT cuts (above) and ET and FT cuts (below). D — NT cut.

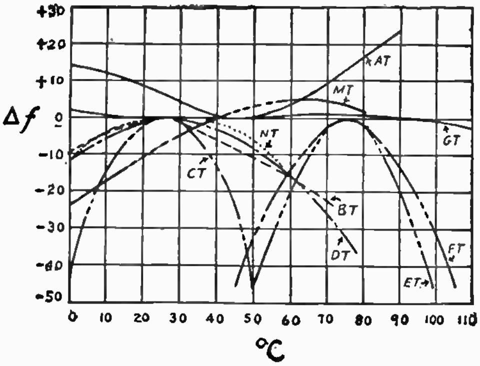


Fig. 266 — Frequency change in parts per million vs. variation in temperature in °C. for various crystal cuts.

has a zero temperature coefficient, making it especially useful as a frequency standard. The MT cut, which also vibrates longitudinally, can be used from 50 to 100 kc. The NT crystal is a flexurally vibrating cut having a low temperature coefficient in the range from 4 to 50 kc. MT and NT cuts are useful for phase-modulated f.m. transmitters.

2-13 Miscellaneous Circuit Details

Combined a.c. and d.c. — There are many practical instances of simultaneous flow of alternating and direct currents in a circuit. When this occurs there is a *pulsating* current, and it is said that an alternating current is *superimposed* on a direct current. As shown in Fig. 267, the maximum value is equal to the d.c. value plus the a.c. maximum, while the minimum value (on the negative a.c. peak) is the difference between the d.c. and the maximum a.c. values. The average value (§ 2-7) of the current is simply equal to the direct-current component alone. The effective value (§ 2-7) of the combination is equal to the square root of the sum of the effective a.c. squared and the d.c. squared:

$$I = \sqrt{I_{ac}^2 + I_{dc}^2}$$

where I_{ac} is the effective value of the a.c. component, I is the effective value of the combination, and I_{dc} is the average (d.c.) value of the combination.

Beats — If two or more alternating currents of different frequencies are present in a normal circuit they will have no particular effect upon one another and can be separated again by the proper selective circuits. However, if two (or more) alternating currents of different frequencies are present in an element having unilateral or one-way current flow properties, not only will the two original frequencies be present in the output but also currents having frequencies equal to the sum, and difference, of the original frequencies. These sum and difference frequencies are called the *beat* frequencies. For example, if frequencies of 2000 and 3000 kc. are present in a normal circuit only those two frequencies exist, but if they are passed through a unilateral element there will be present in the output not only the two original frequencies of 2000 and 3000 kc. but also currents of 1000 (3000 − 2000) and 5000 (3000 + 2000) kc. Suitable circuits can be used to select the desired beat frequency. The human ear has unilateral characteristics and is, therefore, capable of hearing audible beat frequencies. Electronic devices of this nature are called mixers, converters, and detectors.

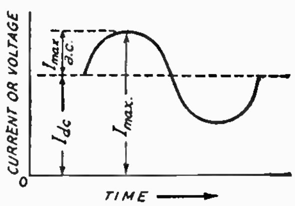
By-passing — In combined circuits, it is frequently necessary to provide a low-impedance path for a.c. around, for instance, a source of d.c. voltage. This can be done by using a *by-pass condenser*, which will not pass direct current but will readily permit the flow of alternating current. The capacity of the condenser should be of such value that its reactance is low (of the order of 1/10th or less) compared to the a.c. impedance of the device being by-passed. The lower the reactance, the more effectively will the alternating current be confined to the desired path.

Similarly, alternating current can be prevented from flowing through a direct-current circuit to which it may be connected by inserting an inductance of high reactance (called a *choke coil*) between the two circuits. This will permit the direct current to flow without hindrance, since the resistance of the choke coil may be made quite low, but will effectively prevent the alternating current from flowing where it is not wanted.

If both r.f and low-frequency (audio or power) currents are present in a circuit, they may be confined to desired paths by similar means, since an inductance of high reactance for radio frequencies will have negligible reactance at low frequencies, while a condenser of low reactance at radio frequencies will have high reactance at low frequencies.

Grounds — The term "ground" is frequently encountered in discussions of circuits. Normally it means the voltage reference point in the circuit. There may or may not be an actual connection to earth, but it is understood that a point in the circuit said to be at *ground potential* could be connected to earth without disturbing the operation of the circuit in any way. In direct-current circuits, the negative side generally is grounded. The ground symbol in circuit diagrams is used for convenience in indicating common connections between various parts of the circuit, as through a metal chassis, and, with respect to actual ground, usually has the meaning indicated above.

Fig. 267 — Pulsating current, composed of an alternating current or voltage superimposed on a steady direct current or voltage.



Chapter Three

Vacuum Tubes

3-1 Diodes

Rectification — Practically all of the vacuum tubes used in radio work depend upon thermionic conduction (§ 2-4) for their operation. The simplest type of vacuum tube is that shown in Fig. 301. It has two elements, a cathode and a plate, and is called a *diode*. When heated by the "A" battery the cathode emits electrons, which are attracted to the plate if the plate is at a positive potential with respect to the cathode.

Because of the nature of thermionic conduction, the tube is a conductor in one direction only. If a source of alternating voltage is connected between the cathode and plate, then electrons will flow only on the positive half-cycles of alternating voltage; there will be no electron flow during the half cycle when the plate is negative with respect to the cathode. Thus the tube can be used as a *rectifier*, to change alternating current to pulsating direct current. This alternating current can be anything from the 60-cycle kind to the highest radio frequencies.

Rectification finds its chief applications in detecting radio signals and in power supplies. These are treated in Chapters Seven and Eight, respectively.

Characteristic curves — The performance of the tube can be reduced to easily understood terms by making use of tube *characteristic curves*. A typical characteristic curve for a diode is shown at the right, in Fig. 301. It shows the current flowing between plate and cathode with different d.c. voltages applied between the elements. The curve of Fig. 301 shows that, with fixed cathode temperature, the plate current increases as the voltage between cathode and plate is raised. For an actual tube the values of plate current and plate voltage would be plotted along their respective axes.

The power consumed in the tube is the product of the plate voltage multiplied by the plate current, just as in any d.c. circuit. In a vacuum tube this power is dissipated in heat developed in the plate and radiated to the bulb.

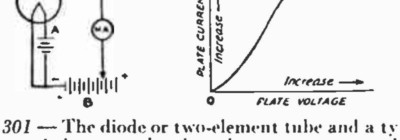


Fig. 301 — The diode or two-element tube and a typical characteristic curve showing plate current vs. voltage.

Space charge — With the cathode temperature fixed the total number of electrons emitted is always the same, regardless of the plate voltage. Fig. 301 shows, however, that less plate current will flow at low plate voltages than when the plate voltage is large. With low plate voltage, only those electrons nearest the plate are attracted to the plate. The electrons in the space near the cathode, being themselves negatively charged, tend to repel the similarly charged electrons leaving the cathode surface and cause them to fall back on the cathode. This is called the *space-charge* effect. As the plate voltage is raised more and more electrons are attracted to the plate, until finally the space charge effect is completely overcome. When this occurs all the electrons emitted by the cathode are attracted to the plate, and a further increase in plate voltage can cause no further increase in plate current. This condition is called *saturation*.

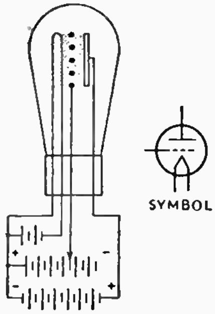
3-2 Triodes

Grid control — If a third element, called the *control grid*, or simply the *grid*, is inserted between the cathode and plate of the diode, the space-charge effect can be controlled. The tube then becomes a *triode* (three-element tube) and is useful for more things than rectification. The grid is usually in the form of an open spiral or mesh of fine wire. If the grid is connected externally to the cathode so that it is at the same potential as the cathode, and a steady voltage from a d.c. supply is then applied between the cathode and plate (the positive of the "B" supply is always connected to the plate), there will be a constant flow of electrons from cathode to plate through the openings of the grid, much as in the diode. However, if the grid is given a positive potential with respect to the cathode, the space charge will be partially neutralized and there will be an increase in plate current. If the grid is made negative with respect to the cathode, the space charge will be reinforced and the current will decrease.

This effect of grid voltage can be shown by curves in which plate current is plotted against grid voltage. At any given value of grid voltage the plate current will still depend upon the plate voltage, so if complete information about the tube is to be secured it is necessary to plot a *series* of curves taken with various values of plate voltage. Such a set of grid voltage vs. plate current curves, typical of a small receiving triode, is shown in Fig. 303.

So long as the grid has a negative potential with respect to the cathode, electrons emitted

Fig. 302 — Illustrating the 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. Battery symbols follow those of the usual schematic diagrams, while the schematic tube symbol is shown at the right.



by the cathode are repelled (§ 2-3) from the grid, with the result that no current flows to the grid. Hence, under these conditions, the grid consumes no power. However, when the grid becomes positive with respect to the cathode, electrons are attracted to it, and a current flows to the grid; when this *grid current* flows, power is dissipated in the grid circuit.

In addition to the set of curves showing the relationship between grid voltage and plate current at various fixed values of plate voltage, two other sets of curves may be plotted to show the characteristics of a triode. These are the plate voltage vs. plate current characteristic, which shows the relationship between plate voltage and plate current for various fixed values of grid voltage, and the constant-current characteristic, which shows the relationship between plate voltage and grid voltage for various fixed values of plate current.

Amplification — The grid evidently acts as a valve to control the flow of plate current, and it is found that it has a much greater effect on plate current flow than does the plate voltage; that is, a small change in grid voltage is just as effective in bringing about a given change in plate current as is a large change in plate voltage.

The fact that a small voltage acting on the grid is equivalent to a large voltage acting on the plate indicates the possibility of *amplification* with the triode tube; that is, the generation of a large voltage by a small one, or the generation of a relatively large amount of power from a small amount. The many uses of the electronic tube nearly all are based upon this amplifying feature. The amplified power or voltage output from the tube is obtained, not 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 device for consuming it, or for transferring it to another circuit, must be connected in the plate circuit, since no particularly useful purpose would be served in having the current merely flow through the tube and the source of e.m.f. Such a device is called the *load*, and may be either a resistance or an impedance. The term "impedance" is frequently used even though the load may be purely resistive.

Amplification factor — The relative effect of the grid and plate voltages on the plate current is measured by the *amplification factor* of the tube, usually represented by the Greek letter μ. Amplification factor is defined as the ratio of the change in plate voltage required to produce a given change in plate current to the change in grid voltage required to produce the same plate-current change. Strictly speaking, very small changes in both grid and plate voltage must be used in determining the amplification factor, because the curves showing the relationship between plate voltage and plate current, and between grid voltage and plate current, are not perfectly straight, especially if the plate current is nearly zero. This indicates that the amplification factor varies at different points along the curves, and different values will be obtained as larger or smaller voltage differences are taken for the purpose of calculating it. The expression for amplification factor can be written:

$$\mu = \frac{\Delta E_p}{\Delta E_g}$$

where ΔE_p indicates a very small change in plate voltage and ΔE_g is the change in grid voltage producing the same plate current change. The symbol Δ (the Greek letter *delta*) indicates a small increment, or small change.

The amplification factor is simply a ratio, and has no unit.

Plate resistance — Since only a limited amount of plate current flows when a given voltage is applied between plate and cathode, it is evident that the plate-cathode circuit of the tube has resistance. However, there is no simple relationship between plate voltage and plate current, so that in general the plate circuit of the tube does not follow Ohm's Law. Under a given set of conditions the application of a given plate voltage will cause a certain plate current to flow, and if the plate voltage is divided by the plate current a "resistance" value will be obtained which frequently is called the "d.c. resistance" of the tube. This "d.c. resistance" will be different for every value of plate voltage and also for different values of grid voltage, since the plate current also depends upon the grid voltage when the plate voltage is fixed.

In applications of the vacuum tube, it is more

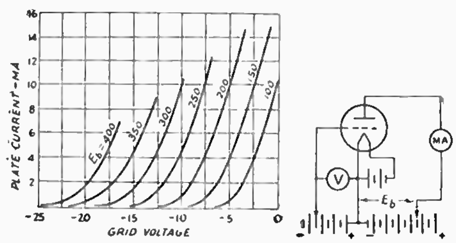


Fig. 303 — Grid voltage vs. plate current curves at various fixed values of plate voltage (E_b) for a typical small triode. Characteristic curves of this type can be taken by varying the battery voltages in the circuit at the right.

important to know how the plate current changes with a *change* in plate voltage than it is to know the relationship between the actual values of plate current and plate voltage. The relationship between plate-current change and plate-voltage change determines the *a.c. plate resistance* of the tube. This resistance, which usually is designated r_p, is significant when there is an a.c. component in the plate current. It can be found from the plate voltage vs. plate current characteristic curves. That is,

$$r_p = \frac{\Delta E_p}{\Delta I_p}$$

where ΔE_p is a small change in plate voltage and ΔI_p the corresponding small change in plate current, the grid voltage being fixed.

Plate resistance is expressed in ohms, since it is the ratio of voltage to current. The value of plate resistance will, in general, change with the particular voltages applied to the plate and grid. It depends as well upon the structure of the tube; low-μ tubes have relatively low plate resistance and high-μ tubes have high plate resistance.

Transconductance — The effect of grid voltage upon plate current is expressed by the *grid-plate transconductance* of the tube. Transconductance is a general term giving the relationship between the voltage applied to one electrode and the current which flows, as a result, in a second electrode. As in the previous two cases, it is defined as the *change* in current through the second electrode caused by a change in voltage on the first. Thus the grid-plate transconductance, commonly called the *mutual conductance*, is

$$g_m = \frac{\Delta I_p}{\Delta E_g}$$

where g_m is the mutual conductance, ΔI_p the change in plate current, and ΔE_g the change in grid voltage, the plate voltage being fixed. As before, the sign Δ indicates that the changes must be small. Transconductance is measured in mhos, since it is the ratio of current to voltage. The unit usually employed in connection with vacuum tubes is the *micromho* (one millionth of a mho), because the conductances are small. By combining with the two preceding formulas, it can be shown that $g_m = \mu / r_p$.

The mutual conductance of a tube is a rough indication of its merit as an amplifier, since it includes the effects of both amplification factor and plate resistance. Its value varies with the voltages applied to the plate and grid. With the plate voltage fixed, the mutual conductance decreases when the grid is made increasingly negative with respect to the cathode. This characteristic frequently can be used to advantage in the control of amplification, since the amount of amplification can be varied over wide limits simply by adjusting the value of a steady voltage applied to the grid.

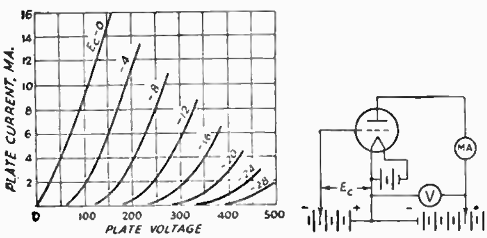


Fig. 304 — Plate voltage vs. plate current curves at various fixed values of negative grid voltage for the same triode as that used to obtain the curves in Fig. 303.

Static and dynamic curves — Curves of the type shown in Figs. 301 and 303 are called *static* curves. They show the current which flows when various voltages are applied directly to the tube electrodes. Another useful set of static curves is the "plate family," or plate voltage vs. plate current characteristic. A typical set of curves of this type is shown in Fig. 304.

A curve showing the relationship between grid voltage and plate current when a load resistance is connected in the plate circuit is called a *dynamic* characteristic curve. Such a curve includes the effect of the load resistance, and hence is more indicative of the performance of the tube as an amplifier. With a fixed value of plate-supply voltage the actual value of voltage between the plate and cathode of the tube will depend upon the amount of plate current flowing, since the plate current also flows through the load resistance and therefore results in a voltage drop which must be subtracted from the plate-supply voltage. The dynamic curve includes the effect of this voltage drop. Consequently, the plate current always is lower, for a given value of grid bias and plate-supply voltage, with the load resistance in the circuit than it is without it.

Representative dynamic characteristics are shown in Fig. 305. These were taken with the same type of tube whose static curves are shown in Fig. 303. Different curves would be obtained with different values of plate-supply voltage, E_b; this set is for a plate-supply voltage of 300 volts. Note that increasing the value of the load resistance reduces the plate current at a given bias voltage, and also that the curves are straighter with the higher values of load resistance. Zero plate current always occurs at the same value of negative grid bias, since at zero plate current there is no voltage drop in the load resistance and the full supply voltage is applied to the plate.

Fig. 306 shows how the plate current responds to an alternating voltage (*signal*) applied to the grid. If the plate current is to have the same waveshape as that of the signal, it is necessary to confine the operation to the straight section of the curve. To do this, it is necessary to select an *operating point* near the middle of the straight portion; this operating point is determined by the fixed voltage (*bias*) applied to the grid. The alternating signal voltage then adds to or subtracts from the grid bias, depending upon whether the instantaneous signal voltage is negative or positive with respect to the cathode, and causes a corresponding variation in plate current. The maximum departure of instantaneous grid voltage or plate current from the operating point is called the *swing*. The varying plate current flows through the load resistance, causing a varying voltage drop which constitutes the useful output voltage of the tube.

The point at which the plate current is reduced to zero is called the *cut-off point*. The value of negative grid voltage at which cut-off occurs depends upon the amplification factor of the tube and the plate voltage. It is approximately equal to the plate-supply voltage divided by the amplification factor.

Interelectrode capacities — Any pair of elements in a tube forms a miniature condenser (§ 2-3), and, although the capacities of these condensers may be only a few micromicrofarads or less, they must frequently be taken into account in vacuum-tube circuits. The capacity from grid to plate (*grid-plate capacity*) has an important effect in many applications. In triodes, the other capacities are the *grid-cathode* and *plate-cathode*. In multi-element tubes (§ 3-6), similar capacities exist between these and other electrodes. With screen-grid tubes, the terms "input" and "output" capacity mean, respectively, the capacity measured from grid to all other elements connected together and from plate to all other elements connected together. The same terms are used with triodes but are not so easily defined, since the effective capacities existing depend upon the operating conditions (§ 3-3).

Tube ratings — Specifications of suitable operating voltages and currents are called *tube ratings*. Ratings include proper values for filament or heater voltage and current, plate voltage and current, and similar operating specifications for other elements. An important rating in power tubes is the *maximum safe plate dissipation*, or the maximum power that can be dissipated continuously in heat on the plate (§ 3-1).

ℂ 3-3 Amplification

Principles — The operation of a simple amplifier, which was described briefly in the preceding section, is shown in more detail in Fig. 307. The load in the plate circuit is the resistor, R_p. For the sake of example, it is assumed that the plate-supply voltage is 300 volts, the negative grid bias is 5 volts, and the plate current at this bias when R_p is 50,000 ohms is 2 milliamperes (0.002 ampere). If no signal is applied to the grid circuit, the voltage drop in the load resistor is 50,000 × 0.002, or 100 volts, leaving 200 volts between the plate and cathode.

If 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 and, as shown by the graph, 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

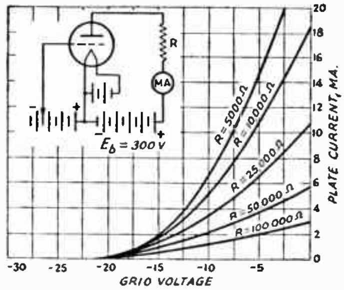


Fig. 305 — Dynamic characteristics of a small triode with various load resistances from 5,000 to 100,000 ohms.

ma. At intermediate values of grid voltage, intermediate plate-current values will occur. The instantaneous voltage between the plate and cathode of the tube also is shown on the graph. When the plate current is maximum the instantaneous voltage drop in R_p is 50,000 × 0.00265 or 132.5 volts, and when the plate current is minimum the instantaneous voltage drop in R_p is 50,000 × 0.00135 or 67.5 volts. The actual voltage between plate and cathode is therefore the difference between the plate-supply voltage, 300 volts, and these voltage drops in the load resistance, or 167.5 and 232.5 volts, respectively.

The varying plate voltage is an a.c. voltage superimposed (§ 2-13) on the steady plate-cathode voltage of 200 volts, which was 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, or 32.5 volts. 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 volt-

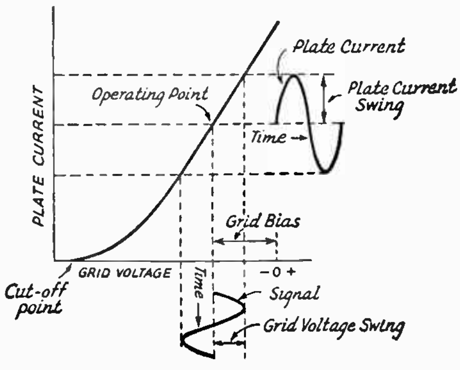


Fig. 306 — Behavior of the plate current of a vacuum tube in response to an alternating signal voltage superimposed on a steady negative grid voltage or bias.

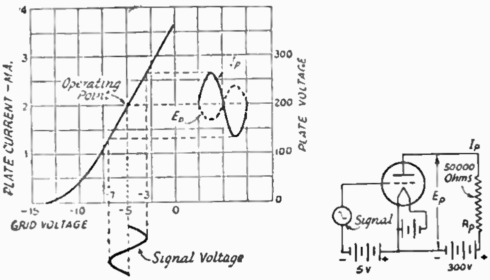


Fig. 307 — Amplifier operation. When the plate current varies in response to the signal applied to the grid, a varying voltage drop appears across the load, R_p, as shown by the dashed curve, E_p. I_p is the plate current.

age will be obtained from the plate circuit as is applied to the grid circuit.

It will be observed that only the alternating plate and grid voltages enter into the calculation of the amplification ratio. The d.c. plate and grid voltages are of course essential to the operation of the tube, since they set the operating point, but otherwise their presence may be ignored. This being the case, it is possible to show that the tube can be replaced by an *equivalent generator* which has an internal resistance equal to the a.c. plate resistance of the tube (§ 3-2) at the operating point chosen and which generates a voltage equal to the amplification factor of the tube multiplied by the signal voltage applied to the grid. The equivalent generator, together with the load resistance, R_p, is shown in Fig. 308. This simplification enables ready calculation of the amplification. If the generated voltage is μE_g, then the same current flows through r_p and R_p, and hence the voltage drop across R_p, which is the useful output voltage, is

$$E_o = \mu E_g \frac{R_p}{r_p + R_p}$$

since R_p and r_p together constitute a voltage divider (§ 2-6). The voltage-amplification ratio is given by the output voltage divided by the input voltage, hence dividing the above expression by E_g gives the following formula for the amplification of the tube:

$$\text{Amplification} = \frac{\mu R_p}{r_p + R_p}$$

This expression shows that, to obtain a large voltage-amplification ratio, it is necessary to make the plate load resistance, R_p, large compared to the plate resistance, r_p, of the tube. The maximum possible amplification, obtained when R_p is infinitely larger than r_p, is equal to the μ of the tube. A tube with a large value of μ will, in general, give more voltage amplification than one with a medium or low value. However, the advantage of the high μ is less than might be expected, because a high-μ tube usually also has a correspondingly high value of r_p, so that a high value of load resistance must be used to realize an appreciable part of the possible amplification. This in turn not only requires the use of high values of plate-supply voltage, but has some further disadvantages to be described later.

Amplifiers in which the voltage output, rather than the power output, is the primary consideration are called *voltage amplifiers*.

Power in grid circuit — In the operation depicted in Fig. 306, the grid is always negative with respect to the cathode. If the peak signal voltage is larger than the bias voltage, the grid will be positive with respect to the cathode during part of the signal cycle. Grid current will flow during this time, and the signal source will be called upon to furnish power during the period while grid current is flowing. In many cases the signal source is not capable of furnishing appreciable power, so that care must be taken to avoid "driving the grid positive."

When dealing with small signals the source of signal voltage frequently has high internal resistance, so that a considerable voltage drop occurs in the source itself whenever it is called upon to furnish grid current. Since this voltage drop occurs only during part of the cycle, the voltage applied to the grid undergoes a change in waveshape because of the current flow. This is shown in Fig. 309, where a sine-wave signal is generated but, because of the internal resistance of the source, is *distorted* at the grid of the tube during the time when grid current flows.

If the internal resistance of the signal source is low, so that the internal voltage drop is negligible when current flows, this distortion does not occur. With such a source, it is possible to operate over a greater portion of the amplifier characteristic.

Harmonic distortion — If the operation of the tube is not confined to a straight or linear portion of the dynamic characteristic, the waveshape of the output voltage will not be exactly the same as that of the signal voltage. This is shown in Fig. 310, where the operating point is selected so that the signal voltage swings into the curved part of the characteristic. While the upper half-cycle of plate current reproduces the sine-wave shape of the positive half-cycle of signal voltage, the lower half-cycle of plate current is considerably distorted and bears little resemblance to the upper half-cycle of plate current.

As explained in § 2-7, a non-sinusoidal waveshape can be resolved into a number of sine-wave components or harmonics which are integral multiples of the lowest frequency present. Consequently, this type of distortion is known as *harmonic distortion*. Distortion re-

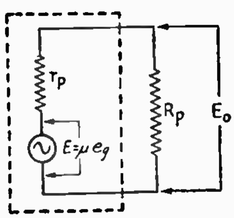


Fig. 308 — Equivalent circuit of the vacuum-tube amplifier. The tube is replaced by an equivalent generator having an internal resistance equal to the a.c. plate resistance of the vacuum tube.

sulting from grid-current flow, described in the preceding paragraph, also is harmonic distortion. Harmonic distortion from either or both causes may arise in the same amplifier.

Harmonic distortion may or may not be tolerable in an amplifier. At audio frequencies it is desirable to keep harmonic distortion to a minimum, but radio-frequency amplifiers are frequently operated in such a way that the r. f. wave is greatly distorted.

Frequency distortion — Another type of distortion, known as *frequency distortion*, occurs when the amplification varies with the frequency of the a.c. voltage applied to the grid circuit of the amplifier. It is not necessarily accompanied by harmonic distortion. It can be shown by a *frequency-response curve* or graph in which the relative amplification is plotted against frequency over the frequency range of interest.

Resistance-coupled amplifiers — An amplifier with a resistance load is known as a "resistance-coupled" amplifier. This type of amplifier is widely used for amplification at audio frequencies. A simplified circuit is shown in Fig. 311, where the amplifier is coupled to a following tube. Since all the power output of a resistance-coupled amplifier is consumed in the load resistor such amplifiers are used almost wholly for voltage amplification, usually working into still another amplifier.

A single amplifier is called a *stage* of amplification, and a number of amplifier stages in succession are said to be in *cascade*.

The purpose of the coupling condenser, C_c, is to transfer to the grid of the following tube the a.c. voltage developed across R_p, and to prevent the d.c. plate voltage on tube *A* from being applied to the grid of tube *B*. The grid resistor, R_g, transfers the bias voltage to the grid of tube *B* and prevents short-circuiting the a.c. voltage through the bias battery. Since no grid current flows, there is no d.c. voltage drop in R_g; consequently the full bias voltage is applied to the grid. In order to obtain the maximum a.c. voltage at the grid of tube *B* the reactance of the coupling condenser must be small compared to the resistance of R_g, so that most of the voltage will appear across R_g rather than across C_c. Also, the resistance of R_g must be large compared to R_p because, so far as the a.c. voltage developed in R_p is concerned, R_g is in parallel with R_p and therefore is just as much a part of R_p as though it were connected directly in parallel with it. (The impedance of the plate-supply battery is assumed to be negligible, so that there is no a.c. voltage drop between the lower end of R_p and the common connection between the two tubes.) In practice the maximum usable value of R_g is limited to from 0.5 to about 2 megohms, depending upon the characteristics of the tube with which it is associated. If the value is made too high, stray electrons collecting on the grid may not "leak off" back to the cathode rapidly enough to prevent the accumulation of a negative charge on the grid. This is equivalent to an increase in the negative grid bias, and hence to a shift in the operating point.

The equivalent circuit of the amplifier now includes C_c, R_g, and a shunt capacity, C_s, which represents the input capacity of tube *B* and the plate-cathode capacity of tube *A*, together with such stray capacity as exists in the circuit. The reactance of C_s will depend upon the frequency of the voltage being amplified, and, since C_s is in parallel with R_p and R_g, it also becomes part of the load impedance for the amplifier. At low frequencies — below 1000 cycles or so — the reactance of C_s usually is so high that it has practically no effect on the amplification, but, since the reactance decreases at higher frequencies, it is found that the amplification drops off rapidly when the reactance of C_s becomes comparable to the resistance of R_p and R_g in parallel. To maintain the amplification at high frequencies, it is necessary that R_p be relatively small if C_s is large, or that C_s be small if R_p is large.

Under the best conditions, in practice C_s will be of the order of 15 $\mu\mu$fd. or more, while it is

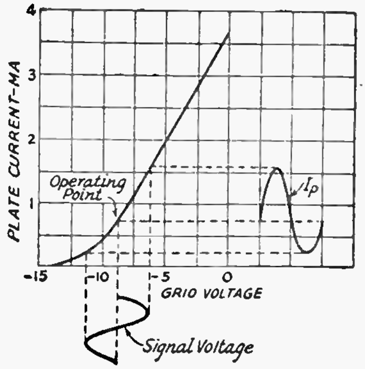


Fig. 310 — Harmonic distortion resulting from choice of an operating point on the curved part of the tube characteristic. The lower half-cycle of plate current does not have the same shape as the grid voltage causing it.

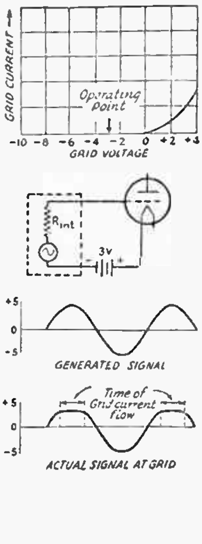


Fig. 309 — Distortion of applied signal because of grid-current flow. With the operating point at 3 volts negative bias, grid current will flow as shown by the curve whenever the applied signal voltage is more than 3 volts positive. If there is appreciable internal resistance, as indicated in the second drawing, there will be a voltage drop in the resistance whenever current is flowing but not during the period when no current flows. The signal will reach the grid unchanged so long as the instantaneous voltage is less than 3 volts positive, but the voltage at the grid will be less than the instantaneous voltage when the latter is above this figure. The shape of the negative half-cycle is unaltered.

possible for it to reach values as high as a few hundred $\mu\mu$fd. The larger values are encountered when tube B is a high-μ triode, as described in a later paragraph. Even with a low value of shunt capacity, the shunt reactance

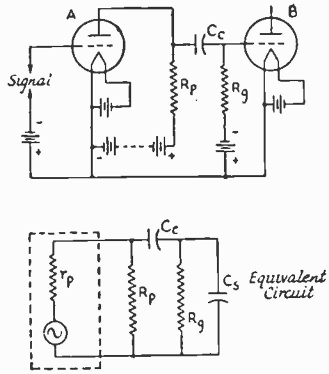


Fig. 311 — Typical resistance-coupled amplifier circuits.

will decrease to a comparatively low value at the upper limit of the audio-frequency range; a shunting capacity of 20 $\mu\mu$fd., for example, represents a reactance of about 0.5 megohm at 15,000 cycles, and hence is of the same order as R_p for the type of tubes with which such a low value of capacity would be associated. In order to secure the same amplification at high as at low frequencies, therefore, it is necessary to sacrifice low-frequency amplification by reducing the value of R_p to the point where the reactance of C_s at the highest frequency of interest is considerably larger than R_p.

At radio frequencies the reactance of C_s becomes so low that the amount of amplification it is possible to realize is negligible compared to that which can be obtained in the audio-frequency range. The resistance-coupled amplifier, therefore, is used principally for audio-frequency work.

Impedance-coupled amplifiers — If either the plate resistor or grid resistor (or both) in the amplifier described in the preceding paragraph is replaced by an inductance, the amplifier is said to be *impedance-coupled*. The inductance or impedance is commonly substituted for the plate load resistor, so that the usual circuit for such an amplifier is as given in Fig. 312.

Considering the operation of the tube from the standpoint of the equivalent circuit of Fig. 308, it is evident that a voltage drop would exist across a reactance of suitable value substituted for the indicated load resistance, R_p, so long as the output of the generator is alternating current. From the physical standpoint, any change in the current flowing through the inductance in Fig. 312 would cause a self-induced e.m.f. having a value proportional to the rate of change of current and to the inductance of the coil. Consequently, if an a.c. signal voltage is applied to the grid of the tube, the resultant variations in plate current cause a corresponding a.c. voltage to appear across the coil terminals. This induced voltage is the useful output voltage of the tube.

The amplitude of the output voltage can be calculated, knowing the μ and plate resistance of the tube and the impedance of the load, in much the same way as in the case of resistance coupling, except that the equation must be modified to take account of the fact that the phase relationship between current and voltage is not the same in an impedance as it is in a resistance. In practice, the plate load inductance is shunted by the tube and stray capacities of the circuit as well as by its own distributed capacity. Since the greatest amplification will be secured when the load impedance is as high as possible, the coil usually is made to have sufficient inductance so that, in combination with these shunting capacities, the circuit as a whole will be parallel-resonant at some frequency near the middle of the audio-frequency range. Under these conditions the load impedance has its highest possible value, and is approximately resistive rather than reactive.

The equation for amplification with resistance coupling shows that, when R_p is several times the plate resistance, r_p, a further increase in R_p results in comparatively little increase in amplification. The load circuit of an impedance-coupled amplifier usually has an impedance value quite high in comparison to the plate resistance of the tube with which it is used, so that the load impedance can vary over a considerable range without much effect on the amplification. This gives the impedance-coupled amplifier an amplification vs. frequency characteristic which is fairly "flat" — that is, the amplification is practically constant with changes in frequency — over a considerable portion of the audio-frequency range. However, the performance of the impedance-coupled amplifier is not as good in this respect as that of a well-designed resistance-coupled amplifier.

If the impedance of the load circuit is high compared to the plate resistance of the tube, which will be the case if the tube is a low-μ triode and normal inductance values (a few hundred henrys) are used in the plate circuit,

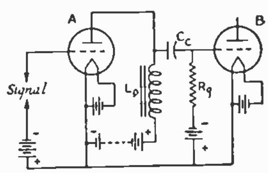


Fig. 312 — Impedance-coupled amplifier.

the amplification in the optimum frequency range will be practically equal to the μ of the tube. At lower frequencies the impedance decreases because of the decreasing reactance of the coil, while at higher frequencies the impedance again decreases because of the decreasing reactance of the shunt capacities. Thus the amplification drops off at both ends of the range, usually more rapidly than with resistance coupling.

The frequency-response characteristic of the impedance-coupled amplifier depends considerably upon the plate resistance of the tube. If impedance coupling is used with tubes of very high plate resistance, the response will be markedly greater at the resonant frequency than at frequencies either higher or lower.

Impedance coupling can be used at radio frequencies, since the inductance can be adjusted to resonate with the shunt capacities at practically any desired frequency.

Transformer-coupled amplifiers — The coupling impedance in Fig. 312 may be replaced by a transformer, connected as shown in Fig. 313. A.c. voltage is developed across the primary of the transformer in the same way as in the case of impedance coupling. The secondary of the transformer serves as a means for transferring the voltage to the grid of the following tube, and if the secondary has more turns than the primary the voltage across the secondary terminals will, in general, be larger than the voltage across the primary terminals.

As in the case of impedance coupling, the effective capacity shunting the primary of an audio-frequency transformer usually causes the primary circuit to be parallel-resonant at some frequency in the middle of the audio-frequency range. At the medium audio frequencies, therefore, the voltage across the primary is practically equal to the applied grid voltage multiplied by the μ of the tube. The voltage across the secondary will be the primary voltage multiplied by the secondary-to-primary turns ratio of the transformer, so that the total voltage amplification is μ times the turns ratio. The amplification at low frequencies depends upon the ratio of the primary reactance to the plate resistance of the tube, as in the case of impedance-coupled amplifiers.

At some high frequency, usually in the range 5000–10,000 cycles with ordinary transformers, the leakage inductance (§ 2-9) of the secondary becomes series resonant with the effective capacity shunting the secondary. At and near this resonant frequency the resonant rise in voltage may increase the amplification considerably, giving rise to a "peak" in the frequency-response curve of the amplifier. At frequencies above this resonance point amplification decreases rapidly, because as the reactance of the shunting capacity decreases it tends to act more and more as a short circuit across the secondary of the transformer. The relative height of the high-frequency peak depends principally upon the effective resistance of the secondary circuit. This effective resistance includes the actual resistance of the secondary coil and the "reflected" (§ 2-9) plate resistance of the tube, this resistance being in parallel with the primary of the transformer. Consequently, the height of the peak is affected by the tube with which the transformer is used. The peak can be reduced by connecting a 0.25 to 1 megohm resistor across the transformer secondary. While this helps to flatten the frequency response curve, it also reduces the amplification at medium and low frequencies.

Transformer coupling is most suitable for triodes of low or medium μ and having medium values of plate resistance. This is because the primary inductance required for good amplification at low frequencies is proportional to the plate resistance of the tube with which the transformer is to be used, and in practice it is difficult to obtain high primary inductance, a large secondary-to-primary turns ratio ("step-up ratio"), and low distributed capacity in the windings all at the same time. Increasing the primary inductance usually means that the turns ratio must be reduced, because the increase in distributed capacity as the coils are made larger tends to bring the resonant peak down to a relatively low frequency unless the secondary inductance is decreased to compensate for the increase in capacity. The step-up ratio seldom is more than 3 to 1 in transformers designed for good frequency response.

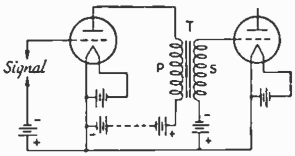


Fig. 313 — Transformer-coupled amplifier.

Transformer coupling can be used at radio frequencies if the transformers are properly designed for the purpose. In such transformers either the primary or secondary (or both) is made resonant at the frequency to be used, so that maximum amplification will be secured.

Phase relations in plate and grid circuits — When the exciting voltage on the grid has its maximum positive instantaneous value, the plate current also is maximum (§ 3-2), so that the voltage drop across the resistance connected in the plate circuit of a resistance-coupled amplifier likewise has its greatest value. The actual instantaneous voltage between plate and cathode is therefore minimum at the same instant, because it is equal to the d.c. supply voltage (which is unvarying) minus the voltage drop across the load resistance. When the signal voltage is at its negative peak the plate current has its least value, with the result that the voltage drop in the load resistance is less than at any other part of the cycle. At this instant, therefore, the voltage between plate and cathode is maximum.

These variations in plate-cathode voltage constitute the a.c. output of the tube, superimposed on the mean or no-signal plate-cathode voltage. Since the alternating plate-cathode voltage is decreasing when the instantaneous grid voltage is increasing (becoming more positive with respect to the cathode), the output voltage is less than the mean value, or negative, when the signal voltage is positive. Likewise, when the signal voltage is negative the output voltage is positive, or greater than the mean value. In other words, the alternating plate voltage is 180 degrees out of phase with the alternating grid voltage. Thus there is a *phase reversal* through the amplifier. The relationships should become clear from the behavior of the signal voltage and E_p in Fig. 307.

The same phase relationship between signal and output voltages holds when the amplifier is impedance- or transformer-coupled, in the frequency region where the load acts like a parallel-resonant circuit. However, if the load is reactive the phase relationship is not exactly 180 degrees but depends upon the kind of reactance present and the relative amounts of reactance and resistance. (This is true also of the resistance-coupled amplifier at low frequencies where the reactance of the coupling condenser affects the amplification, or at high frequencies where the reactance of the shunting capacities becomes important.) Since the reactance varies with the applied signal frequency, the phase relationship between signal voltage and output voltage depends upon the frequency in such cases.

Input capacity and resistance — When an alternating voltage is applied between the grid and cathode of an amplifier tube, an alternating current flows through the small condenser formed by these elements (§ 3-2) just as it would in any other condenser. Similarly, an alternating current also flows in the condenser formed by the grid and plate, since there is an alternating difference of potential between these elements. When the tube is amplifying, the alternating plate voltage and signal voltage are effectively applied in series across the grid-plate condenser, as indicated in Fig. 314. As described in the preceding paragraph, in the resistance-coupled amplifier the two voltages are out of phase with respect to the cathode, but inspection of the circuit shows that they are in phase so far as the grid-plate condenser is concerned. Consequently, the voltage applied to the grid-plate capacity is the sum of the alternating grid and plate voltages, or $E_g + E_p$. Since E_p is equal to $A \times E_g$, where A is the voltage amplification of the tube and circuit, the a.c. voltage between the grid and plate is E_g $(1 + A)$. The current, I, flowing in the grid-plate capacity is E_g $(1 + A)$ divided by the reactance of the grid-plate condenser, and thus is proportional to the grid-plate capacity.

The signal voltage must help in causing this relatively large current to flow, and, since the reactance as viewed from the input circuit

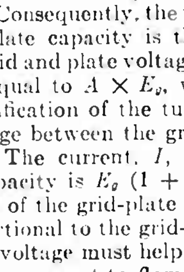


Fig. 314 — The a.c. voltage appearing between the grid and plate of the amplifier is the sum of the signal voltage and the output voltage, as shown by this simplified circuit. Instantaneous polarities are indicated.

is $X_g = E_g/I$, the input reactance becomes smaller as the current becomes larger. That is, the *effective* input capacity of the amplifier is increased when the tube is amplifying. From the above, the increase in input capacity is approximately proportional to the voltage amplification of the circuit and to the grid-plate capacity of the tube. The total input capacity is the sum of the grid-cathode capacity and this additional effective capacity. The total input capacity of an amplifier may reach values ranging from 50 to a few hundred micromicrofarads, if the voltage amplification is high and the grid-plate capacity relatively large. Both usually are true in a high-μ triode.

When the load is reactive the a.c. grid and plate voltages still act in series across the grid-plate condenser, but since they are not exactly 180 degrees out of phase with respect to the cathode they are not exactly in phase with respect to the grid-plate capacity. The lack of exact phase relationship indicates that resistance as well as capacity is introduced into the input circuit. Analysis shows that, when the reactance of the load circuit is capacitive, the resistance component is positive — that is, it represents a loss of power in the input circuit — and that when the load circuit has inductive reactance the resistance component is negative. Negative resistance indicates that power is being supplied to the grid circuit from the plate.

Feed-back — If some of the amplified energy in the plate circuit of an amplifier is coupled back into the grid circuit, the amplifier is said to have *feed-back*. If the voltage fed from the plate circuit to the grid circuit is in such phase that, when it is added to the signal voltage already existing, the sum of the two voltages is larger than the original signal voltage, the feed-back is said to be *positive*. Positive feed-back usually is called *regeneration*. If regeneration exists in a circuit the total amplification is increased because the feed-back increases the amplitude of the signal at the grid and this larger signal is amplified in the same ratio, giving a greater output voltage than would exist if the signal voltage alone were present in the grid circuit. Many types of circuits can be used to secure positive feed-back. A simple one is shown in Fig. 315. The feed-back coil, L, a third winding on the grid-circuit transformer, is connected in series with the primary of the transformer in the plate circuit, so that some of the amplified voltage appears across its terminals. This induces a voltage in the secondary, S, of the grid-circuit transformer which, if the winding directions of the two coils are correct, will increase the value of signal voltage applied to the grid.

Positive feed-back is accompanied by a tendency to give maximum amplification at only one frequency, since the feed-back voltage will tend to be highest at the frequency at which the original amplification is greatest. It therefore increases the selectivity of the amplifier, and hence is used chiefly where high gain and sharpness of resonance both are wanted.

If the phase of the voltage fed back to the grid circuit is such that the sum of the feed-back voltage and the original signal voltage is less than the latter alone, the feed-back is said to be *negative*. Negative feed-back frequently is called *degeneration*. In this case the total amplification is decreased, since the grid signal has been made smaller, and hence the amplified output voltage is smaller for a given original signal than it would be without feed-back.

The amount of voltage fed back will depend upon the actual amplification of the tube and circuit, and if the amplification ratio tends to change, as it may at the extreme high or low frequencies in the audio-frequency range, the feed-back voltage will be reduced when the amplification decreases. For example, suppose that an amplifier has a voltage gain of 20 and that it is delivering an output voltage of 50 volts. Without feed-back, the grid signal voltage required to produce 50 volts output is 50/20 or 2.5 volts. But suppose that 10 per cent of the output voltage (5 volts) is fed back to the grid circuit in opposite phase to the applied grid voltage. Then, since it is still necessary to have a 2.5-volt signal to produce 50 volts output, the applied voltage must be 2.5 + 5 or 7.5 volts. Now suppose that at some other frequency the voltage gain drops to 10. Then for the same 50-volt output a 5-volt signal is required, but since the feed-back voltage is still 5 volts the total required signal is now 10 volts. With feed-back the gain in the first case was 50/7.5 volts or 6.66 and in the second case 50/10 or 5, the gain in the second case being 75 per cent as high as in the first. Without feed-back the gain in the second case was 50 per cent as high as in the first. The effect of feed-back therefore is to make the resultant gain more uniform, despite the tendency of the amplifier itself to discriminate against certain frequencies.

Negative feed-back also tends to decrease harmonic distortion arising in the plate circuit of the amplifier. This distortion is present in the amplified output voltage, but not in the original signal voltage applied to the grid. The voltage fed back to the grid circuit contains the distortion but in opposite phase to the distortion components in the plate circuit, hence the two tend to cancel each other. For similar reasons, the over-all amplification is less dependent upon the value of load impedance used in the plate circuit; in fact, if a large amount of negative feed-back is used in an amplifier it is even possible to substitute tubes of rather widely different characteristics without much effect on the over-all performance.

Both positive and negative feed-back may be applied over several stages of an amplifier, rather than being applied directly from the plate circuit to the grid circuit of a single stage.

Power amplification — In the types of amplifiers previously described, the chief consideration was that of securing as much voltage gain as possible within the permissible limits of harmonic distortion and frequency response characteristic. Such amplifiers are principally used to furnish an amplified signal voltage, which in turn can be supplied to a succeeding amplifier. If the succeeding amplifier is operated in such a way that its grid is never driven positive with respect to its cathode, grid current does not flow, and hence the power requirements are negligibly small. However, if an amplifier is used to actuate some power-consuming device, such as a loudspeaker or a succeeding amplifier in which it is permissible to drive the grid into the positive region, the primary consideration is that of obtaining the maximum power output consistent with the permissible distortion. In such a case the voltage at which the power is secured is of little consequence, since a transformer may be used to change the voltage to any desired value, within reasonable limits. Hence, the voltage gain of a power amplifier is of little importance.

In power-amplifier operation the grid may or may not be driven into the positive region, depending upon the particular application. The present discussion will be confined to the triode amplifier operating without grid current; other types are considered in § 3-4. The principles upon which such a power amplifier operates are practically identical with those already described. The chief differences between a voltage amplifier and a power amplifier lie in the selection of tubes and in the choice of the value of load resistance. As previously described, if voltage gain is the primary consideration the load resistance should be as large as possible in comparison to the plate resistance of the tube. It can be shown that, in any electrical circuit, maximum *power* output is secured when the resistance of the load is made equal to the internal resistance of the source of power. This is true whether the power source is a battery, a generator or a vacuum tube. In the case of the vacuum tube the internal resistance is the plate resistance of the tube, so that for maximum power output the load resistance should be made equal to the plate resistance. However, when the tube is operated with so low a value of load resistance there is considerable harmonic distortion, and *optimum* power output, representing an acceptable compromise between distortion and the power obtainable, is secured when the load resistance is approximately twice the plate resistance.

Fig. 315—An elementary form of feed-back circuit. The feed-back may be either positive or negative, depending upon how the coil *L* is connected in the circuit. This type of circuit illustrates the principle of feed-back, but it is not practical for use in an actual audio-frequency amplifier.

Power-amplifier circuits — The plate or output circuit of a power amplifier almost invariably is transformer-coupled to the power-consuming device or load with which it is associated. This is because the impedance of the desired load seldom is the proper value for obtaining optimum power output from the amplifier. Consequently, the load impedance must be changed to a value suitable for the plate circuit of the amplifier tube. This can be done by the use of transformers, as described in § 2-9.

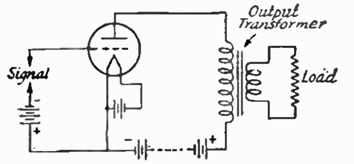


Fig. 316 — An elementary power-amplifier circuit in which the power-consuming load is coupled to the plate circuit through an impedance-matching transformer.

A basic power-amplifier circuit is shown in Fig. 316. So long as the amplifier is operated entirely in the negative-grid region and no grid current flows, any of the previously described types of coupling may be used between the grid of the power amplifier and the preceding amplifier. If there is no preceding amplifier, the method of coupling will depend principally on the characteristics of the source of the signal.

In Fig. 316 the load is represented as a resistance. An actual load may have a reactance as well as a resistance component, but only the resistance will consume power (§ 2-8).

Power amplification ratio — The ratio of a.c. output power to the a.c. power consumed in the grid circuit (*driving power*) is called the *power amplification ratio* or simply *power amplification* of the amplifier. If the amplifier operates without grid current the a.c. power consumed in the grid circuit is negligibly small, so that the power amplification ratio of such an amplifier is extremely large. With other types of operation the power amplification ratio may be relatively small, as described in § 3-4.

Plate efficiency — The ratio of a.c. output power to the d.c. power supplied to the plate circuit is called the *plate efficiency* of the amplifier. It is expressed as a percentage:

$$\% \text{ plate efficiency} = \frac{P_o}{EI} \times 100$$

where P_o is the a.c. output power, E the plate voltage and I the plate current, the latter two being d.c. values.

The plate efficiency of amplifiers designed for minimum distortion and a high power amplification ratio (operation without grid current) is relatively low — of the order of 15 to 30 per cent. For minimum distortion the operation must be confined to the region where the waveshape of the alternating plate current is substantially identical with that of the signal on the grid, and, as previously explained, this requirement can be met only by limiting the plate-current variations (that is, the alternating component of plate current) to the straight portion of the dynamic grid voltage vs. plate current characteristic. Since with a given load resistance the power output is proportional to the square of the alternating component of plate current, it follows that limiting the plate-current variation also limits the power output in comparison to the d.c. plate power input.

Higher plate efficiency can be secured by increasing the alternating component of plate current, but this is accompanied by increased distortion. Special types of amplifiers have been devised to compensate for this distortion, as described in the next section. In some applications, as in r.f. power amplification, the fact that the signal applied to the grid is greatly distorted is of no consequence, so that such amplifiers can have high plate efficiency.

Power sensitivity — The ratio of a.c. power output to alternating grid voltage is called the *power sensitivity* of an amplifier. It provides a convenient measure for comparing power tubes, especially those designed for audio-frequency amplification where the operation is to be without grid current, since it expresses the relationship between power output and the amount of signal voltage required to produce the power.

The term power sensitivity also is used in connection with radio-frequency power amplifiers, in which case it has the same meaning as power amplification ratio. A tube which delivers its rated output power with a relatively small amount of power consumed in the grid circuit is said to have high power sensitivity.

Parallel operation — When it is necessary to obtain more power output than one tube is capable of giving, two or more tubes may be connected in *parallel*. In this case the similar elements in all tubes are connected together. This method is shown in Fig. 317 for a transformer-coupled amplifier. The power output of a parallel stage will be in proportion to the number of tubes used; the exciting voltage required, however, is the same as for one tube.

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

Push-pull operation — An increase in power output can be secured by connecting two tubes in *push-pull*, the grids and plates of the two tubes being connected to opposite ends of the circuit as shown in Fig. 317. A "balanced" circuit, in which the cathode returns are made to the midpoint of the input and output devices, is necessary with push-pull operation. At any instant the ends of the secondary winding of the input transformer, T_1, will be at opposite potentials with respect to the cathode connection, so that 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 stage the voltages and currents of one tube are out of phase with those of the other tube. The plate current of one tube is rising while the plate current of the other is falling, hence the name "push-pull." In push-pull operation the even-harmonic (second, fourth, etc.) distortion is cancelled in the symmetrical plate circuit, so 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 stage is twice that taken by either tube alone.

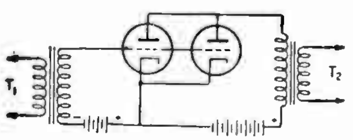
The decibel — The ratio of the power levels at two points in a circuit such as an amplifier can be expressed in terms of a unit called the *decibel*, abbreviated *db.* The number of decibels is 10 times the logarithm of the power ratio, or

$$\text{db.} = 10 \log \frac{P_1}{P_2}$$

The decibel is a particularly useful unit because it is logarithmic, and thus corresponds to the response of the human ear to sounds of varying loudness. One decibel is approximately the power ratio required to make a just noticeable difference in sound intensity. Within wide limits, changing the power by a given ratio produces the same apparent change in loudness regardless of the power level; thus if the power is doubled the increase is 3 db., or three steps of intensity; if it is doubled again the increase is again 3 db., or three further distinguishable steps. Successive amplifications expressed in decibels can be added to obtain the over-all amplification.

A power loss also can be expressed in decibels. A decrease in power is indicated by a minus sign (e.g., − 7 db.), and an increase in power by a plus sign (e.g., + 4 db.). Negative and positive quantities can be added numerically. Zero db. indicates the reference power level, or a power ratio of 1.

Applications of amplification — The major uses of vacuum-tube amplifiers in radio work are for amplifying at audio and radio frequencies (§ 2-7). The audio-frequency amplifier generally is used to amplify without discrimination at all frequencies in a wide range (say from 100 to 3000 cycles for voice communication), and therefore is associated with non-resonant or untuned circuits which offer a uniform load over the desired range. The radio-frequency amplifier, on the other hand, generally is used to amplify selectively at a single radio frequency, or over a small band of frequencies at most, and therefore is associated with resonant circuits tunable to the desired frequency.



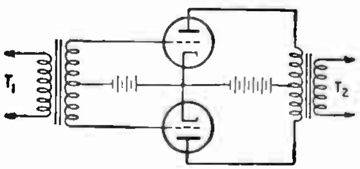


Fig. 317 — Parallel and push-pull a.f. amplifier circuits.

An audio-frequency amplifier may be considered a *broad-band amplifier;* most radio-frequency amplifiers are designed to have relatively narrow bandwidths.

In audio circuits the power tube or output tube in the last stage usually is designed to deliver a considerable amount of audio power, while requiring but negligible power from the input or exciting signal. To get the alternating voltage (*grid swing*) required for the grid of such a tube, voltage amplifiers are used employing high-μ tubes which greatly increase the voltage amplitude of the signal. Voltage amplifiers are used in the radio-frequency stages of receivers as well as in audio amplifiers; power amplifiers are used in the radio-frequency stages of transmitters.

3-4 Classes of Amplifiers

Reason for classification — It is convenient to divide amplifiers into groups according to the work they are intended to perform, as related to the operating conditions necessary to accomplish the purpose. This makes identification easy and obviates the necessity for giving a detailed description of the operation when *specific* operating data are not required.

Class A — An amplifier operated as shown in Fig. 306 or 307, in which the output waveshape is a faithful reproduction of the input waveshape, is known as a *Class-A* amplifier.

As generally used, the grid of a Class-A amplifier never is driven positive with respect to the cathode by the exciting signal, and never is driven so far negative that plate-current cut-off is reached. The plate current is constant both with and without grid excitation. The chief characteristics of the Class-A amplifier are low distortion, relatively low power output for a given size of tube, and a high power-amplification ratio. The plate efficiency is relatively low (§ 3-3).

Class-A power amplifiers find application as output amplifiers in audio systems and as drivers for Class-B power amplifiers. Class-A voltage amplifiers are found in the stages preceding the power stage or stages in such applications, and as r.f. amplifiers in receivers.

Class B — The *Class-B* amplifier is primarily one in which the output current, or alternating component of the plate current, is proportional to the amplitude of the exciting grid voltage. Since power is proportional to the square of the current, the power output of a Class-B amplifier is proportional to the square of the exciting grid voltage.

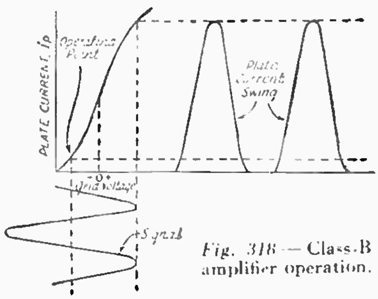


Fig. 318 — Class-B amplifier operation.

In Class-B service the grid bias is set so that the plate current is relatively low without grid excitation; the exciting signal amplitude is made such that the entire linear portion of the characteristic is used. Fig. 318 illustrates operation with the tube biased practically to cut-off. In this condition plate current flows only during the positive half-cycle of excitation. No plate current flows during the negative half-cycle. The shape of the plate current pulse is essentially the same as that of the positive swing of the signal voltage. Since the plate current is driven up toward the saturation point, it is usually necessary for the grid to be driven positive with respect to the cathode during part of the grid swing. Grid current flows, therefore, and the driving source must furnish power to supply the grid losses.

Class-B amplifiers are characterized by medium power output, medium plate efficiency (50 to 60 per cent at maximum signal), and a moderate ratio of power amplification. At radio frequencies they are used as *linear amplifiers* to raise the output power level in radiotelephone transmitters after modulation.

For Class-B audio-frequency amplification two tubes must be used, the second tube working alternately with the first so that both halves of the cycle will be present in the output. A typical method of achieving this is shown in Fig. 319. The signal is fed to a transformer, T_1, whose secondary is divided into two equal parts, with the tube grids connected to the outer terminals and the grid bias fed in at the center. A transformer, T_2, with a similarly divided primary, is connected to the plates of the tubes. When the signal voltage in the upper half of T_1 is positive with respect to the center connection (*center tap*), the upper tube draws plate current while the lower tube is idle; when the lower half of T_1 becomes positive, the lower tube draws plate current while the upper tube is idle. The voltages induced in the primary of T_2 combine in the secondary to produce an amplified reproduction of the signal.

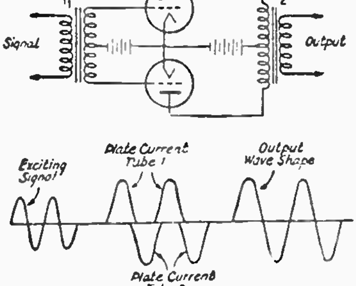


Fig. 319 — Showing how the outputs of the two tubes in push-pull are combined in the Class-B audio amplifier.

Class AB — The similarity between the *Class-AB* amplifier, Fig. 319, and the ordinary push-pull circuit (Fig. 317) will be noted. Actually, the only difference lies in the method of operation. If the bias is adjusted so that the tubes draw a moderate value of plate current with no signal, the amplifier will operate Class A at low signal voltages and more nearly Class B at high signal voltages. This method gives low distortion at moderate signal levels and high plate efficiency at high signal levels, making possible the use of relatively small tubes in audio power amplifiers.

A further distinction can be made between amplifiers which draw grid current and those which do not. The *Class-AB*$_1$ amplifier draws no grid current and thus consumes no power from the driving source. The *Class-AB*$_2$ amplifier draws grid current at higher signal levels, and power must be supplied to its grid circuit.

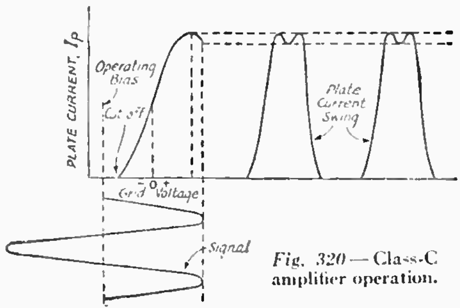


Fig. 320 — Class-C amplifier operation.

Class C — The *Class-C* amplifier is one operated so that the alternating component of the plate current is directly proportional to the plate voltage. The output power is therefore proportional to the square of the plate voltage. Other characteristics inherent to Class-C operation are high plate efficiency, high power output, and relatively low power amplification.

The grid bias is set at a value at least twice that required for plate-current cut-off without excitation. Thus plate current flows during only a fraction of the positive excitation cycle. The exciting signal should be of sufficient amplitude to drive the plate current to the saturation point, as shown in Fig. 320. Since the grid must be driven far into the positive region to cause saturation, considerable numbers of electrons are attracted to the grid at the peak of the cycle, robbing the plate of some that it would normally attract. This causes the droop at the upper bend of the characteristic, and also may cause the plate-current pulse to be indented at the top. The output wave-form is badly distorted, but at radio frequencies the distortion is largely eliminated by the flywheel effect of the tuned output circuit.

3-5 Cathodes; Grid Bias

Types of cathodes — There are two general types of cathodes, known as *directly heated* and *indirectly heated*. In the former the heating current is passed directly through the electron-emitting material, usually a fine wire or filament. In the latter the electrons are emitted from a sleeve or thimble raised to the proper temperature by an electrically-separate heating element as shown in Fig. 321.

Directly-heated or filament-type cathodes may be of pure tungsten, tungsten having a small amount of thorium dissolved in it, or tungsten coated with rare earths (*oxide-coated* type). The latter give the largest amount of electron emission per watt of heating power. Thoriated tungsten filaments are intermediate in electron-emitting efficiency, and are used universally in small and medium-power transmitting tubes. Indirectly-heated cathodes are invariably of the oxide-coated type.

When directly-heated cathodes are operated on alternating current, the cyclic variation of current causes the plate current of the tube to vary at the supply-frequency rate, producing hum in the output. Hum from this source is eliminated in the indirectly heated cathode. This type is also known as the *equi-potential* cathode since all of it is at the same potential, in contrast to the directly heated filament where a voltage drop occurs along the wire.

The source of filament power for a directly heated cathode — battery or transformer — necessarily is directly connected to the tube circuit. With an indirectly heated cathode the source of heating power can be entirely independent of the tube circuit.

The operating temperature of a thoriated tungsten filament is fairly critical, and the specified filament voltage should be maintained within a few per cent. These filaments, as well as oxide-coated cathodes, eventually "lose emission"; that is, the emission efficiency of the cathode decreases until sufficient electron emission for satisfactory tube operation cannot be obtained without raising the cathode temperature to an unsafe value.

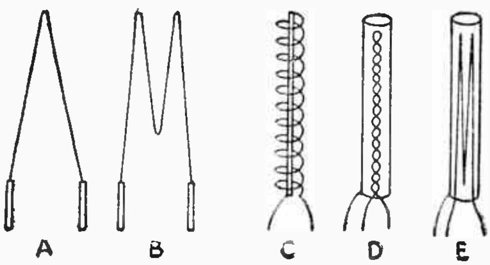


Fig. 321 — Types of cathode construction. Directly heated cathodes or filaments are shown at A, B, and C. The inverted V filament is used in small receiving tubes, the M in both receiving and transmitting tubes. The spiral filament is a transmitting-tube type. The indirectly heated cathodes at D and 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.

Cathode circuits; filament center tap — When a filament-type cathode is heated by a.c., hum can be minimized by making the two ends of the filament have equal and opposite potentials with respect to a center point, usually grounded (§ 2-13), to which the grid and

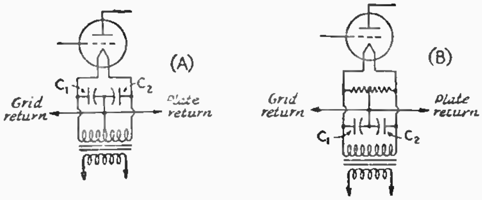


Fig. 322 — Filament transformer center-tap connections.

plate return circuits are connected. The filament transformer winding may be *center-tapped* for this purpose, as shown in Fig. 322-A. With an untapped winding, a center-tapped resistor of 10 to 50 ohms is used, as at B. The by-pass condensers, C_1 and C_2, are used in r.f. circuits to avoid having the r.f. current flow through the transformer or resistor.

The heater supply for tubes with indirectly heated cathodes sometimes is center-tapped for the same purpose; more frequently, however, one side of the heater is grounded.

Methods of obtaining grid bias — Grid bias may be obtained from a source of voltage especially provided for that purpose, such as a battery or other type of d.c. power supply. This is indicated in Fig. 323-A. A second method, utilizing a *cathode resistor*, is shown at B; d.c. plate current flowing through the resistor causes a voltage drop which, with the connections shown, has the right polarity to bias the grid negatively with respect to the cathode. The value of the resistor is determined by the bias required and the plate current which flows at that value of bias, as found from the tube characteristic curves; with the voltage and current known, the resistance can be determined by Ohm's Law (§ 2-6):

$$R_c = \frac{E \times 1000}{I_c}$$

where R_c = cathode bias resistor in ohms
E = desired bias voltage
I_c = total d.c. cathode current in milliamperes.

If the tube is a multi-element type, the screen- and suppressor-grid currents should be added to the plate current to obtain the total cathode current. The control-grid current also should be included if the control grid is driven positive.

The a.c. component of plate current flowing through the cathode resistor will cause an a.c. voltage drop which gives negative feed-back (§ 3-3) into the grid circuit, and thus reduces the amplification. To prevent this, the resistor usually is by-passed (§ 2-13), C_c being the *cathode by-pass condenser*. To be effective, the reactance of the by-pass condenser must be small compared to R_c at the frequency being amplified. This condition generally is satisfied if the reactance is 10 percent or less of the cathode resistance. In audio-frequency amplifiers, the *lowest* frequency at which full amplification must be secured should be used in calculating the required capacity.

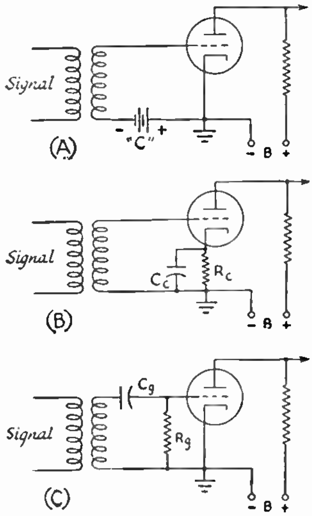


Fig. 323 — The three basic methods of obtaining grid bias. A, fixed bias; B, cathode bias; C, grid-leak bias.

A third biasing method is by use of a *grid leak*, R_g in Fig. 323-C. This requires that the exciting voltage be positive with respect to the cathode during part of the cycle, so that grid current will flow. The flow of grid current through the grid leak causes a voltage drop across the resistor, which gives the grid a negative bias. The time constant (§ 2-6) of the grid leak and grid condenser should be large in comparison to the time of one cycle of the exciting voltage, so that the grid bias will be substantially constant and will not follow the variations in a.c. grid voltage. For grid-leak bias,

$$R_g = \frac{E \times 1000}{I_g}$$

where R_g is the grid-leak resistance in ohms, E the desired bias voltage and I_g the d.c. grid current in ma.

For two tubes operated in push-pull or parallel with a common cathode- or grid-leak resistor, the required resistance becomes one-half that for a single tube. In push-pull Class-A circuits operating at audio frequencies, it is unnecessary to by-pass the cathode resistor. In this case the a.c. component of cathode current in one tube is out of phase with the a.c. component in the other, so that the two cancel each other.

The choice of a biasing method depends upon the type of operation. Fixed bias usually is required where the d.c. plate current of the amplifier varies in operation, as in Class-B audio-frequency amplifiers; if cathode bias is used the bias voltage would vary with the plate current. Since the plate current of a Class-A amplifier is constant with or without signal, such amplifiers almost invariably have cathode bias. Grid-leak bias cannot be used with amplifiers operated so that the grid is always negative with respect to the cathode, since in such a case there is no grid current and hence no voltage drop in the grid leak. Grid-leak bias is chiefly used for r.f. power amplifiers and for certain types of detectors. In power amplifiers, a combination of two or even all three types of bias may be used on one tube.

3-6-A Multi-Grid Tubes

Radio-frequency amplification — As described in § 3-4, the reactances of the grid-to-cathode and plate-to-cathode capacities (together with unavoidable stray capacities) in a vacuum tube become very low at frequencies higher than the audio-frequency range. As a result, ordinary resistance, impedance or transformer coupling cannot be used at radio frequencies because these capacities act as low-reactance by-passes across the input and output circuits. Hence the total impedance in either the plate or the grid circuit is too low for appreciable voltage to be developed.

This situation can be overcome by using resonant circuits as impedances for radio-frequency amplification. As described in § 2-10, the parallel impedance of a resonant circuit can reach quite high values when the Q is high. Values of parallel-resonant impedance suitable for effective amplification are readily obtainable with reasonably well-designed circuits. The tube and stray capacities become part of the tuning capacity and thus are made to serve a useful purpose. However, the circuits have maximum impedance at the resonant frequency only, hence the amplification will decrease at frequencies somewhat removed from resonance. Thus a radio-frequency amplifier must be designed for a specific frequency.

An elementary circuit illustrating the principles of r.f. amplification is shown in Fig. 324. The grid circuit, L_1C_1, and the plate circuit, L_2C_2, must be tuned to the same frequency for maximum amplification. But if the plate circuit is tuned slightly to the high-frequency side of resonance it will show inductive reactance, and as described in § 3-3 energy will be transferred from the plate circuit to the grid circuit under such conditions. If enough energy is transferred the tube will generate a self-sustaining r.f. current, in which case it is said to be *oscillating*. When oscillation commences the circuit ceases to amplify incoming signals, since it is generating a signal of its

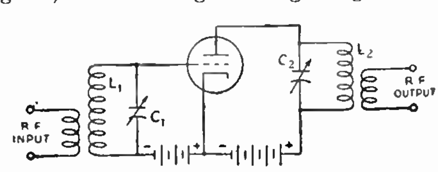


Fig. 324 — Elementary radio-frequency amplifier.

own. Unfortunately, it is almost impossible to prevent such oscillation in a simple triode amplifier such as is shown in Fig. 324.

Special "neutralizing" circuits (§ 4-7) have been devised to prevent oscillation with triode amplifiers, but most of these are more suitable for use in transmitting applications, where the amplifier does not have to be tunable over a wide range of frequencies, than in receivers. However, oscillation can be avoided by using a circuit in which the feed-back is negative rather than positive, as indicated in the next paragraph.

Grounded-grid amplifier — In the circuit of Fig. 325 the grid of the tube is connected to ground and the cathode is connected to the

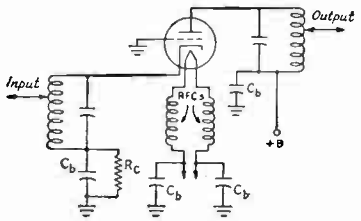


Fig. 325 — Grounded-grid amplifier circuit.

high-potential side of the input resonant circuit, reversing the usual connections. The output circuit is connected in the customary way between plate and ground. Since the alternating component of plate current must flow through the tuned input circuit to return to the cathode there is feed-back from the plate to the grid circuit, but it is negative rather than positive feed-back. Hence this coupling between the two circuits will not cause oscillation.

However, it is still possible for the circuit to oscillate if there is capacity coupling between the plate and cathode. The grounded grid prevents this coupling by acting as a shield between the other two elements (§ 2-11). The circuit is most successful with tubes having very low plate-to-cathode capacity. It is used principally at ultra-high frequencies (where the screen-grid tubes described in the next paragraph become ineffective as amplifiers) with tubes designed especially for the purpose.

The r.f. chokes in the cathode circuit are used to isolate the heater from ground and thus eliminate the effect of the capacity between cathode and heater. This capacity tends to short-circuit the tuned input circuit and thus prevents the amplifier from operating properly.

Screen-grid tubes — The grid-plate capacity can be eliminated, or at least reduced to a negligible value, by inserting a second grid between the control grid and the plate as indicated in Fig. 326. The second grid, called the *screen grid* or *shield grid*, acts as an electrostatic shield (§ 2-11) between the control grid and plate. It is made in the form of a grid or coarse screen rather than as a solid metal sheet, so that electrons can pass through it to the plate; a solid shield would entirely prevent the flow of plate current. The screen grid is connected to the cathode through a by-pass condenser, which has low impedance at the radio frequency being amplified. The electric lines of force from the plate terminate on the screen grid, very little of the field getting through to the control grid; similarly, the field set up by the control grid does not penetrate past the screen grid. Thus there is no common field between the control grid and plate; hence no capacity between these two tube elements.

Since the electric field from the plate does not penetrate into the region occupied by the control grid, which is the region in which most of the space charge is concentrated, the plate is unable to exert an attraction upon the electrons in this region. Consequently, the plate voltage cannot control the flow of plate current as it does in a triode. In order to get electrons to the plate, it is necessary to apply a positive potential (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 velocity, so that most of them shoot between the screen wires into the field from the plate. Those that pass through and are attracted to the plate constitute the plate current of the tube. A certain proportion do strike the screen, however, with the result that some current also flows to the screen grid. The screen current will be low compared to the plate current in a *tetrode*, or four-element tube, however.

Secondary emission — When an electron traveling at appreciable velocity through a tube strikes the plate it dislodges other electrons. These "splash" from the plate into the

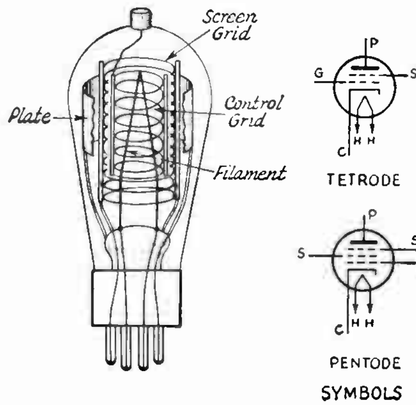


Fig. 326 — Representative arrangement of elements in a screen-grid tube, with front part of plate and screen-grid cut away. The screen grid usually is made longer than either the control grid or plate, so that the shielding will be as effective as possible. In this drawing the control grid connection is made through a cap on the top of the tube, thus eliminating the capacity which would exist between the plate and grid lead wires if both passed through the base. Some modern tubes which have both leads going through the base use special shielding and construction to eliminate capacity. Symbols for pentode and tetrode tubes: H, heater; C, cathode; G, control grid; P, plate; S, screen grid; Sup., suppressor grid.

interelement space, a phenomenon called *secondary emission.* In a triode ordinarily operated with the grid negative with respect to cathode, secondary electrons are repelled back into the plate and 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. The effect is particularly marked when the plate and screen potentials are nearly equal, which may be the case during the part of the a.c. cycle when the instantaneous plate current is large and the plate voltage low (§ 3-3).

Pentode tubes — 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 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.

Although the screen grid in either the tetrode or pentode greatly reduces the influence of the plate upon plate-current flow, it is quite obvious that the control grid still can control the plate current in essentially the same way that it does in a triode, since the control grid is still in the space-charge region. Consequently, the grid-plate transconductance (or mutual conductance) of a tetrode or pentode will be of the same order of value as in a triode of corresponding structure. On the other hand, since the plate voltage has very little effect on the plate-current flow, both the amplification factor and plate resistance of a pentode or tetrode are very high, as is apparent from the definitions of these constants (§ 3-2). 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. In resistance-coupled audio-frequency amplifiers, voltage amplification or gain of 100 to 200 is typical.

A typical set of characteristic curves for a small pentode is shown in Fig. 327. That the plate voltage has little effect on the plate current is indicated by the fact that the curves are practically horizontal once the plate voltage is

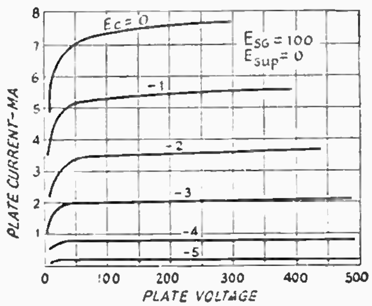


Fig. 327 — Plate voltage vs. plate current curves for a small receiving pentode. Screen-grid voltage, E_{sg}, is 100 volts and suppressor-grid voltage, E_{sup}, is zero.

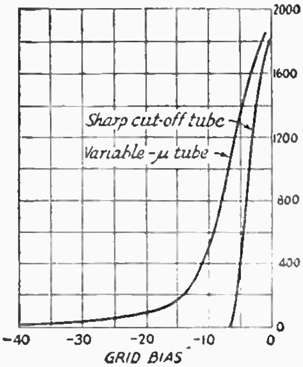


Fig. 328 — Curves showing the relationship between mutual conductance vs. negative grid bias for two small receiving pentodes, one being a sharp cut-off type and the other a variable-μ type.

high enough to prevent the electrons in the space between the screen grid and the plate from being attracted back to the screen. The plate potential at which this occurs is less than the screen potential, because the electrons entering the space have considerable velocity and hence tend to move away from the screen despite the fact that it has a positive charge.

In addition to their applications as radio-frequency amplifiers, pentode or tetrode screen grid tubes also can be constructed for audio-frequency power amplification. In tubes designed for this purpose the shielding effect of the screen grid is not so important; the chief function of the screen is to serve as an accelerator of the electrons, so that large values of plate current can be drawn at relatively low plate voltages. Such tubes have quite high power sensitivity (§ 3-4) compared to triodes of the same power output, because the amplification factor of an equivalent triode has to be made quite low in order to secure the same plate current at the same plate voltage. Because of the low μ, the triode requires a relatively large signal voltage for full output, hence has low power sensitivity. The harmonic distortion is somewhat greater with pentodes and tetrodes than with triodes, however.

Variable-mu and sharp cut-off tubes — Receiving screen-grid tetrodes and pentodes for radio-frequency voltage amplification are made in two types, known as *sharp cut-off* and *variable-μ* or "super-control" types. In the sharp cut-off type the amplification factor is practically constant regardless of grid bias, while in the variable-μ type the amplification factor decreases as the negative bias is increased. The purpose of this design is to permit the tube to handle large signal voltages without distortion in circuits in which grid-bias control is used to vary the mutual conductance, and hence the amplification.

The way in which mutual conductance varies with grid bias in two typical small receiving pentodes, similar except in that one is a sharp cut-off type and the other a variable-μ type, is shown in Fig. 328. Obviously, the variable-μ type can handle a much larger signal voltage without swinging beyond either the point of zero grid bias or of plate-current cut-off (zero mutual conductance), if the bias is properly chosen.

Beam tubes — A "beam"-type tube is a tetrode with grids so constructed as to form the electrons traveling to the plate into concentrated beams, resulting in higher plate efficiency and power sensitivity. Suitable design also overcomes the effects of secondary emission without the necessity for a suppressor grid. Tubes constructed on the beam principle are used in receivers as both r.f. and audio amplifiers, and are built in larger sizes for transmitting circuits.

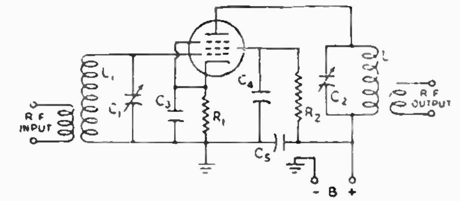


Fig. 329 — Pentode r.f. amplifier circuit. L_1C_1 and L_2C_2 are tuned to the same frequency. R_1 is the cathode resistor, by-passed for r.f. by C_3. R_2 is the screen voltage-dropping resistor, by-passed by C_4. C_5 is the plate by-pass.

ℂ 3-6-B Pentode Amplifiers

R.F. amplification — A fundamental circuit for radio-frequency amplification with a pentode tube is shown in Fig. 329. The grid and plate circuits may be tuned to the same frequency, thus obtaining maximum amplification, without danger of oscillation provided there is no feed-back coupling between the tuned circuits themselves. Practical variations of this circuit and their application to receivers are discussed in § 7-6 and § 7-11.

A.F. amplification — Receiving-type pentodes frequently are used as voltage amplifiers for audio frequencies, using the circuit shown in basic form in Fig. 330. In this application they are capable of much higher voltage gain than can be obtained from triodes, and have the advantage that since there is no coupling from plate to grid there is no increase in input capacity with amplification (§ 3-3). For the latter reason it is possible to obtain high gain, in resistance-coupled amplifiers, at considerably higher frequencies than is possible with a triode.

The discussion of amplification in § 3-3 applies equally to pentodes and triodes, with the exception that the plate resistance of a pentode is so high that the amplification is

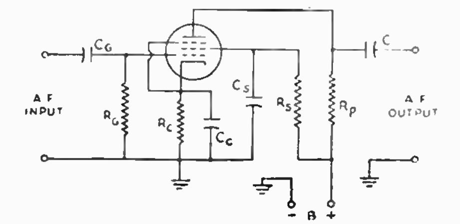


Fig. 330 — Typical pentode audio-frequency amplifier.

usually considered to be proportional to the plate load resistance alone. For maximum voltage gain, R_p should have as high resistance as possible without causing too great a voltage drop. Values range from 0.1 to 0.5 megohm. The value of R_c depends upon R_p, which principally determines the plate current. Values for the screen resistor, R_s, may vary from 0.25 to 2 megohms. A screen by-pass condenser (C_s) of 0.1 μfd. will be adequate in most cases.

Table I in Chapter Fourteen shows suitable values for the more popular types of amplifier tubes. The calculated stage gain and peak undistorted output voltage also are given.

Plate and screen voltage — Since the d.c. plate current flows through any resistance placed in the plate circuit of a tube as a load or coupling medium (§ 3-3), the actual voltage at the plate is less than the supply voltage by the voltage drop across the total resistance.

With transformer coupling this effect is not ordinarily of great importance, because the inductance of the transformer primary provides a high-impedance load at audio frequencies, while the d.c. resistance of the winding causes only a small drop in d.c. plate voltage.

In a resistance-coupled or parallel-fed stage the operating voltage is less than the supply voltage by the drop through the load resistor, R_p. Thus, in Fig. 331-A, $E_p = E_b - (I_p \times R_p)$.

Screen voltage is determined in the same way, using the screen current, I_s, to calculate the drop across the screen dropping resistor, R_s.

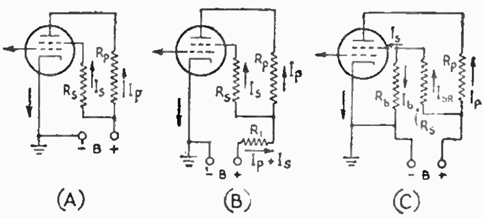


Fig. 331 — Calculation of plate and screen voltages.

In Fig. 331-B both plate and screen current flows through a common filter resistor, so that both currents must be added in calculating the voltage drop across R_f. Thus

$$E_p = E_b - (I_p + I_s)(R_1) - I_pR_p$$
$$E_s = E_b - (I_p + I_s)(R_1) - I_sR_s.$$

In Fig. 331-C, the screen voltage, E_s, is obtained from a tap on a voltage divider consisting of R_s and R_b. Assume a value of bleeder current, I_b (§ 8-4). Then $R_b = E_s/I_b$, where E_s is the rated screen voltage. The total current, I_{sr}, is the sum of I_b and I_s. The voltage across R_s is the difference between the supply voltage and E_s. Hence $R_s = (E_b - E_s)/I_{sr}$. E_p is determined as above.

The resistance-capacity filter (§ 2-11) in Fig. 332, C_fR_f, is a *decoupling circuit* which isolates the stage from the power supply, to eliminate unwanted coupling between stages through the common impedance of the power supply. Although shown in connection with a triode amplifier in the diagram, the same type of filter is used with pentodes.

Wide-band amplifiers — Amplification of audio frequencies, which extend from about 50 to 15,000 cycles, presents no particularly difficult problems so long as the design points discussed in § 3-3 are observed. However, for amplifying signals such as television signals or pulses having a time duration of only a few millionths of a second it is necessary to extend the frequency response of the amplifier well beyond the audio frequency range — and even well into the medium radio-frequency range. At the same time it is frequently necessary to extend the *lower* frequency limit of the amplifier as well. This extension of range is made possible by the use of *compensating* circuits.

Low-frequency compensation — While the amplitude response of a resistance-coupled amplifier usually is satisfactory at low frequencies, the phase angle introduced by the output coupling condenser and the next-stage grid resistor is sufficient to prevent proper reproduction of low-frequency square waves unless very large values are employed. Yet such large values increase the shunt capacity to ground, introduce grid-current difficulties in the following stage, and may even induce relaxation oscillations (motorboating).

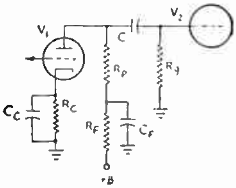


Fig. 332 — Decoupling in a resistance-coupled amplifier.

The effect of the time constant of $C_{G2}R_{G2}$, Fig. 333, may be compensated for by proper design of the amplifier plate circuit. The design equation is $C_F R_P = C_{G2}R_{G2}$ provided the resistance of the decoupling resistor, R_F, is at least 10 times the reactance of C_F at the lowest frequency to be amplified.

High-frequency compensation — It was brought out in § 3-3 that the capacities shunting the plate load resistor are responsible for loss of amplification at the high frequencies in a resistance-coupled amplifier. If the plate load resistor is made low enough in value, the effect of the shunting capacities will be minimized and the upper frequency range will be extended, but at the expense of gain at the lower frequencies. Reducing the plate load resistance to a value low enough to extend the range of uniform amplification to a few megacycles would be impractical with ordinary tubes, since there would be little or no voltage gain, but it is quite practicable with special high-transconductance pentodes such as the 6AC7 and 6AG5. These tubes will give voltage gains of 10 to 15 with plate load resistances as low as a few thousand ohms.

A further extension of high frequency response can be secured by special compensating circuits. The most widely-used method is the *shunt-peaking* circuit, with a resonating (peaking) inductance in parallel with the circuit capacity, as shown in Fig. 333. By resonance effects this raises the impedance to an extent and over a frequency range determined by the Q of the circuit consisting of L, R_P and C_t. Since R_P is relatively large for a resonant circuit, the Q is fairly low and the resonance curve is quite broad. This is desirable for an amplifier intended for wide-band applications. The design values of L and R_P are based on the shunt capacity, C_t, and the maximum required frequency, f_{max}. C_t can be estimated by adding 3 to 5 $\mu\mu$fd. (for socket and wiring) to the sum of the tube input and output capacities.

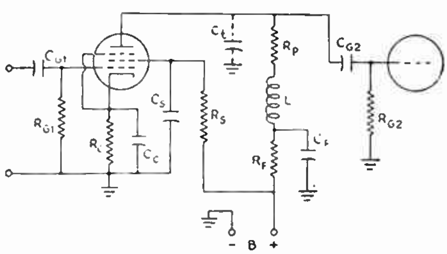


Fig. 333 — Wide-band frequency-compensated amplifier.

The reactance of L is made one-half the reactance of C_t at f_{max}. This is equivalent to making the resonant frequency between L and C_t equal to 1.41 times f_{max}.

Simplified design equations for shunt peaking compensation are as follows:

$$R_P = \frac{1}{2\pi f_{max} C_t}$$

$$L = 0.5 C_t R_p^2$$

Typical values of R_P are from 2000 to 10,000 ohms; of L, from 25 to 100 μh.

Cathode follower — The *cathode-coupler* or *cathode follower* shown in Fig. 334, differs from a conventional amplifier in that output is taken from the cathode circuit rather than from the plate. The circuit is applicable wherever matching to a low value of load impedance (fifty to several hundred ohms) is required and the use of a transformer is impracticable, as in wide-band amplifiers. Because the cathode follower is inherently degenerative, it is particularly useful wherever equalized frequency response and minimum phase shift are important. Power amplification comparable to that of an equivalent plate-coupled stage may be secured, but the voltage gain is always less than unity.

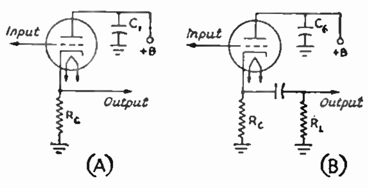


Fig. 334 — Cathode follower or inverted amplifier circuits. A, direct-coupled output; C, resistance-capacity coupling to load. R_c is the usual cathode-bias resistor.

3-6-C Special-Purpose Tubes

Multi-purpose types — A number of combination types of tubes have been constructed to perform multiple functions, particularly in receiver circuits. For the most part these are multi-unit tubes made up of individual tube element structures, combined in a single bulb for compactness and economy. Among the simplest are full-wave rectifiers, combining two diodes in one envelope, and twin triodes, consisting of two triodes in one bulb for Class-B audio amplification. More complex types include duplex-diode triodes, duplex-diode pentodes, converters and mixers (for superheterodyne receivers), combination power tubes and rectifiers, and so on. In many cases the nature can be identified by the name.

Mercury-vapor rectifiers — For a given value of plate current, the power lost in a diode rectifier (§ 3-1) will be lessened if it is possible to decrease the plate-cathode voltage at which the current is obtained. If a small amount of mercury is put in the tube, the mercury will vaporize when the cathode is heated, and, further, will ionize (§ 2-4) when plate voltage is applied. The positive ions neutralize the space charge and reduce the plate-cathode voltage drop to a practically constant value of about 15 volts, regardless of the value of plate current. Since this voltage drop is smaller than can be attained with purely thermionic conduction, there is less power loss in the rectifier. Voltage drop 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, but only if the bias is present before plate voltage is applied. If the bias is lowered to the point where plate current can flow, the mercury vapor will ionize and the grid will lose control of plate current, since the space charge disappears when ionization occurs. It can assume control again only after the plate voltage is reduced below the ionizing potential. The same phenomenon also occurs in triodes filled with other gases which ionize at low pressure. Grid-control rectifiers or thyratrons find considerable application in "electronic switching."

3-7-A Oscillators

Self-oscillation — An amplifier tube can be made to generate a sustained radio-frequency current (§ 3-6-A) because more energy is developed in the plate circuit than is required in the grid circuit. If enough energy is fed back from the plate to the grid, the feed-back process becomes independent of any applied signal voltage. The tube supplies its own grid excitation and continuous oscillations are generated. The actual energy required to overcome the grid losses is, in the end, taken from the d.c. plate supply.

The process of oscillation may also be considered from the standpoint of *negative resistance*. As previously described (§ 3-3), positive feed-back is equivalent to shunting a negative resistance across the input circuit of the tube. When the value of negative resistance becomes lower than the positive resistance of the circuit (if the circuit is parallel resonant the positive resistance will be the resonant impedance of the circuit) the net resistance is negative, indicating that the circuit can be looked upon as a *source* of energy. Such a source is capable of maintaining a constant voltage which can be amplified by the tube. The actual energy, of course, comes from the plate circuit of the tube, so that the two viewpoints are equivalent.

A circuit having the property of generating continuous oscillations is called an *oscillator*. It is not necessary to apply external excitation to such a circuit, since any random variation in current will be amplified to cause oscillation. The frequency of oscillation will be that at which the feed-back voltage has the proper phase and amplitude. Where resonant circuits are associated with oscillators, the oscillation frequency is very nearly that of the tuned circuit.

Excitation and bias — The excitation voltage required depends upon the characteristics of the tube and the losses in the grid circuit. In practically all oscillators the grid is driven positive during part of the cycle, so that power is consumed in the grid circuit (§ 3-2). This power must be supplied from the plate circuit. With insufficient excitation, the tube will not oscillate; with over-excitation, the *grid losses* (power consumed in the grid circuit) will be excessive.

Oscillators customarily are grid-leak biased (§ 3-5). This takes advantage of the grid-current flow and gives better operation, the bias adjusting itself to the excitation voltage.

Tank circuit — The resonant circuit associated with the oscillator is commonly called the *tank circuit*, a name derived from the storage of energy associated with a resonant circuit (§ 2-10). The term is applied to any resonant circuit in transmitting applications, whether in an oscillator or in an amplifier.

Plate efficiency — The *plate efficiency* (§ 3-3) of an oscillator depends upon the load resistance, excitation and other operating factors. Usually it is around 50 per cent. It is not as high as in an amplifier, since the oscillator must supply its own grid losses. These may represent 10 to 20 per cent of the output power.

Power output — The *power output* of an oscillator is the useful a.c. power consumed in any load connected to the oscillator. The load may be coupled as described in § 2-11.

Frequency stability — The frequency stability of an oscillator is its ability to maintain constant frequency. The more important factors which may 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 capacities (§ 3-2). 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 capacity slightly, again causing a shift in the resonant frequency. These effects are relatively slow in operation, and the frequency change caused by them is called *drift.*

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

Plate-voltage variations will cause a corresponding instantaneous shift in frequency; this type of frequency shift is called *dynamic* instability. Dynamic instability can be reduced by using a tuned circuit of high effective Q. Since the tube and load represent a relatively low resistance in parallel with the circuit, this means that a low L/C ratio ("high-C") must be used (§ 2-10) and that the circuit should be lightly loaded. Dynamic stability also can be improved by using a high value of grid leak, which gives high grid bias and raises the effective resistance of the tube as seen by the tank circuit, and by using relatively high plate voltage and low plate current. Drift can be minimized by keeping the d.c. input low for the size of tube, by using coils of large wire to prevent undue temperature rise, and by providing good ventilation to carry off heat rapidly. A low L/C ratio in the tank circuit is desirable, because the interelectrode capacity variations have proportionately less effect on the frequency when shunted by a large condenser.

Mechanical variations, usually caused by vibration, cause changes in inductance and/or capacity which in turn cause the frequency to "wobble" in step with the vibration.

Mechanical instability can be minimized by using well-designed components and by insulating the oscillator from mechinical vibration.

3-7-B Feed-Back Oscillators

Magnetic feed-back — One form of feed-back is by electromagnetic coupling between plate (output) and grid (input) circuits. Two representative circuits of this type are shown in Fig. 335. That at A is called the *tickler* circuit. The amplified current flowing in the "tickler," L_2, induces a voltage in L_1 in the proper phase when both coils are wound in the same direction and connected as shown in the diagram. The feed-back can be adjusted by adjusting the coupling between L_1 and L_2.

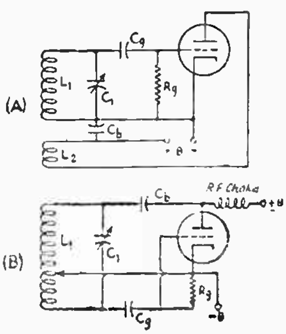


Fig. 335 — Two types of oscillator circuits with magnetic feed-back. A, grid tickler; B, Hartley.

The *Hartley* circuit, B, is similar in principle. There is only one coil, but it is divided so that part of it is in the plate circuit and part in the grid circuit. The magnetic coupling between the two sections provides the feed-back, which can be adjusted by moving the tap on the coil.

Capacity feed-back — The feed-back can also be obtained through capacity coupling, as shown in Fig. 336. In A, the *Colpitts* circuit, the voltage across the resonant circuit is divided, by means of the series condensers, into two parts. The instantaneous voltages at the ends of the circuit are opposite in polarity with respect to the cathode, hence in the right phase to sustain oscillation. The tuned-grid tuned-plate circuit at B utilizes the grid-plate capacity of the tube to provide feed-back coupling. There should be no magnetic coupling between the two tuned-circuit coils. Feed-back can be adjusted by varying the tuning of either the grid or plate circuit. The circuit with the higher Q (§ 2-10) determines the frequency of oscillation. The plate circuit must be tuned to a slightly higher frequency than the grid circuit, so that it will have inductive reactance and hence give positive feed-back (§ 3-3). The amount of detuning is so small it is customary to assume that the circuits are tuned to the same frequency.

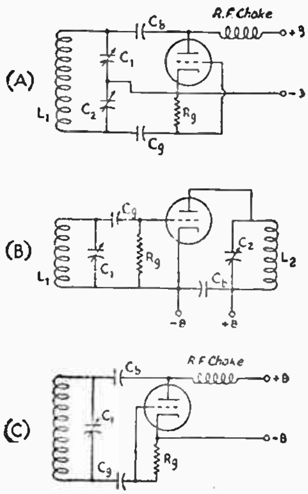


Fig. 336 — Capacity feed-back oscillators. A, Colpitts; B, tuned-plate tuned-grid; C, ultraudion.

The *ultraudion* circuit at C is equivalent to the Colpitts, with the voltage division for oscillation brought about through the grid-to-filament and plate-to-filament capacities of the tube. In this and in the Colpitts circuit, the feed-back can be controlled by varying the ratio of the two capacities. In the ultraudion circuit, this can be done by connecting a small variable condenser between grid and cathode. Feed-back decreases with increasing capacity.

The electron-coupled oscillator — The effects of loading and coupling to the next stage can be greatly reduced by use of the *electron-coupled* circuit, in which a screen-grid tube (§ 3-5) is so connected that its screen grid is used as a plate, in conjunction with the control grid and cathode, in an ordinary triode oscillator circuit. The screen is operated at ground r.f. potential (§ 2-13) to act as a shield between the actual plate and the cathode and control grid; the latter two elements therefore must be above ground potential. The out-

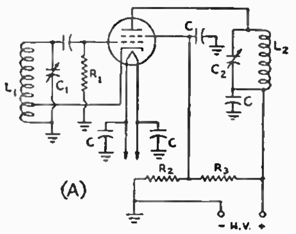


Fig. 337 — Electron-coupled oscillator circuit.

put is taken from the plate circuit. Under these conditions the capacity coupling (§ 2-11) between the plate and other ungrounded tube elements is quite small, hence the output power is secured almost entirely by variations in the plate current caused by the varying potentials on the grid and cathode. Since in a screen-grid tube the plate voltage has a relatively small effect on the plate current, the reaction on the oscillator frequency for different conditions of loading is small.

A Hartley circuit is used in the frequency-determining portion of the oscillator shown in Fig. 337, where L_1C_1 is the oscillator tank circuit. The screen is grounded for r.f. through a by-pass condenser (§ 2-13), but has the usual d.c. potential. The cathode connection is made to a tap on the tank coil to provide feed-back. The resonant plate circuit, L_2C_2, is tuned either to the oscillation frequency or to a harmonic. Untuned output coupling also may be used; the output voltage and power are considerably lower, but better isolation between oscillator and amplifier is secured.

If the oscillator tube is a pentode having an external suppressor connection the suppressor grid should be grounded. This provides additional internal shielding and further isolates the plate from the frequency-determining circuit.

Franklin oscillator — The Franklin oscillator circuit of Fig. 338, popular abroad, has characteristics similar to the e.c.o. A high-gain feed-back amplifier is very loosely coupled to a tank circuit, *LC*, via two condensers, C_1 and C_2, of extremely small capacity. So weak is the coupling that the tube circuit has negligible effect upon the frequency-controlling tank.

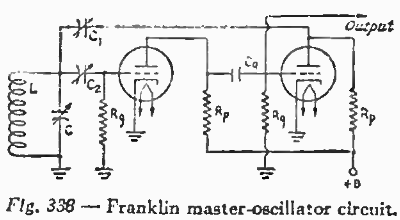


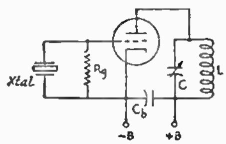
Fig. 338 — Franklin master-oscillator circuit.
C_1, C_2 — Approximately 1 to 2 μμfd. (adjustable).
C — 0.001 μfd. R_g, R_p — 50,000 ohms.

Crystal oscillators — Since a properly cut quartz crystal is equivalent to a high-*Q* tuned circuit (§ 2-10), it may be substituted for a conventional tuned circuit in an oscillator to control the frequency of oscillation. A simple crystal oscillator circuit is shown in Fig. 339. It is similar to the tuned-plate tuned-grid circuit except that a crystal is substituted for the resonant grid circuit. Detailed information on crystal oscillators is given in Chapter Four.

Series and parallel feed — A circuit such as the tickler circuit of Fig. 335-A is said to be *series fed* because the source of plate voltage and the r.f. plate circuit (the tickler coil) are connected in series; hence the d.c. plate current flows through the coil to the plate. A by-pass (§ 2-13) condenser, C_b, is connected across the plate supply to shunt the r.f. current around the power source. Other examples of series plate feed are shown in Figs. 336-B and 337.

In some cases the source of plate power must be connected in parallel with the tuned circuit to provide a direct-current path to the plate. This is illustrated in Fig. 335-B, where it would be impossible to feed the plate current through the coil because there is a direct connection between the coil and cathode. Hence the voltage is applied to the plate through a radio-frequency choke, which prevents the r.f. current

Fig. 339 — Simple crystal oscillator circuit. Many variations of this basic circuit are used in practice.



from flowing to the plate supply and thus short-circuiting the oscillator. The blocking condenser, C_b, provides a low-impedance path for radio-frequency current flow but is an open circuit for direct current (§ 2-13). Other examples of parallel feed are shown in Figs. 336-A and 336-C.

Values for the r.f. chokes, by-pass and blocking condensers shown will be determined by the considerations outlined in § 2-13.

3-7-C Negative Resistance Oscillators

Negative-resistance oscillations — In addition to its ability to simulate negative resistance by feed-back (§ 3-7-A), a vacuum tube can in itself be made to show negative resistance by a number of arrangements of electrode potentials. When a tube so operated is connected to a parallel-resonant circuit, oscillation will be established if the negative resistance is less than the parallel impedance of the resonant circuit. Typical oscillator circuits are shown in Fig. 340.

The circuit of Fig. 340-A is that of the *dynatron* oscillator, which functions because of the secondary emission from the plate occurring in certain types of screen-grid tetrodes. The simplest but also the least stable of the negative-resistance or two-terminal oscillators.

it makes use of the fact that the plate current of a screen-grid tetrode decreases when the plate voltage is increased at certain values of screen voltage, giving a negative plate-resistance characteristic.

In the negative-transconductance or *transitron* circuit shown in Fig. 340-B, negative resistance is produced by virtue of the fact that, if the suppressor grid of a pentode is given negative bias, electrons which normally would pass through to the plate are turned back to the screen, thus increasing the screen current and reversing normal tube action (§ 3-2). The negative resistance produced between the screen and suppressor grids is sufficiently low so that ordinary tuned circuits will oscillate readily up to 15 Mc. or so.

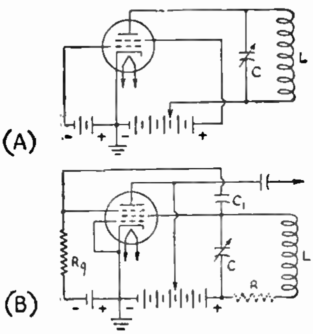


Fig. 340 — Negative-resistance oscillator circuits. A, dynatron; B, transitron.

3-7-D Other Types of Oscillators

Resistance-capacity tuning — It is possible to replace the *LC* resonant circuit in an oscillator by a resistance-capacity combination having an appropriate time constant, in which case $f = 1/2\pi RC$. Moreover, by varying either *R* or *C* the circuit can be tuned over a wide range in the same manner as an *LC* circuit.

The two more common circuits of this type are shown in Fig. 341. The single-stage *RC*-tuned oscillator at A has a three-section phase-shifting network connected between output and input, so arranged that just enough signal is fed back 180° out of phase at the desired frequency to sustain oscillation. By careful feed-back adjustment, excellent sine-wave form with good frequency stability may be obtained.

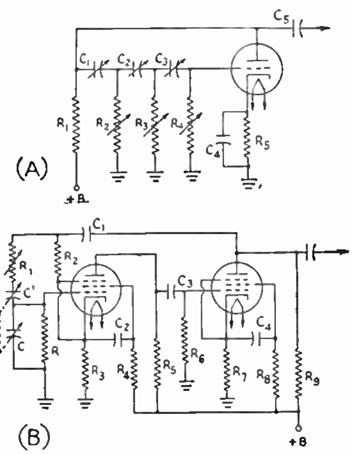


Fig. 341 — Resistance-capacity oscillators. A, phase-shift. B, negative feed-back.

The two-tube *RC*-tuned circuit at B is derived from a two-stage cascade resistance-coupled amplifier with pentode tubes, the second tube constituting the phase-shifting element supplying a regenerative signal to the adjustable *C*, C_1 and R_1 combination at the desired frequency, while at all other frequencies the circuit is degenerative.

Phase-shift oscillators are most useful at audio frequencies, although they can be made to operate up to about 50 kc.

Relaxation oscillators — There is another basic category of oscillators, the *relaxation* type, in which the oscillation frequency is controlled not by a resonant circuit but by the reciprocating change of a current or voltage through the charging or discharging of a condenser when a certain critical value is reached. Relaxation oscillation requires, first, a means for charging a condenser (or other reactive element) at a uniform rate and, second, means for rapidly discharging this condenser once a predetermined voltage has been built up across it. The action is characterized by a period of rapid change or instability followed by a period of relative quiescence or stability during which the stored-up energy transferred or otherwise dissipated in the circuit.

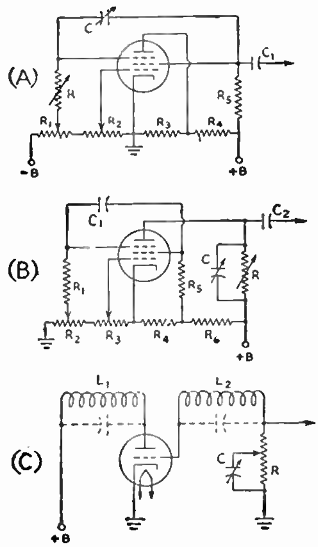


Fig. 342 — Typical relaxation oscillators. A, "dynatron"-type pentode circuit. B, high-frequency pentode circuit. C, squegging oscillator.

Relaxation oscillators have high harmonic content (nonsinusoidal output) and are inherently unstable, permitting ready synchronization with an external controlling voltage.

In the circuit of Fig. 342-A, the operation is based on the reversed screen-current or dynatron characteristic of a pentode tube, the frequency being determined by the rate at which the feed-back condenser, *C*, discharges through the tube. Apart from the frequency-controlling mechanism, this circuit resembles that of the transitron oscillator (Fig. 340-B).

The alternative pentode circuit at B has the frequency-controlling elements, *C* and *R*, in the plate circuit. It is capable of operation at frequencies up to several hundred kilocycles, and affords greater control of wave form.

Operation of the squegging oscillator at C is based on the tendency of any oscillator with excessive feed-back to produce relatively low-frequency intermittent oscillations, controlled by the rate of charge and discharge of L_2, C and R through the tube grid resistance, if the time constant of the combination is large compared to the normal period of oscillation.

The most versatile relaxation oscillator circuit of all, shown in Fig. 343, is known as the *multivibrator.* Two tubes are used with resistance coupling, the output of one tube being fed to the input circuit of the other. The frequency of the resulting oscillation is determined by the time constants (§ 2-6) of the resistance-capacity combinations. The principle of oscillation is that of alternately switching conduction from one tube to the other, with one grid at cut-off and the other at zero bias, so that continuous oscillation is maintained, the second tube being necessary to obtain the proper phase relationship (§ 3-3) for oscillation when the energy is fed back.

Although the multivibrator is a very unstable oscillator, its frequency can be controlled readily by a small signal of steady frequency introduced into the circuit. This phenomenon is called *locking* or *synchronization.* The output waveshape of the multivibrator is highly distorted, hence has high harmonic content (§ 2-7). A useful feature is that the multivibrator can be locked at its fundamental frequency by a frequency corresponding to one of its higher harmonics (the tenth harmonic is frequently used), and thus the circuit can be used as a *frequency divider.*

3-8 Cathode-Ray Tubes

Principles — The cathode-ray tube is a vacuum tube in which the electrons emitted from a hot cathode are first accelerated to give them considerable velocity, then formed into a beam, and finally allowed to strike a special translucent screen which *fluoresces,* or gives off light at the point where the beam strikes. A narrow beam of moving electrons is analogous to a wire carrying current (§ 2-4) and, as in the wire, is accompanied by electrostatic and electromagnetic fields. Hence the beam can be moved laterally, or deflected, by electric or magnetic fields. Such fields exert a force on the beam in much the same way as on charged bodies or on wires carrying current (§ 2-3, 2-5).

Since the cathode-ray beam consists only of moving electrons, its weight and inertia are negligibly small. For this reason, it can be made to follow instantly the variations in periodically changing fields even at radio frequencies.

Electron gun — The electrode arrangement which forms the electrons into a beam is called the *electron gun.* In the simple tube structure shown in Fig. 344, 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 (§ 3-2). Anode No. 1 is operated at a positive potential with respect to the cathode, thus accelerating the electrons which 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, comparable to an optical lens, which makes the electron paths converge to a point at the fluorescent screen in much the same way that a glass lens takes parallel rays of light and brings them to a point focus. Focusing of the electron beam is accomplished by varying the potentials on the anodes, the potential in turn determining the strength of the field. 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.*

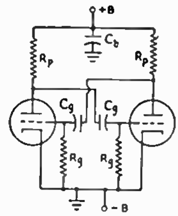


Fig. 343 — The multivibrator, or relaxation oscillator.

Sharpest focus is obtained when the electrons of the beam have high velocity, so that relatively high d.c. potentials are common with cathode-ray tubes. However, the current required is small, so that the power consumption is negligible. A second grid may be placed between the control grid and anode No. 1, for additional acceleration of the electrons.

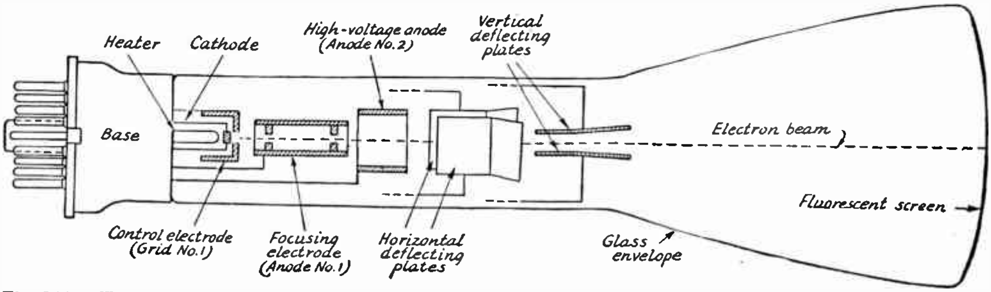


Fig. 344 — Typical construction for a modern cathode-ray tube of the electrostatic-deflection type. The envelope is made of glass, with the fluorescent screen at one end. Leads for the high-voltage anode, the deflection plates, and other electrodes are insulated low-capacity conductors carried inside the envelope to the base.

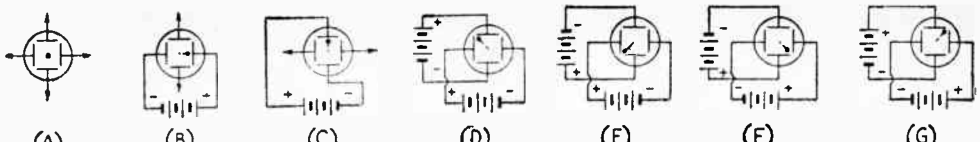


Fig. 345 — Spot diagrams showing the position of the cathode-ray beam on the fluorescent screen for different deflector potentials. A — Both deflectors at zero potential. B — Positive potential on right horizontal deflector. C — Positive potential on upper vertical deflector. D, E, F, G — Equal positive potentials on adjacent plates.

Methods of deflection — When focused, the beam from the gun produces only a small spot on the screen, as described above. However, if after leaving the gun the beam is deflected by either magnetic or electrostatic fields, the spot will move across the screen in accordance with the force exerted on the beam. If the motion is rapid, the path of the spot (*trace*) appears as a continuous line.

Electrostatic deflection, the type generally used in the smaller tubes, is produced by *deflecting plates*. Two sets of plates are placed at right angles to each other, as indicated in Fig. 344. 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 (§ 2-3) of the vertical and horizontal fields with respect to the beam and to each other.

Tubes for magnetic deflection use the same type of electron gun, but have no deflection plates. Instead, the deflecting fields are set up by means of coils corresponding to the plates used in tubes having electrostatic deflection. The coils are external to the tube, as shown in Fig. 346, but are mounted close to the glass envelope in the relative positions occupied by electrostatic deflection plates. Coils A_1 and A_2 are connected so their fields aid and their axes are on the same line through the tube. Coils B_1 and B_2 likewise are connected with fields aiding and are aligned along the same axis through the tube, but perpendicularly to the A_1A_2 axis.

Fluorescent screens — The fluorescent screen materials used have varying characteristics, according to the type of work for which the tube is intended. The spot color is green, white, yellow or blue, depending upon the screen material. The *persistence* of the screen is the time duration of the after-glow which exists when the excitation of the electron beam is removed. Screens are classified as long-, medium- and short-persistence types. Small tubes for oscilloscope use usually have medium-persistence screens of greenish fluorescence.

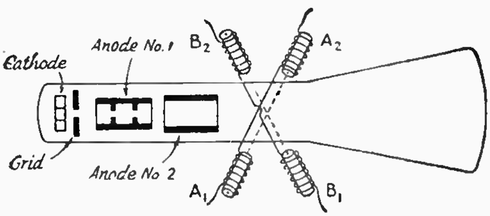


Fig. 346 — A cathode-ray tube with magnetic deflection. The gun is the same as in the electrostatic-deflection tube shown in Fig. 341, but the beam is deflected by magnetic instead of electric fields. Actual deflection coils fit closely to the neck of the tube, so that the field will be as strong as possible for a given coil current.

Tube circuits — A representative cathode-ray tube circuit with electrostatic deflection is shown in Fig. 347. One plate of each pair of deflecting plates is connected to anode No. 2. Since the voltages required normally are rather high, the positive terminal of the supply is usually grounded (§ 2-13) so that the common deflection plates will be at ground potential. This places the cathode and other elements at high potentials above ground, hence these elements must be well insulated. The various electrode voltages are obtained from a voltage divider (§ 2-6) across the high-voltage d.c. supply. R_3 is a variable divider or "potentiometer" for adjusting the negative bias on the control grid and thereby varying the beam current; it is called the *intensity* or *brightness* control. The *focus*, or sharpness of the luminous spot formed on the screen by the beam, is controlled by R_2, which changes the ratio of the anode No. 2 and anode No. 1 voltages. The focusing and intensity controls interlock to some extent, and the sharpest focus is obtained by keeping the beam current low.

Deflecting voltages for the plates are applied to the terminals marked "vertical" and "horizontal." R_4 and R_5 drain off any accumulation of charge on the deflecting plates. Usually some provision is made to place an adjustable d.c. voltage on each set of plates, so that the spot can be "centered" when stray electrostatic or magnetic fields are present; the adjustable d.c. voltage neutralizes the effect of such fields.

The tube is mounted so that one set of plates produces a horizontal line when a varying voltage is applied to it, while the other set of plates produces a vertical line under similar conditions. They are called, respectively, the "horizontal" and "vertical" plates, but which set of actual plates produces which line is simply a matter of how the tube is mounted. It is usually necessary to provide a mounting which can be rotated to some extent, so that the lines will actually be horizontal and vertical.

Power supply — The d.c. voltage required for operation of the tube may vary from 500 volts for the miniature type (1-inch diameter screen) to several thousand volts for the larger tubes. The current, however, is very small, so that the power required likewise is small. Because of the low current drain, a power supply with half-wave rectification (§ 8-3) and a single 0.5- to 2-μfd. filter condenser is satisfactory.

3-9 The Oscilloscope

Description — An oscilloscope is essentially a cathode-ray tube in the basic circuit of Fig. 347, but with provision for supplying a suitable deflection voltage on one set of plates (ordinarily those giving horizontal deflection). The deflection voltage is the *time base* or *sweep*. Oscilloscopes frequently are also equipped with vacuum-tube amplifiers for increasing the amplitude of small a.c. voltages to values suitable for application to the deflecting plates. These amplifiers ordinarily are limited to operation in the audio- or video-frequency range.

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* which is stationary so long as the amplitude and phase relationships of the voltages remain unchanged. Fig. 348 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.

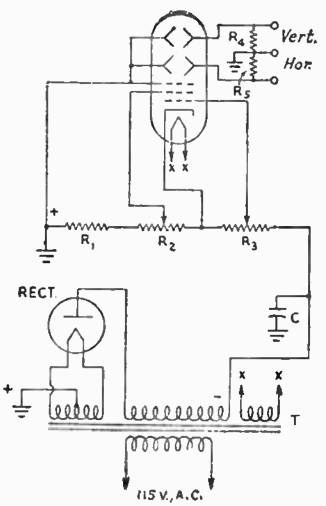


Fig. 347 — Cathode-ray tube circuit. Typical values for a 3-inch (screen-diameter) tube such as the 3AP1/906: R_4, R_5 — 1 to 10 megohms. R_3 — 20,000 ohms. R_2 — 0.2 megohm. R_1 — 0.5 megohm.

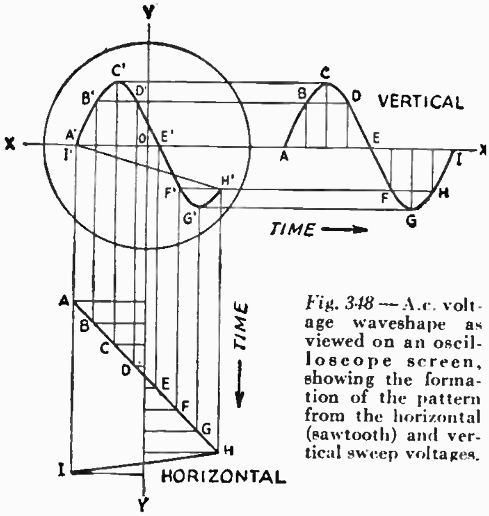


Fig. 348 — A.c. voltage waveshape as viewed on an oscilloscope screen, showing the formation of the pattern from the horizontal (sawtooth) and vertical sweep voltages.

Types of sweeps — A sawtooth sweep-voltage waveshape, such as is shown in Figs. 348 and 350 is called a *linear sweep*, because the deflection in the horizontal direction is directly proportional to time. If the sweep were perfect the "fly-back" time, or time taken for the spot to return from the end (H) to the beginning (I or A) of the horizontal trace, would be zero, so that the line HI would be perpendicular to the axis Y–Y'. Although the fly-back time cannot be made zero in practicable sweep-voltage generators it can be made quite small in comparison to the time of the desired trace AH, at least at most frequencies within the audio range. The fly-back time is somewhat exaggerated in Fig. 345, to show its effect on the pattern. The line $H'I'$ is called the *return trace;* with a linear sweep it is less brilliant than the pattern, because the spot is moving much more rapidly during the fly-back time than during the time of the main trace. If the fly-back time is short enough, the return trace will be invisible.

The linear sweep has the advantage that it shows the shape of the wave applied to the vertical plates in the same way in which it is usually represented graphically (§ 2-7). If the time of one cycle of the a.c. voltage applied to the vertical plates is a fraction of the time taken to sweep horizontally across the screen, several cycles of the vertical or signal voltage will appear in the pattern. The shape of only the last cycle (or the last few cycles, depending upon the number in the pattern and the characteristics of the sweep) to appear will be affected by the fly-back in such a case.

Although the linear sweep generally is most useful, other sweep waveshapes may be desirable for certain purposes. 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. If two sinusoidal voltages of the same frequency are applied simultaneously to both sets of plates, the resulting pattern may be a straight line, an ellipse or a circle, depending upon the

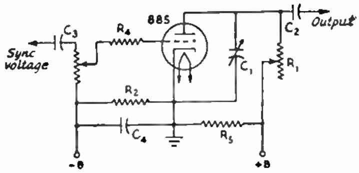


Fig. 349 — A linear-sweep oscillator using a gas triode.
C_1 — 0.001 to 0.25 μfd.
C_2 — 0.5 μfd.
R_1 — 0.3 to 1.5 megohms.
R_2 — 2000 ohms.
R_3 — 0.25 megohm.
C_3 — 0.1 μfd.
C_4 — 25 μfd. 25-volt electrolytic.
R_4 — 25,000 ohms.
R_5 — 0.1 megohm.
The "B" supply should deliver 300 volts. C_1 and R_1 are proportioned to give a suitable sweep frequency; the higher the time constant (§ 2-6), the lower the frequency. R_4 limits grid-current flow during the deionizing period, when positive ions are attracted to the negative grid.

amplitude and phase relationships. If the frequencies are harmonically related (§ 2-7) a stationary pattern will result, but if one frequency is not an exact harmonic of the other the pattern will show continuous motion. This is also the case when a linear sweep circuit is used; the sweep frequency and the frequency under observation must be harmonically related or the pattern will not be stationary.

The sweep generator does not ordinarily function as a self-controlled oscillator but rather as an externally controlled or synchronized oscillator which supplies voltage of the required waveform at the same frequency as the signal under study, or a sub-multiple thereof.

Sweep circuits — A sinusoidal sweep is easiest to obtain, since it is possible to apply a.c. voltage from the power line, either directly or through a suitable transformer, to the horizontal plates. A variable voltage divider or potentiometer may be used to regulate the width of the horizontal trace.

A typical circuit for a linear sweep generator is shown in Fig. 349. The tube is a *gas triode* or grid-control rectifier (§ 3-6-C). The *striking* or breakdown voltage, which is the plate voltage at which the tube ionizes or *fires* and starts conducting, is determined by the grid bias.

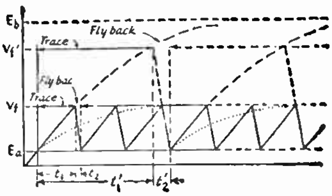


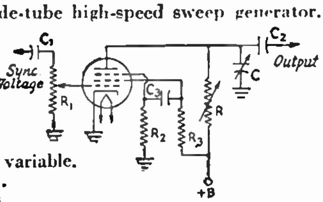
Fig. 350 — Condenser charging curves showing how a sawtooth wave is produced by a gaseous-tube linear sweep oscillator.

When plate voltage, E_b, is applied, the condenser, C_1, acquires a charge through R_1. As shown in Fig. 350, the charging voltage rises relatively slowly, as shown by the solid line, until the breakdown or flashing point, V_f, is reached. Then the condenser discharges rapidly through the comparatively low plate-cathode resistance of the tube. When the voltage drops to a value too low to maintain plate-current flow, E_a, the ionization is extinguished and C_1 once more charges through R_1. If R_1 is large enough, the voltage across C_1 rises linearly with time, t_1, up to the breakdown point. This linear voltage change is used for the sweep, being applied to the cathode-ray tube plates through C_2. The fly-back time, t_2, is the time required for discharge through the tube; to keep this time small, the resistance during discharge must be low.

To obtain a stationary pattern, the "sawtooth" rate is controlled by varying C_1 and R_1 and synchronized by introducing some of the voltage being observed on the vertical plates into the grid circuit of the 884 tube. This voltage "triggers" the tube into operation in synchronism with the signal frequency. Synchronization will occur so long as the signal frequency is nearly the same as, or a multiple of, the sweep frequency, provided the circuit constants and the amplitude of the synchronizing voltage are properly adjusted.

The upper frequency limit of gaseous-tube sweep oscillators is in the vicinity of 50,000 cycles, even with the most careful design, because of the fly-back time limitations imposed by the gaseous content of the tube.

Fig. 351 — Pentode-tube high-speed sweep generator.
C — 0.001 to 0.1 μfd.
C_1, C_2 — 0.1 μfd.
C_3 — 1.0 μfd.
R — 25,000 ohms to 5 megohms.
R_1 — 0.5 megohm variable.
R_2 — 0.1 megohm.
R_3 — 25,000 ohms.



To attain a higher-frequency sweep, a "hard"-tube oscillator such as that shown in Fig. 351 must be used. This circuit may be recognized as being similar to that of the pentode relaxation oscillator of Fig. 342-B. With suitable constants it is capable of an upper frequency limit of 100 to 200 kc. or more. If a tube is used which has a high ratio of plate current to screen current, the screen voltage will rise to a very high value during the plate discharge and thus aid in reducing the fly-back time.

A variety of waveshapes may be obtained from this circuit, ranging from the sawtooth or triangular waves which occur at the plate to the rectangular waveform of the screen-grid voltage. The plate-circuit waveforms are those most often employed for oscilloscope work.

The sweep rate is controlled by R and C, but it is influenced also by the value of R_2. R_3 determines the output waveshape by regulating the ratio of charge to discharge time, thus determining the part of the cycle occupied by the rectangular-shaped screen-voltage wave.

The *blocking-tube oscillator* in Fig. 352 is also capable of high-frequency operation, chiefly because the oscillator portion generates a very short, sharp pulse which charges *C* almost instantaneously. Because of its superiority in this respect, this circuit has received considerable application in television work. Its operation is distinguished from that of the squegging oscillator (Fig. 342-C) in that the intermittent high-frequency oscillations are almost instantly blocked as the bias built up by the grid-leak and condenser, *C* and *R*, goes far beyond cut-off. With suitable constants, the build-up time for this blocking bias can be limited to a single high-frequency cycle, resulting in a very short, abrupt pulse of plate current (I_p). Because of the large time constant of *C* and *R*, the discharge time is very much slower. Until the charge again leaks off through *R*, the circuit is paralyzed. When *C* is discharged, the cycle repeats.

L_1 and L_2 are tightly coupled and designed to be self-resonant at perhaps ten times the maximum sweep frequency.

In the practical form, shown in Fig. 352, the blocking oscillator itself is the left-hand section of the dual triode. The second triode section is used as a discharge tube, the rate of discharge being controlled by the C_2R_1 combination. By giving this combination the proper time constant, the output wave can be made to have almost any desired form. *R* exercises limited control over the frequency range, while the value of R_1 determines the output amplitude.

Vacuum-tube switching circuits — In contrast to time-base circuits which deliver recurrent output impulses, certain applications in oscilloscope and other electronic work call for what are termed *vacuum-tube* or *electronic* switching circuits.

A *keying* circuit is a non-locking electronic switch which closes (or opens) a circuit when a control voltage is applied and returns the circuit to normal when the control voltage is removed. The keying voltage is usually applied as control-grid bias, although screen- and suppressor-grid voltage also are employed.

A *trigger circuit*, also called a *flip-flop* circuit, may also be operated in this manner, but more strictly it is a type of locking or holding electronic switch, wherein a second impulse is required to restore the circuit. After the initiating control pulse the circuit remains closed, despite removal of the control voltage, until a second releasing impulse is received. Circuits in which values of current or voltage change abruptly from one stable condition to another at some critical value of voltage or resistance, and then change back abruptly at a different critical value of the controlling voltage or resistance, are used for this purpose.

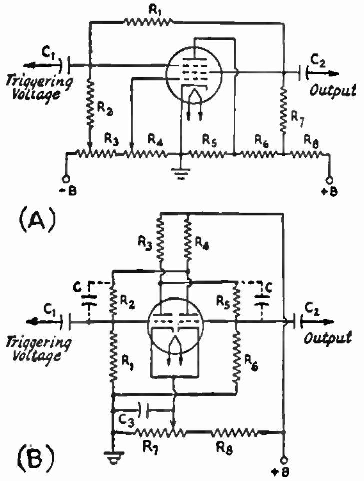


Fig. 353 — Typical vacuum-tube trigger circuits.

Fig. 353-A shows the basic pentode form of trigger circuit. In this circuit d.c. coupling between the screen and suppressor grids causes the suppressor voltage to change with screen voltage. With a high value of resistance in series with the screen, abrupt changes in these currents occur when the supply voltages or the screen-circuit resistance are varied. For example, by proper choice of voltage and circuit constants the plate current corresponding to a given value of screen current may be made zero. Triggering impulses may be introduced in series with any of the electrodes, but the control grid is the most sensitive. The values of the supply voltages are not critical, but the proper relation must be maintained between them.

In the two-tube trigger circuit of Fig. 353-B, a positive impulse applied to the grid of the first tube will increase its plate current. This causes an increased voltage drop across R_3, which in turn makes the bias on the second tube more negative. Consequently the plate current of the second tube decreases, decreasing the voltage drop across R_4. This makes the grid bias on the first tube more positive, causing a further increase in the plate current of this tube and a resultant further decrease in the plate current of the second tube. The process continues until the second tube is cut off, when only the first tube takes current. This condition will continue until a negative pulse is applied to the first grid, or a positive pulse to the second grid, when the action will be reversed. The initial operating point is established by the variable tap on the cathode resistor, R_7.

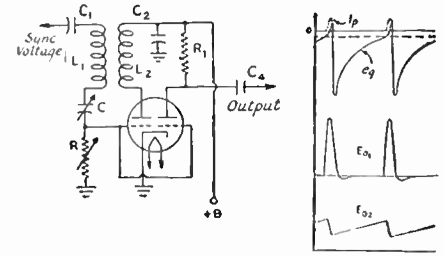


Fig. 352 — Dual-triode blocking-tube oscillator and discharge tube, with characteristic waveforms at the right.

C — 0.001–0.01-μfd. mica. R — 0.25 megohm variable.
C_1, C_3 — 0.005–0.5 μfd. R_1 — 0.1–2 megohm.
C_2 — 0.1 μfd. L_1, L_2 — See text.

3-10 Pulse Technique

In pulse transmission and reception (§ 1-4), specialized means are employed to generate and shape characteristic pulses on the transmitting end and to recreate and interpret these pulses on the receiving end. One is a process of wave-shaping and injection; the other of separation and selection. Certain basic circuit elements are common to both; elementary examples of such circuits will be discussed in this section.

Waveshaping — The primary waveforms employed in pulse transmission, apart from the basic sine wave, are the rectangular wave (from narrow pulse to square wave), trapezoidal wave, triangular wave (from isosceles to right-angle sawtooth), exponential and sawtooth waves.

The nonsinusoidal waveforms obtainable from certain oscillators, particularly those of the relaxation type, approximate the general shapes required. To trim such waves to the ideal form required, auxiliary waveshaping cir-

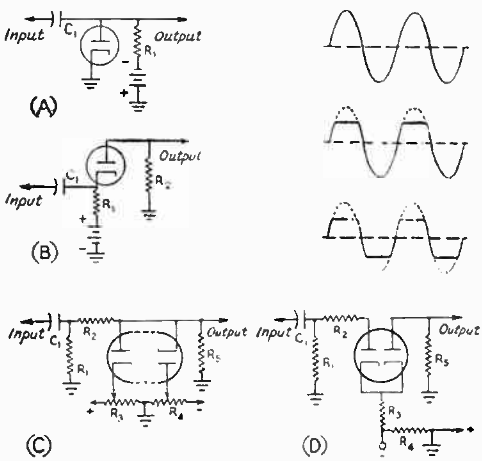


Fig. 351 — Shaping of sine wave to square wave by diode clipping action. The waveforms at the upper right illustrate, progressively, the sinusoidal input wave, the positive peak clipped by the diode parallel limiter (A), and the negative peak clipped by the diode series limiter (B). These are performed jointly in the double-diode parallel limiter (C) and double-diode series limiter (D).

cuits are employed. The basic categories are (1) limiter circuits, which utilize the voltage-limiting action of vacuum tubes, and (2) peaking circuits, which employ RC (or LC) time-constant circuits.

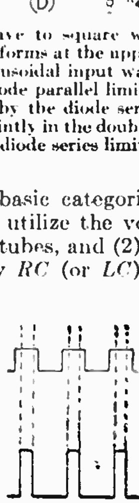


Fig. 354 shows the use of biased-diode limiters in clipping a sine wave to create a square or trapezoidal waveshape by limiting action.

The diode parallel limiter at A does not limit the output until the input voltage attains a value more positive than that of the negative biasing voltage applied in series

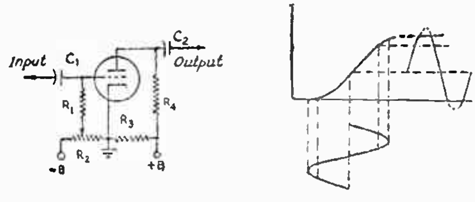


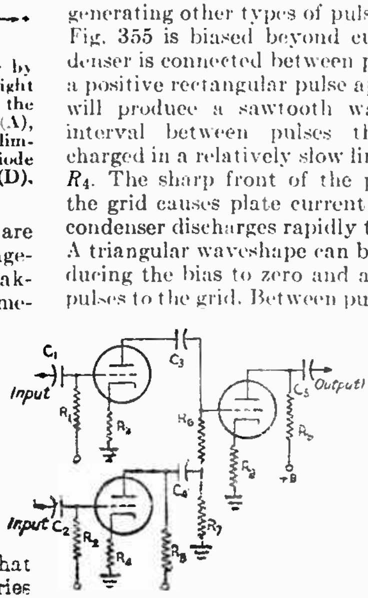
Fig. 355 — Triode limiter action in generating square or trapezoidal wave by clipping peaks of a sinusoidal wave.

with R_1. In the diode series limiter at B, conduction can occur only when the input is more positive than the biasing voltage inserted in series with R_1. Thus there can be no increase in output during the most negative period of the cycle. The series limiter produces a more squarely clipped wave than the parallel type. The operation of either type can be reversed by reversing the diode connections and the polarity of the biasing voltage.

In the double-diode parallel limiter at C, the left-hand diode removes positive peaks while that at the right clips the negative. The degree of limiting is adjusted by varying the fixed bias by means of R_3 and R_4. The double series limiter at D functions in a similar manner but is more critical of adjustment.

Triode limiters may be operated at cut-off or at saturation. In Fig. 355, the tube is biased near the center of its characteristic. When the signal voltage goes negative, at cut-off plate current ceases to flow and the bottom of the sine wave is clipped. On the positive peak the plate current is limited by saturation and the top of the sine curve is squared off. The input signal should be 20 or 30 times the grid bias for the sine wave to be squared off sharply.

Limiter circuits may also be employed for generating other types of pulses. If the tube in Fig. 355 is biased beyond cut-off and a condenser is connected between plate and ground, a positive rectangular pulse applied to the grid will produce a sawtooth wave. During the interval between pulses the condenser is charged in a relatively slow linear rate through R_4. The sharp front of the positive pulse on the grid causes plate current to flow, and the condenser discharges rapidly through the tube. A triangular waveshape can be obtained by reducing the bias to zero and applying negative pulses to the grid. Between pulses plate current



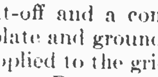


Fig. 356 — Pulse mixer or injector circuit, illustrating how two rectangular pulses of different bases and amplitudes are combined into one complex pulse before transmission.

will flow, but each negative pulse biases the tube beyond cut-off, making it nonconducting. The condenser charges through R_1 for the duration of the pulse, then discharges through R_1. The result is a symmetrical triangular pulse.

Pulse selection — Pulse selectivity is based on the following characteristics: (1) polarity; (2) amplitude; (3) shape; and (4) duration (including both "mark" and "space" intervals).

The diode separator functions much like the diode limiters of Fig. 354, except that the action is reversed. Selection by polarity is based on the unilateral conductivity of the diode rectifier, and requires only that the diode be so con-

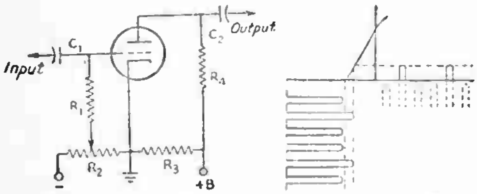
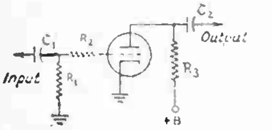


Fig. 357 — Cut-off biased triode amplitude separator.
C_1 — 0.1 µfd. R_1 — 1 megohm. R_3 — 50,000 ohms.
C_2 — 0.5 µfd. R_2 — 2000 ohms. R_4 — 25,000 ohms.

nected as to pass positive or negative pulses, as desired. For amplitude separation the diode is so biased that only pulses having an amplitude exceeding the bias voltage will be passed.

The same resemblance applies in the case of triode amplitude separators. In the cut-off separator of Fig. 357, the grid normally is biased beyond cut-off. When a positive voltage of sufficient amplitude is applied, plate current flows. There will be no response to voltages of lesser amplitude, or to negative pulses.



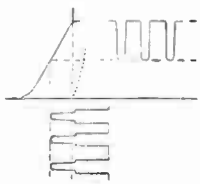


Fig. 358 — Zero-bias or positive-grid limiter-separator.
C_1 — 0.1 µfd. R_1, R_2 — 1 megohm.
C_2 — 0.5 µfd. R_3 — 0.1 megohm

The positive-grid or blocked-grid separator, Fig. 358, operates at saturation and is characterized by a series resistor in the grid circuit. Positive pulses drive the tube into the positive-grid region, where grid-current flow increases bias and limits plate-current to a steady value regardless of signal level. Since this circuit passes only negative pulses, it is selective as to polarity.

Differentiation and integration — If the front of a rectangular wave is applied to an *RC* circuit with series capacity and shunt resistance, as in Fig. 359, the voltage across the load resistor will equal the applied voltage at the instant of application. Then, as the condenser acquires charge the voltage across the resistor will decrease exponentially (§ 2-6). If the time

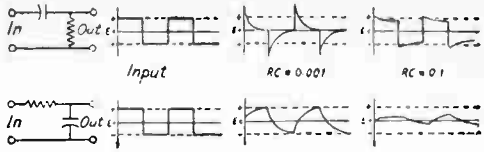


Fig. 359 — With square wave input, the voltage waveshapes across *R* and *C* respectively in an *RC* circuit have the shapes shown. Note the variation in waveshapes for different time constants. (Time constant values given are in terms of fractions of the period of the input wave.)

constant of the circuit is very small, the charging period will be very short. Thus the voltage across the resistor will have the shape of a short pulse, sharply peaked at the front.

Following this initial pulse, no current flows through the resistor because the condenser is charged to the maximum voltage of the applied square wave. Hence the voltage across the resistor is zero so long as the input voltage is unchanging. At the trailing edge of the input wave the process is repeated, except that the resultant pulse has the opposite polarity since the condenser is now discharging.

By altering the steepness of either the ascending or descending slopes of the input wave the amplitude of the output pulse can be controlled. This is the principle upon which pulse selection by waveshape is based, as illustrated in Fig. 360. A steep front produces a sharp pulse having an amplitude equal to the applied voltage, while a sloping front produces a pulse of correspondingly greater length and lesser amplitude. For sharp pulses the time constant must be considerably shorter than one-half cycle of the input wave. With a longer time constant the charging period becomes correspondingly longer, while retaining a logarithmic shape, and approaches the duration and form of the wave. Such a network is called a *differentiating circuit*.

In a circuit with the resistor in series and the condenser in shunt, also shown in Fig. 359, the action is such that with a very short time constant the output wave resembles that of the input except for a slight curvature at the beginning because of the exponential charging characteristic. The amplitude is, however, greatly reduced because of the voltage divider effect of the reactance-resistance combination. Increasing the time constant to a value comparable to the duration of the constant-amplitude portion of the input wave increases the amplitude but accentuates also both the ascending and descending slopes of the wave.

Increasing the time constant to a value very long compared with the base of the input wave, results in what is called an *integrating circuit*. In this circuit discrimination or selection is

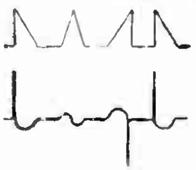


Fig. 360 — Pulse selection based on the discriminating action of a differentiating circuit with inputs of different wavefront shapes. Typical input waves are shown above and the resulting output pulses below.

based on the duration or frequency of the input wave. For example, if a series of short pulses is applied, the energy stored in the condenser by each individual pulse will be small and will be discharged before the next pulse arrives. If, however, a series of pulses with longer bases and shorter intervals is applied, only a portion of the energy from each pulse will be discharged before the next begins charging. Energy is therefore accumulated on the condenser until a predetermined amplitude is established. Thus long-base pulses can be separated from shorter pulses.

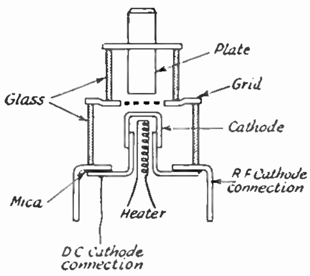


Fig. 361 — Sectional view of the "lighthouse" tube's construction. Close electrode spacing reduces transit time while the disc electrode connections reduce lead inductance.

ℂ 3-11 V.H.F. and U.H.F. Tubes

Negative-grid tubes — At very high frequencies, interelectrode capacities and the inductance of internal leads determine the highest possible frequency to which a vacuum tube can be tuned. The tube usually will not oscillate up to this limit, however, because of dielectric losses, grid emission, and "transit-time" effects. In low-frequency operation, the actual time of flight of electrons between the cathode and the anode is negligible in relation to the duration of the cycle. At 1000 kc., for example, transit time of 0.001 microsecond, which is typical of conventional tubes, is only 1/1000 cycle. But at 100 Mc., this same transit time represents 1/10 of a cycle, and a full cycle at 1000 Mc. These limiting factors establish about 3000 Mc. as the upper frequency limit for negative-grid tubes.

With tubes of ordinary construction, the upper limit of oscillation is about 150 Mc. For higher frequencies, v.h.f. tubes of special construction are used. The "acorn" and "doorknob" types and the special v.h.f. "miniature" tubes, in which the grid-cathode spacing is made as little as 0.005 inch, are capable of operation up to about 700–800 Mc. The normal frequency limit is around 600 Mc., although output may be obtained up to 800 Mc.

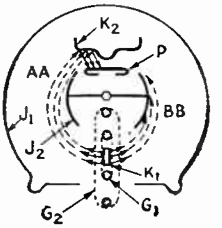


Fig. 362 — Schematic cross-section of the orbital-beam secondary-electron multiplier tube.

Very low interelectrode capacities and lead inductance have been achieved in the newer tubes of modified construction. In multiple-lead types the electrodes are provided with up to three separate leads which, when connected in parallel, have considerably reduced effective inductance. In double-lead types the plate and grid elements are supported by heavy single wires which run entirely through the envelope, providing terminals at either end of the bulb. When a resonant circuit is connected to each pair of leads, the shunting capacity 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. 361, instead of coaxially. The uniform coplanar electrode design and disc-seal terminals permit very low interelectrode capacities.

In the orbital-beam tube, Fig. 362, a small electrode structure is used in combination with a secondary-electron emitter to raise the effective transconductance. Electrons emitted from the cathode, K_1, are accelerated through the control grid, G_1, by a positive grid, G_2, and enter a radial electrostatic field established by the cylindrical electrodes, J_1 and J_2, causing the electrons to move in a circular path and driving them against the secondary-emitter electrode, K_2. About ten secondary electrons are emitted for each primary electron; thus the ultimate electron flow to the plate, P, is considerably greater than the original current emitted. As a result, high over-all transconductance (15,000 at 500 Mc. in an experimental tube) is obtained without increasing transit-time losses or internal capacities.

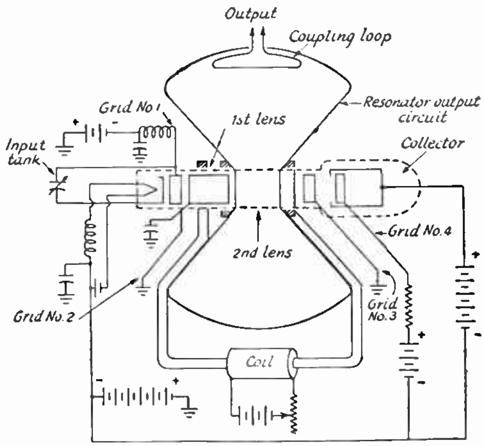


Fig. 363 — Schematic of the inductive output amplifier.

Inductive output amplifier — In the inductive-output tube shown in Fig. 363 a high-velocity electron beam is intensity-modulated by the control grid (grid No. 1). After being accelerated and focused by the combined action of the first and second lenses in the magnetic circuit and the sleeve electrodes (grids

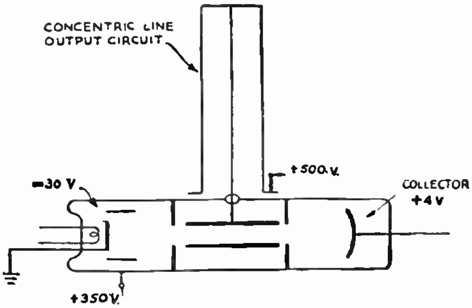


Fig. 364 — Simple form of cylindrical-grid velocity-modulated tube with retarding-field collector and coaxial-line output circuit, used as a superheterodyne high-frequency oscillator or as a superregenerative detector. Similar tubes can also be used as r.f. amplifiers and frequency converters in the 5–50-cm. region.

No. 2 and 3), the beam moves past a small aperture in the "dimpled sphere" cavity resonator. The potential difference across this gap slows down the electrons and thereby causes the resonant cavity to absorb power from the beam. Electrons passing through the structure are decelerated by a suppressor electrode (grid No. 4) before reaching the final anode or collector. The control-grid structure gives sharp cut-off and large transconductance, while the high accelerating potentials and small apertures result in very short transit time and consequently low input conductance. The inductive-output tube is useful for wide-band operation above 500 Mc., giving efficiencies of 25 per cent or better.

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 proportion to the variation in velocity, the output becoming 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 in ordinary tubes.

A simple form of velocity-modulation oscillator tube is shown in Fig. 364. 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 discs at the ends of the larger tube portion. With r.f. voltage applied across these gaps, which are small compared to the distance traveled by the electrons in one half cycle, electrons entering the tube will be accelerated on positive half cycles and decelerated on the negative half cycles. The length of the tube is made equal to the distance covered by the electrons in one-half cycle, so that the electrons will be further accelerated or decelerated as they leave the tube.

As the beam approaches the collector electrode, which is at nearly zero potential, the electrons are retarded, brought to rest, and ultimately turned back by the attraction of the positive sleeve electrode. The collector electrode is, therefore, also termed a *reflector.* The point at which electrons are returned depends on their velocity. Thus the velocity modulation is again translated into current modulation.

Velocity-modulated tubes operate satisfactorily up to 6000 Mc. (5 cm.) and higher, with outputs of 100 watts or more.

The klystron — In the *klystron* velocity-modulated tube, the electrons emitted by the cathode are accelerated or retarded during their passage through an electric field established by two grids in a cavity resonator, or *rhumbatron,* called the "buncher." The high-frequency electric field between the grids is parallel to the electron stream. This field accelerates the electrons at one moment and retards them at another, in accordance with the variations of the r.f. voltage applied.

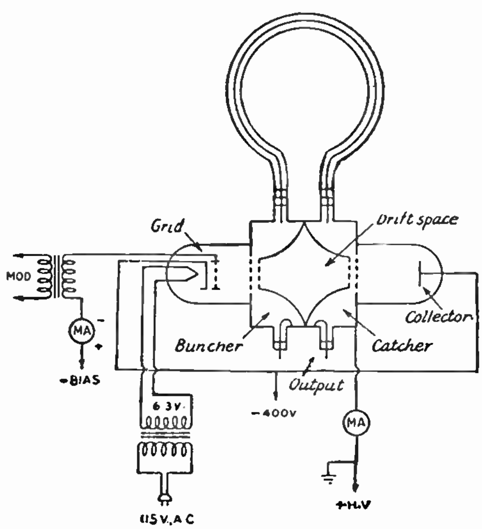


Fig. 365 — Circuit diagram of the klystron oscillator, showing the feed-back loop coupling the frequency-controlling rhumbatrons and the output loop in the catcher.

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, and as the beam passes through two parallel grids, the r.f. current created by the bunching of the electron beam induces an r.f. voltage between the grids. The catcher cavity is made resonant at the frequency of the velocity-modulated electron beam, so that an oscillating field is set up within it by the passage of the electron bunches through the grid aperture.

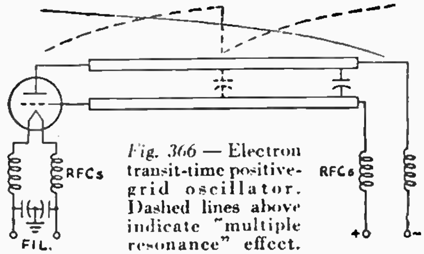


Fig. 366 — Electron transit-time positive-grid oscillator. Dashed lines above indicate "multiple resonance" effect.

If a feed-back loop is provided between the two rhumbatrons, as shown in Fig. 365, 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.

Positive-grid electron oscillators — A triode in which the grid rather than the plate is positive with respect to the cathode will oscillate at frequencies higher than those at which transit-time effects cause the tube to be inoperative as a normal negative-grid oscillator. Oscillators of the positive-grid type are known as "brakefield" or "electron transit-time" oscillators. Successful performance is most readily achieved with tube structures having cylindrical grids and plates.

This type of operation makes use of the transit time of electrons from the cathode to the grid and plate regions. Electrons emitted by the cathode are accelerated toward the positive grid, some striking it and some passing through. Those that pass through are repelled by the negative plate and turn around, passing between the grid wires once more. In the process, the electrons induce a.c. voltages in the grid at a frequency depending upon the transit time. Some electrons may pass back and forth between the grid wires several times, while others may strike the grid after a single round trip. Those which remain free in the tube for several oscillations lose energy, but those which make only one trip gain energy. However, since the former are free for a longer time there is a net transfer of energy which can be used to maintain oscillations.

In this type of oscillator, shown in Fig. 366, the frequency is controlled primarily by the grid voltage and the tube element spacing. The resonant circuit must be tuned to approximately the oscillation frequency for maximum output.

Positive-grid oscillators can be operated at frequencies up to 10,000 Mc. (3 cm.), but the efficiency is usually only 2 or 3 per cent. Since most of the power is dissipated in the grid, the tube is not capable of delivering much power.

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

Magnetron oscillators are operated in two different ways. Electrically the circuits are similar, the difference being in the relation between electron transit time and the frequency of oscillation.

In the negative-resistance or dynatron type of magnetron oscillator, the element dimensions and anode voltage are such that the transit time is short compared with the period of the oscillation frequency. Electrons omitted from the cathode are driven towards both halves of the anode. If the potentials of the two halves are unequal, the effect of the magnetic field is such that the majority of the electrons travel to that half of the anode which 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 (§ 3-7). Negative-resistance magnetron oscillators are useful between 100 and 1000 Mc. Under the best operating conditions efficiencies of 20 to 25 per cent may be obtained. Since the power loss in the tube appears as heat in the anode, where it is readily dissipated, relatively large power-handling capacity can be obtained.

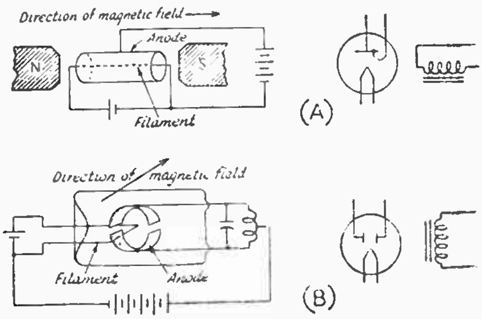


Fig. 367 — Conventional magnetrons, with equivalent schematic symbols at the right. A, simple cylindrical magnetron. B, split-anode negative-resistance magnetron.

In the transit-time magnetron the frequency is determined primarily by its dimensions and by the electric and magnetic field intensities rather than by the tuning of the tank circuits. The efficiency is much better than that of a positive-grid oscillator and good power output can be obtained even on the superhighs.

In a nonoscillating magnetron with a weak magnetic field, electrons traveling from the cathode to the anode move almost radially, their trajectories being bent only slightly by the magnetic field. With increased magnetic field the electrons tend to spiral around the filament, their radial component of velocity being much smaller than the angular component. Under critical conditions of magnetic field strength, a cloud of electrons rotates about the filament. It extends up to the anode but does not actually reach it.

The nature of these electron trajectories is shown in Fig. 368. Cases A, B, and C correspond to the non-oscillating condition. For a

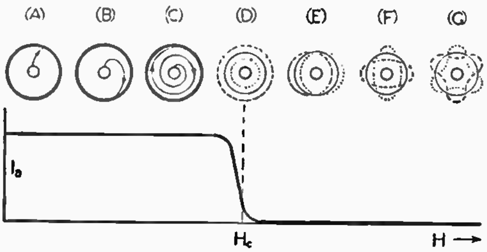


Fig. 368 — Electron trajectories for increasing values of magnetic field strength, *H*. Below is shown the corresponding curve of plate current, *I*. Oscillations commence when *H* reaches a critical value, H_c; progressively higher order modes of oscillation occur beyond this point.

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.

Figs. 368-E, -F and -G depict 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

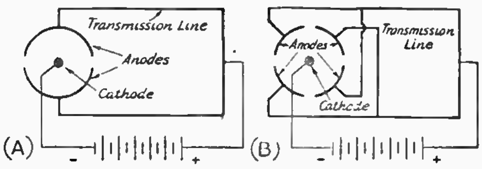
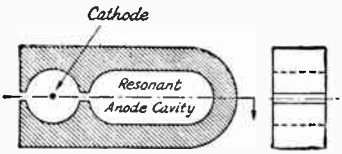


Fig. 369 — U.h.f. magnetron circuits. A, split-anode type. B, four-anode type with opposite electrodes paralleled.

fail to reach the anode. All electrons are therefore deflected back to the cathode, and the anode current is zero. When an alternating voltage is applied between the two halves of the anode, causing the potentials of these halves to vary about their average positive values, the conditions in the tube become analogous to those in a positive-grid oscillator. If the period of the alternating voltage is made equal to the time required for an electron to make one complete rotation in the magnetic field, the a.c. component of the anode voltage reverses direction twice with each electron rotation. Some electrons will lose energy to the electric field, with the result that they are unable to reach the cathode and continue to rotate about it. Meanwhile other electrons gain energy from the field and are returned to the cathode. Since those electrons which lose energy remain in the interelectrode space longer than those which 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. 370. 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. 371.

Fig. 370 — Split-anode magnetron with integral resonant anode cavity for use at u.h.f.



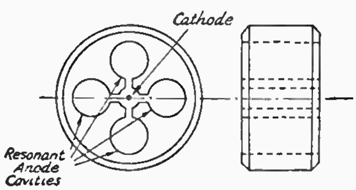


Fig. 371 — Multisegment magnetron with four resonant cavities. This construction is used for extremely high frequencies.

The efficiency of multi-segment 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

Chapter Four

R.-F. Power Generation

4-1 Transmitter Requirements

General requirements — To minimize interference when a large number of stations must work in one frequency band, the power output of a transmitter must be as stable in frequency and as free from spurious radiations as the state of the art permits. The steady r.f. output, called the *carrier* (§ 5-1), must be free from amplitude variations attributable to ripple from the plate power supply (§ 8-4) or other causes, its frequency should be unaffected by variations in supply voltages or inadvertent changes in circuit constants, and there should be no radiation on other than the intended frequency. The degree to which these requirements can be met depends upon the operating frequency.

Design principles — The design of the transmitter depends on the output frequency, the required power output and the type of operation (c.w. telegraphy or 'phone). For c.w. operation at low power on medium-high frequencies (up to 7 Mc. or so), a simple crystal oscillator circuit can meet the requirements satisfactorily. However, the stable power output which can be taken from an oscillator is limited, so that for higher power the oscillator is used simply as a frequency-controlling element, the power being raised to the desired level by means of amplifiers. The requisite frequency stability can be obtained only when the oscillator is operated on relatively low frequencies, so that for output frequencies up to about 60 Mc. it is necessary to increase the oscillator frequency by multiplication (harmonic generation — § 3-3), which usually is done at fairly low power levels and before the final amplification. An amplifier which delivers power on the frequency applied to its grid circuit is known as a *straight amplifier;* one which gives harmonic output is known as a *frequency multiplier*. An amplifier used principally to isolate the frequency-controlling oscillator from the effects of changes in load or other variations in following amplifier stages is called a *buffer amplifier*. A complete transmitter therefore may consist of an oscillator followed by one or more buffer amplifiers, frequency multipliers and straight amplifiers, the number being determined by the output frequency and power in relation to the oscillator frequency and power. The last amplifier is called the *final amplifier*, and the stages up to the last comprise the *exciter*. Transmitters usually are designed to work in a number of frequency bands so that means for changing frequency in harmonic steps usually is provided, generally by means of plug-in inductances.

The general method of designing a transmitter is to decide upon the power output and the highest output frequency required, and also the number of bands in which the transmitter is to operate. The latter usually will determine the oscillator frequency, since it is general practice to set the oscillator on the lowest frequency band to be used. The oscillator frequency seldom is higher than 7 Mc. except in some portable installations where tubes and power must be conserved. A suitable tube (or pair of tubes) should be selected for the final amplifier, and the required grid driving power determined from the tube manufacturer's data. This sets the power required from the preceding stage. From this point the same process is followed back to the oscillator, including frequency multiplication wherever necessary. The selection of a suitable tube complement requires a knowledge of the operating characteristics of the various types of amplifiers and oscillators. These are discussed in the following sections.

Above 100 Mc. and higher frequencies these methods of transmitter design tend to become rather cumbersome, because of the necessity for a large number of frequency multiplier stages. However, in this frequency region less severe stability requirements are imposed because the transmission range is limited (§ 9-5) and the possibility of interference to other communication is reduced. Simple oscillator transmitters, without frequency multiplication or buffer amplifiers, are widely used.

Vacuum tubes — The type of tube used in the transmitter has an important effect on the circuit design. Tubes of high power sensitivity (§ 3-3) such as pentodes and beam tetrodes give larger power amplification ratios per stage than do triodes, hence fewer tubes and stages may be used to obtain the same output power. On the other hand triodes have certain operating advantages, such as simpler power supply circuits and relatively simpler adjustment for modulation (§ 5-3), and in addition are considerably less expensive for the same power output rating. Consequently it is usually more economical to use triodes as output amplifiers, even though an extra low-power amplifier stage may be necessary.

At frequencies in the region of 50 Mc. and above it is necessary to select tubes designed particularly for operation at very-high frequencies, since tubes built primarily for lower frequencies may work poorly or not at all.

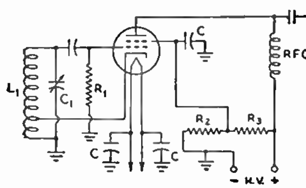
4-2 Self-Controlled Oscillators

Advantages and disadvantages — The chief advantage of a self-controlled oscillator is that the frequency of oscillation is determined by the constants of the tuned circuit, and hence readily can be set to any desired value. However, extreme care in design and adjustment are essential to secure satisfactory frequency stability (§ 3-7). Since frequency stability is generally poorer as the load on the oscillator is increased, the self-controlled oscillator should be used purely to control frequency and not for the purpose of obtaining appreciable power output in transmitters intended for working below 60 Mc.

Oscillator circuits — The inherent stability of all of the oscillator circuits described in § 3-7 is about the same, since stability is more a function of choice of proper circuit values and of adjustment than of the method by which feed-back is obtained. However, some circuits are more convenient to use than others, particularly from the standpoint of feed-back adjustment, mechanical considerations (whether the tuning condenser rotor plates can be grounded or not, etc.), and uniform output over a considerable frequency range. In all simple circuits the power output must be taken from the frequency-determining tank circuit, which means that, aside from the effect of loading on frequency stability, the following amplifier stage can react on the oscillator and cause a change in the frequency.

Factors influencing stability — The causes of frequency instability and the necessary remedial steps have been discussed in § 3-7. These apply to all oscillators. In the case of the electron-coupled oscillator the ratio of plate to screen voltage has marked effect on the stability with changes in supply voltage; the optimum ratio is generally of the order of 3:1, but should be determined experimentally for each case. Since the cathode is above ground potential, means should be taken to reduce the effects of heater-to-cathode capacitance or leakage which, by allowing a small a.c. voltage from the heater supply to develop between cathode and ground, may cause modulation (§ 5-1) at the supply frequency.

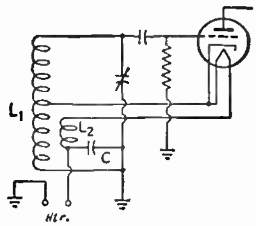
Fig. 401 — Electron-coupled oscillator circuit. R_1 should be 100,000 ohms or more, the grid condenser 100 μμfd. and the other fixed condensers 0.002 to 0.1 μfd.



This effect, which is usually appreciable only at 14 Mc. and higher, may be reduced by by-passing the heater as in Fig. 401 or by operating the heater at the same r.f. potential as the cathode. The latter may be accomplished by the wiring arrangement shown in Fig. 402.

Tank-circuit Q — The most important single factor in determining frequency stability is the Q of the oscillator tank circuit. The effective Q must be as high as possible for best stability. Since oscillation is accompanied by grid-current flow the grid-cathode circuit

Fig. 402 — Method of operating the heater at cathode r.f. potential in an electron-coupled oscillator. L_2 should have the same number of turns as the cathode section of L_1 and should be closely coupled (preferably interwound). Condenser C may be 0.01 to 0.1 μfd.



constitutes a resistance load of appreciable proportions, the effective resistance being low enough to be the determining factor in establishing the effective parallel impedance of the tank circuit. Consequently, if the ends of the tank are connected to plate and grid, as is usual, a high effective Q can be obtained only by decreasing the $L\ C$ ratio and making the inherent resistance in the tank as low as possible. The tank resistance can be decreased by using low-loss insulation and by winding the coil with large wire. With ordinary construction, the optimum tank capacity is of the order of 500 to 1000 μμfd. at a frequency of 3.5 Mc.

The effective circuit Q can be raised by increasing the resistance of the grid circuit and thus decreasing the loading. This can be accomplished through reducing the oscillator grid current, which may be accomplished by using minimum feed-back for stable oscillation, plus a high value of grid-leak resistance.

A high-Q tank circuit can also be obtained with a higher $L\ C$ ratio by "tapping down" the tube connections on the tank (§ 2-10). This is advantageous in that a coil with higher inherent Q can be used; also, the circulating r.f. current in the tank circuit is reduced so that drift from coil heating is decreased. However, under some conditions parasitic oscillations may be set up (§ 4-10).

Plate supply — Since the oscillator frequency will be affected to some extent by changes in plate-supply voltage, it is necessary that the latter be free from ripple (§ 8-4) which would cause frequency variations at the ripple-frequency rate (*frequency modulation*). It is advantageous to use a voltage-stabilized power supply (§ 8-8). Since the oscillator usually is operated at low voltage and current, VR-type gaseous regulator tubes are quite suitable.

Power level — The self-controlled oscillator should be designed purely for frequency control and not to give appreciable power output, hence small tubes of the receiving type may be used. The power input ordinarily is not more than a watt or two, subsequent buffer amplifiers being used to increase the power to the desired level. The use of receiving tubes is advantageous mechanically, since the small elements are less susceptible to vibration and usually are securely braced to the envelope of the tube.

Oscillator adjustment — The adjustment of an oscillator consists principally in observing the design principles outlined in the preceding paragraphs. Frequency stability should be checked with the aid of a stable receiver. An auxiliary crystal oscillator may be used as a standard for checking dynamic stability and drift, the self-controlled oscillator being adjusted to approximately the same frequency so that an audio-frequency beat (§ 2-13) can be obtained. If it is possible to vary the oscillator plate voltage (an adjustable resistor of 50,000 or 100,000 ohms in series with the plate supply lead will give considerable variation), the change in frequency with change in plate voltage may be observed and the operating conditions varied until minimum frequency shift results. The principal factors affecting dynamic stability will be the tank circuit L/C ratio, the grid-leak resistance, and the amount of feed-back. In the electron-coupled circuit the latter may be adjusted by changing the cathode tap on the tank coil; critical adjustment is required for optimum stability.

Drift may be checked by allowing the oscillator to operate continuously from a cold start, the frequency change being observed at regular intervals. Drift may be minimized by using less than the rated power input to the plate of the tube, by construction which prevents tube heat from reaching the tank circuit elements, and by use of large wire in the tank coil to reduce temperature rise from internal heating.

In the electron-coupled oscillator having a tuned plate circuit (Fig. 334), resonance at the fundamental and harmonic frequencies of the oscillator portion of the tube will be indicated by a decrease in plate current as the plate tank condenser is varied. This "dip" is less marked at the fundamental than on harmonics.

ℂ 4-3 Crystal Control

Characteristics — Piezoelectric crystals (§ 2-12-D) are widely used for controlling the frequency of transmitting oscillators, because the extremely high Q of the crystal and the necessarily loose coupling between it and the

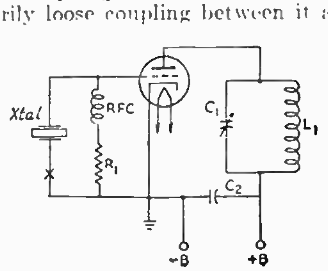


Fig. 403 — Triode crystal oscillator. The tank condenser, C_1, may be a 100-μμfd. variable, with L_1 proportioned so that the tank will tune to the crystal frequency. C_2 should be 0.001 μfd. or larger. The grid leak, R_1, will vary with the type of tube; high-μ tubes take values of 2500 to 10,000 ohms, while medium and low-μ types take values of 10,000 to 25,000 ohms. A small flashlight bulb or r f. milliammeter (§ 4-3) may be inserted at X.

oscillator tube make the frequency stability of a crystal-controlled oscillator very high.

The ability to adhere closely to a known frequency is the outstanding characteristic of a crystal oscillator. This also is a disadvantage, in that a different crystal is required for each frequency on which the transmitter is to operate.

Power limitations — The temperature of a crystal depends not only on the temperature of its surroundings but also on the power it must dissipate while oscillating, since power dissipation causes heating (§ 2-6, 2-8). Consequently, the crystal temperature in operation may be considerably above that of the surrounding air. To minimize heating and frequency drift (§ 3-7), the power dissipated must be kept to a minimum.

If the crystal is made to oscillate too strongly, as when it is used in an oscillator circuit with high plate voltage and excessive feed-back, the amplitude of the mechanical vibration will become great enough to crack or puncture the quartz. An indication of the vibration amplitude (and power dissipated) can be obtained by connecting an r.f. current-indicating device of suitable range in series with the crystal. Safe r.f. crystal currents range from 50 to 200 milliamperes, depending upon the type of crystal cut. A flashlight bulb or dial light of equivalent current rating makes a good current indicator. By choosing a bulb of lower rating than the current specified by the manufacturer as safe for the particular type of crystal used, the bulb will serve as a fuse, burning out before a current dangerous to the crystal is reached. The 60-ma. and 100-ma. bulbs may be used for this purpose.

Crystal mountings — To make use of the crystal, it must be mounted between two metal electrodes. There are two types of mountings, one having a small air-gap between the top plate and the crystal and the other maintaining both plates in contact with the crystal. It is essential that the surfaces of the metal plates in contact with the crystal be perfectly flat. In the air-gap type of holder, the frequency of oscillation depends to some extent upon the size of the gap. By using a holder having a top plate with closely adjustable spacing, a controllable frequency variation can be obtained. A suitable 3.5-Mc. crystal will oscillate without great variation in power output over a range of about 5 kc. X- and Y-cut crystals are not generally suitable for this type of operation; they have a tendency to "jump" in frequency with different air gaps.

A holder having a heavy metal bottom plate with a large surface exposed to the air is advantageous in that it radiates quickly the heat generated in the crystal, thereby reducing temperature effects. Different plate sizes, pressures, etc., will cause slight changes in frequency, so that if a crystal is being ground to an exact frequency it should be tested in the same holder and in the same oscillator circuit with which it will be used in the transmitter.

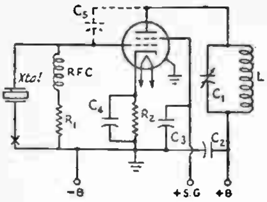


Fig. 404 — Tetrode or pentode crystal oscillator. Typical values: C_1, 100 $\mu\mu$fd., with L wound to suit frequency; C_2, C_3, 0.001 μfd. or larger; C_4, 0.01 μfd.; R_1, 10,000 to 50,000 ohms (value determined by trial); R_2, 250 to 400 ohms.

ℂ 4-4 Crystal Oscillators

Triode oscillators — The triode crystal oscillator circuit (§ 3-7) is shown in Fig. 403. The limit of plate voltage that can be used without endangering the crystal is about 250 volts. With the r.f. crystal current limited to a safe value of about 100 ma., the power output obtainable is about 5 watts. The oscillation frequency is dependent to some extent on the plate tank tuning, because of the change in input capacity with changes in effective amplification (§ 3-3).

Tetrode and pentode oscillators — Since the power output of a crystal oscillator is limited by the permissible r.f. crystal current (§ 4-3), it is advantageous to use an oscillator tube of high power sensitivity (§ 3-3) such as a pentode or beam tetrode (§ 3-5). Thus for a given crystal voltage or current more power output may be obtained than with the triode oscillator, or for a given output the crystal voltage will be lower, thereby reducing crystal heating. In addition, tank-circuit tuning and loading react less on the crystal frequency because of the lower grid-plate capacity (§ 3-3).

Fig. 404 shows a typical pentode or tetrode oscillator circuit. Pentode and tetrode tubes originally designed for audio power work are excellent crystal-oscillator tubes. The screen voltage is generally of the order of half the plate voltage for optimum operation. Small tubes rated at 250 volts for audio work may be operated with 300 volts on the plate and 100–125 on the screen as crystal oscillators. The screen is at ground potential for r.f. and has no part in the operation of the circuit other than to set the operating characteristics of the tube. The larger beam tubes may be operated at 400 to 500 volts on the plate and 250 on the screen for maximum output.

Pentode oscillators operating at 250 to 300 volts will give 4 or 5 watts output under normal conditions. Beam-type tubes such as the 6L6 and 807 will give 15 watts or more at maximum plate voltage.

The grid-plate capacity may be too low to give sufficient feed-back, particularly at the lower frequencies, in which case a feed-back condenser, C_5, may be required. Its capacity should be the lowest value which will give stable oscillation; 1 or 2 $\mu\mu$fd. is generally sufficient. R_2 and C_4 may be omitted, connecting the cathode directly to ground, if plate voltage is limited to 250 volts. C_5 (if needed) may be formed by two metal plates ½-inch square spaced ¼ inch. If the tube has a suppressor grid, it should be grounded. X indicates where a flashlight bulb may be inserted (§ 4-3).

Circuit constants — Typical values for grid-leak resistances and by-pass condensers are given in Figs. 403 and 404. Since the crystal is the frequency-determining element, the Q of the plate tank circuit has a relatively minor effect on the oscillator frequency. A Q of 12 (§ 4-8) is satisfactory for average conditions, but some departure from this figure will not greatly affect the performance of the oscillator.

Adjustment of crystal oscillators — The tuning characteristics and procedure to be followed in tuning are essentially the same for triode, tetrode or pentode crystal oscillators. Using a plate milliammeter as an indicator of oscillation (a 0–100 ma. d.c. meter will have ample range for all low-power oscillators), the plate current will be found to be steady when the circuit is in the non-oscillating state, but will dip when the plate condenser is tuned through resonance at the crystal frequency. Fig. 405 is typical of the behavior of plate current as the tank condenser capacity is varied. An r.f. indicator, such as a small neon bulb touched to the plate end of the tank coil, will show a maximum indication at point A. However, when the oscillator is delivering power to a load it is best to operate in the region B-C since the oscillator will be more stable and there is less likelihood that a slight change in loading will throw the circuit out of oscillation, which is likely to happen when operation is too near the critical point, A. The crystal current also is lower in the B-C region.

When power is taken from the oscillator the dip in plate current is less pronounced, as indicated by the dotted curve. The greater the power output, the smaller the dip in plate current. If the load is made too great, oscillations will not start. Loading is adjusted by varying the coupling to the load circuit (§ 2-11).

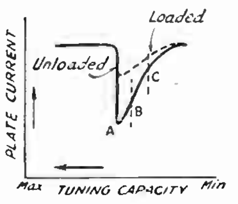


Fig. 405 — Curves showing d.c. plate current vs. plate-circuit tuning in a crystal oscillator, both with and without load. These curves apply equally to the triode, tetrode or pentode crystal oscillator.

The greater the loading, the smaller the voltage fed back to the grid circuit for excitation purposes. This means that the r.f. voltage across the crystal also will be reduced under load, hence there is less crystal heating when the oscillator is delivering power than when it is unloaded.

Failure of a crystal circuit to oscillate may be caused by any of the following:

1) Dirty, chipped or fractured crystal.
2) Imperfect or unclean holder surfaces.
3) Too tight coupling to load.
4) Plate tank circuit not tuning correctly.
5) Insufficient feed-back capacity.

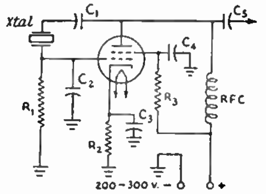


Fig. 406 — Pierce oscillator circuit. R_1 is 25,000 to 50,000 ohms. R_2 is 1000 ohms; R_3, 75,000 ohms for a 6F6; C_1, 0.001 to 0.01 μfd.; C_3 and C_4, 0.01 μfd. For values of C_2 and C_5, see text.

Pierce oscillator — This circuit, Fig. 406, is equivalent to the ultraudion circuit (§3-7), with the crystal replacing the tuned circuit. Although the output is small, it has the advantage that no tuning controls are required. The circuit requires capacitive coupling to a following stage. The amount of feed-back is determined by the condenser, C_2; its capacity must be determined by experiment, usual values being between 50 and 150 μμfd. To sustain oscillation, the net reactance (§ 2-8) of the plate-cathode circuit must be capacitive; this condition is met so long as the inductance of the r.f. choke, together with the inductance of any coils associated with the input circuit of the following stage and the tube and stray capacities, forms a circuit tuned to a lower frequency than that of the crystal.

Tubes such as the triode 6C5 and pentode 6F6 are suitable for use in this circuit. (When a triode is used the screen-voltage dropping resistor, R_3, and by-pass condenser, C_4, in Fig. 406 should, of course, be omitted.) The applied plate voltage should not exceed 300, to prevent crystal fracture. The capacity of the output-coupling condenser, C_5, should be adjusted by experiment so that the oscillator is not overloaded; usually 100 μμfd. is a satisfactory value.

ℂ 4-5 Harmonic-Generating Crystal Oscillators

Tri-tet oscillator — The Tri-tet oscillator circuit is shown in Fig. 407. In this circuit the screen grid is operated at ground potential and the cathode at an r.f. potential above ground. The screen-grid acts as the anode of a triode crystal oscillator, while the plate or output circuit is tuned to the oscillator frequency or, for harmonic output, to a multiple of it.

Besides giving harmonic output, the Tri-tet circuit has the "buffering" feature of electron-coupling between crystal and output circuits (§ 4-2). This makes the crystal frequency less susceptible to changes in loading or tuning, and hence improves the stability.

If the output circuit is to be tuned to the same frequency as the crystal, a tube having low grid-plate capacity (§ 3-2, 3-5) must be used. Otherwise there may be excessive feed-back with consequent danger of fracturing the crystal. The cathode tank circuit, L_1 C_1, is not tuned to the frequency of the crystal, but to a considerably higher frequency. Recommended values for L_1 are given under the diagram. C_1 should be set to as near minimum capacity as is consistent with good output. This reduces the crystal voltage.

With pentode-type tubes having separate suppressor connections, the suppressor may be either connected directly to ground or operated at about 50 volts positive. The latter method will give somewhat higher output.

With transmitting pentodes or beam tubes operated at 500 volts on the plate an output of 15 watts can be obtained on the fundamental and nearly as much on the second harmonic.

Grid-plate oscillator — In the grid-plate oscillator, Fig. 408, the crystal is connected between grid and ground and the cathode tuned circuit. C_2 and *RFC*, is tuned to a frequency lower than that of the crystal. This circuit gives high output on the fundamental crystal frequency with low crystal current. The output on even harmonics (2nd, 4th, etc.) is not so great as that obtainable with the Tri-tet, but on odd harmonics (3rd, 5th, etc.) the output is appreciably better.

If harmonic output is not needed, C_2 may be a fixed capacity of 100 μμfd. The cathode coil, *RFC*, may be a 2.5-mh. choke, since the inductance is not critical.

Output power of 15 to 20 watts at the crystal fundamental may be obtained with a tube such as the 6L6G at plate and screen voltages of 400 and 250, respectively.

Tuning and adjustment — The tuning procedure for the Tri-tet oscillator is as follows: With the cathode tank condenser at about three-quarters scale turn the plate tank condenser until there is a sharp dip in plate cur-

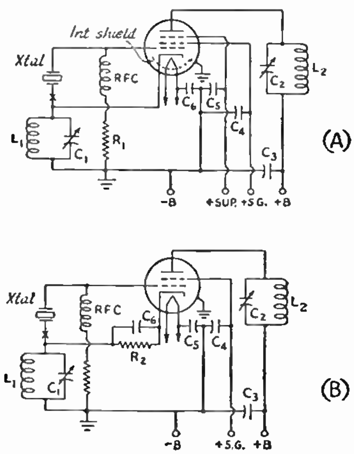


Fig. 407 — Tri-tet oscillator circuit, using pentodes (A) or beam tetrodes (B). C_1 and C_2 are 200-μμfd. variable condensers. C_3, C_4, C_5, C_6, may be 0.001 to 0.01 μfd.; their values are not critical. R_1, 20,000 to 100,000 ohms. R_2 should be 400 ohms for 400- or 500-volt operation. The following specifications for the cathode coils, L_1, are based on a diameter of 1½ inches and a length of 1 inch; turns should be spaced evenly to fill the required length: for 1.75-Mc. crystal, 32 turns; 3.5 Mc., 10 turns; 7 Mc., 6 turns. The screen should be operated at 250 volts or less. Audio beam tetrodes such as the 6L6 and 6L6G should be used only for second-harmonic output. A flash-light bulb may be inserted at the point marked X (§ 1-3). The *L/C* ratio in the plate tank, L_2C_2, should be such that the capacity in use is 75 to 100 μμfd. for fundamental output and about 25 μμfd. for second-harmonic output.

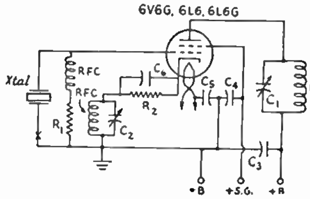


Fig. 408 — Grid-plate crystal oscillator circuit. In the cathode circuit, *RFC* is a 2.5-mh. r.f. choke. Other constants are the same as in Fig. 407. A crystal-current indicator may be inserted at the point marked X (§ 4-3).

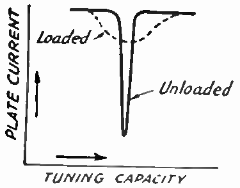
rent, indicating that the plate circuit is in resonance. The crystal should be oscillating continuously, regardless of the setting of the plate condenser. Set the plate condenser so that plate current is minimum. The load circuit may then be coupled and adjusted so that the oscillator delivers power. The minimum plate current will rise; it may be necessary to retune the plate condenser when the load is coupled to bring the plate current to a new minimum. Fig. 409 shows the typical behavior of plate current with plate-condenser tuning.

After the plate circuit is adjusted and the oscillator is delivering power, the cathode condenser should be readjusted to obtain optimum power output. The setting should be as far toward the low-capacity end of the scale as is consistent with good output; it may, in fact, be desirable to sacrifice a little output if so doing lowers the current through the crystal and thus reduces heating.

For harmonic output the plate tank circuit is tuned to the harmonic instead of the fundamental of the crystal frequency. A plate-current dip will occur at the harmonic. If the cathode condenser is adjusted for maximum output at the harmonic, this adjustment will usually serve for the fundamental as well. The crystal should be checked for excessive heating, the most effective remedy being to lower plate and/or screen voltage or to reduce the loading. Maximum r.f. voltage across the crystal is developed at maximum load, so heating should be checked with the load coupled.

When a fixed cathode condenser is used in the grid-plate oscillator the plate tank circuit is simply resonated, as indicated by the plate-current dip, to the fundamental or a harmonic of the output frequency, loading being adjusted to give optimum power output. If the variable cathode condenser is used, it should be set to give, by observation, the maximum power output consistent with safe crystal current. The variable condenser is useful chiefly in increasing the output on the third and higher harmonics; for fundamental operation, the cathode capacity is not critical and the fixed condenser may be used.

Fig. 409 — Curves showing d.c. plate current vs. plate-condenser tuning, both with and without load, for the Tri-tet oscillator. The setting for minimum plate current may shift with loading.



ℂ 4-6 Interstage Coupling

Requirements — The purpose of the interstage coupling system is to transfer, with as little energy loss as possible, the power developed in the plate circuit of one tube (the *driver*) to the grid circuit of the following amplifier tube or frequency multiplier. The circuits in practical use are based on the fundamental coupling arrangements described in § 2-11. In the process of power transfer, impedance transformation (§ 2-9) frequently is necessary so that the proper exciting voltage and current will be available at the grid of the driven tube.

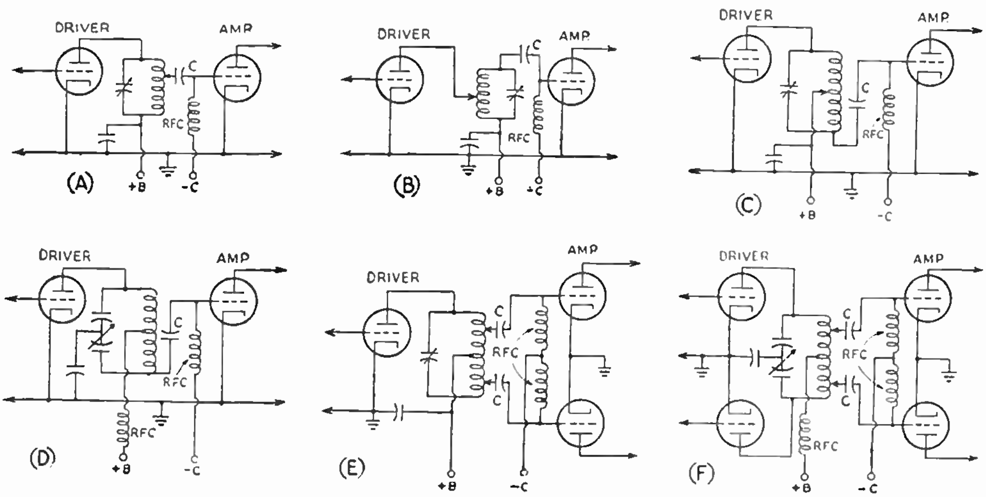


Fig. 410 — Direct- or capacity-coupled driver and amplifier stages. The coupling capacity may be from 50 μμfd. to 0.002 μfd.; it is not critical except where tapping the coils for control of excitation is not possible. Parallel plate feed to the driver and series grid feed to the amplifier may be substituted in any of these circuits (§3-7).

Capacity coupling — Fig. 410 shows several types of capacitive coupling. In each case, *C* is the coupling condenser. The coupling condenser serves also as a blocking condenser (§ 2-13) to isolate the d.c. plate voltage of the driver from the grid of the amplifier. The circuits of C and D are preferable when a balanced circuit is used in the output of the driver; instead of both tubes being in parallel across one side, the output capacity of the driver tube and the input capacity of the amplifier are across opposite sides of the tank circuit, thereby preserving a better circuit balance. The circuits of E and F are designed for coupling to a push-pull stage.

In A, B, E and F, excitation is adjusted by moving the tap on the coil to provide an optimum impedance match. In E and F, the two grid taps should be maintained equidistant from the center-tap on the coil.

While capacitive coupling is simplest from the viewpoint of construction, it has certain disadvantages. The input capacity of the amplifier is shunted across at least a portion of the driver tank coil. When added to the output capacity of the driver tube, this additional capacity may be sufficient, in many cases, to prevent use of a desirable *L C* ratio in circuits for frequencies above about 7 Mc.

Link coupling — At the higher frequencies it is advantageous in reducing the effects of tube capacities on the *L C* ratio to use separate tank circuits for the driver plate and amplifier grid, coupling the two circuits by means of a link (§ 2-11). This method of coupling also has some constructional advantages, in that separate parts of the transmitter may be constructed as separate units without the necessity for running long leads at high r.f. potential.

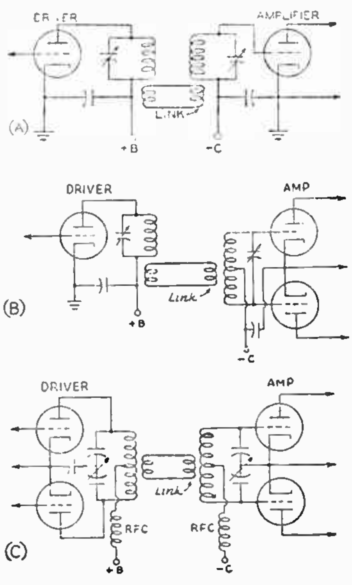


Fig. 411 — Link coupling between driver and amplifier.

Circuits for link coupling are shown in Fig. 411. The coupling ordinarily is by a turn or two of wire closely coupled to the tank inductance at a point of low r.f. potential, such as the center of the coil of a balanced tank circuit or the "ground" end of the coil in a single-ended circuit. The link line usually consists of two closely spaced parallel wires; occasionally the wires are twisted together, but this usually causes undue losses at high frequencies.

It is advisable to have some means of varying the coupling between link and tank coils. The link coil may be arranged to be swung in relation to the tank coil or, when it consists of a large turn around the outside of the tank coil, may be split into two parts which can be pulled apart or closed somewhat in the fashion of a pair of calipers. If the tank coils are wound on forms, the link may be wound close to the main coil.

With fixed coils, some adjustment of coupling usually can be obtained by varying the number of turns on the link. In general, the proper number of turns for the link must be found by experiment.

ℂ 4-7 R.F. Power-Amplifier Circuits

Tetrode and pentode amplifiers — When the input and output circuits of an r.f. amplifier tube are tuned to the same frequency it will oscillate as a tuned-grid tuned-plate oscillator, unless some means is provided to eliminate the effects of feed-back through the plate-to-grid capacity of the tube (§ 3-5). In all transmitting r.f. tetrodes and pentodes, this capacity is reduced to a satisfactory degree by the internal shielding between grid and plate provided by the screen. Tetrodes and pentodes designed for audio use (such as the 6L6, 6V6, 6F6, etc.) are not sufficiently well screened for use as r.f. amplifiers without employing suitable means for nullifying the effect of the grid-plate capacity.

Typical circuits of tetrode and pentode r.f. amplifiers are shown in Fig. 412. The high power sensitivity (§ 3-3) of pentodes and tetrodes, makes them prone to self-oscillate with very small values of feed-back voltage, however, so that particular care must be used to prevent feed-back by means external to the tube itself. This calls for adequate isolation of plate and grid tank circuits to prevent undesired magnetic or capacity coupling between them. The requisite isolation can be secured either by keeping the circuits well separated and mounting the coils so that magnetic coupling is minimized, or by the use of interstage shielding (§ 2-11).

Triode amplifiers — The feed-back through the grid-plate capacity of a triode cannot be eliminated, and therefore special circuit means called *neutralization* must be used to prevent oscillation. A properly neutralized triode amplifier then behaves as though it were operating at very low frequencies, where the grid-plate capacity feed-back is negligible (§ 3-3).

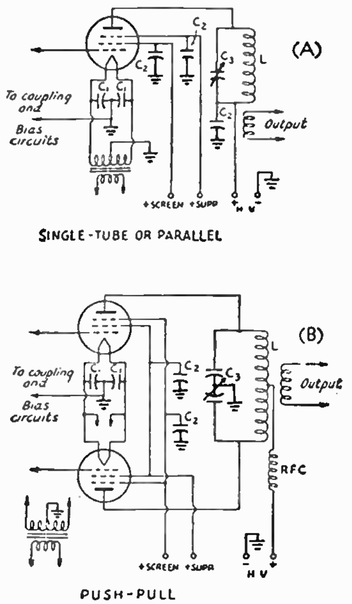


Fig. 412 — Typical tetrode-pentode r.f. amplifier circuits. C_1 — 0.01 μfd. C_2 — 0.001 μfd. C_3-L — See § 4-8. In circuits for tetrodes, the suppressor-grid connection and its associated by-pass condenser are omitted.

Neutralization — Neutralization amounts to taking some of the radio-frequency current from the output or input circuit of the amplifier and introducing it into the other circuit in such a way that it effectively cancels the current flowing through the grid-plate capacity of the tube, thus rendering it impossible for the tube to supply its own excitation. For complete neutralization of the amplifier, the two currents must be opposite in phase (§ 2-7) and equal in amplitude.

The out-of-phase current (or voltage) can be obtained quite readily by using a balanced tank circuit for either grid or plate, taking the neutralizing voltage from the end of the tank opposite that to which the grid or plate is connected. The amplitude of the neutralizing voltage can be regulated by means of a small condenser, the *neutralizing condenser*, having the same order of capacity as the grid-plate capacity of the tube. Circuits in which the neutralizing voltage is obtained from a balanced grid tank and fed to the plate through the neutralizing condenser are *grid-neutralized* circuits, while if the neutralizing voltage is obtained from a balanced plate tank and fed to the grid the circuit is *plate-neutralized.*

Plate-neutralized circuits — The circuits for plate neutralization are shown in Fig. 413 at A, B and C. In A, voltage induced in the extension of the tank coil is fed back to the grid through the neutralizing condenser, C_n, to balance the voltage appearing between grid and plate. In this circuit, the capacity required at C_n increases as the tank coil extension is made smaller; in general, neutralization is satisfactory over only a small range of frequencies since the coupling between the two sections of the tank coil will vary with the amount of capacity in use at C.

In B the tank coil is center-tapped to give equal voltages on either side of the center tap, the tank condenser being across the whole coil. The neutralizing capacity is approximately equal to the grid-plate capacity of the tube, in this case. A disadvantage of the circuit, when used with the single tank condenser shown, is that the rotor of the condenser is above ground potential, and hence small capacity changes caused by bringing the hand near the tuning control (*hand capacity*) cause detuning. In general, neutralization is complete at only one

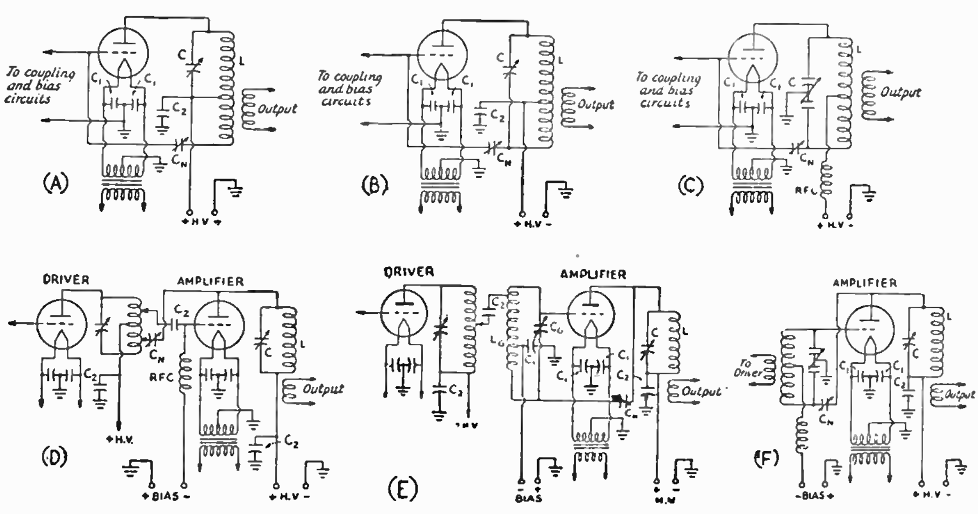


Fig. 413 — Neutralized triode amplifier circuits. Plate neutralization is shown in A, B and C, while D, E and F show types of grid neutralization. Either capacitive or link coupling may be used with the circuits of A, B or C. C-L — See § 4-8. C_g-L_g — Grid tank circuit. C_n — Neutralizing condensers. C_1 — 0.01 μfd. C_2 — 0.001 μfd.

frequency since the plate-cathode capacity of the tube is across only half the tank coil; also, it is difficult to secure an exact center-tap. Both of these factors cause unbalance, which in turn causes the voltages across the two halves of the coil to differ when the frequency is changed.

The circuit of C also uses a center-tapped tank circuit, the voltage division being secured by use of a balanced (split-stator) tank condenser, the two condenser sections being identical. C_n is approximately equal to the grid-plate capacity of the tube. In this circuit the upper section of the tank condenser is in parallel with the output capacity of the tube, hence the circuit can be completely neutralized at only one setting of the tank condenser unless a

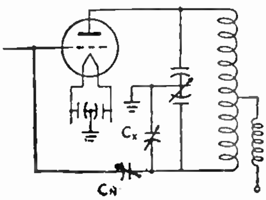


Fig. 414 — Compensating for unbalance in the single-tube neutralizing circuit. C_x, the balancing condenser, has a maximum capacity somewhat larger than the output capacity of the tube.

compensating capacity (Fig. 414) is connected across the lower section. It is adjusted so the neutralizing condenser need not be changed when frequency is shifted. In practice, if the capacity in use in the tank circuit is large compared to the plate-cathode capacity the unbalancing effect is not serious.

Grid-neutralized circuits — Typical circuits employing grid neutralization are shown in Fig. 413 at D, E and F. The principle of balancing out the feed-back voltage is the same as in plate neutralization. However, in these circuits the neutralizing voltage may be either in phase or out of phase with the excitation voltage on the grid side of the input tank circuit depending upon whether the tank is divided by means of a balanced condenser or a tapped coil. Circuits such as those at D and E, neutralized by ordinary procedure (described below), will be regenerative when the plate voltage is applied; the circuit at F will be degenerative. In addition the normal unbalancing effects previously described are present, so that grid neutralizing is less satisfactory than the plate method.

Inductive neutralization — With this type of neutralization, inductive coupling between the grid and plate circuits is provided in such a way that the voltage induced in the grid coil by magnetic coupling from the plate coil opposes the voltage fed back through the grid-plate capacity of the tube. A representative circuit arrangement, using a coupling link to provide the mutual inductance (§ 2-11), is shown in Fig. 415-A. The link coils are of one or two turns coupled to the grounded ends of the tank coils. Neutralization is adjusted by moving the link coils in relation to the tank coils. Reversal of connections to one coil may be required for proper phasing. Ordinary inductive coupling between the two coils also could be used, but it is less convenient. Inductive neutralization is complete only at one frequency since the effective mutual inductance changes to some extent with tuning, but is useful in cases where the grid-plate capacity of the tube is very small and suitable circuit balance cannot be obtained by using neutralizing condensers.

Another form of neutralization, known as "coil" or "shunt" neutralization, is shown at B. Its operation is based on making the inductance of L_n such that, together with the grid-plate capacity of the tube, it resonates at the operating frequency. C_2 is merely a plate-voltage blocking condenser. If the Q of the coil is sufficiently high, the parallel resonant impedance between grid and plate is much higher than the grid-cathode circuit impedance. Because the system is difficult to adjust and functions satisfactorily only at one frequency, it is used chiefly in fixed-frequency transmitters. The variation in Fig. 414-C is useful for v.h.f. In this arrangement the coil is replaced by a parallel line, the effective length of which is adjusted until it is resonant when loaded by the grid-plate capacity.

Push-pull neutralization — With push-pull circuits two neutralizing condensers are used, as shown in Fig. 416. In these circuits, the grid-plate capacities of the tubes and the neutralizing capacities form a capacity bridge (§ 2-11) which is independent of the grid and plate tank circuits. The neutralizing capacities are approximately the same as the tube grid-plate capacities. With electrically similar tubes and symmetrical construction (stray capacities to ground equal on both sides of the circuit), the neutralization is complete and independent of frequency. A circuit using a balanced condenser, as at B, is preferred, since it is an aid in obtaining good circuit balance.

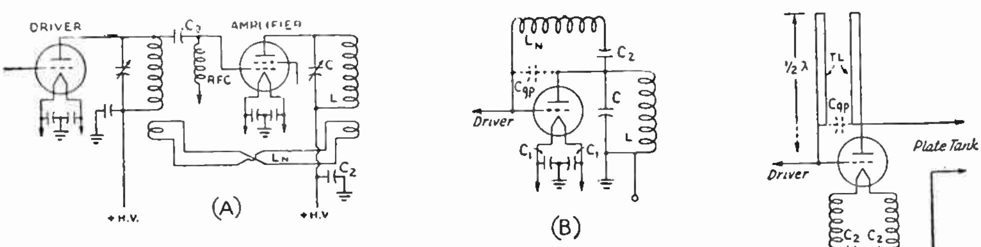


Fig. 415 — Inductive neutralization circuits. A, link neutralization. B, "coil" or shunt neutralization. C, modified shunt neutralizing circuit for v.h.f. using a half-wave line.

Frequency effects — The effects of slight dissymmetry in a neutralized circuit become more important as the frequency is raised, and may be sufficient at the very-high frequencies (or even lower) to prevent good neutralization. At these frequencies the inductances and stray capacities of even short leads become important elements in the circuit, while input loading effects (§ 7-6) may make it impossible to get proper phasing, particularly in single-tube circuits. In such cases the use of a push-pull amplifier, with its general freedom from the effects of dissymmetry, is not only much to be preferred but may be the only type of circuit which can be satisfactorily neutralized.

Neutralizing condensers — In most cases the neutralizing voltage will be equal to the r.f. voltage between the plate and grid of the

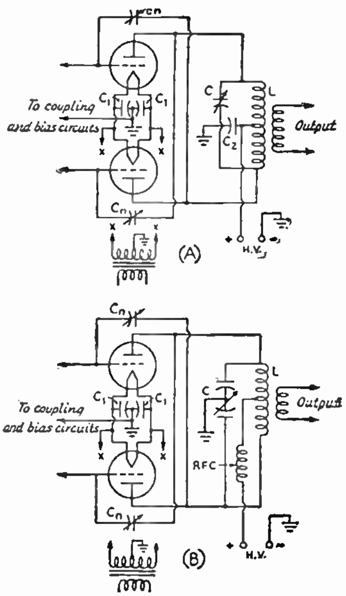


Fig. 416 — "Cross-neutralized" push-pull r.f. amplifier circuits. Either capacitive or link coupling may be used.
C-L — See § 4-8. C_n — Neutralizing condensers.
C_1 — 0.01 μfd. C_2 — 0.001 μfd. or larger.

tube, so that for perfect balance the capacity required in the neutralizing condenser theoretically will be equal to the grid-plate capacity. If, in the circuits having tapped tank coils, the tap is more than half the total number of turns from the plate end of the coil, the required neutralizing capacity will increase approximately in proportion to the relative number of turns in the two sections of the coil.

With tubes having grid and plate connections brought out through the bulb, a condenser having at about half-scale or less a capacity equal to the grid-plate capacity of the tube should be chosen. If the grid and plate leads are brought through a common base the capacity needed is greater, because the tube socket and its associated wiring adds some capacity to the actual interelement capacities. When two or more tubes are connected in parallel, the neutralizing capacity required will be in proportion to the number of tubes.

The voltage rating of neutralizing condensers must at least equal the r.f. voltage across the condenser plus the sum of the d.c. plate voltage and the grid-bias voltage.

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

The grid-circuit milliammeter makes a good neutralizing indicator. If the circuit is not completely neutralized, tuning of the plate tank circuit through resonance will change the tuning of the grid circuit and affect its loading, causing a change in the rectified d.c. grid current. The setting of the neutralizing condenser which leaves the grid current unaffected as the plate tank is tuned through resonance is the correct one. If the circuit is out of neutralization, the grid current will drop perceptibly as the plate tank is tuned through resonance. As the point of neutralization is approached, by adjusting the neutralizing capacity in small steps the dip in grid current as the plate condenser is swung through resonance will become less and less pronounced, until, at exact neutralization, there will be no dip at all. Further change of the neutralizing capacity in the same direction will bring the grid-current dip back. The neutralizing condenser should always be adjusted with a screwdriver of insulating material to avoid hand-capacity effects.

Adjustment of the neutralizing condenser may affect the tuning of the grid tank or driver plate tank, so both circuits should be retuned each time a change is made in neutralizing capacity. In neutralizing a push-pull amplifier the neutralizing condensers should be adjusted together, step by step, keeping their capacities as equal as possible.

With single-ended circuits having split-stator neutralizing, the behavior of the grid meter will depend somewhat upon the type of tube used. If the tube output capacity is not great enough to upset the balance, the action of the meter will be the same as in other circuits. With high-capacity tubes, however, the meter usually will show a gradual rise and fall as the plate tank is tuned through resonance, reaching a maximum right at resonance when the circuit is properly neutralized.

When an amplifier is not neutralized a neon bulb touched to the plate of the amplifier tube or to the plate side of the tuning condenser will glow when the tank circuit is tuned through resonance, providing the driver has sufficient power. The glow will disappear when the amplifier is neutralized. However, touching the neon bulb to such an ungrounded point in the circuit may introduce enough stray capacity to unbalance the circuit slightly, thus upsetting the neutralizing.

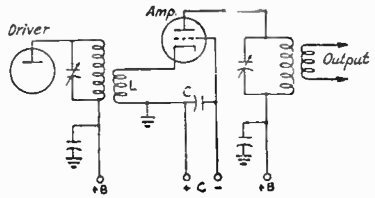


Fig. 417 — Inverted amplifier. The number of turns at *L* should be adjusted by experiment to give optimum grid excitation. By-pass condenser *C* is 0.001 μfd. or larger.

A flashlight bulb connected in series with a single-turn loop of wire 2½ or 3 inches in diameter, with the loop coupled to the tank coil, also will serve as a neutralizing indicator. Capacitive unbalance can be avoided by coupling the loop to the low-potential part of the tank coil.

Incomplete neutralization — If a setting of the neutralizing condenser can be found which gives minimum r.f. current in the plate tank circuit without completely eliminating it, there may be magnetic or capacity coupling between the input and output circuits external to the tube itself. Short leads in neutralizing circuits are highly desirable, and the input and output inductances should be so placed with respect to each other that magnetic coupling is minimized. Usually this requires that the axes of the coils must be at right angles to each other. In some cases it may be necessary to shield the input and output circuits from each other. Magnetic coupling can be detected by disconnecting the plate tank from the remainder of the circuit and testing for r.f. in it (by means of the flashlight lamp and loop) as the tank condenser is tuned through resonance. The driver stage must be operating while this is done, of course.

With single-ended amplifiers there are many stray capacities left uncompensated for in the neutralizing process. With large tubes, especially those having relatively high interelectrode capacities, these commonly neglected stray capacities can prevent perfect neutralization. Symmetrical arrangement of a push-pull stage is about the only way to obtain practically perfect balance throughout the amplifier.

The neutralization of tubes with extremely low grid-plate capacity, such as the 6L6, is often difficult, since it frequently happens that the wiring itself will introduce sufficient capacity between the right points to "over-neutralize" the grid-plate capacity. The use of a neutralizing condenser only aggravates the condition. Inductive or link neutralization, as shown in Fig. 415, has been used successfully with such tubes.

The inverted amplifier — The circuit of Fig. 417 avoids the necessity for neutralization by operating the control grid of the tube at ground potential, thus making it serve as a shield between the input and output circuits. It is particularly useful with tubes of low grid-plate capacity, which are difficult to neutralize by ordinary methods. Excitation is applied between grid and cathode through the coupling coil, *L*; since this coil is common to both the plate and grid circuits the amplifier is degenerative with the circuit constants normally used, hence more excitation voltage and power are required for a given output than is the case with a neutralized amplifier. The tube used must have low plate-cathode capacity (of the order of 1 μμfd. or less) since larger values will give sufficient feed-back to permit it to oscillate, the circuit then becoming the ultraudion (§ 3-7). Tubes having sufficiently low plate-cathode capacity (audio pentodes, for example) can be used without danger of oscillation at frequencies up to perhaps 30 Mc. or so.

4-8 Power Amplifier Operation

Efficiency — An r.f. power amplifier is usually operated Class-C (§ 3-4) to obtain a reasonably high value of plate efficiency (§ 3-3). The higher the plate efficiency the higher the power input that can be applied to the tube without exceeding the plate dissipation rating (§ 3-2), up to the limits of other tube ratings (plate voltage and plate current). Plate efficiencies of the order of 75 per cent are readily obtainable at frequencies up to the 30–50-Mc. region. The *over-all* efficiency of the amplifier will be lower by the power lost in the tank and coupling circuits, so that the actual efficiency is less than the plate efficiency.

Operating angle — The operating angle is the proportionate part of the exciting grid-voltage cycle (§ 2-7) during which plate current flows, as shown in Fig. 418. For Class-C operation, it is usually in the vicinity of 120–150 degrees. With other operating considerations, this angle results in an optimum relationship between plate efficiency and grid driving power.

Load impedance — The load impedance (§ 3-3) for an r.f. power amplifier is adjusted, by tuning the plate tank circuit to resonance, to represent a pure resistance at the operating frequency (§ 2-10). Its value, which usually is in the neighborhood of a few thousand ohms, is

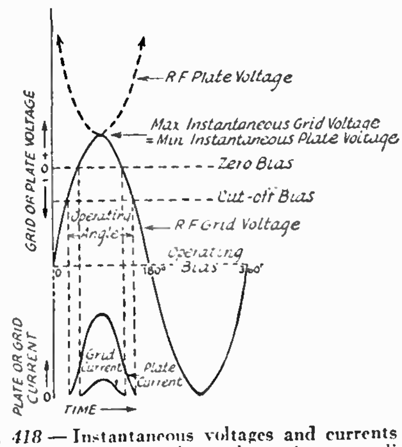


Fig. 418 — Instantaneous voltages and currents in a Class-C amplifier operating under optimum conditions.

adjusted by varying the loading on the tank circuit, closer coupling to the load giving lower values of load resistance and vice versa (§ 2-11). The load may be either the grid circuit of a following stage or the antenna circuit.

For highest efficiency the value of load resistance should be relatively high, but if only limited excitation voltage is available greater power output will be secured by using a lower value of load resistance. The latter adjustment is accompanied by a decrease in plate efficiency. The optimum load resistance is that which, for the maximum permissible peak plate current, causes the minimum instantaneous plate voltage (Fig. 418) to be equal to the maximum instantaneous grid voltage required to cause the peak plate current to flow; this gives the optimum ratio of plate efficiency to required grid driving power.

R.f. grid voltage and grid bias — For most tubes optimum operating conditions result when the minimum instantaneous plate voltage is 10 to 20 per cent of the d.c. plate voltage, so that the maximum instantaneous positive grid voltage must be approximately the same figure. Since plate current starts flowing when the instantaneous voltage reaches the cut-off value (§ 3-2), the d.c. grid voltage must be considerably higher than cut-off to confine the operating angle to 150 degrees or less (with grid bias at cut-off, the angle would be 180 degrees). For an angle of 120 degrees, the r.f. grid voltage must reach 50 per cent of its peak value (§ 2-7) at the cut-off point. The corresponding figure for an angle of 150 degrees is 25 per cent. Hence, the operating bias required is the cut-off value plus 25 to 50 per cent of the peak r.f. grid voltage. These relations are shown in Fig. 418. The grid bias should be at least twice cut-off if the amplifier is to be plate modulated, so that the operating angle will be not less than 180 degrees when the plate voltage rises to twice the steady d.c. value (§ 5-3). Because of their relatively high amplification factors, with most modern tubes Class-C operation requires considerably more than twice cut-off bias to make the operating angle fall in the region mentioned above. Suitable operating conditions are usually given in the data accompanying the type of tube used.

Grid bias may be secured either from a bias source (*fixed bias*), a grid leak (§ 3-6) of suitable value, or from a combination of both. When a bias supply is used, its voltage regulation should be taken into consideration (§ 8-9).

Driving power — As indicated in Fig. 418, grid current flows only during a small portion of the peak of the r.f. grid voltage cycle. The power consumed in the grid circuit therefore is approximately equal to the peak r.f. grid voltage multiplied by the average rectified grid current as read by a d.c. milliammeter. The peak r.f. grid voltage, if not included in the tube manufacturer's operating data, can be estimated roughly by adding 10 to 20 per cent of the plate voltage to the operating grid bias, assuming the operating conditions are as described above.

At frequencies up to 30 Mc. or so, the grid losses are practically entirely those resulting from grid-current flow. At the very-high frequencies, however, dielectric losses in the glass envelope and base materials become appreciable, together with losses caused by transit-time effects (§ 7-6), and may necessitate supplying several times the driving power indicated above. At any frequency, the driving stage should be capable of a power output two to three times the power it is expected the grid circuit of the amplifier will consume. This is necessary because losses in the tank and coupling circuits must also be supplied, and also to provide reasonably good regulation of the r.f. grid voltage. Good voltage regulation (see § 8-1 for general definition) insures that the waveform of the excitation voltage will not be distorted because of the changing load on the driver during the r.f. cycle.

Grid impedance — During most of the r.f. grid-voltage cycle no grid current is flowing, as

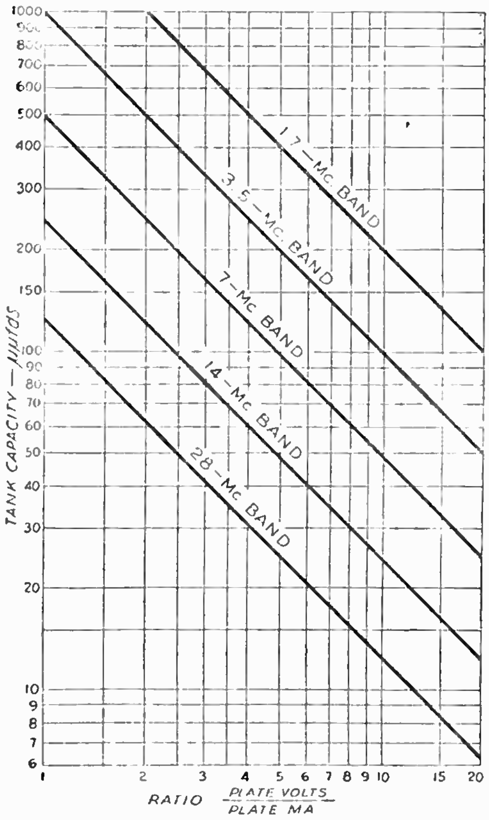


Fig. 419 — Chart showing tank capacities required for a Q of 12 with various ratios of plate voltage to plate current, for various frequencies. In circuits F, G, H (Fig. 120), the capacities shown in the graph may be divided by four. In circuits C, D, E, I, J and K, the capacity of each section of the split-stator condenser may be one-half that shown by the graph. The values given by the graph should be used for circuits A and B.

indicated in Fig. 418, hence the grid impedance is infinite. During the peak of the cycle, however, the impedance may drop to very low values (of the order of 1000 ohms), depending upon the type of tube. Both the minimum and average values of grid impedance depend to a considerable extent on the amplification factor of the tube, being lower with tubes having large amplification factors.

The average grid impedance is equal to E^2/P, where E is the r.m.s. (§ 2-7) value of r.f. grid voltage and P is the grid driving power. Under optimum operating conditions, values of average grid impedance ranging from 2000 ohms for high-μ tubes to four or five times as much for low-μ types are representative. Values in the vicinity of 4000 to 5000 ohms are typical of modern triodes with amplification factors of 20 to 30.

Because of the large change in impedance during the cycle, it is necessary that the tank circuit associated with the amplifier grid have fairly high Q. This is essential to provide sufficient storage capacity so that the voltage regulation over the cycle will be good. The requisite Q may be obtained by adjusting the L/C ratio or by tapping the grid circuit across only part of the tank (§ 4-6).

Tank-circuit Q — Besides serving as a means for transforming the actual load resistance to the required value of plate load impedance for the tube, the plate tank circuit also should suppress the harmonics present in the tube output as a result of the non-sinusoidal plate current (§ 2-7, 3-3). For satisfactory harmonic suppression, a Q of 12 or more (with the circuit fully loaded) is desirable. A Q of this order also is helpful from the standpoint of securing adequate coupling to the load or antenna circuit (§ 2-11). The proper Q can be obtained by suitable selection of L/C ratio in relation to the optimum plate load resistance for the tube (§ 2-10).

For a Class-C amplifier operated under optimum conditions as described above, the plate load impedance is approximately proportional to the ratio of d.c. plate voltage to d.c. plate current. For a given effective Q the tank capacity required at a given frequency will be inversely proportional to the parallel resistance (§ 2-10), so that it will also be inversely proportional to the plate-voltage/plate-current ratio.

The tank capacity required on various amateur bands for a Q of 12 is shown in Fig. 419 as a function of this ratio. The capacity given is for single-ended tank circuits, as shown in Fig. 420 at A and B. When a balanced tank circuit is used the total tank capacity required is reduced to one-fourth this value, because the tube is connected across only half the circuit (§ 2-9). Thus, if the plate-voltage/plate-current ratio calls for a capacity of 200 $\mu\mu$fd. in a single-ended circuit at the desired frequency, only 50 $\mu\mu$fd. would be needed in a balanced circuit. If a split-stator or balanced tank condenser is used each section should have a capacity of 100 $\mu\mu$fd., the total capacity of the two in series being 50 $\mu\mu$fd. These are "in use" capacities; not simply the rated maximum capacity of the condenser. Larger values may be used with an increase in the effective Q.

To reduce energy loss in the tank circuit, the inherent Q of the coil and condenser should be high. Since transmitting coils usually have Qs ranging from 100 to several hundred, the tank transfer efficiency generally is 90 per cent or more. An unduly large C/L ratio is not advisable since it will result in large circulating r.f. tank current and hence relatively large losses in the tank, with a consequent reduction in the power available for the load.

Tank constants — When the capacity necessary for a Q of 12 has been determined from Fig. 419, the inductance required to resonate at the given frequency can be found by means

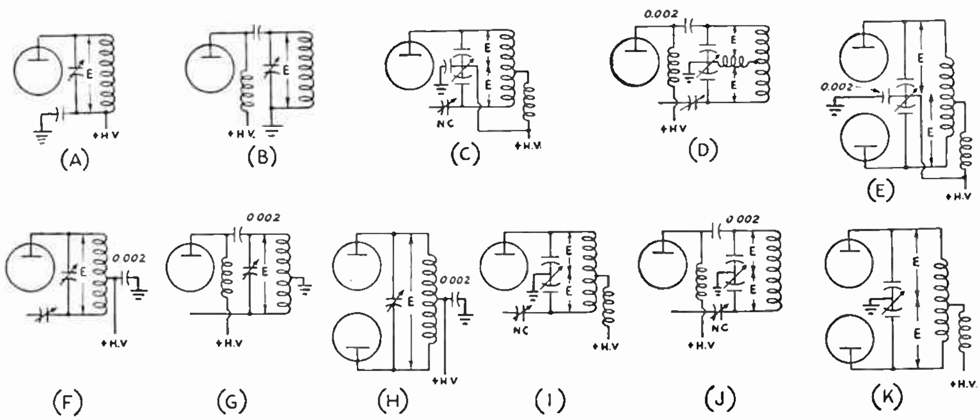


Fig. 420 — In circuits A, B, C, D and E, the peak voltage E will be approximately equal to the d.c. plate voltage applied for c.w. or twice this value for 'phone. In circuits F, G, H, I, J and K, E will be twice the d.c. plate voltage for c.w. or four times the plate voltage for 'phone. The circuit is assumed to be fully loaded. Tubes in parallel in any of the circuits will not affect the peak voltage. Circuits A, C, E, F, G and H require that the tank condenser be insulated from chassis or ground and that it be provided with a suitably insulated shaft coupling for tuning.

of the formula in § 2-10. Alternatively, the required number of turns on coils of various construction can be found from the charts of Figs. 421 and 422.

Fig. 421 is for coils wound on receiving-type forms having a diameter of 1½ inches and ceramic forms having a diameter of 1¾ inches and winding length of 3 inches. Such coils would be suitable for oscillator and buffer stages where the power is not over 50 watts. In all cases, the number of turns given must be wound to fit the length indicated and the turns should be evenly spaced.

Fig. 422 gives data on coils wound on transmitting-type ceramic forms. In the case of the smallest form, extra curves are given for double spacing (winding turns in alternate grooves). This is sometimes advisable in the case of 14- and 28-Mc. coils when only a few turns are required. In all other cases, the specified number of turns should be wound in the grooves without any additional spacing.

Ratings of components — The peak voltage to be expected between the plates of a tank condenser depends upon the arrangement of the tank circuit as well as the d.c. plate voltage. Peak voltage may be determined from Fig. 420, which shows all of the commonly used tank-circuit arrangements. These estimates assume that the amplifier is fully loaded; the voltage will rise considerably should the amplifier be

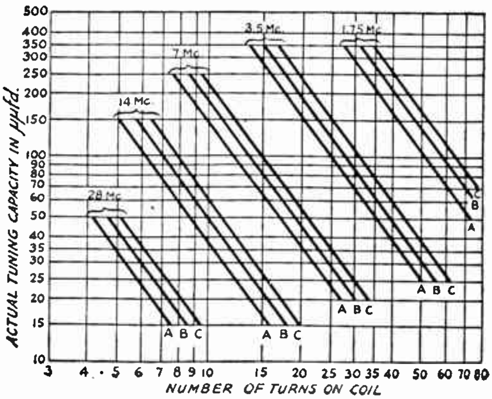


Fig. 421 — Coil-winding data for receiving-type forms, diameter 1½ inches. Curve A — winding length, 1 inch; Curve B — winding length, 1½ inches; Curve C — winding length, 2 inches. Curve C is also suitable for coils wound on 1¾-inch diameter transmitting-type ceramic forms with 3 inches of winding length.

operated without load. The figures include a reasonable factor of safety.

The condenser plate spacing required to withstand any particular voltage will vary with the construction. Most manufacturers specify peak-voltage ratings in describing their condensers.

Plate or screen by-pass condensers of 0.001 μfd. should be satisfactory for frequencies as low as 1.7 Mc. Cathode-resistor and filament by-passes in r.f. circuits should be not less than 0.01 μfd. Fixed condensers used for these pur-

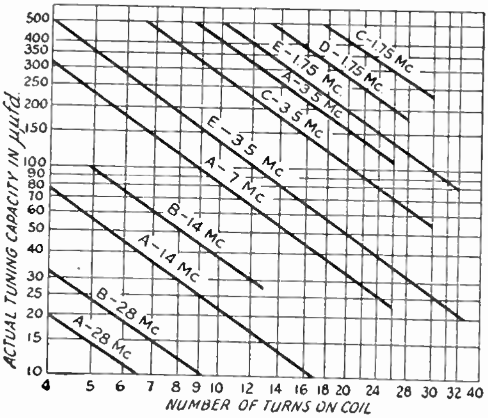


Fig. 422 — Coil-winding data for ceramic transmitting-type forms. Curve A — ceramic form 2½-inch effective diameter, 26 grooves, 7 per inch; Curve B — same as A, but with turns wound in alternate grooves; Curve C — ceramic form 2⅞-inch effective diameter, 32 grooves, 7.1 turns per inch, approximately; Curve D — ceramic form 4-inch effective diameter, 28 grooves, 5.85 turns per inch, approximately; Curve E — ceramic form 5-inch effective diameter, 26 grooves, 7 per inch. Coils may be wound with either No. 12 or No. 14 wire.

poses should have voltage ratings 25 to 50 per cent greater than the maximum d.c. or a.c. voltage across them.

Interstage coupling condensers should have voltage ratings 50 to 100 per cent greater than the sum of the driver plate and amplifier grid-biasing voltages.

4-9 Adjustment of Power Amplifiers

Excitation — The effectiveness of adjustments to the coupling between the driver plate and amplifier grid circuits can be gauged by the relative values of amplifier rectified grid current and driver plate current, the object being to obtain maximum grid current with minimum driver loading. The amplifier grid circuit represents the load on the driver stage, and the average grid impedance must therefore be transformed to the value for optimum driver operation (§ 4-8).

With capacity coupling, either the driver plate or amplifier grid must be tapped down on the driver tank coil, as shown in Fig. 410 at A and B, unless the grid impedance is approximately the right value for the driver plate load, when it will be satisfactory to connect both elements to the end of the tank. If the grid impedance is lower than the required driver plate load, Fig. 410-A is used; if higher, Fig. 410-B. In either case, the coupling which gives the desired grid current with minimum driver loading should be determined experimentally by moving the tap. Should both plate and grid be connected to the end of the circuit it is sometimes possible to control the loading, when the grid impedance is low, by varying the capacity of the coupling condenser, C, but this method is not altogether satisfactory since it is simply an expedient to prevent driver overloading without giving suitable impedance matching.

In push-pull circuits the method of adjustment is similar, except that the taps should be kept symmetrically located with respect to the center of the tank circuit.

With link coupling, Fig. 411, the object of adjustment is the same. The two tanks are first tuned to resonance, as indicated by maximum grid current, and the coupling adjusted by means of the links (§ 4-6) to give maximum grid current with minimum driver plate current. This usually will suffice to load the driver to its rated output, provided the driver plate and amplifier grid tank circuits have reasonable values of Q. If the Q of one or both of the circuits is too low, it may not be possible to load the driver fully with any adjustment of link turns or coupling at either tank. In such a case, the Qs of the tank circuits must be increased to the point where adequate coupling is secured. If the driver plate tank is designed to have a Q of 12, the difficulty almost invariably is in the amplifier grid tank. The Q can be increased to a suitable value either by adjustment of the L/C ratio or by tapping the load across part of the coil (§ 2-10).

Whatever the type of coupling, a preliminary adjustment should be made with the proper bias voltage and/or grid leak, but with the amplifier plate voltage off; then the amplifier should be carefully neutralized. After neutralization the driver-amplifier coupling should be readjusted for optimum power transfer, after which plate voltage may be applied and the amplifier plate circuit adjusted to resonance and coupled to its load. Under actual operating conditions the grid current decreases below the value obtained without plate voltage on the amplifier and the effective grid impedance rises, hence the final adjustment is to re-check the coupling to take care of this shift.

With recommended bias, the grid current obtained before plate voltage is applied to the amplifier should be 25 to 30 per cent higher than the value required for operating conditions. If this value is not obtained, and the driver plate input is up to rated value, the reason may be either improper matching of the amplifier grid to the driver plate or simply insufficient power output from the driver to take care of all losses. Driver operating voltages should be checked to assure they are up to rated values. If batteries are used for bias and are not strictly fresh, they should be replaced, since batteries which have been in use for some time often develop high internal resistance which effectively acts as additional grid-leak resistance. If a rectified a.c. bias supply is used, the bleeder or voltage-divider resistances should be checked to make certain that low grid current is not caused by greater grid-circuit resistance than is recommended. In this connection it is helpful to measure the actual bias when grid current is flowing, by means of a high-resistance d.c. voltmeter. There is also the possibility of loss of filament emission of the amplifier tube, either from prolonged service or from operating the filament under or over the rated voltage.

Plate tuning — In preliminary tuning, it is desirable to use low plate voltage to avoid possible damage to the tube. With excitation and plate voltage applied, rotate the plate tank condenser until the plate current dips. Then set the condenser at the minimum plate-current point (resonance). When the resonance point has been found, the plate voltage may be increased to its normal value.

With adequate excitation, the off-resonance plate current of a triode amplifier may be two or more times the normal operating value. With screen-grid tubes the off-resonance plate current may not be much higher than the normal operating value, since the plate current is principally determined by the screen rather than the plate voltage.

Under reasonably efficient operating conditions the minimum plate current with the amplifier unloaded will be a small fraction of the rated plate current for the tube (usually a fifth or less), since with no load the parallel impedance of the tank circuit is high. If the excitation is low the "dip" will not be very marked, but with adequate excitation the plate current at resonance without loading will be just high enough so that the d.c. plate power input supplies all the losses in the tube and circuit. As an indication of probable efficiency, the minimum plate current value should not be taken too seriously, because without load the Q of the circuit is high and the tank current relatively large. When the amplifier is delivering power to a load, the circulating current drops considerably and the tank losses correspondingly decrease. High minimum unloaded plate current is chiefly encountered at 28 Mc. and above, where tank losses are higher and the tank L/C ratio is usually lower than normal because of irreducible tube capacities. The effect is particularly noticeable with screen-grid tubes, which have relatively high output capacity. Because of the decrease in tank r.f. current with loading, however, the actual efficiency under load is reasonably good.

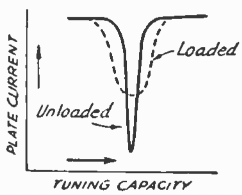


Fig. 423 — Typical behavior of d.c. plate current vs. tuning capacity in the plate circuit of an amplifier.

With the load (antenna or following amplifier grid circuit) connected, the coupling between plate tank and load should be adjusted to make the tube take rated plate current, keeping the tank always tuned to resonance. As the output coupling is increased the minimum plate current also will increase, about as shown in Fig. 423. Simultaneously the tuning becomes less sharp, because of the increase in effective resistance of the tank. If the load circuit simulates a resistance, the resonance setting of the tank condenser will be practically unchanged with loading; this is generally the case, since the load circuit usually is also tuned to resonance. A reactive load (such as an antenna or feeder system not tuned exactly to resonance) may cause the tank condenser setting to change with loading, since reactance as well as resistance is coupled into the tank (§ 2-11).

Power output — As a check on the operation of an amplifier, its power output may be measured by the use of a load of known resistance, coupled to the amplifier output as shown in Fig. 424. At A a thermoammeter, M, and a noninductive (ordinary wire-wound resistors are not satisfactory) resistance, R, are connected across a coil of a few turns coupled to the amplifier tank coil. The higher the resistance of R, the greater the number of turns required in the coupling coil. A resistor used in this way is generally called a "dummy antenna," since its use permits the transmitter to be adjusted without actually radiating power. The loading may readily be adjusted by varying the coupling between the two coils, so that the amplifier draws rated plate current when tuned to resonance. The power output is then calculated from Ohm's Law:

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

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

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

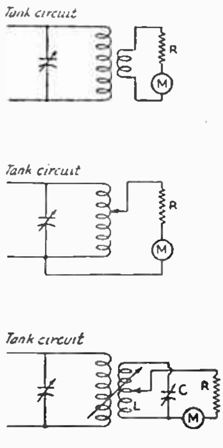
Incandescent light bulbs can be used to replace the special resistor and thermoammeter. The lamp should be equipped with a pair of leads, preferably soldered to the terminals on the lamp base. The coupling should be varied until the greatest brilliance is obtained for a given plate input. In using lamps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

The circuit of Fig. 424-B is for resistors or lamps of relatively high resistance. In using this circuit, care should be taken to avoid accidental contact with the plate tank when the power is on. This danger is avoided by circuit C, in which a separate tank circuit, LC, tuned to the operating frequency, is coupled to the plate tank circuit. The loading is adjusted by varying the number of turns across which the dummy antenna is connected on L and by changing the coupling between the two coils. With push-pull amplifiers, the dummy antenna should be tapped equally on either side of the center of the tank when the circuit of Fig. 424-B is used.

Harmonic suppression — The most important step in the elimination of harmonic radiation (§ 4-8, 2-12) is to use an output tank circuit having a Q of 12 or more. Beyond this it is desirable to avoid any considerable amount of over-excitation of a Class-C amplifier, since excitation in excess of that required for normal Class-C operation further distorts the plate-current pulse and increases the harmonic content in the output of the amplifier even though the proper tank Q is used. If the antenna system in use will accept harmonic frequencies they will be radiated when distortion is present, and consequently the antenna coupling system preferably should be selected with harmonic transfer in mind (§ 10-6).

Harmonic content can be reduced to some extent by preventing distortion of the r.f. grid-voltage wave-shape. This can be done by using a grid tank circuit with high effective Q. Link coupling between the driver and final amplifier are helpful, since the two tank circuits provide more attenuation than one at the harmonic frequencies. However, the advantages of link coupling in this respect may be nullified unless the Q of the grid tank is high enough to give good voltage regulation, which minimizes harmonic transfer and thus prevents distortion in the grid circuit.

The stray capacity between the antenna coupling coil and the tank coil may be sufficient to couple harmonic energy into the antenna system. This coupling may be eliminated by the use of electrostatic shielding (*Faraday shield*) between the two coils. Fig. 425 shows the construction of such a shield, while Fig. 426 illustrates the manner in which it is installed. The construction shown in Fig. 425 prevents current flow in the shield, which would occur if the wires formed closed circuits since the shield is in the magnetic field of the tank coil.



(A)

(B)

(C)

Fig. 424 — "Dummy antenna" circuits for checking power output and making operating adjustments under load without applying power to the actual antenna.

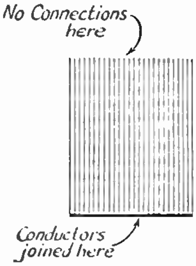


Fig. 425 — The Faraday electrostatic shield for eliminating capacitive transfer of harmonic energy. It is made of parallel conductors, insulated from each other except at one end where all are joined. Stiff wire or small diameter rod may be used, spaced about the diameter of the wire or rod. The shield should be larger than the diameter of the coil.

Should this occur, there would be magnetic shielding as well as electrostatic; in addition, there would be a power loss in the shield.

Improper operation — Inexact neutralization or stray coupling between plate and grid circuits may result in regeneration. This effect is most evident with low excitation, when the amplifier will show a sudden increase in output when the plate tank circuit is tuned slightly to the high-frequency side of resonance. It is accompanied by a pronounced increase in grid current.

Self-oscillation is apt to occur with tubes of high power sensitivity, such as the r.f. pentodes and tetrodes. In event of either regeneration or oscillation, circuit components should be arranged so that those in the plate circuit are well isolated from those of the grid circuit. Plate and grid leads should be made as short as possible and the screen should be by-passed as close to the socket terminal as possible. A cylindrical shield surrounding the lower portion of the tube up to the lower edge of the plate is sometimes required.

"Double resonance," or two tuning spots on the plate-tank condenser, one giving minimum plate current and the other maximum power output, may occur when the tank circuit Q is too low (§ 2-10). A similar effect also occurs at times with screen-grid amplifiers when the screen-voltage regulation (§ 8-1) is poor, as when the screen is supplied through a dropping resistor. The screen voltage decreases with a decrease in plate current, because the screen current increases under the same conditions. Thus the minimum plate-current point causes the screen voltage, and hence the power output, to be less than when a slightly higher plate current is drawn.

A phenomenon known as "grid emission" may occur when the amplifier tube is operated at higher than rated power dissipation on either the plate or grid. It is particularly likely to occur with tubes having oxide-coated cathodes, such as the indirectly heated types. It is caused by the grid reaching a temperature high enough to cause electron emission (§ 2-4). The electrons so emitted are attracted to the plate, further increasing the power input and heating, so that grid emission is characterized by gradually increasing plate current and heat which eventually will ruin the tube if the power is not removed. Grid emission can be prevented by operating the tube within its ratings.

ℂ 4-10 Parasitic Oscillations

Description — If the circuit conditions in an oscillator or amplifier are such that self-oscillation exists at some frequency other than that desired, the spurious oscillation is termed *parasitic*. The energy required to maintain a parasitic oscillation is wasted insofar as useful output is concerned, hence an oscillator or amplifier having parasitics will operate at reduced efficiency. In addition, its behavior at the operating frequency often will be erratic. Parasitic oscillations may be either higher or lower in frequency than the operating frequency.

The parasitic oscillation usually starts the instant plate voltage is applied, or, when the amplifier is biased beyond cut-off, at the instant excitation is applied. In the latter case, the oscillation frequently will be self-sustaining after the excitation has been removed. At other times the oscillation may not be self-sustaining, becoming active only in the presence of excitation. It may be apparent only by the production of abnormal key clicks (§ 6-1) over a wide frequency range, or by the presence of spurious side-bands (§ 5-2) with 'phone modulation.

Low-frequency parasitics — Parasitic oscillations at low frequencies (usually 500 kc. or less) are of the tuned-plate tuned-grid type, the tuned circuits being formed by r.f. chokes and associated by-pass and coupling condensers, with the regular tank tuning condensers having only a minor effect on the oscillation. The operating-frequency tank coil has negligible inductance for such low frequencies and may be short-circuited without affecting the oscillations. The oscillations do not occur when no r.f. chokes are used, hence whenever possible in series-fed circuits such chokes should be omitted. With single-ended amplifiers, it is usually possible to arrange the circuit so that either the grid or plate circuit needs no choke. In push-pull stages having chokes in both plate and grid circuits, it is helpful to connect an unby-passed grid leak from the choke to the bias supply or ground, thus placing the resistance in the parasitic circuit and tending to prevent oscillation. When the driver plate circuit has parallel feed and the amplifier grid circuit series feed (§ 3-7) this type of oscillation cannot occur if no choke is used in the series grid circuit, since the grid is grounded through the tank coil for the parasitic frequency.

Parasitics near operating frequency — In circuits utilizing a tap on the plate tank coil to establish a ground for a balanced neutralizing circuit, such as Fig. 413-B, a parasitic oscillation may be set up if the amplifier grid is tapped down on the grid (or driver plate) tank circuit for adjustment of driver-amplifier coupling (§ 4-6). In this case the turns between grid and ground and between plate and ground form, with the stray and other capacities present, a t.p.t.g. circuit (§ 3-7) which oscillates at a frequency somewhat higher than the nominal operating frequency. Such an oscillation can be prevented by dispensing with the taps in either the plate or grid circuit. Balancing the plate circuit by means of a split-stator condenser (Fig. 413-C) is recommended.

Very-high-frequency parasitics — Parasitics in the v.h.f. region are likely to occur with any amplifier having a balanced tank circuit, particularly when associated with neutralizing connections. The parasitic resonant circuit, formed by the leads connecting the various components, may be of either the t.p.t.g. or the ultraudion type.

The frequency of such oscillations may be determined by connecting a tuned circuit in series with the grid lead to the tube. A variable condenser (50 or 100 μμfd.) may be used, in conjunction with three or four self-supporting turns of heavy wire wound into a coil an inch or so in diameter. With the amplifier oscillating at the parasitic frequency, the condenser is slowly tuned through its range until oscillations cease. If this point is not found on the first trial, the turns of the coil may be spread apart or a turn removed and the process repeated. The use of such a tuned circuit as a trap is an almost certain remedy if the frequency can be determined, and introduces little if any loss at the operating frequency.

An alternative cure, which is feasible when the oscillation is of the t.p.t.g. type, is to detune the parasitic circuit in either the plate or grid circuit. Since this type of oscillation occurs most frequently with push-pull amplifiers, it may often be cured by making the grid and plate leads to their respective tank circuits of considerably different length. Similar considerations apply to neutralizing connections in push-pull circuits. The extra wire length may be coiled up in the form of a so-called "choke," which in this case is simply additional inductance for detuning the parasitic circuit.

Testing for parasitic oscillations — An amplifier always should be tested for parasitic oscillations before being considered ready for service. The preferable method is first to neutralize the amplifier, then apply sufficient fixed bias to permit a moderate value of plate current to flow without excitation. (The plate current should not be large enough to cause the power input to exceed the rated plate dissipation of the tube.) If the amplifier is free from self-starting parasitics, the plate current will remain steady as the tank condensers are varied; also, there will be no grid current and a neon bulb touched either to the plate or grid will show no glow. Extreme care must be used not to let the hand come into contact with any metal parts of the transmitter when using the neon bulb.

If any of these effects are present, the frequency of the parasitic must first be determined. If r.f. chokes are used in both the plate

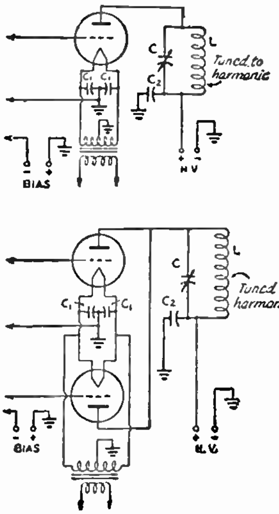


Fig. 427 — Frequency-multiplying circuits. A is for triodes, used either singly or in parallel. The push-push doubler is shown at B. Any type of coupling may be used between the grid circuit and the driver. C_1 should be 0.01 μfd. or larger; C_2, 0.001 μfd. or larger.

and grid circuits, one of them should be short-circuited to determine if the oscillation is at a low frequency; if so, it may be eliminated by the methods outlined above. If the test indicates that the parasitic is not a low-frequency oscillation, the grid trap described above should be tried for the v.h.f. type. The type which occurs near the operating frequency will not exist unless the plate and grid tank coils are both tapped, hence may be eliminated from consideration if this is not the case in the circuit used. When such an oscillation is present its existence can be detected by moving the grid tap to include the whole tank circuit, whereupon the oscillation will cease.

Some indication of the frequency of the parasitic can be obtained from the color of the glow in the neon bulb. Usually it will be yellowish with low-frequency oscillations and violet with v.h.f. oscillations.

If the amplifier is stable under the conditions described above, excitation should be applied and then removed to ascertain if a self-sustaining oscillation is set up with excitation. If the plate current does not return to the previous value when the excitation is cut off, the same tests should be applied to determine the parasitic frequency.

As a final test, the transmitter should be put on the air and a near-by receiver tuned over as wide a frequency range as possible, to locate

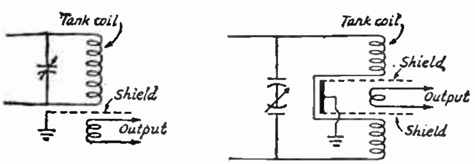


Fig. 426 — Methods of using Faraday shields. Two are required with a push-pull or balanced tank circuit.

any off-frequency signals associated with the radiation. Parasitics usually can be recognized by their poor stability as contrasted to the normal harmonics of the signal, which will have the same stability as the fundamental signal as well as the usual harmonic relationship. Harmonics should be quite weak compared to the output at the fundamental frequency, whereas parasitic oscillations may have considerable strength.

4-11 Frequency Multiplication

Circuits — A frequency multiplier is an amplifier having its plate tank circuit tuned to a multiple (harmonic) of the frequency applied to its grid. The difference between a straight amplifier (§ 4-1) and a frequency multiplier is in the way in which it is operated, rather than in the circuit. However, since the grid and plate tank circuits are tuned to different frequencies a triode frequency multiplier will not self-oscillate, hence does not need neutralization. A typical circuit arrangement is shown in Fig. 427-A. For screen-grid multipliers, the circuit is the same as in Fig. 412-A. Under usual conditions the plate efficiency of a frequency multiplier drops off rapidly with an increase in the number of times the frequency is multiplied. For this reason most multipliers are used as *frequency doublers*, giving second harmonic output.

A special circuit for frequency doubling ("push-push" doubler) is shown in Fig. 427-B. The grids of the tubes are in push-pull and the plates in parallel, thus the plate tank receives two pulses of plate current for each cycle of excitation frequency. The circuit is similar to that of a full-wave rectifier (§ 8-3), where the output ripple frequency is twice the applied frequency.

Push-pull amplifiers are suitable for frequency multiplication at odd harmonics, particularly the third, but they are unsuited to even-harmonic multiplication because the even harmonics are largely balanced out in the push-pull tank circuit (§ 3-3).

Operating conditions and circuit constants — To obtain good efficiency the operating angle at the harmonic frequency must be 180 degrees or less, preferably in the vicinity of 150–120 degrees (§ 4-8). In a doubler, this means that plate current should flow during only half this angle of fundamental frequency. Consequently the r.f. grid voltage, operating bias, and grid driving power must be increased considerably beyond the values obtaining for normal Class-C amplification. For comparable plate efficiency the bias will ordinarily be four to five times the normal Class-C bias, and the r.f. grid voltage must be considerably larger to drive the tube to the same peak plate current. Since the plate and grid current pulses under these conditions have the same peak amplitudes but only half the time duration as in a straight amplifier, the average d.c. values should be one-half those for normal Class-C operation. That is, a tube operated in this way will have the same plate efficiency as a Class-C amplifier but can be operated at only half the plate input, so that the output power also is halved. The driving power required usually is about twice that necessary with straight-through amplification to obtain the same plate efficiency.

Greater output can be secured by using a larger operating angle (lower grid bias) or a lower plate load resistance, to increase the plate current; but this is accompanied by a decrease in efficiency. Since operation of the tube as described in the preceding paragraph is below its maximum plate dissipation rating, the decreased efficiency usually can be tolerated in the interests of securing more power output. In practice, an efficiency of 40 to 50 per cent is about average.

The tank circuit should have reasonably high Q (12 is satisfactory) to give good output voltage regulation (§ 4-9), since a plate-current pulse occurs only once for every two cycles of the output frequency. A low-Q circuit (high L/C ratio) is helpful chiefly when the operating angle is greater than 180 degrees at the second harmonic. Such a tank circuit will have relatively high impedance to the fundamental-frequency component of plate current which is present with large operating angles, and thus will aid in reducing the average d.c. plate current.

The grid impedance of a frequency multiplier is considerably higher than that of a straight-through amplifier, because of the high bias voltage. The average impedance can be calculated as previously described (§ 4-8). The L/C ratio of the grid tank circuit may be higher, therefore, for a given Q. Often it is advantageous to use a fairly high ratio, since a large r.f. voltage must be developed between grid and cathode. However, it must not be made too high (Q too low) to permit adequate coupling between the grid tank circuit and the preceding driver stage.

It may prove necessary to step up the driver output voltage to obtain sufficient r.f. grid voltage for the doubler; this can be done by tapping the driver plate on its tank circuit, when capacity coupling is used, or by similar tapping down or the use of a higher C/L ratio in the driver plate tank when the stages are link-coupled (§ 4-6).

Tubes for frequency multiplication — There is no essential difference between tubes of various characteristics in their performance as frequency doublers. Tubes having high amplification factors will require somewhat less bias for equivalent operation but the grid driving power needed is almost independent of the μ, assuming tubes of otherwise similar construction and characteristics. Pentodes and tetrodes will, as in normal amplifier operation, require less driving power than triodes for efficient doubling, although more power will be needed than for straight amplification.

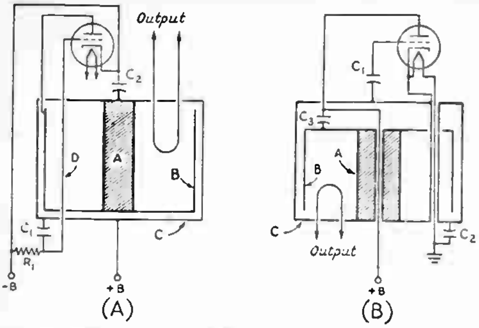


Fig. 428 — High-*Q* "pot"-type lumped-constant tank circuit as used in v.h.f. oscillators. The tank, shown in cross-section, is made of concentric closed cylinders.

4-12 Very-High-Frequency Oscillators

High-Q circuits with lumped constants — To obtain reasonably high effective *Q* when a low resistance is connected across the tank circuit, it is necessary to use a high *C/L* ratio and a tank of inherently high *Q* (§ 2-10). At low frequencies the inherent *Q* of any well-designed circuit will be high enough so that it may be neglected in comparison to the effective *Q* when loaded, so that no special precautions have to be taken with respect to the resistance of coils and condensers. At the very-high frequencies these internal resistances are too large to be ignored, however.

Reduction of the *L/C* ratio will not increase the effective *Q* unless the internal resistance of the tank can be made very small. This resistance can be reduced by use of large conducting surfaces and elimination of radiation. In such cases special lumped-constant tank circuits (§ 2-12) are used. The oscillator shown in Fig. 428-A uses a "pot"-type tank in the tickler circuit (§ 3-7), with the feed-back coil in the grid circuit; this inductance is the wire *D* in the diagram. Output is taken from the tank by means of a hairpin coupling loop.

Fig. 428-B corresponds to the shunt-fed Hartley circuit. Such a tank also may be used in the ultraudion circuit. A variable condenser may be connected across the tank for tuning, although the *Q* may be reduced if a considerable portion of the tank r.f. current flows through it.

Linear Circuits — A quarter-wave or half-wave line, either of the parallel-conductor open type or of the coaxial type, is equivalent to a resonant circuit (§ 2-12) and can be used as the tank circuit (§ 3-7) in an oscillator.

The resonant line is usually constructed of thin-walled copper tubing, rather than wire, since this reduces resistance and provides a mechanically stable circuit, particularly at the lower frequencies. At frequencies above 100 Mc. flat copper strip conductor of equivalent cross-section may be used for parallel-line circuits with comparable efficiency. Frequency can be changed by moving a shorting bar or condenser to change the effective line length, or by reducing its length and loading it to resonance by connecting a low-capacity variable condenser across the open end of the line. The added capacity makes it necessary to shorten the line considerably for a given frequency. This, together with the additional loss in the condenser, causes a decrease in *Q*. These effects will be less if the condenser is connected down on the line. Tapping down also gives greater bandspread effect (§ 7-7).

At very high frequencies an adequate ground connection for the cathode circuit becomes a problem because of the inductance of the cathode lead. Special tubes are available

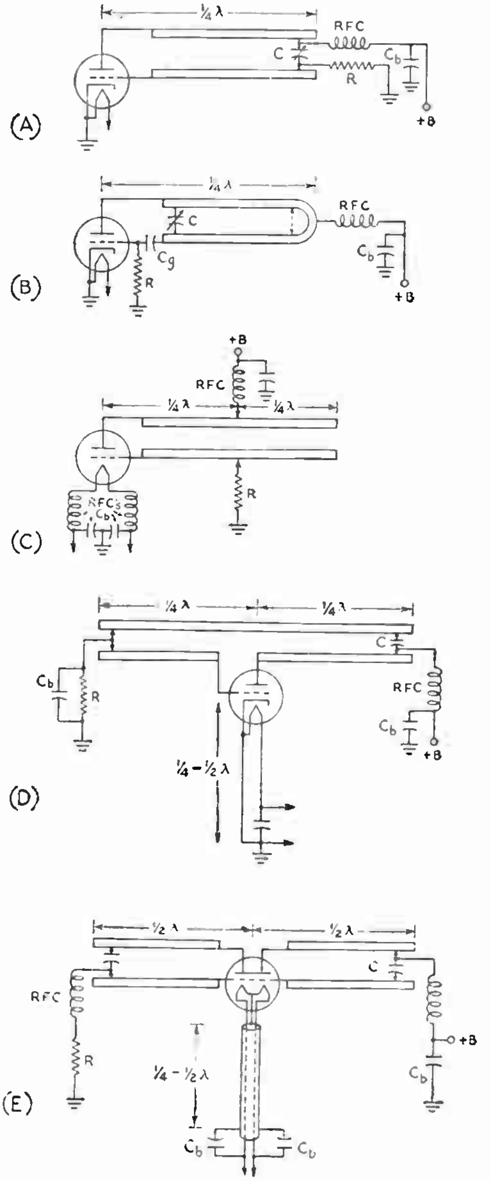


Fig. 429 — Typical single-tube parallel-line oscillators. Constants and applications are discussed in the text.

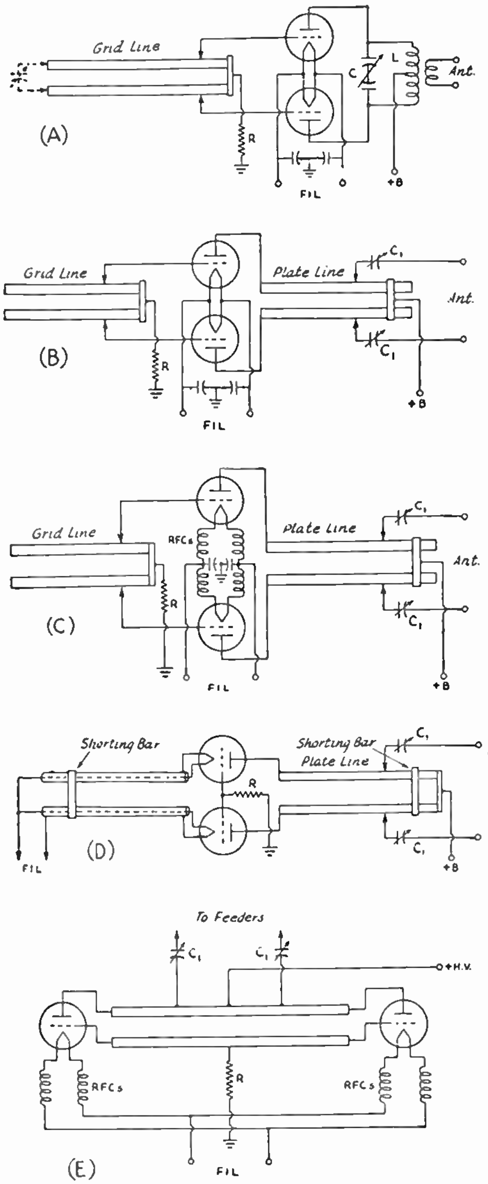


Fig. 430 — Push-pull parallel-line oscillator circuits.

with two or three cathode leads (§ 3-6); connected in parallel, these reduce the effective inductance. With ordinary tubes, coils may be inserted in the filament circuit to compensate for the effects of the internal inductance. The effective length of the filament circuit should be one-half wavelength, to bring the cathode filament to the same potential as the shorted ends of the tank lines. The added inductance required must be determined by experiment, the coils being adjusted for optimum stability and power output.

Another method is to use a tuned line in the filament circuit, adjusting its length so that the electrical length of the line plus that of the filament is one-half wavelength. A convenient arrangement is the use of a coaxial (or trough) line with an initial length of about 3/8 wavelength. A shorting disc in the form of a movable plunger equipped with an extension handle may be provided for ease of adjustment. With filament-type tubes one such line will be required for each filament lead. In the case of cathode-type tubes only one line is necessary, the cathode and one side of the filament being connected to the outer conductor and the other filament connection being made by an insulated lead running through a hollow-tubing inner conductor. The return lead should be by-passed where it emerges from the line.

The antenna or other load may be connected through blocking condensers direct to the line (the correct point being determined experimentally). Alternatively, a hair-pin coupling link or, in the case of an oscillator-amplifier system, direct inductive coupling to the grid line of the amplifier may be used.

For highest-frequency operation separate lines must be used for each electrode — grid, plate and cathode. This places all of the interelectrode capacities in series, reducing the loading effect. Still higher frequencies can be reached by using double-lead tubes (Fig. 429-E), in which case the leads form an integral part of the line and the interelectrode capacities are divided between the two quarter-wave sections.

Parallel-line oscillators — Typical parallel-line oscillator circuits are shown in Fig. 429. In A, a shorting condenser (which may be either a fixed blocking condenser or a small variable which will provide a limited tuning range) is used to bridge the line at the voltage node; the frequency can also be changed by sliding the shorting condenser along the line.

The circuit at B eliminates the need for a blocking condenser at the voltage node, where the r.f. current reaches its maximum value. An r.f. choke may be inserted between the grid and the associated grid resistor, *R*. This circuit also can be resonated either by a variable condenser, *C*, or by a sliding bar as indicated by the dashed line.

Fig. 429-C uses a half-wave open-ended line. The grid and plate feed connections are made at nodal points on the line. As indicated on the diagram, these do not occur at the physical center of the line because of the loading effect of the tube. In practice, the position of the taps, as well as the over-all length of the line, are adjusted to obtain maximum grid current. Using this circuit, a 955 acorn or a 9002 can be made to oscillate up to 600 or 700 Mc.

Fig. 429-D is a variation of the above preferable for use with tubes having grid and plate terminals at opposite ends of the envelope. The circuit of Fig. 429-E is most useful with double-lead tubes. To attain high output at the maximum operating frequency, the desirable arrangement is to use two or more double-lead tubes, each in a circuit such as this, with the lines connected end to end.

Push-pull parallel-line oscillators — It is often advantageous to use push-pull oscillator circuits at the very-high frequencies, not only as a means to secure more power output but also for better circuit symmetry. In addition, the interelectrode capacities of the push-pull tubes are in series across the point of connection to the tank circuit, hence have less capacity-loading effect than is experienced with a single tube.

Fig. 430 shows typical push-pull circuits of this type. Figs. 430-A, -B and -C all employ the same circuit — the t.p.t.g. type (§ 3-7). The grid line is usually operated as the frequency-controlling circuit, since it is not associated with the load and hence its Q can be kept high. The same adjustment considerations apply as in the case of single-tube oscillators. Grid taps in particular should be tapped down as far as possible, to improve the frequency stability.

In Fig. 430-A, a conventional coil-and-condenser tank is used in the plate circuit where the lower Q does not have so great an effect on frequency stability. For maximum efficiency the use of a linear output circuit is desirable at the higher frequencies, however. This is shown at B, and at C with isolating r.f. chokes in the filament circuit.

Fig. 430-D shows a push-pull oscillator having tuned plate and cathode lines, the cathode circuit being tuned with a quarter-wave line which controls excitation and, to some extent, tuning. The grids are connected together and grounded through the grid leak, R_1; ordinarily no by-pass condenser is needed across R_1. This circuit gives good power output at very-high frequencies, but is not especially stable unless the plates are tapped down on the plate tank circuit to avoid too great a reduction in Q. Tapping on the cathode line is not feasible for mechanical reasons. With ordinary tubes this oscillator is capable of higher-frequency operation than the conventional t.g.t.p. type, and it has been found particularly useful on 224 Mc.

The symmetrical circuit at E is preferable above 200 Mc. Coaxial or equivalent lines may be used instead of r.f. chokes in the filament circuits for ultrahigh-frequency operation. With this modification, and (assuming the use of double-lead tubes) by the addition of quarter-wave sections at each end, this circuit may be considered equivalent to the center section of a double linear oscillator as discussed in connection with Fig. 429-E.

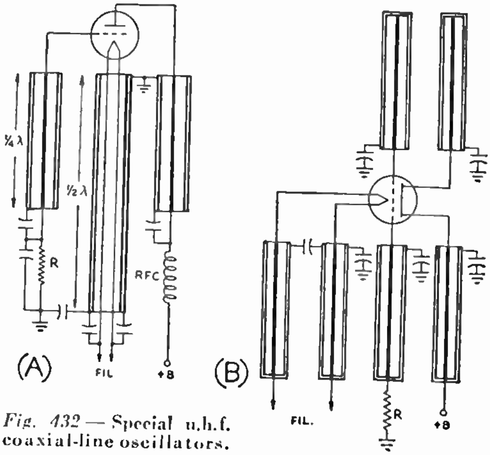


Fig. 432 — Special u.h.f. coaxial-line oscillators.

Coaxial-line circuits — At frequencies in the neighborhood of 300 Mc. the radiation loss (§ 2-12) from open lines greatly reduces the Q, because the conductor spacing unavoidably becomes an appreciable fraction of a wavelength. Consequently, these frequencies and higher coaxial lines, in which the field is confined inside the line so that radiation is negligible, are used. A further advantage is that the outside of the line is "cold"; that is, no r.f. potentials develop between points on the outer surface. While the coaxial line is also advantageous at lower frequencies, it is more complicated to construct and adjust than parallel lines.

For ease of construction, the coaxial line sometimes is modified into a "trough," in which the cross-section of the outer conductor is in the shape of a square U, one side being left open for tapping and adjustment of the inner conductor. Some radiation takes place with this type of construction, although not so much as with open lines.

The conventional coaxial-line oscillator circuits shown in Fig. 431 illustrate the application of two basic circuits — the Hartley and the t.g.t.p. — to both cathode-type and filamentary tubes. The tube loads the line, as previously described; hence the actual length is always shorter than a quarter wavelength. The length can be adjusted by a short-circuiting sliding plunger, a close-fitting low-resistance contact being necessary to avoid losses. The inner conductor may also have a short tight-

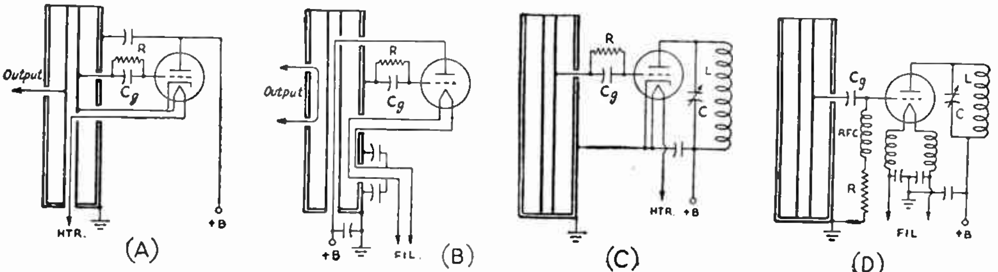


Fig. 431 — Single-tube v.h.f. coaxial-line oscillators. A and B use Hartley circuits; C and D are t.g.t.p. equivalents.

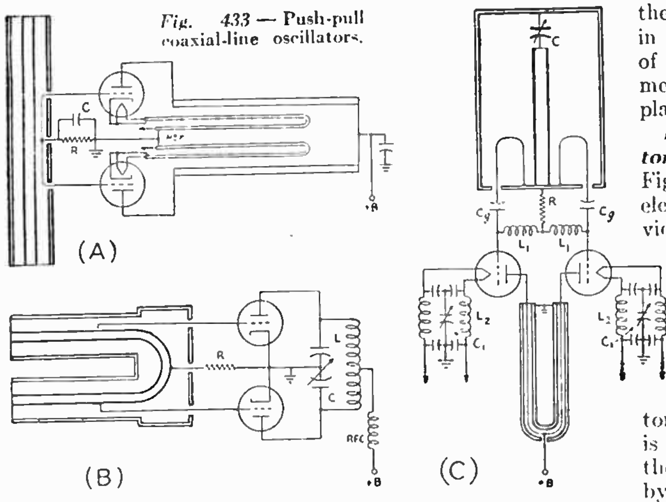


Fig. 433 — Push-pull coaxial-line oscillators.

fitting extension tube which is slid in or out to change the effective conductor length.

The t.g.t.p. circuits are somewhat easier to adjust and load as well as to construct, but are not as satisfactory from the standpoint of frequency stability because of reaction on the frequency-controlling grid line by the tuning of the output circuit. The grid tap should be as far down on the line as will permit reliable oscillation under load. Under some conditions the addition of a small adjustable feed-back capacity between grid and plate not only permits a lower tap location but also increases the upper frequency limit obtainable by advancing the phase of the grid excitation to compensate partially for transit-time lag in the tube.

In the Hartley circuit at A, an output tap is provided on the inner conductor. At B inductive output coupling by means of a half-turn "hairpin" is shown; loading can be changed to some extent by varying its position.

Fig. 432 shows two types of coaxial-line oscillator circuits designed particularly for operation near the upper frequency limits for negative-grid tubes. The circuit at A, with quarter-wave grid and plate lines and a half-wave filament line, is convenient for use with single-lead tubes such as the 955 and 316-A. With the three lines arranged in the form of a triangle, so that their inner conductors attach directly to the tube terminals for minimum lead length, this oscillator will function satisfactorily up to 700–800 Mc.

The circuit of Fig. 432-B is designed to take maximum advantage of the u.h.f. capabilities of double-lead and ring-electrode tube types. Interelectrode capacities are divided between each pair of grid and plate lines, and separate parallel-resonant filament lines complete the isolation. Frequencies as high as 1500–1700 Mc. have been attained with this arrangement.

The by-pass condensers shown in the two circuits of Fig. 432 are made of copper plates insulated by sheet mica. Flanges soldered to the ends of the outer conductor in each line constitute one plate of the condenser; a grounded metal sheet serves as the other plate.

Push-pull coaxial-line oscillators— The push-pull circuits of Fig. 433 employ the same basic elements as the arrangements previously described. At A, a half-wave open-ended line is used in the grid circuit, the grids of the tubes being "tapped" down on the line by coupling them inductively through a small balanced loop running inside the outer conductor. A conventional parallel line is used in the plate circuit, with the cathodes balanced to ground by means of closed half-wave lines. The cathode lines may be small-diameter copper tubing, folded to conserve space, through which rubber-insulated wire is run for the return circuit. These lines may be shielded from the plate line by running them underneath the chassis or separated by a shielding partition.

A folded half-wave grid line is used at B. The copper-tubing inner conductor is bent into the shape of a U. The outer conductor may be either a square-section double trough of sheet copper or two short sections of pipe soldered to a rectangular box of sheet copper which forms the "closed" end. Where even more compact construction is required, the dimensions of the grid line may be still further reduced by using sections of folded coaxial line (§ 2-12). A conventional coil-and-condenser output circuit is shown; at the comparatively low frequencies where this type of construction would be advantageous in the interest of compactness, such an output circuit should be satisfactory.

The arrangement at C has certain modifications which make it particularly suitable for use with higher-powered tubes. The quarter-wave capacity-loaded coaxial line in the grid circuit is of relatively large dimensions and consequently has high Q. Coupling to the tube grids, which is made very loose to preserve the Q of the line, is by means of twin hairpin loops. The inductance of the shunt choke coils, L_1, is adjusted for maximum grid current.

To minimize radiation loss and preserve circuit symmetry, a coaxial line is used in the plate tank circuit. If desired this line may be tuned by a balanced split-stator condenser of the type which has the rotor connection at the center, connected across the plate terminals.

Parallel resonant circuits in the filament leads, tuned to resonance at the operating frequency by the variable condensers, C_1, isolate the filament from ground. The fixed by-pass condensers must have low reactance at the operating frequency. The filament coils, which are in parallel for r.f., are of copper tubing.

Chapter Five

Radiotelephony

5-1 Modulation

The carrier — The steady radio-frequency power generated by transmitting circuits cannot alone result in the transmission of an intelligible message to a receiving point. The continuous wave from the transmitter itself serves only as a "carrier" for the message; the intelligence is conveyed by *modulation* (a change) of the carrier. In radiotelephony, this modulation reproduces electrically the sounds it is intended to convey in a form which can be correctly interpreted or demodulated at the receiving end.

Sound and alternating currents — Sounds are caused by vibrations of air particles. The pitch of the sound depends upon the rate of vibration; the more rapid the vibration, the higher the pitch. Most sounds consist of complex combinations of vibrations of differing rates or frequencies; the human voice, for instance, generates frequencies from about 100 cycles per second to several thousand per second. The problem of transmitting speech by radio, therefore, is one of varying the r.f. carrier in a way which corresponds to the air-particle vibrations. The first step in doing this is to change the sound vibrations into alternating electrical currents of the same frequency and relative intensity; the electromechanical device which achieves this translation is the *microphone*. These audio-frequency currents then may be amplified and used to vary or modulate the normally steady r.f. output of the transmitter.

Methods of modulation — The carrier may be made to vary in accordance with the speech current by using the current to change the phase (§ 2-7), frequency or amplitude of the carrier. *Amplitude modulation* of a constant-frequency carrier is by far the most common system, and is used exclusively on all frequencies below the very-high-frequency region (§ 2-7). *Frequency modulation* of a constant-amplitude carrier, which has special characteristics which make its use desirable under certain conditions, is used to a considerable extent on the very-high frequencies. *Phase modulation*, which is closely related to frequency modulation, has had little or no direct application in practical communication.

Other specialized varieties of modulation, developed for other applications of radio transmission, have been proposed for voice communication. Thus far none of these has achieved practical utilization, however.

5-2 Amplitude Modulation

Carrier requirements — For proper amplitude modulation, the carrier should be completely free from inherent amplitude variations such as might be caused by insufficient filtering of a rectified-a.c. power supply (§ 8-4). It is also essential that the carrier *frequency* be entirely unaffected by the application of modulation. If modulating the amplitude of the carrier also causes a change in the carrier frequency the signal wobbles back and forth with the modulation, introducing distortion and widening the channel taken by the signal. This causes unnecessary interference to other transmissions. In practice, this undesirable frequency modulation is prevented by applying the modulation to an r.f. amplifier stage which is isolated from the frequency-controlling oscillator by a "buffer" amplifier. Amplitude modulation of an oscillator almost always is accompanied by frequency modulation. Under existing regulations it is permitted, therefore, only on frequencies above 112 Mc., because the

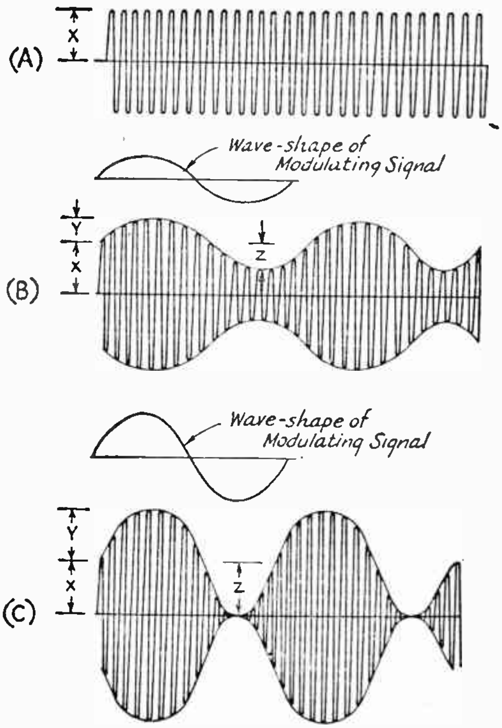


Fig. 501 — Graphical representation of (A) carrier unmodulated, (B) modulated 50%, (C) modulated 100%.

problem of interference is less acute in this region than on lower frequencies.

Percentage of modulation — In the amplitude-modulation system the audible output at the receiver depends entirely upon the amount of variation — termed *depth of modulation* — in the carrier wave, and not upon the strength of the carrier alone. It is desirable therefore to obtain the largest permissible variations in the carrier wave. This condition is reached when the carrier amplitude during modulation is at times reduced to zero and at other times increased to twice its unmodulated value. Such a wave is said to be *fully modulated*, or *100 per cent modulated*. Any desired degree of modulation can be expressed as a percentage, using the unmodulated carrier as a base. Fig. 501 shows, at A, an unmodulated carrier wave; at B, the same wave modulated 50 per cent, and at C, the wave with 100 per cent modulation, using a sine-wave (§ 2-7) modulating signal. The outline of the modulated r.f. wave is called the *modulation envelope*.

The percentage modulation can be found by dividing either *Y* or *Z* by *X* and multiplying the result by 100. If the modulating signal is not symmetrical, the larger of the two (*Y* or *Z*) should be used.

Power in modulated wave — The amplitude values correspond to current or voltage, so that the drawings may be taken to represent instantaneous values of either. Since power varies as the square of either the current or voltage (so long as the resistance in the circuit is unchanged), at the peak of the modulation up-swing the instantaneous power in the wave of Fig. 501-C is four times the unmodulated carrier power. At the peak of the down-swing the power is zero, since the amplitude is zero. With a sine-wave modulating signal, the *average* power in a 100 per cent modulated wave is one and one-half times the unmodulated carrier power; that is, the power output of the transmitter increases 50 per cent with 100 per cent modulation.

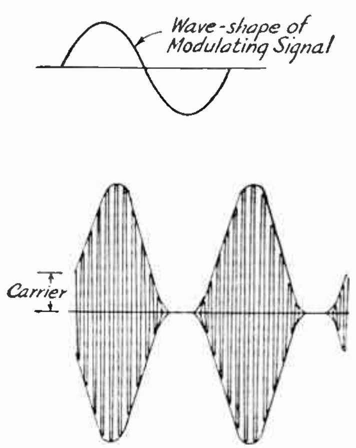


Fig. 502 — An overmodulated r.f. carrier wave.

Linearity — Up to the limit of 100 per cent modulation, the amplitude of the carrier should follow faithfully the amplitude variations of the modulating signal. When the modulated r.f. amplifier is incapable of meeting this condition, it is said to be *non-linear*. The amplifier may not, for instance, be capable of quadrupling its power output at the peak of 100 per cent modulation. A non-linear modulated amplifier causes distortion of the modulation envelope.

Modulation characteristic — A graph showing the relationship between r.f. amplitude and instantaneous modulating voltage is called the *modulation characteristic* of the modulated amplifier. This graph should be a straight line (linear) between the limits of zero and twice carrier amplitude. Curvature of the line between these limits indicates non-linearity in the amplifier.

Modulation capability — The *modulation capability* of the transmitter is the maximum percentage of modulation that is possible without objectionable distortion from non-linearity. The maximum capability is, of course, 100 per cent. The modulation capability should be as high as possible, so that the most effective signal can be transmitted for a given carrier power.

Overmodulation — If the carrier is modulated more than 100 per cent, a condition such as is shown in Fig. 502 occurs. Not only does the peak amplitude exceed twice the carrier amplitude, but actually there may be a considerable period during which the output is entirely cut off. The modulated wave is therefore distorted (§ 3-3), with the result that harmonics of the audio modulating frequency appear. The carrier should never be modulated more than 100 per cent.

Sidebands — The combining of the audio frequency with the r.f. carrier is essentially a heterodyne process, and therefore gives rise to beat frequencies equal to the sum and difference of the a.f. and r.f. frequencies involved (§ 2-13). Therefore, for each audio frequency appearing in the modulating signal, two new radio frequencies appear, one equal to the carrier frequency plus the audio frequency, the other equal to the carrier minus the audio frequency. These new frequencies are called *side frequencies*, since they appear on each side of the carrier, and the groups of side frequencies representing a band or group of modulation frequencies are called *sidebands*. Hence a modulated signal occupies a group of radio frequencies, or *channel*, rather than a single frequency as in the case of the unmodulated carrier. The *channel width* is twice the highest modulation frequency.

To accommodate the largest number of transmitters in a given part of the r.f. spectrum it is apparent that the channel width should be as small as possible. On the other hand it is necessary, for speech transmission of reasonably good quality, to use modulating frequencies up to a minimum of about 3000 or 4000 cycles. This calls for a channel width of 6 to 8 kilocycles.

Spurious sidebands — Besides the normal sidebands required by speech frequencies, unwanted sidebands may be generated by the transmitter. These usually lie outside the normally required channel, and hence cause it to be wider without increasing the useful modulation. By increasing the channel width, these spurious sidebands cause unnecessary interference to other transmitters. The quality of transmission also is adversely affected when spurious sidebands are generated.

The chief causes of spurious sidebands are harmonic distortion in the audio system, overmodulation, unnecessary frequency modulation, and lack of linearity in the modulated r.f. system.

Types of amplitude modulation — The most widely used type of amplitude-modulation system is that in which the modulating signal is applied in the plate circuit of a radio-frequency power amplifier (*plate modulation*). In a second type the audio signal is applied to a control-grid (*grid-bias modulation*). A third system, involving variation of both plate and grid voltages, is called *cathode modulation*.

5-3 Plate Modulation

Transformer coupling — In Fig. 503 is shown the most widely used system of plate modulation. A balanced (push-pull Class-A, Class-AB or Class-B) modulator is transformer-coupled to the plate circuit of the modulated r.f. amplifier. The audio-frequency power generated in the modulator plate circuit 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.

Modulator power — The average power output of the modulated stage must increase 50 per cent for 100 per cent modulation (§ 5-2), so that the modulator must supply to the modulated r.f. stage audio power equal to 50 per cent of the d.c. plate input. For example, if the d.c. plate power input to the r.f. stage is 100 watts, the sine-wave audio power output of the modulator must be 50 watts.

Modulating impedance; linearity — The *modulating impedance*, or load resistance presented to the modulator by the modulated r.f. amplifier, is equal to

$$\frac{E_b}{I_p} \times 1000$$

where E_b is the d.c. plate voltage and I_p the d.c. plate current in milliamperes, both measured without modulation.

Since the power output of the r.f. amplifier must vary as the square of the plate voltage (the r.f. voltage must be proportional to the applied plate voltage) in order for the modulation to be linear, the amplifier must operate under Class-C conditions (§ 3-4). The linearity then depends upon having sufficient grid excitation and proper bias, and upon the adjustment of circuit constants to the proper values (§ 4-8).

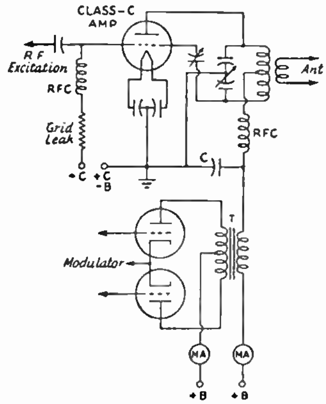


Fig. 503 — **Plate modulation of a Class-C r.f. amplifier. The r.f. plate by-pass condenser, *C*, in the amplifier stage should have high reactance at audio frequencies. A capacity of 0.002 μfd. or less usually is satisfactory.**

Power in speech waves — The complex waveform of a speech sound translated into alternating current does not contain as much power, on the average, as there is in a pure tone or sine wave of the same peak (§ 2-7) amplitude. That is, with speech waveforms the ratio of peak to average amplitude is higher than in the sine wave. For this reason, the previous statement that the power output of the transmitter increases 50 per cent with 100 per cent modulation, while true for tone modulation, is not true for speech. On the average, speech waveforms will contain only about half as much power as a sine wave, both having the same peak amplitude. The average power output of the transmitter therefore increases only about 25 per cent with 100 per cent speech modulation. However, the *instantaneous* power output must quadruple on the peak of 100 per cent modulation (§ 5-2) regardless of the modulating waveform. Therefore, the peak output power capacity of the transmitter must be the same for any type of modulating signal.

Adjustment of plate-modulated amplifiers — The general operating conditions for Class-C operation have been described (§ 3-4, 4-8). 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 (§ 4-8) of about 120 degrees at carrier plate voltage, and the excitation should be sufficient to maintain the plate efficiency constant when the plate voltage is varied over the range from zero to twice the d.c. plate voltage applied to the amplifier. For best linearity, the grid bias should be obtained partly from a fixed source of about the cut-off value, supplemented by grid-leak bias to supply the remainder of the required operating bias.

The maximum permissible d.c. plate power input for 100 per cent modulation is twice the sine-wave audio-frequency power output of the modulator. This input is obtained by varying the loading on the amplifier (keeping its tank circuit tuned to resonance) until the product of d.c. plate voltage and plate current is the desired power. The modulating impedance under these conditions will be the proper value for the modulator, if the proper output-transformer turns ratio (§ 2-9) is used.

Neutralization, when triodes are used, should be as nearly perfect as possible, since regeneration may cause non-linearity. The amplifier also should be free from parasitic oscillations (§ 4-10).

Although the *effective* value (§ 2-7) of power input increases with modulation, as described above, the *average* plate input to a plate-modulated amplifier does not change, since each increase in plate voltage and plate current is balanced by an equivalent decrease in voltage and current. Consequently, the d.c. plate current to a properly modulated amplifier is always constant, with or without modulation.

Screen-grid amplifiers — Screen-grid tubes of the pentode or beam tetrode type can be used as Class-C plate-modulated amplifiers provided the modulation is applied to both the plate and screen grid. The method of feeding the screen grid with the necessary d.c. and modulation voltage is shown in Fig. 504. 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.

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 is the power-input figure which is used as the basis for determining the audio power required from the modulator.

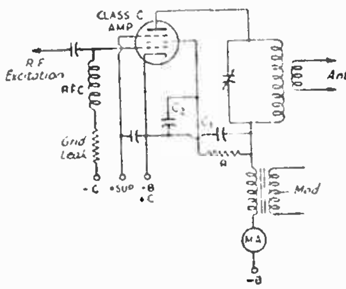


Fig. 504 — Plate and screen modulation of a Class-C r.f. amplifier using a pentode tube. The plate and screen r.f. by-pass condensers, C_1 and C_2, should have high reactance at all audio frequencies (0.002 μfd. or less).

Choke coupling — In Fig. 505 is shown the circuit of the choke-coupled system of plate modulation. The plate power for the modulator tube and modulated amplifier is furnished from a common source through the modulation choke, *L*, which has high impedance for audio frequencies. The modulator operates as a power amplifier with the plate circuit of the r.f. amplifier as its load, the audio output of the modulator being superimposed on the d.c. power supplied to the amplifier. For 100 per cent modulation, the audio voltage applied to the r.f. amplifier plate circuit across the choke, *L*, must have a peak value equal to the d.c. voltage on the modulated amplifier. To obtain this without distortion the r.f. amplifier must be operated at a d.c. plate voltage less than the

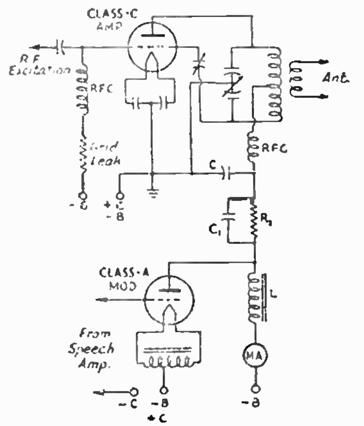


Fig. 505 — Choke-coupled plate modulation.

modulator plate voltage, the extent of the voltage difference being determined by the type of modulator tube used. The necessary drop in voltage is provided by the resistor, R_1, which is by-passed for audio frequencies by the by-pass condenser, C_1.

This type of modulation seldom is used except in very low-power portable sets, because a single-tube Class-A (§ 3-4) modulator is required. The output of a Class-A modulator is very low compared to that obtainable from a pair of tubes of the same size operated Class B, hence only a small amount of r.f. power can be modulated.

Absorption modulation — Absorption or "loss" modulation, in its basic form the oldest and simplest method of all, recently has been revived for wide-band modulation (such as television) at ultrahigh frequencies. In the system shown in Fig. 506, the modulating tubes are connected to the antenna feed line through a quarter-wave stub line, located a quarter-wavelength from the transmitter tank circuit. With no modulation (i.e., no conduc-

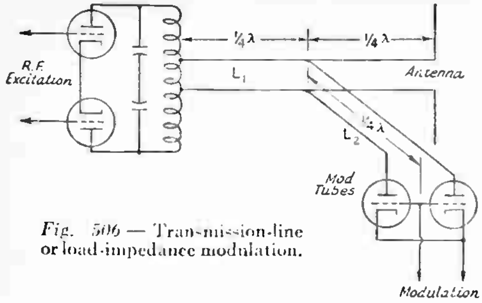


Fig. 506 — Transmission-line or load-impedance modulation.

tion through the modulating tubes) the stub appears as a short circuit across the line and little or no power reaches the antenna. When modulating voltage is applied to the grids of the modulator tubes, however, their conductance serves to increase the effective impedance of the quarter-wave shunt, permitting a proportionate amount of energy to reach the antenna. At maximum modulation the stub approaches an open circuit, allowing maximum r.f. output to the antenna.

ℂ 5-4 Grid-Bias Modulation

Circuit — Fig. 507 is the diagram of a typical arrangement for grid-bias modulation. In this system, the secondary of an audio-frequency output transformer, the primary of which is connected in the plate circuit of the modulator tube, is connected in series with the grid-bias supply for the modulated amplifier. The audio voltage thus introduced varies the grid bias, and thus the power output of the r.f. stage, when suitable operating conditions are chosen. The r.f. stage is operated as a Class-C amplifier, with the d.c. grid bias considerably beyond cut-off.

Operating principles — In this system the plate voltage is constant, and the increase in power output with modulation is obtained by making the plate current and plate efficiency vary with the modulating signal. For 100 per cent modulation, both plate current and efficiency must, at the peak of the modulation upswing, be twice their carrier values, so that the peak power will be four times the carrier power. Since the peak efficiency in practicable circuits is of the order of 70 to 80 per cent, the carrier efficiency ordinarily cannot exceed about 35 to 40 per cent. For a given r.f. tube, the carrier output is about one-fourth the power obtainable from the same tube plate-modulated. Grid bias, r.f. excitation, plate loading and the audio voltage in series with the grid must be adjusted to give a linear modulation characteristic.

Modulator power — Since the increase in average carrier power with modulation is secured by varying the plate efficiency and d.c. plate input of the amplifier, the modulator need supply only such power losses as may be occasioned by connecting it in the grid circuit. These are quite small, hence a modulator capable of only a few watts output will suffice for transmitters of considerable power. Since the load on the modulator varies over the a.f. cycle as the rectified grid current of the modulated amplifier changes, the modulator should have good voltage regulation (§ 5-6).

Grid-bias source — The change in bias voltage with modulation causes the rectified grid current of the amplifier also to vary, the r.f. excitation being fixed. If the bias source has appreciable resistance, the change in grid current also will cause a change in bias in a direction opposite to that caused by the modulation. It is necessary, therefore, to use a grid-bias source having low resistance, so that these bias variations will be negligible. Battery bias is satisfactory. If a rectified a.c. bias supply is used, the type having regulated output (§ 8-9) should be chosen. Grid-leak bias for a grid-modulated amplifier is unsatisfactory, and its use should not be attempted.

Driver regulation — The load on the driving stage varies with modulation, and a linear modulation characteristic may not be obtained if the r.f. voltage from the driver does not stay constant with changes in load. Driver regulation (ability to maintain constant output voltage with changes in load) may be improved by using a driving stage having two or three times the power output necessary for excitation of the amplifier (this is somewhat less than the power required for ordinary Class-C operation), and by dissipating the extra power in a constant load such as a resistor. The load variations are thereby reduced in proportion to the total load.

Adjustment of grid-bias modulated amplifiers — This type of amplifier should be adjusted with the aid of an oscilloscope, to obtain optimum operating conditions. The oscilloscope should be connected as described in § 5-10, the wedge pattern being preferable. A tone source for modulating the transmitter will be convenient. The fixed grid bias should be two or three times the cut-off value (§ 3-2). The d.c. input to the amplifier, assuming 33

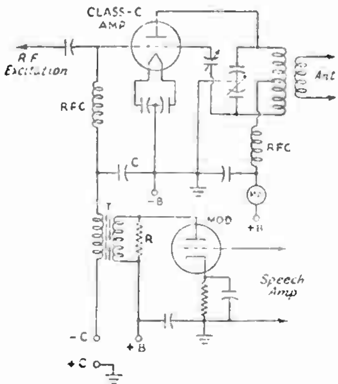


Fig. 507 — Grid-bias modulation of a Class-C amplifier. The r.f. grid by-pass condenser, *C*, should have high reactance at audio frequencies (0.002 μfd. or less).

per cent carrier efficiency, will be 1½ times the plate dissipation rating of the tube or tubes used in the modulated stage. The plate current for this input (in milliamperes, 1000 P/E, where P is the power and E the d.c. plate voltage) must be determined. Apply r.f. excitation

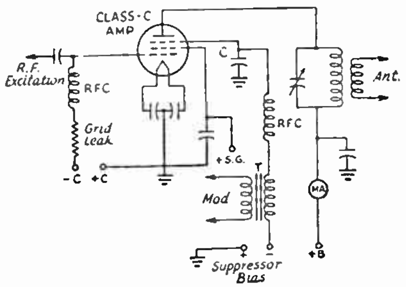


Fig. 508 — Suppressor-grid modulation of an r.f. amplifier using a pentode-type tube. The suppressor-grid r.f. by-pass condenser, C, should be 0.002 μfd. or less

and, without modulation, adjust the plate loading to give the required plate current (keeping the plate tank circuit tuned to resonance). Next, apply modulation and increase the modulating signal until the modulation characteristic shows curvature (§ 5-10). This probably will occur well below 100 per cent modulation, indicating that the plate efficiency is too high. Increase the plate loading and reduce the excitation to maintain the same plate current; then apply modulation and check the characteristic again. Continue this process until the characteristic is linear from the axis to twice the carrier amplitude. It is advantageous to use the maximum permissible plate voltage on the tube, since it is usually easier to obtain a more linear characteristic with high plate voltage and low current (carrier conditions) than with relatively low plate voltage and high plate current.

The amplifier can be adjusted without an oscilloscope by determining the plate current as described above, then setting the bias to the cut-off value (or slightly beyond) for the d.c. plate voltage used and applying maximum excitation. Adjust the plate loading, keeping the tank circuit at resonance, until the amplifier draws twice the carrier plate current, and note the antenna current. Decrease the excitation until the output and plate current just start to drop. Then increase the bias, leaving the excitation and plate loading unchanged, until the plate current drops to the proper carrier value. The antenna current should be just half the previous value; if it is larger, try somewhat more loading and less excitation; if smaller, less loading and more excitation. Repeat until the antenna current drops to half its maximum value when the plate current is biased down to the carrier value. Under these conditions the amplifier should modulate properly, provided the plate supply has good voltage regulation (§ 8-1) so that the plate voltage is practically the same at both values of plate current during the initial testing. The d.c. plate current should be substantially constant with or without modulation (§ 5-3).

Suppressor modulation — The circuit arrangement for suppressor-grid modulation of a pentode tube is shown in Fig. 508. The operating principles are the same as for grid-bias modulation. However, the r.f. excitation and modulating signals are applied to separate grids, which gives the system a simpler operating technique since best adjustment for proper excitation requirements and proper modulating circuit requirements are more or less independent. The carrier plate efficiency is approximately the same as for grid-bias modulation, and the modulator power requirements are similarly small. With tubes having suitable suppressor-grid characteristics, linear modulation up to practically 100 per cent can be obtained with negligible distortion.

The method of adjustment is essentially the same as that described in the preceding paragraph. Apply normal excitation and bias to the control grid and, with the suppressor bias at zero or the positive value recommended for c.w. telegraph operation with the particular tube used, adjust the plate loading to obtain twice the carrier plate current (on the basis of 33 per cent carrier efficiency). Then apply sufficient negative bias to the suppressor to bring the plate current to the carrier value, leaving the loading unchanged. Simultaneously, the antenna current also should drop to half its maximum value. The amplifier is then ready for modulation. Should the plate current not follow the antenna current in the same proportion when the suppressor bias is made negative, the loading and excitation should be readjusted to make them coincide.

ℂ 5-5 Cathode Modulation

Circuit — The fundamental circuit for cathode or "center-tap" modulation is shown in Fig. 509. This type of modulation is a com-

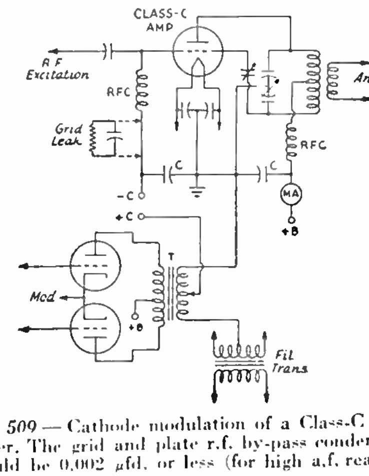


Fig. 509 — Cathode modulation of a Class-C r.f. amplifier. The grid and plate r.f. by-pass condensers, C, should be 0.002 μfd. or less (for high a.f. reactance).

bination of the plate and grid-bias methods, and permits a carrier efficiency midway between the two. The audio power is introduced in the cathode circuit, and both grid bias and plate voltage vary during modulation.

The cathode circuit of the modulated stage must be independent of other stages in the transmitter; that is, when filament-type tubes are modulated they must be supplied from a separate filament transformer. The filament by-pass condensers should not be larger than about 0.002 μfd., to avoid by-passing the audio-frequency modulation.

Operating principles — Because part of the modulation is by the grid-bias method, the plate efficiency of the modulated amplifier must vary during modulation. The carrier efficiency therefore must be lower than the efficiency at the modulation peak. The required reduction in carrier 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. 510. In these curves the performance of the cathode-modulated r.f. amplifier is plotted in terms of the tube ratings for plate-modulated telephony, with the percentage of plate modulation as a base. As the percentage of plate modulation is decreased, it is assumed that the grid-bias modulation is increased to make the over-all percentage of modulation reach 100 per cent. The limiting condition, 100 per cent plate modulation and no grid-bias modulation, is at the right (A); pure grid-bias modulation is represented by the left-hand ordinate (B and C).

As an example, assume that 40 per cent plate modulation is to be used. Then the modulated r.f. amplifier must be adjusted for a carrier plate efficiency of 56 per cent, the permissible plate input will be 65 per cent of the ratings of the same tube with pure plate modulation, the power output will be 48 per cent of the rated output of the tube with plate modulation, and the audio power required from the modulator will be 20 per cent of the d.c. input to the modulated amplifier.

Modulating impedance — The modulating impedance of a cathode-modulated amplifier is approximately equal to

$$m\frac{E_b}{I_b}$$

where m is the percentage of plate modulation expressed as a decimal, E_b is the plate voltage and I_b the plate current of the modulated r.f. amplifier. This figure for the modulating impedance is used in the same way as the corresponding figure for pure plate modulation, in determining the proper modulator operating conditions (§ 5-6).

Conditions for linearity — R.f. excitation requirements for the cathode-modulated amplifier are midway between those for plate modulation and grid-bias modulation. More excitation is required as the percentage of plate modulation is increased. Grid bias should be considerably beyond cut-off; fixed bias from a supply having good voltage regulation (§ 8-9) 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.

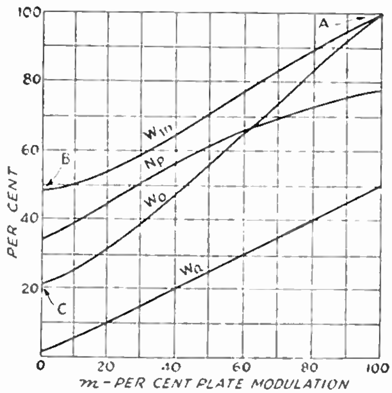


Fig. 510 — Cathode-modulation performance curves, in terms of percentage of plate modulation plotted against percentage of Class-C telephony tube ratings.
W_{in} — D.c. plate input watts in terms of percentage of plate-modulation rating.
W_o — Carrier output watts in per cent of plate-modulation rating (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.

Adjustment of cathode-modulated amplifiers — In most respects, the adjustment procedure is similar to that for grid-bias modulation (§ 5-4). The critical adjustments are those of 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 (§ 5-10). 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 (indicating downward modulation), as also will too-high excitation (§ 5-10). The cathode current will be practically constant with or without modulation when the proper operating conditions have been established (§ 5-3).

5-6 Class-B Modulators

Modulator tubes — In the case of plate modulation, the relatively large audio power needed (§ 5-3) practically dictates the use of a Class-B (§ 3-4) modulator, since the power can be obtained most economically with this type of amplifier. A typical circuit is given in Fig. 511. A pair of tubes must be chosen which is capable of delivering sine-wave audio power equal to half the d.c. input to the modulated Class-C amplifier. It is sometimes convenient to use tubes which will operate at the same plate voltage as that applied to the Class-C stage, since one power supply of adequate current capacity may then suffice for both stages. Available components do not always permit this, however, and better over-all performance and economy may result from the use of separate power supplies.

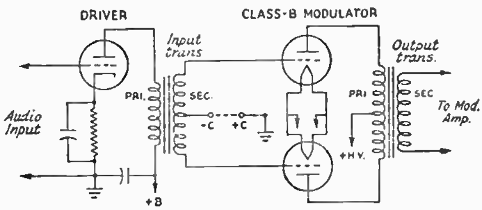


Fig. 511 — Class-B audio modulator and driver circuit.

Matching to load — In giving Class-B ratings on power tubes, manufacturers specify the plate-to-plate load impedance (§ 3-3) into which the tubes must operate to deliver the rated audio power output. This load impedance seldom is the same as the modulating impedance (§ 5-3) of the Class-C r.f. stage, so that a match must be brought about by adjusting the turns ratio of the coupling transformer. The required turns ratio, primary to secondary, is

$$\sqrt{\frac{Z_p}{Z_m}}$$

where Z_m is the Class-C modulating impedance and Z_p is the plate-to-plate load impedance specified for the Class-B tubes.

Commercial Class-B output transformers usually are rated to work between specified primary and secondary impedances and are designed for specific Class-B tubes. In such a case, the turns ratio can be found by substituting the given impedances in the formula above. Many transformers are provided with primary and secondary taps, so that various turns ratios can be obtained to meet the requirements of various tube combinations.

Driving power — Class-B amplifiers are driven into the grid-current region, so that power is consumed in the grid circuit (§ 3-3). The preceding stage (*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, since the grid current does not increase directly with the grid voltage. To prevent distortion, therefore, it is necessary to have a driving source which has good *regulation* — that is, which will maintain the waveform of the signal without distortion even though the load varies. This can be brought about by using a driver capable of delivering two or three times the actual power consumed by the Class-B grids, and by using an input coupling transformer having a turns ratio giving the largest step-down in the voltage between the driver plate or plates and the Class-B grids that will permit obtaining the specified grid-to-grid a.f. voltage.

Driver coupling — A Class-A or Class-AB (§ 3-4) driver is used to excite a Class-B stage. Tubes for the driver preferably should be triodes having low plate resistance, since these will have the best regulation. Having chosen a tube or tubes capable of ample power output from tube data sheets, the peak output voltage will be, approximately,

$$E_o = 1.4\sqrt{PR}$$

where P is the power output and R the load resistance. The input transformer ratio, primary to secondary, will be

$$\frac{E_o}{E_g}$$

where E_o is as given above and E_g is the peak grid-to-grid voltage required by the modulator tubes.

Commercial transformers normally are designed for specific driver-modulator combinations, and usually are adjusted to give as good driver regulation as the conditions will permit.

Grid bias — Modern Class-B audio tubes are intended for operation without fixed bias. This lessens the variable grid-circuit loading effect and eliminates the need for a grid-bias supply.

When a grid-bias supply is required, it must have low internal resistance so that the flow of grid current with excitation of the Class-B tubes does not cause a continual shift in the actual grid bias and thus cause distortion. Batteries or a regulated bias supply (§ 8-9) should be used.

Plate supply — The plate supply for a Class-B modulator should be sufficiently well filtered (§ 8-3) to prevent hum modulation of the r.f. stage (§ 5-2). An additional requirement is that the output condenser of the supply should have low reactance (§ 2-8) at 100 cycles or less compared to the load into which each tube is working, which is one-fourth the plate-to-plate load resistance. A 4-μfd. output condenser with a 1000-volt supply, or a 2-μfd. condenser with a 2000-volt supply, usually will be satisfactory. With other plate voltages, condenser values should be in inverse proportion to the plate voltage.

Overexcitation — When a Class-B amplifier is overdriven in an attempt to secure more than the rated power, distortion in the output waveshape increases rapidly. The high-frequency harmonics which result from the distortion (§ 3-3) modulate the transmitter, producing spurious sidebands (§ 5-2) which readily can cause serious interference over a band of frequencies several times the channel width required for speech. This may happen even though the transmitter is not being overmodulated, as in the case where the modulator is incapable of delivering the power required to modulate the transmitter fully, or when the Class-C amplifier is not adjusted to give the proper modulating impedance (§ 5-3).

The tubes used in the Class-B modulator should be capable of somewhat more than the power output nominally required (50 per cent of the d.c. input to the modulated amplifier) to take care of losses in the output transformer. These usually run from 10 per cent to 20 per cent of the tube output. In addition, the Class-C amplifier should be adjusted to give the proper modulating impedance and the correct output transformer turns ratio should be used. Such high-frequency harmonics as may be generated in these circumstances can be reduced by connecting condensers across the primary and secondary of the output transformer (about 0.002 μfd. in the average case), to form, with the transformer leakage inductance (§ 2-9) a low-pass filter (§ 2-11) which cuts off just above the maximum audio frequency required for speech transmission (about 4000 cycles). The condenser voltage ratings should be adequate for the peak a.f. voltages appearing across them.

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 load resistance of the same value as the modulating impedance, and capable of dissipating the full power output of the modulator, should be connected across the transformer secondary.

5-7 Low-Level Modulators

Selection of tubes — Modulators for grid-bias and suppressor modulation can be small audio power tubes, since the audio power required usually is small. A triode such as the 2A3 is preferable because of its low plate resistance, but pentodes will work satisfactorily.

Matching to load — Since the ordinary Class-A receiving power tube will develop about 200 to 250 peak volts in its plate circuit, which is ample for most low-level modulator applications, a 1:1 coupling transformer is generally used. If more voltage is required, a step-up ratio must be provided in the transformer. It is usual practice to load the primary of the output-coupling transformer with a resistance equal to or slightly higher than the rated load resistance for the tube, to stabilize the voltage output and thus improve the regulation. This is indicated in Fig. 507.

5-8 Microphones

Sensitivity — The *level* of a microphone is its electrical output for a given speech intensity input. Level varies greatly with microphones of different basic types, and also varies between different models of the same type. The output is also greatly dependent on the character of the individual voice (that is, the audio frequencies present in the voice) and the distance of the speaker's lips from the microphone, decreasing approximately as the square of the distance. Hence, only approximate values based on averages of "normal" speaking voices can be attempted. The values given in the following paragraphs are based on close talking; that is, with the microphone less than an inch from the speaker's lips.

Frequency response — The *frequency response* or *fidelity* of a microphone is its relative ability to convert sounds of different frequencies into alternating current. With fixed sound intensity at the microphone, the electrical output may vary considerably as the sound frequency is varied. For understandable speech transmission only a limited frequency range is necessary, and natural-sounding speech can be obtained if the output of the microphone does not vary more than a few decibels (§ 3-3) at any frequency within a range of about 200 cycles to 4000 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 — Fig. 512-A and B show connections for single- and double-button carbon microphones, with a rheostat included in each circuit for adjusting the button current to the correct value as specified with each microphone. The single-button 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 back-plate the other. 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 (§ 2-9). The double-button type is similar, but with two buttons in push-pull.

Good quality single-button 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 across 100,000 ohms or so can be assumed available at the grid of the first tube. The usual button current is 50 to 100 ma.

The level of good-quality double-button microphones is considerably less, ranging from 0.02 volt to 0.07 volt across 200 ohms. With this type of microphone and the usual push-pull input transformer, a peak voltage of 0.4 to 0.5 across 100,000 ohms or so can be assumed available at the first speech-amplifier grid. The button current with this type of microphone ranges from 5 to 50 ma. per button.

Crystal microphones — The input circuit for a piezoelectric or crystal type of microphone is shown in Fig. 512-F. The element in this type consists of a pair of Rochelle salts crystals cemented together, with plated electrodes. In the more sensitive types, the crystal is mechanically coupled to a diaphragm. Sound waves actuating the diaphragm cause the crystal to vibrate mechanically and, by piezoelectric action (§ 2-10), to generate a corresponding alternating voltage between the electrodes, which are connected to the grid circuit of a vacuum-tube amplifier, as shown. The crystal type requires no separate source of current or voltage.

Although the level of crystal microphones varies with different models, an output of 0.01 to 0.03 volt 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 being attenuated as the shunt resistance becomes less. A grid-resistor value of 1 megohm or more should be used for reasonably flat response, 5 megohms being a customary figure.

Condenser microphones — The condenser microphone of Fig. 512-C consists of a two-plate capacity, with one plate stationary. The other, which is separated from the first by about a thousandth of an inch, is a thin metal membrane serving as a diaphragm. This condenser is connected in series with a resistor and a d.c. voltage source. When the diaphragm vibrates, the change in capacity causes a small charging current to flow through the circuit. The resulting audio voltage which appears across the resistor is fed to the grid of the tube through the coupling condenser.

The output of condenser microphones varies with different models, the high-quality type being about one-hundredth to one-fiftieth as sensitive as the double-button carbon microphone. The first speech-amplifier stage must be built into the microphone, since the capacity of a connecting cable would impair both output and frequency range.

Velocity and dynamic microphones — In a velocity or "ribbon" microphone, the element acted upon by the sound waves is a thin corrugated metallic ribbon suspended between the poles of a magnet. When vibrating, the ribbon cuts the lines of force between the poles, first in one direction and then the other, thus generating an alternating voltage. The movement of the ribbon is proportional to the velocity of the sound-energized air particles. 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. 512-E). 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. 512-D.

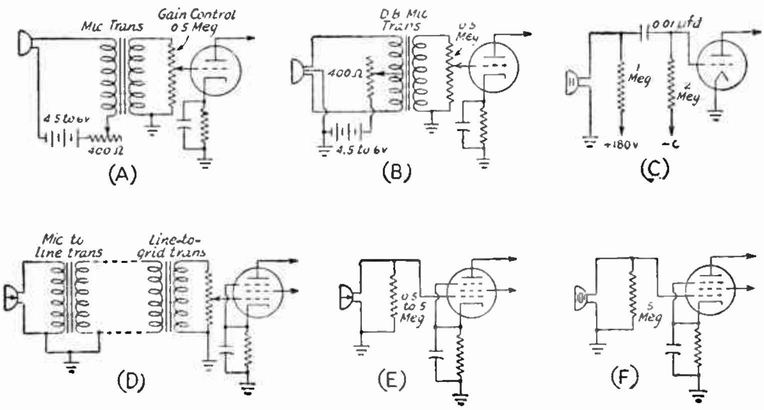


Fig. 512 — Speech input circuits of five commonly used types of microphones. A, single-button carbon; B, double-button carbon; C, condenser; D, low-impedance velocity; E, high-impedance velocity; F, crystal.

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 coupling transformer secondary.

The dynamic microphone somewhat resembles a dynamic loud speaker in principle. A light-weight voice coil is rigidly attached to a diaphragm, the coil being placed between the poles of a permanent magnet. Sound causes the diaphragm to vibrate, thus moving the coil back and forth between the magnet poles and generating an alternating voltage the frequency of which is proportional to the frequency of the impinging sound and the amplitude proportional to the sound pressure. The dynamic microphone usually is built with high-impedance output, suitable for working directly into the grid of an amplifier tube. If the connecting cable must be unusually long a low-impedance type should be used, with a step-up transformer at the end of the cable. A small permanent-magnet speaker can be used as a dynamic microphone, although the fidelity is not as good as is obtainable with a properly designed microphone.

5-9 The Speech Amplifier

Description — The function of the speech amplifier is to build up the weak microphone voltage to a value sufficient to excite the modulator to the required output. It may have from one to several stages. The last stage nearly always must deliver a certain amount of audio power, especially when it is used to excite a Class-B modulator. Speech amplifiers for grid-bias modulation usually end in a power stage which also functions as the modulator.

The speech amplifier frequently is built as a unit separate from the modulator, and in such a case may be provided with a step-down transformer designed to work into a low impedance, such as 200 or 500 ohms (tube-to-line transformer). When this is done, a step-up input transformer intended to work between the same impedance and the modulator grids (line-to-grid transformer) is provided in the modulator circuit. The line which connects the two transformers may be made of any convenient length.

General design considerations — The last stage of the speech amplifier must be selected on the basis of the power output required from it; for instance, the power necessary to drive a Class-B modulator (§ 5-6). It may be either single-ended or push-pull (§ 3-3), the latter generally being preferable because of the higher power output and lower harmonic distortion. Push-pull amplifiers may be either Class A, Class AB_1 or Class AB_2 (§ 3-4), as the power requirements dictate. If a Class-A or AB_1 amplifier is used, the preceding stages all may be voltage amplifiers, but when a Class-AB_2 amplifier is used the stage immediately preceding it must be capable of furnishing the power consumed by its grids at full output. The requirements in this case are much the same as those which must be met by a driver for a Class-B stage (§ 5-6), but the actual power needed is considerably smaller and usually can be supplied by one or two small receiving triodes. All lower-level speech amplifier stages invariably are worked purely as voltage amplifiers.

The minimum amplification which must be provided ahead of the last stage is equal to the peak audio-frequency grid voltage required by the last stage for full output (peak grid-to-grid voltage in the case of a push-pull stage), divided by the output voltage of the microphone or secondary of the microphone transformer if one is used (§ 5-8). The peak a.f. grid voltage required by the output tube or tubes is equal to the d.c. grid bias in the case of a single-tube Class-A amplifier, and approximately twice the grid bias for a push-pull Class-A stage. The requisite information for Class-AB_1 and AB_2 amplifiers can be obtained from the manufacturer's data on the type considered. If the gain is not obtainable in one stage, several stages must be used in cascade. When the output stage is operated Class AB_2, due allowance must be made for the fact that the next-to-the-last stage must deliver power as well as voltage. In such cases, suitable driver combinations usually are recommended by manufacturers of tubes and interstage transformers. The coupling transformer must be designed especially for the purpose.

The total gain provided by a multi-stage amplifier is equal to the product of the individual stage gains. For example, when three stages are used, the first having a gain of 100, the second 20 and the third 15, the total gain is $100 \times 20 \times 15$, or 30,000. It is good practice to provide two or three times the minimum required gain in designing the speech amplifier. This will insure having ample gain available to cope with varying conditions.

When the gain must be fairly high, as when a crystal microphone is used, the speech amplifier frequently has four stages, including the power output stage. The first generally is a pentode, because of the high gain attainable with this type of tube. The second and third stages usually are triodes, the third frequently having two tubes in push-pull when it drives a Class-AB_2 output stage. Two pentode stages seldom are used consecutively, because of the difficulty of getting stable operation when the gain per stage is very high. With carbon microphones less amplification is needed and hence the pentode first stage usually is omitted, one or two triode stages being ample to obtain full output from the power stage.

Stage gain and voltage output — In voltage amplifiers, the *stage gain* is the ratio of a.c. output voltage to a.c. voltage applied to the grid. It will vary with the applied audio frequency, but for speech the variation should be small over the range of 100–4000 cycles. This condition is easily met in practice.

The output voltage is the maximum value which can be taken from the plate circuit without distortion. It is usually expressed in terms of the peak value of the a.c. wave (§ 2-7), since this value is independent of the wave-form. The peak output voltage usually is of interest only when the stage drives a power amplifier, since only in this case is the stage called upon to work near its maximum capabilities. Low-level stages very seldom are worked near their full capacity, hence harmonic distortion is negligible and the voltage gain of the stage is the primary consideration.

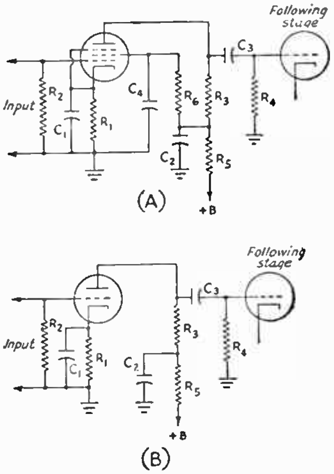


Fig. 513 — Resistance-coupled voltage amplifier circuits. A, pentode; B, triode. Designations are as follows:
C_1 — Cathode by-pass condenser.
C_2 — Plate by-pass condenser.
C_3 — Output coupling condenser (blocking condenser).
C_4 — Screen by-pass condenser.
R_1 — Cathode resistor.
R_2 — Grid resistor.
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 Chapter Fourteen.

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, since with transformers a sufficiently high load impedance (§3-3) cannot be obtained without considerable frequency distortion. Typical circuits are given in Fig. 513 and design data in § 3-6.

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.

Representative circuits for coupling single-ended to push-pull stages are shown in Fig. 514. That at A uses a combination of resistance and transformer coupling, and may be used for exciting the grids of a Class-A or AB_1 following stage. The resistance coupling is used to keep the d.c. plate current from flowing through the transformer primary, thereby preventing a reduction in primary inductance below its no-current value (§ 8-4). 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) as in the case of a following Class-AB_2 stage used as a driver for a Class-B modulator.

Gain control — The over-all gain of the amplifier may be changed to suit the output level of the microphone, which will vary with voice intensity and distance of the speaker from the microphone, by varying the proportion of a.c. voltage applied to the grid of one of the stages.

The gain-control potentiometer should be near the input end of the amplifier, so that there will be no danger of overloading the stages ahead of the gain control. With carbon microphones the gain control may be placed directly across the microphone transformer secondary, but with other types the gain control usually will affect the frequency response of the microphone when connected directly across it. The control therefore usually is placed in the grid circuit of the second stage.

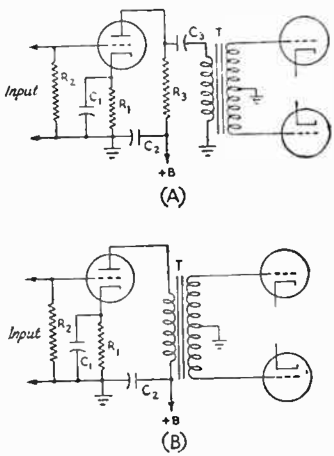


Fig. 514 — Transformer-coupled amplifier circuits for driving a push-pull amplifier. A is for resistance-transformer coupling; B, for transformer coupling. Designations correspond to those in Fig. 513. In A, values can be taken from Table I. In B, the cathode resistor is calculated from the rated plate current and grid bias as given for the particular type of tube used (§ 3-6).

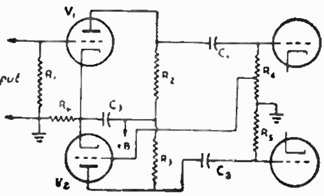


Fig. 515 — Phase-inverter circuit for resistance-coupled push-pull output. With a double-triode tube (such as the 6N7) the following values are typical:
R_1, R_4, R_5 — 0.5 megohm. R_2, R_3 — 0.1 megohm.
R_6 — 1500 ohms. C_1, C_2 — 0.1 μfd.
R_4 should be tapped as described in the text. The voltage gain of a stage using these constants is 22.

Phase inversion — Push-pull output may be secured with resistance coupling by using an extra tube, as shown in Fig. 515. There is a phase shift of 180 degrees through any normally operating resistance-coupled stage (§ 3-3), and the extra tube is used purely to provide this phase shift without additional gain. The outputs of the two tubes are then added to provide push-pull excitation for the following amplifier. The tap on R_4 is adjusted to make V_1 and V_2 give equal voltage outputs so that balanced excitation is applied to the grids of the following stage. The cathode resistor, R_6, commonly is left un-bypassed since this tends to help balance the circuit. For convenience, double-triode tubes frequently are used as phase inverters.

Output limiting — It is desirable to modulate as heavily as possible without overmodulating, yet it is difficult to speak into the microphone at a constant intensity. To maintain reasonably constant output from the modulator in spite of variations in speech intensity, it is possible to use automatic gain control which follows the *average* (not instantaneous) variations in speech amplitude. This is accomplished by rectifying and filtering (§ 8-2, 8-3) some of the audio output and applying the rectified and filtered d.c. to a control electrode in an early stage in the amplifier.

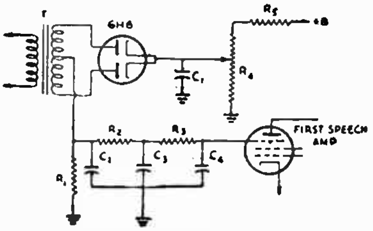


Fig. 516 — Speech amplifier output-limiting circuit.
C_1, C_2, C_3, C_4 — 0.1-μfd. R_1, R_2, R_3 — 0.25 megohm.
R_4 — 25,000-ohm pot. R_5 — 0.1 megohm. T — See text.

A practical circuit for this purpose is shown in Fig. 516. 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 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 variable bias on the rectifier plates, so that the limiting action can be delayed until a desired microphone input level is reached. R_2, R_3, C_2, C_3, and C_4 form the filter (§ 2-11), and the output of the rectifier is connected to the suppressor grid of the pentode first stage of the speech amplifier.

A step-down transformer with a turns ratio such as to give about 50 volts when its primary is connected to the output circuit of the power stage should be used. A half-wave rectifier may be used instead of the full-wave circuit shown, although satisfactory filtering will be more difficult to achieve.

Noise — It is important that the noise level in a speech amplifier be low compared to the level of the desired signal. Noise in the speech amplifier is caused chiefly by hum, which may be the result of insufficient power-supply filtering or may be introduced into the grid circuit of a tube by magnetic or electrostatic means from heater wiring. The plate voltage for the amplifier should be free from ripple (§ 8-4), particularly the voltage applied to the low-level stages. A two-section condenser-input filter (§ 8-5) usually is satisfactory. The decoupling circuits mentioned in the preceding paragraphs also are helpful in reducing plate-supply hum.

Hum from heater wiring may be reduced by keeping the wiring well away from ungrounded components or wiring, particularly in the vicinity of the grid of the first tube. Complete shielding of the microphone jack is advisable, and when tubes with grid caps instead of the single-ended types are used the caps and the exposed wiring to them should be shielded. Heater wiring preferably should run in the corners of a metal chassis, to reduce the magnetic field. A ground should be made either on one side of the heater circuit or to the center-tap of the heater winding. The shells of metal tubes should be grounded; glass tubes require separate shields, especially when used in low-level stages. Heater connections to the tube sockets should be kept as far as possible from the plate and grid prongs, and the heater wiring to the sockets should be kept close to the chassis. A connection to a good ground (such as a cold water pipe) also is advisable. The speech amplifier always should be constructed on a metal chassis, with all ground connections made directly to the metal chassis.

When the power supply is mounted on the same chassis with the speech amplifier, the power transformer and filter chokes should be well separated from audio transformers in the amplifier proper to reduce magnetic coupling, which would cause hum and raise the residual noise level.

5-10 Checking 'Phone Transmitter Operation

Modulation percentage—The most reliable method of determining percentage of modulation is by means of the cathode-ray oscilloscope (§ 3-9). The oscilloscope gives a direct picture of the modulated output of the transmitter, and by its use the waveform errors inherent in other types of measurements are eliminated.

Two types of oscilloscope patterns may be obtained, known as the "wave envelope" and "trapezoid." The former shows the shape of the modulation envelope (§ 5-2) directly, while the latter in effect plots the modulation characteristic (§ 5-2) of the modulated stage on the cathode-ray tube screen. To obtain the wave-envelope pattern, the oscilloscope must have a horizontal sweep circuit. The trapezoidal pattern requires only the oscilloscope, the sweep circuit being supplied by the transmitter itself. Fig. 517 shows methods of connecting the oscilloscope to the transmitter for both types of patterns. The oscilloscope connections for the wave-envelope pattern, Fig. 517-A, are usually simpler than those for the trapezoidal figure. The vertical-deflection plates are coupled to the amplifier tank coil or an antenna coil by means of a pick-up coil of a few turns connected to the oscilloscope through a twisted-pair line. The position of the pick-up coil is varied until a carrier pattern, Fig. 518-B, of suitable height is obtained. The sweep voltage should be adjusted to make the width of the pattern somewhat more than half the diameter of the screen. It is frequently helpful in eliminating r.f. harmonics from the pattern to connect a resonant circuit, tuned to the operating frequency, between the vertical deflection plates, using link coupling between this and the transmitter tank circuit.

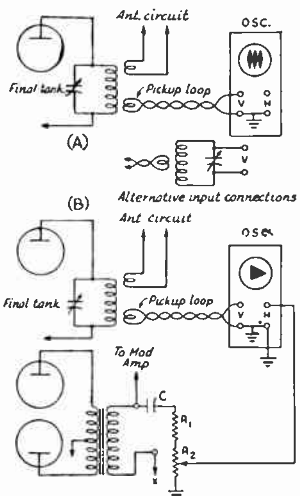


Fig. 517—Methods of connecting an oscilloscope to the modulated r.f. amplifier for checking modulation.

With the application of voice modulation, 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 (§ 5-2). This is illustrated by Fig. 518-D, where the point X represents the sweep line (reference line) alone, YZ is the carrier height, and PQ is the maximum height of the modulated wave. If the height is greater than the distance PQ, as illustrated in E, the wave is overmodulated in the upward direction. Overmodulation in the downward direction is indicated by a gap in the pattern at the reference axis, where a single bright line appears on the screen. Overmodulation in either direction may take place even when the modulation in the other direction is less than 100 per cent. Assuming that the modulation is symmetrical, however, any modulation percentage can be measured directly from the screen by measuring the maximum height with modulation and the height of the carrier alone; calling these two heights h_1 and h_2 respectively, the modulation percentage is

$$\frac{h_1 - h_2}{h_2} \times 100$$

Connections for the trapezoidal pattern are shown in Fig. 517-B. The vertical plates are similarly coupled to the transmitter tank circuit through a pick-up loop; the tuned input circuit to the oscilloscope may also be used. The horizontal plates are coupled to the output of the modulator through a voltage divider (§ 2-6). R_1R_2, the resistance of R_2 being variable to permit adjustment of the audio voltage to a suitable value to give a satisfactory horizontal sweep on the screen. R_2 may be a 0.25-megohm volume control resistor. The value of R_1 will depend upon the audio output voltage of the modulator. This voltage is equal to $\sqrt{PR}$, where P is the audio power output of the modulator and R is the modulating impedance of the modulated r.f. amplifier. In the case of grid-bias modulation with a 1:1 output transformer, it will be satisfactory to assume that the a.c. output voltage of the modulator is equal to $0.7E$ for a single tube or $1.4E$ for a push-pull stage, where E is the d.c. plate voltage on the modulator. If the transformer ratio is other than 1:1, the voltage so calculated should be multiplied by the actual secondary-to-primary turns ratio.

The total resistance of R_1 and R_2 in series should be 0.25 megohm for every 150 volts of modulator output: for example, if the modulator output voltage is 600, the total resistance should be four (600/150) times 0.25 megohm, or 1 megohm. Then, with 0.25 megohm at R_2, R_1 should be 0.75 megohm. The blocking condenser, C, should be 0.1 μfd or more, and its voltage rating should be greater than the maximum voltage in the circuit. With plate modulation, this is twice the d.c. voltage applied to the plate of the modulated amplifier.

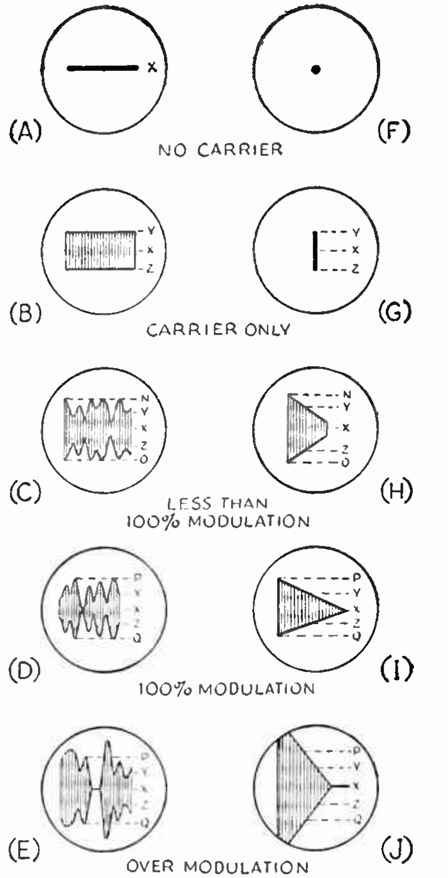


Fig. 518 — Wave-envelope and trapezoidal patterns encountered under different conditions of modulation.

The trapezoidal patterns are shown in Fig. 518 at F to J, each alongside the corresponding wave-envelope pattern. With no signal, only the cathode-ray spot appears on the screen. When the unmodulated carrier is applied, a vertical line appears; the length of the line should be adjusted, by means of the pick-up coil coupling, to a convenient value. When the carrier is modulated, the wedge-shaped pattern appears; the higher the modulation percentage, the wider and more pointed the wedge becomes. At 100 per cent modulation it just makes a point on the axis, X, at one end, and the height, PQ, at the other end is equal to twice the carrier height, YZ. Overmodulation in the upward direction is indicated by increased height over PQ, and in the downward direction by an extension along the axis X at the pointed end. The modulation percentage may be found by measuring the modulated and unmodulated carrier heights, in the same way as with the wave-envelope pattern.

Non-symmetrical waveforms — In voice waveforms the average maximum amplitude in one direction from the axis frequently is greater than in the other direction, although the average *energy* on both sides is the same. For this reason the percentage of modulation in the "up" direction frequently differs from that in the "down" direction. With a given voice and microphone, this difference in modulation percentage is usually always in the same direction. Since overmodulation in the downward direction causes more out-of-channel interference than overmodulation upward because of the steeper wavefront (§ 6-1), it is advisable to "phase" the modulation so that the side of the voice waveform having the larger excursions causes the instantaneous carrier power to increase and the smaller excursions to cause a power decrease. This reduces the likelihood of overmodulation on the "down" peak. The direction of the larger excursions can readily be found by careful observation of the oscilloscope pattern. The phase can be reversed by reversing the connections of one winding of any transformer in the speech amplifier or modulator.

Modulation monitoring — While it is desirable to modulate as fully as possible, 100 per cent modulation should not be exceeded, particularly in the downward direction, because harmonic distortion will be introduced and the channel width increased (§ 5-2), thus causing unnecessary interference to other stations. The oscilloscope may be used to provide a continuous check on the modulation, but simpler indicators may be used for the purpose, once calibrated. A convenient indicator, when a Class-B modulator (§ 5-6) is used, is the plate milliammeter in the Class-B stage, since plate current fluctuates with the voice intensity. Using the oscilloscope, determine the gain-control setting and voice intensity which gives 100 per cent modulation on voice peaks, and simultaneously observe the maximum Class-B plate-milliamnneter reading on the peaks. When this maximum reading is obtained, it will suffice in regular operation to adjust the gain so that it is not exceeded.

A sensitive rectifier-type voltmeter (copper-oxide type) also can be used for modulation monitoring. It should be connected across the output circuit of an audio driver stage where the power level is a few watts, and similarly calibrated against the oscilloscope to determine the reading which represents 100 per cent modulation.

The plate milliammeter of the modulated r.f. stage may also be used as an indicator of overmodulation. Since the average plate current is constant (§ 5-3, 5-4, 5-5) when the amplifier is linear, the reading will be the same with or without modulation. 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 non-linearity. However, it is possible that the average plate current will remain constant with considerable overmodulation under some operating conditions, so that an indicator of this type is not wholly reliable unless it has been checked previously against an oscilloscope.

Linearity — The linearity (§ 5-2) of a modulated amplifier may readily be checked with the oscilloscope. The trapezoidal pattern is more easily interpreted than the wave envelope pattern, and less auxiliary equipment is required. The connections are the same as for measuring modulation percentage (Fig. 517). If the amplifier is perfectly linear, the sloping sides of the trapezoid will be perfectly straight from the point at the axis up to at least 100 per cent modulation in the upward direction. Nonlinearity will be shown by curvature of the sides. Curvature near the point, extending the point farther along the axis than would occur with straight sides, indicates that the output power does not decrease rapidly enough in this region; it may also be caused by imperfect neutralization (a push-pull amplifier is recommended because better neutralization is possible than with single-ended amplifiers) or by r.f. leakage from the exciter through the final stage. The latter condition can be checked by removing the plate voltage from the modulated stage, when the carrier should disappear, leaving only the beam spot remaining on the screen (Fig. 518-F). If a small vertical line remains, the amplifier should be re-neutralized; if this does not eliminate the line, it is an indication that r.f. is being picked up from lower-power stages, either by coupling through the final tank or via the oscilloscope pick-up loop.

Inward curvature at the large end of the pattern is caused by improper operating conditions of the modulated amplifier, usually improper bias or insufficient excitation, or both, with plate modulation. In grid-bias and cathode-modulated systems, the bias, excitation and plate loading are not correctly proportioned when such curvature occurs, usually because the amplifier has been adjusted to have too-high carrier efficiency without modulation (§ 5-4, 5-5).

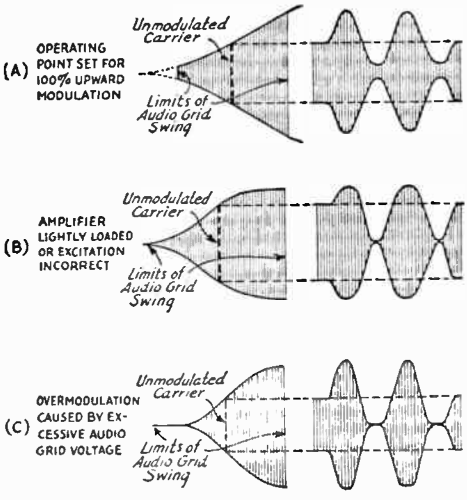


Fig. 519 — Oscilloscope patterns representing proper and improper adjustments for grid-bias or cathode modulation. The pattern obtained with a correctly adjusted amplifier is shown at A. The other drawings indicate non-linear modulation from typical causes.

For the wave-envelope pattern, it is necessary to have a linear horizontal-sweep circuit in the oscilloscope and a source of sine-wave audio signal voltage (such as an audio oscillator or signal generator) which can be synchronized with the sweep circuit. The linearity can be judged by comparing the wave envelope with a true sine wave. Distortion in the audio circuits will affect the pattern in this case (such distortion has no effect on the trapezoidal pattern, which shows the modulation characteristic of the r.f. amplifier alone), and it is also readily possible to misjudge the shape of the modulation envelope, so that the wave envelope is less useful than the trapezoid for checking linearity of the modulated amplifier.

Fig. 519 shows typical patterns of both types. The cause of the distortion is indicated for grid-bias and suppressor modulation. The patterns at A, although not truly linear, are representative of properly operated grid-bias modulation systems. Better linearity can be obtained with plate modulation of a Class-C amplifier.

Faulty patterns — The drawings of Figs. 518 and 519 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 is probable that the pattern is faulty rather than the transmitter. It is important that only r.f. from the modulated stage be coupled to the oscilloscope, and then only to the vertical plates. The effect of stray r.f. from other stages in the transmitter has been mentioned in the preceding paragraph. If r.f. is present also on the horizontal plates, the pattern will lean to one side instead of being upright. If the oscilloscope cannot be moved to a spot where the unwanted pick-up disappears, a small by-pass condenser (10 $\mu\mu$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 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. 517-B.

Plate-current shift — As mentioned above, the d.c. plate current of a modulated amplifier will be the same with and without modulation so long as the amplifier operation is perfectly linear and other conditions remain unchanged. This also assumes that the modulator is working within its capabilities. Because there is usually some curvature of the modulation characteristic with grid-bias modulation there is normally a slight upward change in plate current of a stage so modulated, but this occurs only at high modulation percentages and is barely detectable under the usual conditions of voice modulation.

With plate modulation, a downward shift in plate current may indicate one or more of the following:

1) Insufficient excitation to the modulated r.f. amplifier.
2) Insufficient grid bias on the modulated stage.
3) Wrong load resistance for the Class-C r.f. amplifier.
4) Insufficient output capacity in the filter of the modulated-amplifier plate supply.
5) Heavy overloading of the Class-C r.f. amplifier tube or tubes.

Any of the following may cause an upward shift in plate current:

1) Overmodulation (excessive audio power, audio gain too great).
2) Incomplete neutralization of the modulated amplifier.
3) Parasitic oscillation in the modulated amplifier.

When a common plate supply is used for both a Class-B (or Class-AB) modulator and a modulated r.f. amplifier, the plate current of the latter may "kick" downward because of poor power-supply voltage regulation (§ 8-1) with the varying additional load of the modulator on the supply. The same effect may occur with high-power transmitters because of poor regulation of the a.c. supply mains, even when a separate power-supply unit is used for the Class-B modulator. Either condition may be detected by measuring the plate voltage applied to the modulated stage; in addition, poor line regulation also may be detected by observing if there is any downward shift in filament or line voltage.

With grid-bias modulation, any of the following may be the cause of a plate current shift greater than the normal mentioned above:

Downward kick: Too much r.f. excitation; insufficient operating bias; distortion in modulator or speech amplifier; too-high resistance in bias supply; insufficient output capacity in plate-supply filter to modulated amplifier; amplifier plate circuit not loaded heavily enough; plate-circuit efficiency too high under carrier conditions.

Upward kick: Overmodulation (excessive audio voltage); distortion in audio system; regeneration because of incomplete neutralization; operating grid bias too high.

A downward kick in plate current will accompany an oscilloscope pattern like that of Fig. 519-B; the pattern with an upward kick will look like Fig. 519-A, with the shaded portion extending farther to the right and above the carrier, for the "wedge" pattern.

Noise and hum on carrier — These may be detected by listening to the signal on a receiver sufficiently removed from the transmitter to avoid overloading. The hum level should be low compared to the voice at 100 per cent modulation. Hum may come either from the speech amplifier and modulator or from the r.f. section of the transmitter. Hum from the r.f. section can be detected by completely shutting off the modulator; if hum remains when this is done, the power-supply filters for one or more of the r.f. stages have insufficient smoothing (§ 8-4). 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 which can be checked by removing the microphone from the circuit but leaving the first speech-amplifier grid circuit otherwise unchanged. A good ground on the microphone and speech system usually is essential to hum-free operation.

Hum can be checked with the oscilloscope, where it appears as modulation on the carrier in the same way as the normal modulation. While the percentage usually is rather small, if the carrier shows modulation with no speech input hum is the likely cause. The various parts of the transmitter may be checked through as described above.

Spurious sidebands — A superheterodyne receiver having a crystal filter (§ 7-8, 7-11) is needed for checking spurious sidebands outside the normal communication channel (§ 5-2). 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 (§ 7-8). With the crystal filter in its sharpest position and the beat oscillator turned on, tune through the region outside the normal channel limits (3 to 4 kilocycles each side of the carrier) while another person talks into the microphone. Spurious sidebands will be observed as intermittent beat notes coinciding with voice peaks, or, in bad cases of distortion or overmodulation, as "clicks" or crackles well away from the carrier frequency. Sidebands more than 4 kilocycles from the carrier should be of negligible strength in a properly modulated 'phone transmitter. The causes are overmodulation or non-linear operation (§ 5-3).

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 are necessary to prevent r.f. pick-up, and a ground connection separate from that to which the transmitter is connected is advisable. Unsymmetrical or capacity coupling to the antenna (single-wire feed, feeders tapped on final tank circuit, etc.) may be responsible in that these systems sometimes cause the transmitter chassis to take an r.f. potential above ground. Inductive coupling to a two-wire transmission line is advisable. This antenna effect can be checked by disconnecting the antenna and dissipating the power in a dummy antenna (§ 4-9), when it usually will be found that the r.f. feed-back disappears. If it does not, the speech amplifier and microphone shielding are at fault.

5-11 Frequency Modulation

Principles — In frequency modulation the carrier amplitude is constant and the output frequency of the transmitter is made to vary about the carrier or mean frequency at a rate corresponding to the audio frequencies of the speech currents. The extent to which the frequency changes in one direction from the unmodulated or carrier frequency is called the *frequency deviation.* It corresponds to the change of carrier amplitude in the amplitude-modulation system (§ 5-2). Deviation is usually expressed in kilocycles, and is equal to the difference between the carrier frequency and either the highest or lowest frequency reached by the carrier in its excursions with modulation. There is no modulation percentage, in the usual sense; with suitable circuit design the deviation may be made as large as desired without encountering any effect equivalent to overmodulation in the amplitude-modulated system.

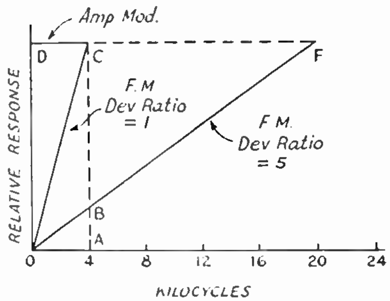


Fig. 520 — Triangular spectrum showing the noise response in an f.m. receiver compared with amplitude modulation. Deviation ratios of 1 and 5 are shown.

Deviation ratio — The ratio of the maximum frequency deviation to the audio frequency of the modulation is called the *deviation ratio.* It also is called the modulation index. Unless otherwise specified, it is taken as the ratio of the maximum frequency deviation to the *highest* audio frequency to be transmitted.

Advantages of f.m. — The chief advantage of frequency modulation over amplitude modulation is noise reduction at the receiver. All electrical noises in the radio spectrum, including those originating in the receiver, are r.f. oscillations which vary in amplitude, this variation causing the noise response in amplitude-modulation receivers. If the receiver does not respond to amplitude variations but only to frequency changes, noise can affect it only by causing a phase shift which appears as frequency modulation on the signal. The effect of such frequency modulation by the noise can be made small by making the frequency change (deviation) in the signal large.

A second advantage is that the power required for modulation is inconsequential, since there is no power variation in the modulated output of the transmitter.

Triangular spectrum — The way in which noise is reduced by a large deviation ratio is illustrated by Fig. 520. In this figure the noise is assumed to be evenly distributed over the channel used, an assumption which is almost always true. It is also assumed that audio frequencies above 4000 cycles (4 kc.) are not necessary to voice communication, and that the audio system in the receiver has no response above this frequency. Then, if an amplitude modulation receiver is used and its selectivity is such that there is no attenuation of sidebands (§ 5-2) below 4000 cycles, the noise components of all frequencies within the channel will produce equal response when they beat with a carrier centered in the channel. The response under these conditions is shown by the line *DC*.

In the f.m. receiver the output amplitude is proportional to the frequency deviation, and noise components in the channel can be considered to frequency-modulate the steady carrier with a deviation proportional to the difference between the actual frequency of the component and the frequency of the carrier, and also to give an audio-frequency beat of the same frequency difference. This leads to a rising response characteristic, such as the line *OC*, where the noise amplitude is proportional to the audio beat frequency. The average noise power output is proportional to the square root of the sum of the squares of all the amplitude values (§ 2-7), so that the noise power with frequency modulation having a deviation ratio of 1 is only one-third that with amplitude modulation, or an improvement of 4.75 db.

If the deviation ratio is increased to 5, the noise response is represented by the line *OF*. Since only frequencies up to 4000 cycles are reproduced in the output, however, the audible noise is confined to the triangle OAB. These relations hold only when the carrier is strong compared to the noise. For reception of stations with weak signal strength, the signal-to-noise ratio is better with a deviation ratio of 1.

Linearity — A transmitter in which frequency deviation is directly proportional to the amplitude of the modulating signal is said to be *linear*. It is essential also that the carrier amplitude remain constant under modulation, which in turn requires that the transmitter tuned circuits, as well as the antenna, have broad enough response to handle without discrimination the entire range of audio frequencies transmitted. This requirement is easily met under ordinary conditions.

Sidebands — In frequency modulation there is a series of sidebands on either side of the carrier frequency for each audio-frequency component in the modulation. In addition to the usual sum and difference frequencies (§ 5-2) there are also beats at harmonics of the fundamental modulating frequency, even though the latter may be a pure tone. This occurs because of the necessity for maintaining the proper phase relationships between the carrier and sidebands to keep the power output constant. Hence, a frequency-modulated signal inherently occupies a wider channel than an amplitude-modulated signal. Because of the necessity for conserving space in the usual communication spectrum, the use of f.m. by amateurs is confined to the very-high frequencies in the region above 28 Mc.

The number of sidebands for a single modulating frequency increases with the frequency deviation. When the deviation ratio is of the order of 5 the sidebands beyond the maximum frequency deviation are usually negligible, so that the channel required is approximately twice the frequency deviation.

5-12 Methods of Frequency Modulation

Requirements and methods — At present there are no fixed standards of frequency deviation in amateur work. Since a deviation ratio of 5 is considered high enough in any case, the maximum deviation necessary is 15 to 20 kc. for an upper audio-frequency limit of 3000 or 4000 cycles (§ 5-2), or a channel width of 30 to 40 kc. The permissible deviation is determined by the receiver (§ 7-18), since deviation beyond the limits of the receiver pass-band causes distortion. If the transmitter is designed to be linear (§ 5-11) with a deviation of about 15 kc., it can be used at a lower deviation ratio simply by reducing the gain in the speech amplifier. Thereby it can be made to conform to the requirements of the receiver in use.

The several possible methods of frequency modulation include mechanical modulation (for instance, varying condenser plate spacing in accordance with voice vibrations), initial phase-shift modulation which later is transformed into frequency modulation, and direct frequency modulation of an oscillator by electronic means. The latter, in the form of the *reactance-tube modulator*, is the simplest system.

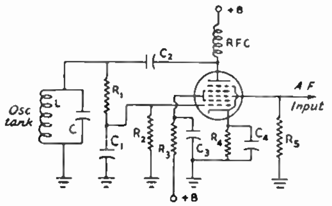


Fig. 521 — Reactance modulator circuit using a 6L7 tube.
C — Tank capacity. C_1 — 3–10 μμfd. C_2 — 250 μμfd.
C_3 — 8-μfd. electrolytic (a.f. by-pass) in parallel with 0.01-μfd. paper (r.f. by-pass).
C_4 — 0.01 μfd. L — Oscillator tank inductance.
R_1 — 50,000 ohms. R_2, R_5 — 0.5 megohm.
R_3 — 30,000 ohms. R_4 — 300 ohms.

The reactance modulator — The reactance modulator consists of 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 capacity, of a value dependent upon the instantaneous a.f. voltage applied to its grid. Fig. 521 is a representative circuit. The control grid circuit of the 6L7 tube is connected across the small capacity, C_1, which is in series with the resistor, R_1, across the oscillator tank circuit. Any type of oscillator circuit (§ 3-7) may be used. R_1 is large compared to the reactance (§ 2-8) of C_1, so the r.f. current through R_1C_1 will be practically in phase (§ 2-7) with the r.f. voltage appearing at the terminals of the tank circuit. However, the voltage across C_1 will lag the current by 90 degrees (§ 2-8). The r.f. current in the plate circuit of the 6L7 will be in phase with the grid voltage (§ 3-3), and consequently is 90 degrees behind the current through C_1, or 90 degrees behind the r.f. tank voltage. This lagging current is drawn through the oscillator tank, giving the same effect as though an inductance were connected across the tank (in an inductance the current lags the voltage by 90 degrees — § 2-8). The frequency increases in proportion to the lagging plate current of the modulator, as determined by the a.f. voltage applied to the No. 3 grid of the 6L7; hence the oscillator frequency varies with the audio signal voltage.

If, on the other hand, C_1 and R_1 are reversed and the reactance of C_1 is made large compared to the resistance of R_1 the r.f. current in the 6L7 plate circuit will lead the oscillator tank r.f. voltage, making the reactance capacitive rather than inductive.

Other circuit arrangements to produce the same effect may be employed. It is convenient to use a tube (such as the 6L7) in which the r.f. and a.f. voltages can be applied to separate control grids; however, both voltages may be applied to the same grid provided precautions are taken to prevent r.f. from flowing in the external audio circuit, and vice versa (§ 2-13).

The modulated oscillator usually is operated on a relatively low frequency, so that a high order of carrier stability can be secured. Frequency multipliers are used to raise the frequency to the final frequency desired. The frequency deviation increases with the number of times the initial frequency is multiplied; for instance, if the oscillator is operated on 7 Mc. and the output frequency is to be 112 Mc., an oscillator frequency deviation of 1000 cycles will be raised to 16,000 cycles at the output frequency.

Design considerations — The sensitivity of the modulator (frequency change per unit change in grid voltage) increases when C_1 is made smaller, for a fixed value of R_1, and also increases with an increase in L/C ratio in the oscillator tank circuit. Since the carrier stability of the oscillator depends on the L/C ratio (§ 3-7), it is desirable to use the highest tank capacity which will permit the desired deviation to be secured while keeping within the limits of linear operation. When the circuit of Fig. 521 is used in connection with a 7-Mc. oscillator, a linear deviation of 2000 cycles above and below the carrier frequency can be secured when the oscillator tank capacity is approximately 200 μμfd. A peak a.f. input of two volts is required for full deviation. At 56 Mc. the maximum deviation would be 8×2000, or 16 kc.

Since 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, 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 (§ 8-8).

Speech amplification — The speech amplifier preceding the modulator follows ordinary design (§ 5-9), except that no power is required from it and the a.f. voltage taken by the modulator grid usually is small — not more than 10 or 15 volts, even with large modulator tubes. Because of these modest requirements, only a few speech-amplifier stages are needed; a two-stage amplifier consisting of a pentode followed by a triode, both resistance-coupled, will suffice for crystal microphones (§ 5-8).

R.f. amplifier stages — The frequency multiplier and output stages following the modulated oscillator may be designed and adjusted in accordance with ordinary principles. No special excitation requirements are imposed, since the amplitude of the output is constant. Enough frequency multiplication must be used to give the desired maximum deviation at the final frequency; this depends upon the maximum linear deviation available from the modulator-oscillator. All stages in the transmitter should be tuned to resonance, and careful neutralization (§ 4-7) of any straight amplifier stages is necessary to prevent r.f. phase shifts which might cause distortion.

Checking operation — The two quantities to be checked in the f.m. transmitter are linearity and frequency deviation. With a modulator of the type shown in Fig. 521, both the r.f. and a.f. voltages are small enough to make the operation Class A (§ 3-4), so that the plate current of the modulator is constant so long as operation is over the linear portions of the No. 1 and No. 3 grid characteristics. Hence, non-linearity will be indicated by a change in plate current as the a.f. modulating voltage is increased. The distortion will be within acceptable limits, with the tube and constants given in Fig. 521, when the plate current does not change more than 5 per cent with signal.

Non-linearity is accompanied by a shift in the carrier frequency, so it also can be checked by means of a selective receiver such as one with a crystal filter (§ 7-11). A tone source is convenient for the test. Set the receiver for high selectivity, switch on the beat oscillator, and tune to the oscillator carrier frequency. (The check does not need to be made at the output frequency and the oscillator frequency usually is more convenient, since it will fall within the tuning range of a communications receiver.) Increase the modulating signal until a definite shift in carrier frequency is observed; this indicates the point at which non-linearity starts. The modulating signal should be kept below the level at which carrier shift is observed, for minimum distortion.

A selective receiver also can be used to check frequency deviation, again at the oscillator frequency. A source of tone of known frequency is required, preferably a continuously variable calibrated audio oscillator or signal generator. Tune in the carrier as described above, using the beat oscillator and high selectivity, and adjust the modulating signal to the maximum level at which linear operation is secured. Starting with the lowest frequency available, slowly raise the tone frequency while listening closely to the carrier beat note. As the tone frequency is raised the beat note first will decrease in intensity, then disappear entirely at a definite frequency, and finally come back and increase in intensity as the tone frequency is raised still more. The frequency at which the beat note disappears, multiplied by 2.4, is the frequency deviation at that level of modulating signal; for example, if the beat note disappears with an 800-cycle tone, the deviation is 2.4×800, or 1920 cycles. The deviation at the output frequency is the oscillator deviation multiplied by the number of times the frequency is multiplied; in this example, if the oscillator is on 7 Mc. and the output on 56 Mc., the final deviation is 1920×8, or 15.36 kc.

The output of the transmitter can be checked for amplitude modulation by observing the antenna current. It should not change from the unmodulated carrier value when the transmitter is modulated. Where 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.

Chapter Six

Keying

6-1 Keying Principles and Characteristics

Requirements — The keying of a transmitter can be considered satisfactory if the method employed reduces the power output to zero when the key is open, or "up," and permits full power to reach the antenna when the key is closed, or "down." Furthermore, the keying system should accomplish this without producing keying transients or "clicks," which cause interference with other amateur stations and with local broadcast reception, and the keying process should not affect the frequency of the emitted wave.

Back-wave — From various causes, some energy may get through to the antenna during keying spaces. The effect then is as though the dots and dashes were only louder portions of a continuous carrier; in some cases, in fact, the *back-wave*, or signal heard during the keying spaces, may seem to be almost as loud as the keyed signal. Under these conditions the keying is hard to read. A pronounced back-wave often results when the amplifier stage feeding the antenna is keyed; it may be present because of incomplete neutralization (§ 4-7) of the final stage, allowing some energy to get to the antenna through the grid-plate capacity of the tube, or because of magnetic coupling between antenna coupling coils and one of the low-power stages.

A back-wave also may be radiated if the keying system does not reduce the input to the keyed stage to zero during keying spaces. This trouble will not occur in keying systems which cut off the plate voltage when the key is open, but may be present in grid-blocking systems (§ 6-3) if the blocking voltage is not great enough and in power-supply primary keying systems (§6-3) if only the final-stage power-supply primary is keyed.

Keying waveform and sidebands — A keyed c.w. signal can be considered equivalent to a modulated signal (§ 5-1), except that, instead of being modulated by sinusoidal waves and their harmonics, it is modulated by a rectangular wave, as in Fig. 601-A. If it were modulated by a sinusoidal wave of single frequency, as in Fig. 601-B, the only sidebands would be those equal to the carrier frequency plus and minus the modulation frequency (§ 5-2). A keying speed of 50 words per minute, sending sinusoidal dots, would give sidebands only 20 cycles either side of the carrier. However, when harmonics are present in the modulation the sidebands will extend out on both sides of the signal as far as the frequency of the highest harmonic. The rectangular wave form contains an infinite number of harmonics of the keying frequency, so a carrier modulated by truly rectangular dots would have sidebands covering the entire spectrum. Actually, the high-order harmonics are eliminated because of the selectivity of the tuned circuits (§ 2-10) in the transmitter, but there still is enough energy in the lower harmonics to extend the sidebands considerably. Considered from another viewpoint, whenever a pulse of current has a steep front (or back) high frequencies are certain to be present. If the pulse can be slowed down, or caused to *lag*, through a suitable filter circuit, the highest-order harmonics are filtered out.

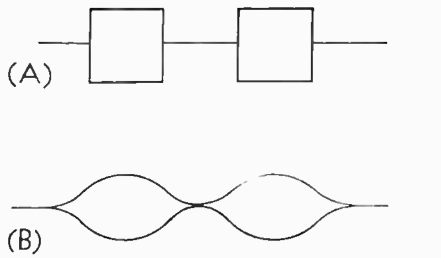


Fig. 601 — Extremes of possible keying wave-shapes A, rectangular characters; B, sine-wave characters.

Key clicks — Because the high-order harmonics exist only during the brief interval when the keying character is started or ended (when the amplitude of the keying wave is building up or dying down), their effects outside the normal communication channel are observed as pulses of very short duration. These pulses are called *key clicks*.

Tests have shown that practically all operators prefer to copy a signal which is "solid" on the "make" end of each dot or dash; i.e., one that does not build up too slowly but just slowly enough to have a slight click when the key is closed. The same tests indicate that the most pleasing and least difficult signal to copy, particularly at high speeds, is one that has a fairly soft "break" characteristic; i.e., one that has practically no click as the key is opened. A signal with heavy clicks on both make and break is difficult to copy at high speeds (and also causes considerable interference), but if it is too "soft" the dots and dashes will tend to run together. It is relatively simple to adjust the keying of a transmitter so that for all normal hand speeds (15 to 40 w.p.m.) the readability will be satisfactory while the keying still will not cause interference to reception of other signals near the frequency of the transmitter.

Break-in keying — In code transmission, there are definite intervals, between dots and dashes and between words, when no power is being radiated by the transmitter. It is possible, therefore, to allow the receiver to operate continuously and thus be capable of receiving incoming signals during the keying intervals.

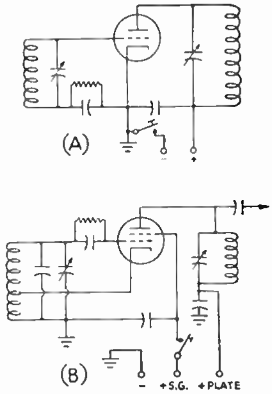


Fig. 602 — A, shows plate keying; B, screen grid keying. Oscillator circuits are shown in both cases, but the same keying methods can be used with amplifier circuits.

This practice facilitates communication, because the receiving operator can signal the transmitting operator, by holding down the key of his transmitter, whenever he has failed to copy part of the message, and thus obtain a repetition of the part that is missing without waiting until the end of the message. This is called *break-in* operation.

Frequency stability — Keying should have no effect upon the output frequency of a properly designed and adjusted transmitter. However, in many instances keying will cause a "chirp," or small frequency change, at the instant of closing or opening the key, which makes the signal difficult to read. Multistage transmitters keyed in a stage subsequent to the oscillator usually are free from this condition, unless the keying causes line-voltage changes which in turn affect the frequency of the oscillator. When the oscillator is keyed for break-in operation, special care must be taken to insure that the signal does not have keying chirps.

Selecting the stage to key — It is advantageous from an operating standpoint to design the c.w. transmitter for break-in operation. In ordinary cases this dictates that the oscillator be keyed, since a continuously running oscillator will create interference in the receiver and thus prevent break-in operation on or near the transmitter frequency. On the other hand, it is easier to avoid a chirpy signal by keying a buffer or amplifier stage. In either case, the tubes following the keyed stage must be provided with sufficient fixed bias to limit the plate currents to safe values when the key is up and the tubes are not being excited (§ 8-9). Complete cut-off reduces the possibility of a back-wave if a stage other than the oscillator is keyed, but the keying waveform is not as well preserved and some clicks can be introduced even though the keyed stage itself produces no clicks. It is a good general rule to bias the tubes so that they draw a key-up plate current equal to about 5 per cent of the normal key-down value.

Keyed power — The power broken by the key is an important consideration, both from the standpoint of safety for the operator and that of arcing at the key contacts. Keying the oscillator or a low-power stage is favorable in both respects. The use of a keying relay is highly recommended when a high-power circuit is keyed.

6-2 Keying Circuits

Plate-circuit keying — Any stage of the transmitter can be keyed by opening and closing the plate power circuit. Two methods are shown in Fig. 602. In A the key is in series with the negative lead from the plate power supply to the keyed stage. It could also be placed in the positive lead, although this is to be avoided whenever possible because the key is necessarily at the plate voltage above ground, and there is danger of shock unless a keying relay is used.

Fig. 602-B shows the key in the screen-supply lead of an electron-coupled oscillator. This can be considered to be a variation of plate keying.

Both the plate and screen-grid keying circuits, A and B of Fig. 602, respond well to the use of key-click filters, and are particularly suitable for use with crystal and self-controlled oscillators which are operated at low plate voltage and power input.

Power-supply keying — A variation of plate keying, in which the keying is introduced in the power-supply system itself, rather than in

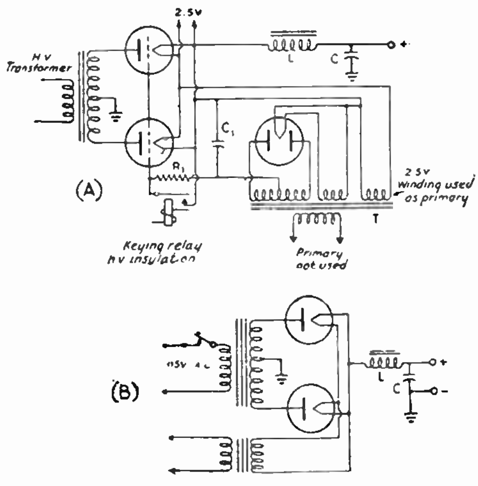


Fig. 603 — Power-supply keying. Grid-control rectifiers are used in A. Transformer *T* is a small multiple-secondary unit of the type used in receiver power supplies, and is used in conjunction with the full-wave rectifier tube to develop bias voltage for the grids of the high-voltage rectifiers. R_1 limits the load on the bias supply when the keying relay is closed: 50,000 ohms is a suitable value. C_1 may be 0.1 µfd. or larger. *L* and *C* constitute the smoothing filter for the high-voltage supply in both circuits. B shows direct keying of the transformer primary.

the connections between the power supply and transmitter, is illustrated by the diagrams in Fig. 603.

Fig. 603-A shows the use of grid-controlled rectifier tubes (§ 3-5) in the power supply. Keying is accomplished by applying suitable bias to the grids to cut off plate current flow when the key is open, and by removing the bias when the key is closed. Since in practice this circuit is used only with high-powered high-voltage supplies, a well-insulated keying relay is a necessity.

Direct keying of the primary of the plate power transformer for the keyed stage or stages is shown in Fig. 603-B. This and the method at A inherently have a keying lag because of the time constant (§ 2-6) of the smoothing filter. If enough filter is provided to reduce ripple to a low percentage (§ 8-4) the lag (§ 6-1) is too great to permit crisp keying at speeds above about 25 words per minute, although this type of keying is very effective in eliminating key clicks. A single-section plate-supply filter (§ 8-6) is about the most elaborate type that can be used if a reasonably good keying characteristic is to be achieved.

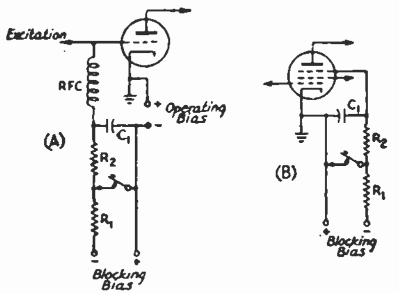


Fig. 604 — Blocked-grid keying. R_1, the current-limiting resistor, should have a value of about 50,000 ohms. C_1 may have a capacity of 0.1 to 1 μfd., depending upon the keying characteristic desired. R_2 also depends on the performance characteristic desired, values being of the order of 5000 to 10,000 ohms in most cases.

Blocked-grid keying — Keying may be accomplished by applying sufficient negative bias voltage to a control or suppressor grid to cut off plate current flow when the key is open, and by removing this *blocking* bias when the key is closed. The blocking bias voltage must be sufficient to overcome the r.f. grid voltage, in the case where the bias is applied to the control grid, and hence must be considerably higher than the nominal cut-off value for the tube at the operating d.c. plate voltage. The fundamental circuits are shown in Fig. 604.

In both circuits the key is connected in series with a resistor, R_1, which limits the current drain on the blocking-bias source when the key is closed. R_2C_1 is a resistance-capacity filter (§ 2-11) for controlling the lag on make and break of the key circuit. The lag increases as the time constant (§ 2-6) of this circuit is made larger. Since grid current flows through R_2 when the key is closed in Fig. 604-A, additional operating bias is developed, hence somewhat less bias is needed from the regular bias supply. The operating and blocking biases can be obtained from the same supply, if desired, by

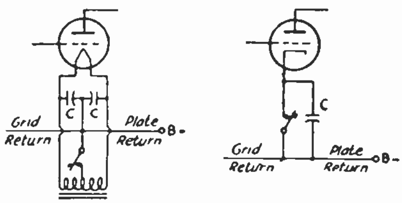


Fig. 605 — Center-tap and cathode keying. The condensers, C, are r.f. by-pass condensers. Their capacity is not critical, values of 0.001 to 0.01 μfd. ordinarily being used.

utilizing suitable taps on a voltage divider (§ 8-10). For circuits in which no fixed bias is used R_2 can be the regular grid leak (§ 3-6) for the stage.

With blocked-grid keying a relatively small direct current is broken as compared to other systems. Thus any sparking at the key is reduced. The keying characteristic (lag) readily can be controlled by a suitable choice of values for C_1 and R_2.

Cathode keying — Opening the d.c. circuits of both plate and grid simultaneously is called *cathode keying*. It is usually called *center-tap* keying with a directly heated filament-type tube, since in this case the key is placed in the filament-transformer center-tap lead. Typical circuits for this type of keying are shown in Fig. 605.

Cathode keying results in less sparking at the key contacts, for the same plate power, as compared with keying in the plate-supply lead. When used with an oscillator it does not respond as readily to key-click filtering (§ 6-3) as does plate keying, but there is little difference in this respect between the two systems when an amplifier is keyed.

ℂ 6-3 Key-Click Reduction

R.f. filters — A spark at the key contacts, even though minute, will cause a damped oscillation to be set up in the keying circuit which may modulate the transmitter output or may simply be radiated by the wiring in the keying circuit. Interference from the latter source is usually confined to the immediate vicinity of the transmitter, and is similar in nature and effects to the click which is frequently heard in a receiver when an electric light is turned on or off. It can be minimized by isolating the key from the wiring by means of a low-pass filter (§ 2-11), which usually consists of an r.f. choke in each key lead, placed as close as possible to the key, and by-passed on the keying-line side by a condenser, as shown in Fig. 606. Suitable values must be determined by experiment. Choke values may range from 2.5 to 80 millihenrys, and condenser capacities from 0.001 to 0.1 μfd.

This type of r.f. filter is required in nearly every keying installation, in addition to the lag circuits which are discussed in the next paragraph.

Lag circuits — A filter used to give a desired shape to the keying character, to eliminate unnecessary sidebands and consequent interference, is called a *lag circuit*. In one form, suitable for the circuits of Figs. 602 and 605, it consists of a condenser across the key terminals and an inductance in series with one of the leads. This is shown in Fig. 607. The optimum values of capacity and inductance must be found by experiment, but are not especially critical. If a high-voltage low-current circuit is being keyed a small condenser and large inductance will be necessary, while if a low-voltage high-current circuit is keyed the capacity required will be high and the inductance small. For example, a 300-volt 6-ma. circuit will require about 30 henrys and 0.05 μfd., while a 300-volt 50-ma. circuit needs about 1 henry and 0.5 μfd. For any given circuit and fixed values of current and voltage, increasing the inductance will reduce the clicks on "make" and increasing the capacity will reduce the clicks on "break."

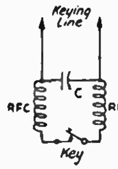


Fig. 606 — R.f. filter used for eliminating the effects of sparking at key contacts. Suitable values for best results with individual transmitters must be determined by experiment. Values for *RFC* range from 2.5 to 80 millihenries and for *C* from 0.001 to 0.1 μfd.

Blocked-grid keying is adjusted by changing the values of resistors and condensers in the circuit. In Fig. 604, the click on "make" is reduced by increasing the capacity of C_1, and the click on break is reduced by increasing C_1 and/or R_2. The values required for individual installations will vary with the amount of blocking voltage and the grid current. The constants given in Fig. 604 will serve as a first approximation.

Tube keying — A tube keyer is a convenient adjunct to the transmitter, because it allows the keying characteristic to be adjusted easily without necessitating condenser and inductance values which may not be readily available. It uses the plate resistance of a tube (or tubes in parallel) to replace the key in a plate or cathode circuit, the keyer tube (or tubes) being keyed by the blocked-grid method (§ 6-2). A typical circuit is shown in Fig. 608. Type 45 tubes are suitable because of their low plate resistance and consequent small voltage drop between plate and cathode. When a tube keyer is used to replace the key in a plate or cathode circuit, the power output of the stage will be somewhat reduced because of the voltage drop across the keyer tube, but this can be compensated for by a slight increase in the supply voltage. The use of a tube keyer makes the key itself entirely safe to handle, since the high resistance in series with the key and blocking voltage prevents possible danger of shock through contact with high-voltage circuits.

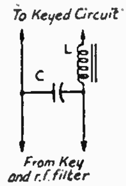
ℭ 6-4 Checking Transmitter Keying

Clicks — Transmitter keying can be checked by listening to the signal on a superheterodyne receiver. The antenna should be disconnected, so that the receiver does not overload, and, if necessary, the r.f. gain may be reduced as well. Listening with the beat oscillator and a.v.c. off, the keying should be adjusted so that a slight click is heard as the key is closed but practically none can be heard when the key is released. When the keying constants have been adjusted to meet this condition, the clicks will be about optimum for all normal amateur work. If the clicks are too pronounced, they will cause interference with other amateur transmissions, and possibly to nearby broadcast receivers.

Chirps — Keying chirps (instability) may be checked by tuning in the signal or one of its harmonics on the highest frequency range of the receiver and listening with the b.f.o. on and the a.v.c. off. The gain should be sufficient to give moderate signal strength, but it should be low enough to preclude the possibility of overloading. Adjust the tuning to give a low-frequency beat note and key the transmitter. Any chirp introduced by the keying adjustment will be readily apparent. Listening to a harmonic will magnify the effect of any instability by the order of the harmonic, and thus make it more perceptible.

Oscillator keying — The keying of an amplifier is relatively straightforward and requires no special treatment, but a few additional precautions will be found necessary with oscillator keying. Any oscillator, either self-excited or crystal, will key well if it will oscillate at low plate voltages (of the order of one or two volts) and if its change in frequency with plate-voltage change is negligible. A crystal oscillator will oscillate at low plate voltages if a regenerative type of circuit such as the Tri-tet or grid-plate (§ 4-5) is used and if an r.f. choke is connected in series with the grid leak, to reduce loading on the crystal. Crystal oscillators of this type generally are free from chirp unless there is a relatively large air-gap between the crystal and top plate of the crystal holder, as is the case with a variable-frequency crystal set at the high-frequency end of its range.

Fig. 607 — Lag circuit used for shaping the keying character to eliminate unnecessary sidebands. Actual values for any given circuit must be determined by experiment, and may range from 1 to 30 henries for *L* and from 0.05 to 0.5 μfd. for *C*, depending on the keyed current.



Self-controlled oscillators can be made to meet the same requirements by using a high *C/L* ratio in the tank circuit, low plate and screen currents, and judicious feed-back adjustment (§ 3-7). A self-controlled oscillator intended to be keyed should be designed for good keying rather than maximum output.

Stages following keying — When a keying filter is being adjusted, the stages following the keyed tube should be made inoperative by removing the plate voltage. This facilitates monitoring the keying without the introduction of additional effects. The following stages should then be added, one at a time, checking the keying after each addition. An increase in click intensity (for the same carrier strength) indicates that the clicks are being added in the stages following the one being keyed. The fixed bias on such stages should be sufficient to reduce the idling plate current (no excitation) to a low value, but not to zero. Under these conditions, any instability or tendency toward parasitic oscillations, either of which can adversely affect the keying characteristic, usually will evidence itself.

Monitoring of keying — Most operators find a keying monitor helpful in developing and maintaining a good "fist," especially if a "bug" or semi-automatic key is used. While several types have been devised, the most popular consists of an audio oscillator the output of which is coupled to the receiver loud speaker or headphones, and which is keyed simultaneously with the transmitter. Fig. 609 shows the circuit diagram of a simple keying-monitor oscillator. The plate voltage, as well as the heater voltage, is supplied by a 6.3-volt filament transformer. One section of the 6F8G dual triode is used as the rectifier to supply d.c. for the plate of the second section, which is used as the oscillator. A change in the value of R_1 will alter the output tone. The output terminal labeled *Gnd* should be connected directly to the receiver chassis, while P_1 should be connected to the "hot" side of the headphones. Shunting of the 'phones by the oscillator may cause some loss of volume on received signals, unless the coupling capacity, C_3, is made sufficiently small. However, the capacity should be made large enough to provide good transfer of the oscillator signal.

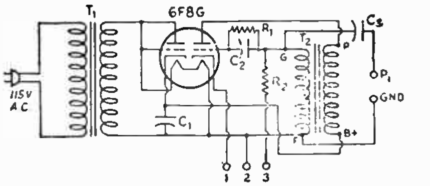


Fig. 609 — Circuit diagram of a keying monitor of the audio-oscillator type, with self-contained power supply

C_1 — 25-μfd. 25-volt electrolytic.
C_2 — 250-μμfd. mica.
C_3 — Approximately 0.01 μfd. (see text).
R_1 — 0.15 megohm, ½ watt.
R_2 — Approximately 0.1 megohm, 1 watt (see text).
T_1 — 6.3-volt 1-ampere filament transformer.
T_2 — Small audio transformer, interstage type.

If the transmitter oscillator is keyed for break-in, the keying terminals of the oscillator may be connected in parallel with those of the transmitter. With cathode keying, terminals 1 and 2 will be connected across the key, with terminal 2 going to the ground side of the key. With blocked-grid keying, terminals 2 and 3 go to the key and a resistance of 0.1 megohm or so is inserted in series with terminal 3.

Electronic keys — Several electronic circuits have been devised for producing automatic dots and dashes. A typical example is shown in Fig. 610. The values provide for a maximum speed of 60 w.p.m. with a 300-volt supply. R_1 and R_2 should be of the same type and ganged to form the speed control. To adjust for proper operation, ground the right cathode and adjust R_7 until the left plate current is zero. Do the same thing with the sections reversed, biasing the right section to cut-off temporarily. Adjust R_5 until the plate voltages are equal. Return the circuit to normal and check the average plate voltages with the key on the "dot" side. If they are unequal, adjust a fixed resistor connected in series with R_1 or R_2 until they are equal. On dashes, the plate voltage of the right section should drop one-third and that of the left section should increase by one-third. Adjust the size of C_3 until this condition is met. (See *QST* for March, 1944.)

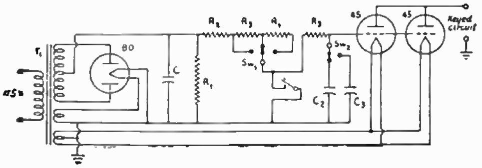
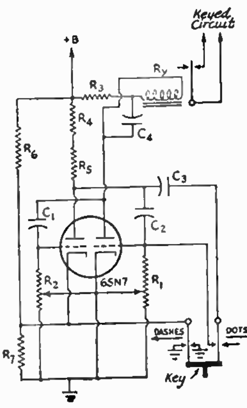


Fig. 608 — Vacuum-tube keyer circuit. The voltage drop across the tubes will be approximately 90 volts with the two Type 45 tubes shown, when the keyed current is 100 milliamperes. More tubes can be connected in parallel to reduce the drop. Suggested values are as follows:

C_1 — 2-μfd. 600-volt paper.
C_2 — 0.003-μfd. mica.
C_3 — 0.005-μfd. mica.
R_1 — 0.25 megohm, 2 watt.
R_2 — 50,000 ohms, 10 watt.
R_3, R_4 — 5 megohms, ½ watt.
R_5 — 0.5 megohm, ½ watt.
Sw_1, Sw_2 — 1-circuit 3-position rotary switch.
T_1 — Power transformer, 325 volts each side of center-tap, with 5-volt and 2.5-volt filament windings.

A wider range of lag adjustment can be obtained by using additional resistors and condensers. Suggested values of capacity, in addition to C_2 and C_3, are 0.001 and 0.002 μfd. Resistors in addition to R_2 could be 2, 2, 3 and 5 megohms. More switch positions will be required.

Fig. 610 — A multivibrator-type electronic key.



C_1, C_2 — 0.005-μfd. mica.
C_3 — 0.01-μfd. 400-volt paper.
C_4 — 0.01 — μfd., approximately.
R_1, R_2 — 2-megohm variable (see text).
R_3, R_4 — 50,000 ohms, ½ watt.
R_5 — 3000 ohms (or resistance equal to resistance of R_y).
R_6 — 0.25 megohm, ½ watt.
R_7 — 75,000 ohms, ½ watt.
R_y — Sensitive relay (Eby).

Chapter Seven

Receiver Principles & Design

7-1 Elements of Receiving Systems

Basic requirements — The purpose of a radio receiving system is to abstract energy from passing radio waves and convert it into a form which conveys the intelligence contained in the transmitted signal. The receiver also must be able to select a desired signal and eliminate those not wanted. The fundamental processes involved are those of amplification and detection.

Detection — The high frequencies used for radio signaling are well beyond the audio-frequency range (§ 2-7), and therefore cannot be used to actuate a loudspeaker directly. Neither can they be used to operate other devices, such as relays, by means of which a message might be transmitted. The process of converting a modulated radio-frequency wave to a usable low frequency, called *detection* or *demodulation*, is essentially that of rectification (§ 3-1). The modulated carrier (§ 5-1) is thereby converted to a unidirectional current, the amplitude of which will vary at the same rate as the modulation. These low-frequency variations are readily amplified, and can be applied to the headphones, loudspeaker or other form of electromechanical device.

Code signals — The dots and dashes of code (c.w.) transmissions are rectified as described, but in themselves can produce no audible tone in the headphones or loudspeaker because they are of constant amplitude. For aural reception 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 (§ 2-13). The frequency difference, and hence the *beat note*, is generally of the order of 500 to 1000 cycles, since these tones are within the range of optimum response of both the ear and the headset. If the source of the second radio frequency is a separate oscillator, the system is known as *heterodyne* reception; if the detector itself is made to oscillate and produce the second frequency, it is known as an *autodyne* detector.

Amplification — To build up weak signals to usable output level, modern receivers employ considerable amplification — often of the order of hundreds of thousands of times. Amplifiers are used at the frequency of the incoming signal (*r.f. amplifiers*), after detection (*a.f. amplifiers*), and, in superheterodyne receivers, at one or more intermediate radio frequencies (*i.f. amplifiers*). R.f. and i.f. amplifiers practically always employ tuned circuits.

Types of receivers — Receivers may vary in complexity from a simple detector with no amplification to multi-tube arrangements having amplification at several different radio frequencies as well as at audio frequency. A regenerative detector (§ 7-4) with or without audio-frequency amplification (§ 7-5) is known as a *regenerative receiver;* if the detector is preceded by one or more tuned r.f. amplifier stages (§ 7-6), the combination is known as a *t.r.f. (tuned radio frequency) receiver*. The *superheterodyne receiver* (§ 7-8) employs r.f. amplification at a fixed intermediate frequency as well as at the frequency of the signal itself, the latter being converted by the heterodyne process to the intermediate frequency.

At very-high frequencies the superregenerative detector (§ 7-4), usually with audio amplification, is used in the *superregenerative receiver* or *superregenerator*, providing large amplification of weak signals with simple circuit arrangements.

7-2 Receiver Characteristics

Sensitivity — Sensitivity is defined as the strength of the signal (usually expressed in microvolts) which must be applied to the input terminals of the receiver to produce a specified audio-frequency power output at the loudspeaker or headphones (§ 7-5). It is a measure of the amplification or gain of the receiver.

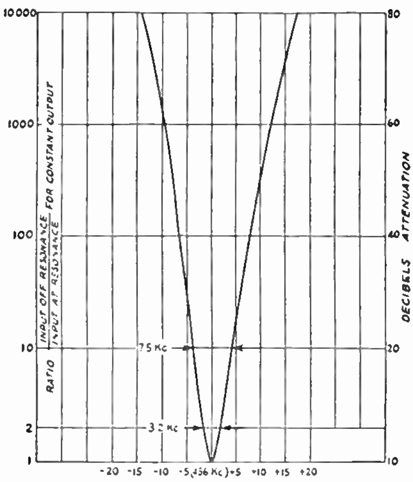


Fig. 701 — Selectivity curve of a modern superheterodyne receiver. Relative response is plotted against deviations above and below the resonance frequency. The scale at the left is in terms of voltage ratios; the corresponding decibel steps are shown at the right.

Signal-to-noise ratio — Every receiver generates some noise of a hiss-like character, and signals weaker than the noise cannot be separated from it no matter how much amplification is used. This relation between noise and a weak signal is expressed by the term *signal-to-noise* ratio. It can be defined in various ways, one simple way being to give it as the ratio of signal power output to noise output from the receiver at a specified value of modulated carrier voltage applied to the input terminals.

The hiss-like noise mentioned above is inherent in the circuits and tubes of the receiver, and its amplitude depends upon the selectivity of the receiver. The greater the selectivity the smaller the noise, other things being equal (§ 7-6). In addition to inherent receiver noise, atmospheric electricity (natural "static") and electrical devices in the vicinity of the receiver also cause noise which adversely affects the signal-to-noise ratio.

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 which gives the ratio of signal strength required at various frequencies off resonance to the signal strength at resonance, to give constant output. A *resonance curve* of this type (taken on a typical communications-type superheterodyne receiver) is shown in Fig. 701. The *band-width* is the width of the resonance curve (in cycles or kilocycles) of a receiver at a specified ratio; in Fig. 701, the band-widths are indicated for ratios of response of 2 and 10 ("2 times down" and "10 times down").

Selectivity for signals within a few kilocycles of the desired-signal frequency is called *adjacent-channel* selectivity, to distinguish it from the discrimination against signals considerably removed from the desired frequency.

Stability — The stability of a receiver is its ability to give constant output, over a period of time, from a signal of constant strength and frequency. Primarily, it means the ability to stay tuned to a given signal. However, a receiver which at some settings of its controls has a tendency to break into oscillation, or "howl," also is said to be unstable.

The stability of a receiver is affected principally by temperature variations, supply-voltage changes, and constructional features of a mechanical nature.

Fidelity — Fidelity is the relative ability of the receiver to reproduce in its output the modulation (keying, 'phone, etc.) carried by the incoming signal. For exact reproduction the band-width must be great enough to accommodate the highest modulation frequency transmitted, and the relative amplitudes of the various frequency components within the band must not be changed in the output.

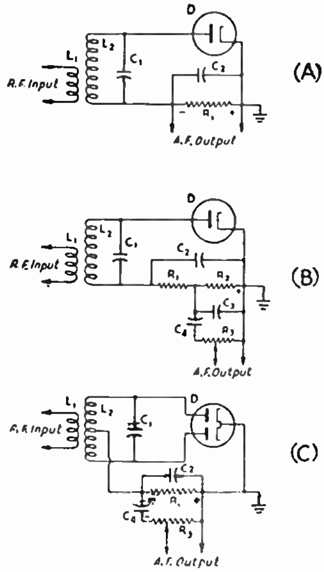


Fig. 702 — Simplified and practical diode detector circuits. A, the elementary half-wave diode detector; B, a practical circuit, with r.f. filtering and audio output coupling; C, full-wave diode detector, with output coupling indicated. The circuit, L_2C_1, is tuned to the signal frequency; typical values for C_2 and R_1 in A and B 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.

ℂ 7-3 Detectors

Characteristics — The important characteristics of a detector are its sensitivity, fidelity or linearity, resistance or impedance, and signal-handling capability.

Detector *sensitivity* is the ratio of audio-frequency output to radio-frequency input. *Linearity* is a measure of the ability of the detector to reproduce, as an audio frequency, the exact form of the modulation on the incoming signal. The *resistance* or *impedance* of the detector is important in circuit design, since a relatively low resistance means that power is consumed in the detector. The *signal-handling capability* means the ability of the detector to accept signals of a specified amplitude without overloading.

Diode detectors — The simplest detector is the diode rectifier. Circuits for both half-wave and full-wave (§ 8-3) diodes are given in Fig. 702. The simplified half-wave circuit at 702-A 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 through R_1 causes a d.c. voltage to develop across its terminals, and this voltage varies with the modulation on the signal. The − and + signs show the polarity of the voltage. The variation in amplitude of the r.f. signal with modulation causes corresponding variations in the value of the d.c. voltage across R_1. The load resistor, R_1, usually

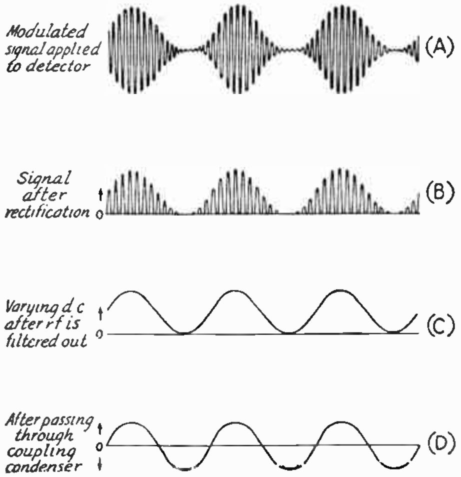


Fig. 703 — Diagrams showing the detection process.

has a rather high value of resistance, so that a fairly large voltage will develop from a small rectified-current flow.

The progress of the signal through the detector or rectifier is shown in Fig. 703. A typical modulated signal as it exists in the tuned circuit is shown at A. When applied to the rectifier tube, current flows from plate to cathode only during the part of the r.f. cycle when the plate is positive with respect to the cathode, so that the output of the rectifier consists of half-cycles of r.f. still modulated as in the original signal. These current "pulses" flow in the load circuit comprised of R_1 and C_2, the resistance of R_1 and the capacity of C_2 being so proportioned that C_2 charges to the peak value of the rectified voltage on each pulse and retains enough charge between pulses so that the voltage across R_1 is smoothed out, as shown in C. C_2 thus acts as a filter for the radio-frequency component of the output of the rectifier, leaving a d.c. component which 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. 702-B), only the *variations* in voltage are transferred, so that the final output signal is a.c., as shown in D.

In the circuit at 702-B, R_1 and C_2 have been divided for the purpose of providing a more effective filter for r.f. It is important to prevent the appearance of any r.f. voltage in the output of the detector, because it may cause overloading of a succeeding amplifier tube. The audio-frequency variations can be transferred to another circuit through a coupling condenser, C_4 in Fig. 702, to a load resistor, R_3, which usually is a "potentiometer" (§ 8-10) so that the volume can be adjusted to a desired level.

The full-wave diode circuit at 702-C 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.

The reactance of C_2 must be small compared to the resistance of R_1 at the radio frequency being rectified, but at audio frequencies must be relatively large compared to R_1 (§ 2-8, 2-13). This condition is satisfied by the values shown. If the capacity of C_2 is too large, response at the higher audio frequencies will be lowered.

Compared with other detectors, the sensitivity of the diode is low. Since the diode consumes power, the Q of the tuned circuit is reduced, bringing about a reduction in selectivity (§ 2-10). The linearity is good, however, and the signal-handling capability is high.

Grid-leak detectors — The grid-leak detector is a combination diode rectifier and audio-frequency amplifier. In the circuit of Fig. 704-A, the grid corresponds to the diode plate and the rectifying action is exactly the same as just described. The d.c. voltage from rectified-current flow through the grid leak, R_1, biases the grid negatively with respect to cathode, and the audio-frequency variations in voltage across R_1 are amplified through the tube just as in a normal a.f. amplifier. In the plate circuit, R_2 is the plate load resistance (§ 3-3) and C_3 is a by-pass condenser to elim-

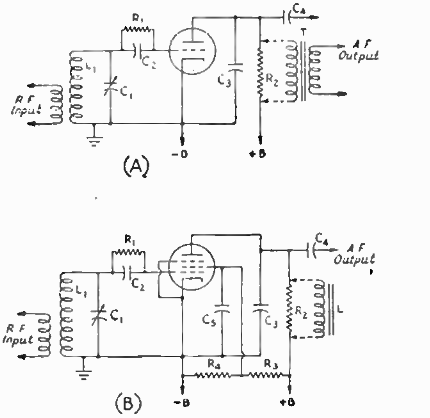


Fig. 704 — Grid-leak detector circuits, A, triode; B, pentode. A tetrode may be used in the circuit of B by neglecting the suppressor-grid connection. Transformer coupling may be substituted for resistance coupling in A, or a high-inductance choke may replace the plate resistor in B. L_1C_1 is a circuit tuned to the signal frequency. The grid leak, R_1, may be connected directly from grid to cathode instead of across the grid condenser as shown. The operation with either connection will be the same. Representative values for components are:

Component	*Circuit A*	*Circuit B*
C_2	100 to 250 μμfd.	100 to 250 μμfd.
C_3	0.001 to 0.002 μfd.	250 to 500 μμfd.
C_4	0.1 μfd.	0.1 μfd.
C_5		0.5 μfd. or larger.
R_1	1 to 2 megohms.	1 to 5 megohms.
R_2	50,000 ohms.	100,000 to 250,000 ohms.
R_3		50,000 ohms.
R_4		20,000 ohms.
T	Audio transformer.	
L		500-henry choke.

The plate voltage in A should be about 50 volts for best sensitivity. In B, the screen voltage should be about 30 volts and the plate voltage from 100 to 250.

inate r.f. in the output circuit. C_4 is the output coupling condenser. With a triode, the load resistor, R_2, may be replaced by an audio transformer, T, in which case C_4 is not used.

Since audio amplification is added to rectification, the grid-leak detector has considerably greater sensitivity than the diode. The sensitivity can be further increased by using a screen-grid tube instead of a triode, as at 704-B. The operation is equivalent to that of the triode circuit. The screen by-pass condenser, C_5, should have low reactance (§ 2-8, 2-13) for both radio and audio frequencies. R_3 and R_4 constitute a voltage divider (§ 8-10) from the plate supply to furnish the proper d.c. voltage to the screen. In both circuits, C_2 must have low r.f. reactance and high a.f. reactance compared to the resistance of R_1; the same applies to C_3 with respect to R_2.

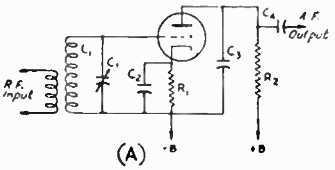
Because of the high plate resistance of the screen-grid tube (§ 3-5), transformer coupling from the plate circuit of a screen-grid detector is not satisfactory. An impedance (L in Fig. 704-B) can be used in place of a resistor, with a gain in sensitivity because a high value of load impedance can be developed with little loss of plate voltage as compared to the voltage drop through a resistor. The coupling coil, L_2, for a screen-grid detector should have an inductance of the order of 300 to 500 henrys.

The sensitivity of the grid-leak detector is higher than that of any other type. Like the diode, it "loads" the tuned circuit and reduces its selectivity. The linearity is rather poor, and the signal-handling capability is limited.

Plate detectors — The plate detector is arranged so that rectification of the r.f. signal takes place in the plate circuit of the tube, as contrasted to the grid rectification just described. Sufficient negative bias is applied to the grid to bring the plate current nearly to the cut-off point, so that the application of a signal to the grid circuit causes an increase in average plate current. The average plate current follows the changes in signal amplitude in a fashion similar to the rectified current in a diode detector.

Circuits for triodes and pentodes are given in Fig. 705. C_3 is the plate by-pass condenser, R_1 is the cathode resistor which provides the operating grid bias (§ 3-6), and C_2 is a by-pass for both radio and audio frequencies across R_1 (§ 2-13). R_2 is the plate load resistance (§ 3-3), across which a voltage appears as a result of the rectifying action described above. C_4 is the output coupling condenser. In the pentode circuit at B, R_3 and R_4 form a voltage divider to supply the proper potential (about 30 volts) to the screen, and C_5 is a by-pass condenser between screen and cathode. C_5 must have low reactance for both radio and audio frequencies.

In general, transformer coupling from the plate circuit of a plate detector is not satisfactory, because the plate impedance even of a triode is very high when the bias is set near the plate-current cut-off point (§ 3-2, 3-3). Impedance coupling may be used in place of the resistance coupling shown in Fig. 705. The same order of inductance is required as with the screen-grid detector described previously.



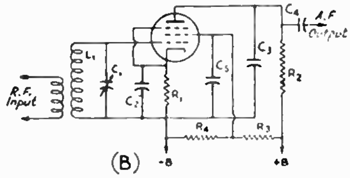


Fig. 705 — Circuits for plate detection. A, triode; B, pentode. The input circuit, L_1C_1, is tuned to the signal frequency. Typical values for the other constants are:

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.

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

The plate detector is more sensitive than the diode since there is some amplifying action in the tube, but less so than the grid-leak detector. It will handle considerably larger signals than the grid-leak detector, but is not quite so tolerant in this respect as the diode. Linearity, with the self-biased circuits shown, is good. Up to the overload point the detector takes no power from the tuned circuit, and so does not affect its Q and selectivity (§ 2-10).

Infinite-impedance detector — The circuit of Fig. 706 combines the high signal-handling capabilities of the diode detector with low distortion (good linearity), and, like the plate detector, does not load the tuned circuit to which it is connected. The circuit resembles that of the plate detector, except that the load resistance, R_1, is connected between cathode and ground and thus is common to both grid and plate circuits, giving negative feed-back for the audio frequencies. The cathode resistor is by-passed for r.f. (C_1) but not for audio (§ 2-13), while the plate circuit is by-passed to ground for both audio and radio frequencies. R_2 forms, with C_3, an RC filter (§ 2-11) to isolate the plate from the "B" supply at a.f.

The plate current is very low at no signal, increasing with signal as in the case of the plate detector. The voltage drop across R_1 similarly increases with signal, because of the increased plate current. Because of this and the fact that the initial drop across R_1 is large, the grid cannot be driven positive with respect to the cathode by the signal, hence no grid current can be drawn.

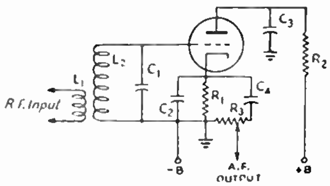


Fig. 706 — The infinite-impedance linear detector. The input circuit, L_2C_1, is tuned to the signal frequency. Typical values for the other constants are:

C_2 — 250 $\mu\mu$fd.	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.

§ 7-4 Regenerative Detectors

Circuits — By providing controllable r.f. feed-back or regeneration (§ 3-3) 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 hence increases the selectivity (§ 2-10) by virtue of the fact that the maximum regenerative amplification takes place only at the frequency to which the circuit is tuned. The grid-leak type of detector is most suitable for the purpose. Except for the regenerative connection, the circuit values are identical with those previously described for this type of detector, and the same considerations apply. The amount of regeneration must be controllable, because maximum regenerative amplification is secured at the critical point where the circuit is just about to oscillate (§ 3-7) and the critical point in turn depends upon circuit conditions, which may vary with the frequency to which the detector is tuned.

Fig. 707 shows the circuits of regenerative detectors of various types. The circuit of A is for a triode tube, with a variable by-pass condenser, C_3, in the plate circuit to control regeneration. When the capacity is small the tube does not regenerate, but as it increases toward maximum its reactance (§ 2-8) becomes smaller until a critical value is reached where there is sufficient feed-back to cause oscillation. If L_2 and L_3 are wound end-to-end in the same direction, the plate connection is to the outside of the plate or "tickler" coil, L_3, when the grid connection is to the outside of L_2.

The circuit of B is for a screen-grid tube, regeneration being controlled by adjustment of the screen-grid voltage. The tickler, L_3, is in the plate circuit. The portion of the control resistor between the rotating contact and ground is by-passed by a large condenser (0.5 μfd. or more) to filter out scratching noise when the arm is rotated (§ 2-11). The feed-back is adjusted by varying the number of turns on L_3 or the coupling (§ 2-11) between L_2 and L_3, until the tube just goes into oscillation at a screen voltage of approximately 30 volts.

Circuit C is identical with B in principle of operation, except that the oscillating circuit is of the Hartley type (§ 3-7). 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.

Adjustment for smooth regeneration — 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 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 which gives maximum sensitivity. Should the tube break into oscillation suddenly as the regeneration control is advanced, making a click, the operation often can be made smoother by changing the grid-leak resistance to a higher or lower value. The wrong grid leak plus too-high plate and screen voltage are the most frequent causes of lack of smoothness in going into oscillation.

Antenna coupling — If the detector is coupled to an antenna, slight changes in the antenna constants (as when the wire swings in a breeze) affect the frequency of the oscillations generated, and thereby the beat frequency when c.w. signals are being received. The tighter the antenna coupling is made, the greater will be the feed-back 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 (§ 2-11) to the grid end of the coil is used, only a very small amount of capacity will be needed to couple to the antenna. Increasing the capacity increases the coupling.

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

Body capacity — A regenerative detector occasionally shows a tendency to change frequency slightly as the hand is moved near the dial. This condition (*body capacity*) can be caused by poor design of the receiver, or by the antenna if the detector is coupled directly to it. If body capacity is present when the antenna is disconnected, it can be eliminated by better shielding, and sometimes by r.f. filtering of the 'phone leads. Body capacity which is present only when the antenna is connected is caused by resonance effects in the antenna, which tend to cause a portion of a standing wave (§ 2-12) of r.f. voltage to appear on the ground lead and thus raise the whole detector circuit above ground potential. A good, short ground connection should be made to the receiver and the length of the antenna varied electrically (by adding a small coil or variable condenser in the antenna lead) until the effect is minimized. Loosening the coupling to the antenna circuit also will help.

Hum — Hum at the power-supply frequency may be present in a regenerative detector, especially when it is used in an oscillating condition for c.w. reception, even though the plate supply itself is free from ripple (§ 8-4). The hum may result from the use of a.c. on the tube heater, but effects of this type normally are troublesome only when the circuit of Fig. 707-C is used, and then only at 14 Mc. and higher frequencies. Connecting one side of the heater supply to ground, or grounding the center-tap of the heater transformer winding, is good practice to reduce hum, and the heater wiring should be kept as far as possible from the r.f. circuits.

House wiring, if of the "open" type, will have a rather extensive electrostatic field which may cause hum if the detector tube, grid lead, and grid condenser and leak are not electrostatically shielded. This type of hum is easily recognizable because of its rather high pitch, a result of harmonics (§ 2-7) in the power-supply system. The hum is caused by a species of grid modulation (§ 5-4).

Antenna resonance effects frequently cause a hum of the same nature as that just described which is most intense at the various resonance points, and hence varies with tuning. For this reason it is called tunable hum. It is prone to occur with a rectified a.c. plate supply (§ 8-1) when a standing wave effect of the type described in the preceding paragraph occurs, and is associated with the non-linearity of the rectifier tube in the plate supply. Elimination of antenna resonance effects as described and by-passing the rectifier plates to cathode (using by-pass condensers of the order of 0.001 µfd.) usually will cure it.

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

The proper adjustment of the regeneration control for best reception of c.w. signals is where the detector just starts to oscillate, when it will be found that c.w. signals can be tuned in and will give a tone with each signal depending on the setting of the tuning control. As the receiver is tuned through a signal the tone first will be heard as a very high pitch, then will go down through "zero beat" (the region where the frequencies of the incoming signal and the oscillating detector are so nearly alike that the difference or beat is less than the lowest audible tone) and rise again on the other side, finally disappearing at a very high pitch. This behavior is shown in Fig. 708. It will be found that a low-pitched beat-note cannot be obtained from a strong signal because the detector "pulls in" or "blocks"; that is, the signal tends to control the detector in such a way that the latter oscillates at the signal frequency, despite the fact that the circuit may not be tuned exactly to resonance. This phenomenon, commonly observed when an oscillator is coupled to a source of a.c. voltage of approximately the

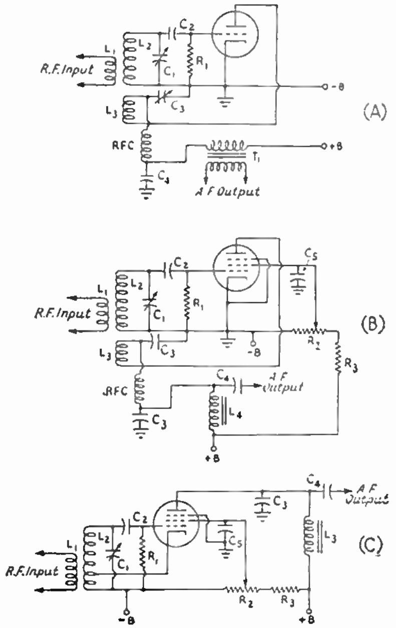


Fig. 707 — Triode and pentode regenerative detector circuits. The input circuit, L_2C_1, is tuned to the signal frequency. The grid condenser, C_2, should have a value of about 100 µµfd. in all circuits; the grid leak, R_1, may range in value from 1 to 5 megohms. The tickler coil, L_3, ordinarily will have from 10 to 25 per cent of the number of turns on L_2; in C, the cathode tap is about 10 per cent of the number of turns on L_2 above ground. Regeneration control condenser C_3 in A should have a maximum capacity of 100 µµfd. or more; by-pass condensers C_3 in B and C are likewise 100 µµfd. C_5 is ordinarily 1 µfd. or more; R_2, a 50,000-ohm potentiometer; R_3, 50,000 to 100,000 ohms. L_4 in B (L_3 in C) is a 500-henry inductance, C_4 is 0.1 µfd. in both circuits. T_1 in A is a conventional audio transformer for coupling from the plate of a tube to a following grid. *RFC* is 2.5 mh. In A, the plate voltage should be about 50 volts for best sensitivity. Pentode circuits require about 30 volts on the screen; plate voltage may be 100 to 250 volts.

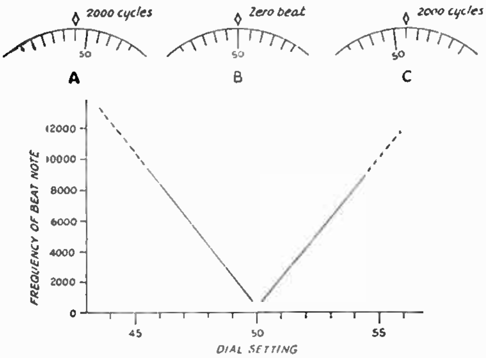


Fig. 708 — **As the tuning dial of a receiver is turned past a c.w. signal, the beat-note varies from a high tone down through "zero beat" (no audible frequency difference) and back up to a high tone, as shown at A, B and C. The curve is a graphical representation of the action. The beat exists past 8000 or 10,000 cycles but usually is not heard because of the limitations of the audio system.**

frequency at which the oscillator is operating, is called "locking-in"; the more stable of the two frequencies assumes control over the other. "Blocking" usually can be corrected by advancing the regeneration control until the beat-note occurs again. If the regenerative detector is preceded by an r.f. amplifier stage, the blocking can be eliminated by reducing the gain of the r.f. stage. If the detector is coupled to an antenna, the blocking condition can be eliminated by advancing the regeneration control or loosening the antenna coupling.

The point just after the receiver starts oscillating is the most sensitive condition for c.w. reception. Further advancing the regeneration control makes the receiver less prone to blocking by strong signals, but also less capable of receiving weak signals.

If the receiver 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.

Superregeneration — The limit to which ordinary regenerative amplification can be carried is the point at which oscillations commence, since at that point further amplification ceases. The *superregenerative* detector overcomes this limitation by introducing into the detector circuit an alternating voltage of a frequency somewhat above the audible range (of the order of 20 to 200 kilocycles), in such a way as to vary the detector's operating point (§ 3-3). As a consequence of the introduction of this *quench* or *interruption* frequency, the detector can oscillate only when the varying operating point is in a region suitable for the production of oscillations. Because the oscillations are constantly being interrupted, the regeneration can be greatly increased, and the amplified signal will build up to tremendous proportions. A one-tube superregenerative detector is capable of an inherent sensitivity approaching the thermal-agitation noise level of the tuned circuit, and may have an antenna input sensitivity of two microvolts or better.

Because of its inherent characteristics, the superregenerative circuit is suitable only for the reception of modulated signals, and operates best on the very-high frequencies. Typical superregenerative circuits for the very-high frequencies are shown in Fig. 709.

The basic regenerative detector circuit is the ultraudion oscillator (§ 3-7). In Fig. 709-A the quench frequency is obtained from a separate oscillator and introduced into the plate circuit of the detector. The quench oscillator, operating at a low radio frequency, alternately allows oscillations to build up in the regenerative circuit and then causes them to die out. In the absence of a signal, the thermal agitation noise in the input circuit produces the voltage that initiates the build-up process. However, when an incoming signal provides the initiating pulse, it has the effect of advancing the starting time of the oscillations. This causes the area within the envelope to increase, as indicated in Fig. 710-C.

If regeneration in an ordinary regenerative circuit is carried sufficiently far, the circuit will break into a low-frequency oscillation simultaneously with that at the operating radio frequency. This low-frequency oscillation has much the same quenching effect as that from a separate oscillator, hence a circuit so operated is called a *self-quenching* superregenerative detector. The frequency of the quench oscillation depends upon the feed-back and upon the time constant of the grid leak and condenser, the oscillation being a "blocking" or "squegging" in which the grid accumulates a strong negative charge which does not leak off rapidly enough through the grid leak to prevent a relatively slow variation of the operating point.

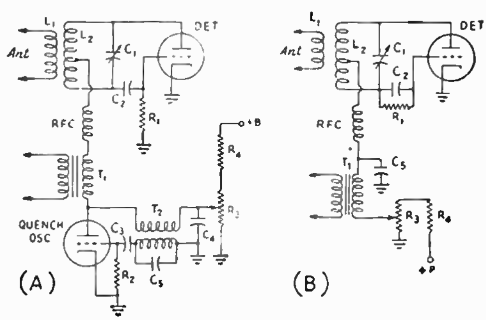


Fig. 709 — **(A) Superregenerative detector circuit using a separate quench oscillator. (B) Self-quenched superregenerative detector circuit. L_2C_1 is tuned to the signal frequency. Typical values for other components are:**

C_2 — 50 μμfd.
C_3 — 500 μμfd.
C_4 — 0.1 μfd.
C_5 — 0.001–0.005 μfd.
R_1 — 2-10 megohms.
R_2 — 50,000 ohms.
R_3 — 50,000-ohm potentiometer.
R_4 — 50,000 ohms.
T_1 — Audio transformer, plate-to-grid type.
RFC — R.f. choke, value depending upon frequency. Small low-capacity chokes are required for v.h.f. operation.

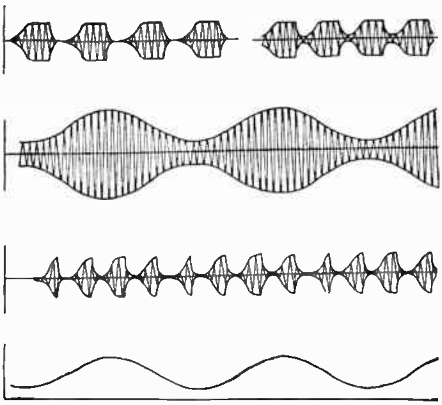


Fig. 710 — R.f. oscillation envelopes in a self-quenched superregenerative detector. Without signal (A at left) oscillations are completely quenched after each period, resuming in random phase depending on momentary noise voltages. At right, when the initiating pulses are supplied by a received signal the starting time of the oscillations is advanced causing the build-up period to begin before damping is complete. This advance is proportional to the carrier amplitude when modulated (B). Since the building-up period varies in accordance with modulation (C), when these wave trains are rectified the average rectified current is proportional to the amplitude of the signal. Amplitude modulation is therefore reproduced as an audio wave in the output circuit (D).

The greater the difference between the quenching and signal frequencies the greater the amplification, because the signal then has a longer period in which to build up during the nonquenching half-cycle when the resistance of the circuit is negative. This ratio should not exceed a certain limit, however, for during the quenched or nonregenerative intervals the input selectivity is merely that of the Q of the tuned circuit alone. The optimum quench frequency is in the neighborhood of 150 kc. for the 60-Mc. band and 250 kc. for 112 Mc.

The superregenerative detector has relatively little selectivity as compared to a regular regenerative detector, but discriminates against noise such as ignition interference. It also has marked a.v.c. action, strong signals being amplified much less than weak signals.

Adjustment of superregenerative detectors — Because of the greater amplification, the hiss noise when a superregenerative detector goes into oscillation is much stronger than with the ordinary regenerative detector. The most sensitive condition is at the point where the hiss first becomes marked. When a signal is tuned in, the hiss will disappear to a degree which depends upon the signal strength.

Lack of hiss indicates insufficient feed-back at the signal frequency, or inadequate quench voltage. Antenna loading effects will cause dead spots which are similar to those in regenerative detectors and can be overcome by the same methods. The self-quenching detector may require critical adjustment of the grid leak and grid condenser values for smooth operation, since these determine the frequency and amplitude of the quench voltage.

7-5 Audio-Frequency Amplifiers

General — The ordinary detector does not produce very much audio-frequency power output — usually not enough to give satisfactory sound volume, even in headphone reception. Consequently, audio-frequency amplifiers are used after the detector to increase the power level. One amplifier usually is sufficient for headphones, but two stages generally are used where the receiver is to operate a loudspeaker. A few milliwatts of a.f. power is sufficient for headphones, but a loudspeaker requires a watt or more for good room volume.

In all except battery-operated receivers, the negative grid bias of audio amplifiers usually is secured from the voltage drop in a cathode resistor (§ 3-6). The cathode resistor must be by-passed by a condenser having low reactance at the lowest audio frequency to be amplified, compared to the resistance of the cathode resistor (10 per cent or less) (§ 2-8, 2-13). In battery-operated receivers, a separate grid-bias battery generally is used.

Headset and voltage amplifiers — The circuits shown in Fig. 711 are typical of those used for voltage amplification and for providing sufficient power for operation of headphones (§ 3-3). Triodes usually are preferred to pentodes because they are better suited to working into an audio transformer or headset, the input impedances of which are of the order of 20,000 ohms.

In these circuits, R_2 is the cathode bias resistor and C_1 the cathode by-pass condenser. The grid resistor, R_1, gives volume control action (§ 5-9). Its value ordinarily is from 0.25 to 1 megohm. C_2 is the input coupling condenser, already discussed under detectors; it is, in fact, identical to C_4 in Figs. 704 and 705, if the amplifier is coupled to a detector.

Power amplifiers — A popular type of power amplifier is the single pentode, operated Class A or AB; the circuit diagram is given in Fig. 711-A. The grid resistor, R_1, may be a potentiometer for volume control, as shown at

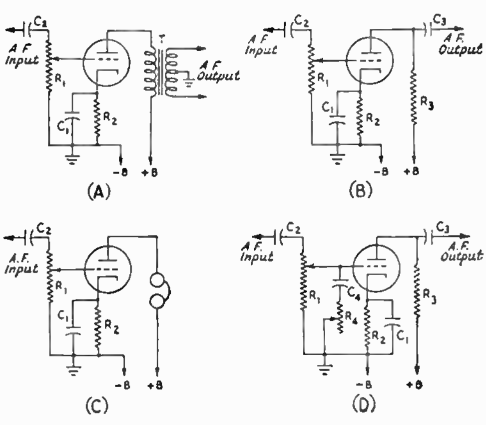


Fig. 711 — Audio amplifier circuits used for voltage amplification and to provide power for headphone output. The tubes are operated as Class-A amplifiers (§ 3-4).

R_1 in Fig. 711. The output transformer, T, should have a turns ratio (§ 2-9) suitable for the loudspeaker used; many of the small loudspeakers now available are furnished complete with output transformer.

When greater volume is needed, a pair of pentodes or tetrodes may be connected in push-pull (§ 3-3), as shown in Fig. 712-B. Transformer coupling to the voltage-amplifier stage is the simplest method of obtaining push-pull input for the amplifier grids. The interstage transformer, T_1, has a center-tapped secondary with a secondary-to-primary turns ratio of about 2 to 1. An output transformer, T_2, with a center-tapped primary must be used. No by-pass condenser is needed across the cathode resistor, R, since the a.f. current does not flow through the resistor as it does in single-tube circuits (§ 3-3).

Tone control — A tone control is a device for changing the frequency response (§ 3-3) of an audio amplifier; usually it is simply a method for reducing high-frequency response. This is helpful in reducing hissing and crackling noises without disturbing the intelligibility of the signal. R_4 and C_4, in Fig. 711-D, together form an effective tone control of this type. The maximum effect is secured when the resistance of R_4 is entirely out of the circuit, leaving C_4 connected directly between grid and ground. R_4 should be large compared to the reactance of C_4 (§ 2-8) so that when its resistance is all in circuit the effect of C_4 on the frequency response is negligible.

Headphones and loudspeakers— Two types of headphones are in general use, the *magnetic* and *crystal* types. They are shown in cross-section in Fig. 713. In the magnetic type the signal is applied to a coil or pair of coils having a great many turns of fine wire wound on a permanent magnet. (Headphones having one coil are known as the "single-pole" type, while those having two coils, as shown in Fig. 713, are called "double-pole.") A thin circular diaphragm of iron is placed close to

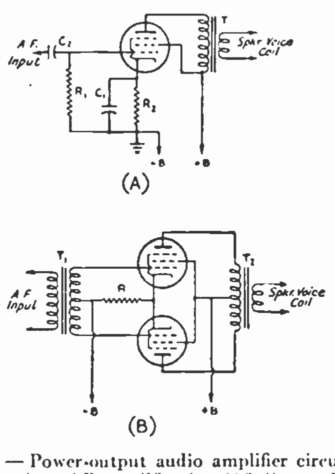


Fig. 712 — Power-output audio amplifier circuits. Either Class A or AB amplification (§ 3-4) may be used.

the open ends of the magnet. It is tightly clamped by the earpiece assembly around its circumference, and the center is drawn toward the permanent magnet under some tension. When an alternating current flows through the windings the field set up by the current alternately aids and opposes the steady field of the permanent magnet, so that the diaphragm alternately is drawn nearer to and allowed to spring farther away from the magnet. Its motion sets the air into corresponding vibration. Although the d.c. resistance of the coils may be of the order of 2000 ohms, the a.c. impedance of a magnetic type headset will be of the order of 20,000 ohms at 1000 cycles.

In the crystal headphone, two piezoelectric crystals (§ 2-10) of Rochelle salts are cemented together in such a way that the pair tends to be bent in one direction when a voltage of a certain polarity is applied and to bend in the other direction when the polarity is reversed. The crystal unit is rigidly mounted to the earpiece, with the free end coupled to a diaphragm. When an alternating voltage is applied, the alternate bending as the polarity of the applied voltage reverses makes the diaphragm vibrate back and forth. The impedance is several times that of the magnetic type.

Magnetic-type headsets tend to give maximum response at frequencies of the order of 500 to 1000 cycles, with a considerable reduction of response (for constant applied voltage) at frequencies both above and below this region. The crystal type has a "flatter" frequency-response curve, and is particularly good at reproducing the higher audio frequencies. The peaked response curve of the magnetic type is advantageous in code reception, since it tends to reduce interference from signals having beat tones lying outside the region of maximum response, while the crystal type is better for the reception of voice and music. Magnetic headsets can be used in circuits in which d.c. is flowing, such as the plate circuit of a vacuum tube, providing the current is not too large to be carried safely by the wire in the coils; the limit is a few milliamperes. Crystal headsets must be used only on a.c. (since a steady d.c. voltage will damage the crystal unit), and consequently must be coupled to the tube through a device, such as a condenser, which isolates the d.c. voltage but permits the passage of an alternating current.

The most common type of loudspeaker is the *dynamic* type, shown in cross-section in Fig. 713. The signal is applied to a small coil (the *voice coil*) which is free to move in the gap between the ends of a magnet. The magnet is made in the form of a cylindrical coil slightly smaller than the form on which the voice coil is wound, with the magnetic circuit completed through a pole piece which fits around the outside of the voice coil leaving just enough clearance for free movement of the coil. The path of the flux through the magnet is as shown by the dotted lines in the figure.

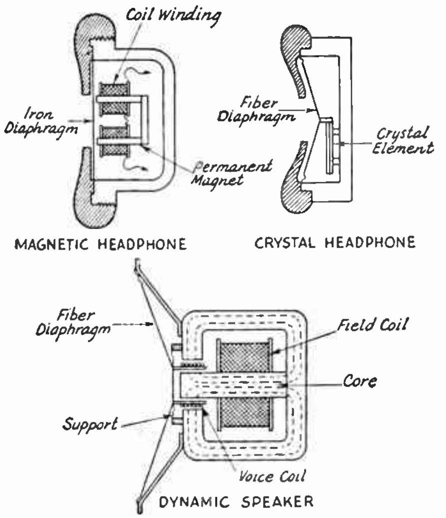


Fig. 713 — Headphone and loudspeaker construction.

The voice coil is supported so that it is free to move along its axis but not in other directions, and is fastened to a fiber or paper conical diaphragm. When current is sent through the coil it moves in a direction determined by the polarity of the current (§ 2-5), and thus moves back and forth when an alternating voltage is applied. The motion is transmitted by the diaphragm to the air, setting up sound waves.

The type of speaker shown in Fig. 713 obtains its fixed magnetic field by electromagnetic means, direct current being sent through the *field coil* for this purpose. Other types use permanent magnets to replace the electromagnet, and hence do not require a source of d.c. power. The voice coils of dynamic speakers have few turns and therefore low impedance, values of 3 to 15 ohms being representative.

ℂ 7-6 Radio-Frequency Amplifiers

Circuits — Although there may be variations in detail, practically all r.f. amplifiers conform to the basic circuit shown in Fig. 714. A screen-grid tube, usually a pentode, is used, since a triode will oscillate when its grid and plate circuits are tuned to the same frequency (§ 3-5). The amplifier operates Class A, without grid current (§ 3-4). The tuned grid circuit, L_1C_1, is coupled through L_2 to the antenna (or, in some cases, to a preceding stage). R_1 and C_2 are the cathode bias resistor and by-pass condenser, C_3 is the screen by-pass condenser, and R_2 is the screen dropping resistor. L_3 is the primary of the output transformer (§ 2-11), tightly coupled to L_4, which, with C_5, constitutes the tuned circuit feeding the detector or following amplifier. The input and output circuits, L_1C_1 and L_4C_5, are both tuned to the signal frequency.

Shielding — The screen-grid construction of the amplifier tube prevents feed-back (§ 3-3) from plate to grid inside the tube, but in addition it is necessary to prevent transfer of energy from the plate circuit to the grid circuit external to the tube. This is accomplished by enclosing the coils in grounded shielding containers and by keeping the plate and grid leads well separated. With "single-ended" tubes, care in laying out the wiring to obtain the maximum possible physical separation between plate and grid leads is necessary to prevent capacity coupling.

The shield around a coil will reduce the inductance and Q of the coil (§ 2-11) to an extent which depends upon the shielding material and the distance it is placed from the coil. Adjustments therefore must be made with the shield in place.

By-passing — In addition to shielding, good by-passing (§ 2-13) is imperative. This is not simply a matter of choosing the proper type and capacity of by-pass condenser. Short *separate* leads from C_3 and C_4 to cathode or ground are a prime necessity. At the higher radio frequencies even an inch of wire will have enough inductance to provide feed-back coupling, and hence cause oscillation, if the wire happens to be common to both the plate and grid circuits.

Gain control — The gain of an r.f. amplifier usually is varied by varying the grid bias. This method works best with variable-μ type tubes (§ 3-5), hence this type usually is found in r.f. amplifiers. In Fig. 714, R_3 and R_4 comprise the gain-control circuit. R_3 is the control resistor (§ 3-6) and R_4 a dropping resistor of such value as to make the voltage across the outside terminals of R_3 about 50 volts (§ 8-10). The gain is maximum with the variable arm on R_3 all the way to the left (grounded), and minimum at the right. R_3 could simply be placed in series with R_1, omitting R_4 entirely, but the range of control with this connection is limited because it depends on the cathode current alone.

In a multi-tube receiver the gain of several stages may be varied simultaneously, a single control sufficing for all. The lower ends of the several cathode resistors (R_1) are then connected together and to the movable contact on R_3 in Fig. 714.

Circuit values — The value of the cathode resistor, R_1, should be calculated for the *minimum* recommended bias for the tube used. The capacities of C_2, C_3 and C_4 must be such that the reactance is low at radio frequencies; this condition is easily met by using 0.01-μfd. condensers at communication frequencies, or 0.001 to 0.002 mica units at very-high fre-

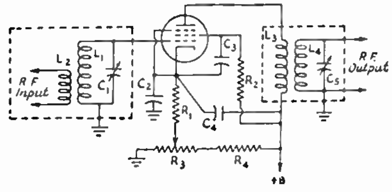


Fig. 714 — Basic circuit of a tuned radio-frequency amplifier. Component values are discussed in the text.

quencies up to 112 Mc. R_2 is found by taking the difference between the recommended plate and screen voltages, then substituting this and the rated screen current in Ohm's Law (§ 2-6). R_3 must be selected on the basis of the number of tubes to be controlled: a resistor must be chosen which is capable of carrying, at its low-resistance end, the sum of all the tube currents plus the bleeder current. A resistor of suitable current-carrying capacity being found, the bleeder current necessary to produce a drop through it of about 50 volts can be calculated by Ohm's Law. The same formula will give R_4, using the plate voltage less 50 volts for E and the bleeder current previously found for I.

The constants of the tuned circuits will depend upon the frequency range, or band, to be covered. A fairly high L/C ratio (§ 2-10) should be used on each band; this is limited, however, by the irreducible minimum capacities. To an allowance of 10 to 20 μμfd. for tube and stray capacities should be added the minimum capacity of the tuning condenser.

If the input circuit of the amplifier is connected to an antenna, the coupling coil, L_2, should be adjusted to provide critical coupling (§ 2-11) between the antenna and grid circuit. This will give maximum energy transfer. The turns ratio of L_1/L_2 will depend upon the frequency, the type of tube used, the Q of the tuned circuit and the constants of the antenna system, and in general is best determined experimentally. The selectivity will increase as the coupling is reduced below this "optimum" value, a consideration which it is well to keep in mind if selectivity is of more importance than maximum gain.

The output-circuit coupling depends upon the plate resistance (§ 3-2) of the tube, the input resistance of the succeeding stage, and the Q of the tuned circuit, L_4C_5. L_3 usually is coupled as closely as possible to L_4 (avoiding the necessity for an additional tuning condenser across L_3) and the energy transfer is maximum when L_3 has ⅔ to ⅘ as many turns as L_4, with ordinary receiving pentodes.

Tube and circuit noise — In any conductor electrons will be moving in random directions simultaneously and, as a result, small irregular voltages are developed across the conductor terminals. The voltage is larger the greater the resistance of the conductor and the higher its temperature. This is known as the *thermal-agitation* effect, and it produces a hiss-like noise voltage distributed uniformly throughout the radio-frequency spectrum. The thermal-agitation noise voltage appearing across the terminals of a tuned circuit will be the same as in a resistor of a value equal to the parallel impedance (§ 2-10) of the tuned circuit, even though the actual circuit resistance is low. Hence, the higher the Q of the circuit, the greater the thermal agitation noise.

Another component of hiss noise is developed in the tube because the rain of electrons on the plate is not entirely uniform. Small irregularities caused by gas in the tube also contribute to the effect. Tube noise varies with the type of tube: in general, the higher the cathode current and the lower the mutual conductance of the tube, the more internal noise it will generate.

To obtain the best signal-to-noise ratio, the signal must be made as large as possible at the grid of the tube, which means that the antenna coupling must be adjusted to that end and also that the Q of the grid tuned circuit must be high. A tube with low inherent noise obviously should be chosen. In an amplifier having good signal-to-noise ratio, the thermal-agitation noise will be greater than the tube noise. This can easily be checked by disconnecting the antenna so that no outside noise is being introduced into the receiver, then grounding the grid through a 0.01-μfd. condenser and observing whether there is a decrease in noise. If there is no change the tube noise is greatly predominant, indicating a poor signal-to-noise ratio in the stage. The test is valid only if there is no regeneration in the amplifier. The signal-to-noise ratio will decrease as the frequency is raised, because it becomes increasingly difficult to obtain a tuned circuit of high effective Q (§ 7-7).

The first stage of the receiver is the important one from the standpoint of signal-to-noise ratio. Noise generated in the second and subsequent stages, while comparable in magnitude to that generated in the first, is masked by the amplified noise and signal from the first stage. After the second stage, further contributions by tubes and circuits to the total noise are inconsequential in any normal receiver.

Tube input resistance — At high radio frequencies the tube may consume power from the tuned grid circuit, even though the grid is not driven positive by the signal. Above 7 Mc. all tubes "load" the tuned circuit to some extent, the amount of loading varying with the type of tube. This effect comes about because of the transit time necessary for electrons to travel from the cathode to the grid becomes comparable to the time of one r.f. cycle, and because of the degenerative effect (§ 3-3) of the cathode lead inductance. It becomes more pronounced as the frequency is increased. Certain types of tubes may have an input resistance of only a few thousand ohms at 28 Mc. and as little as a few hundred ohms at very-high frequencies. The input resistance of the same tubes at 7 Mc. and lower frequencies may be so high as to be considered infinite.

This *input-loading* effect is in addition to the normal decrease in the Q of the tuned circuit alone, because of increased losses in the coil and condenser at the higher frequencies. Thus the selectivity and gain of the circuit both are affected adversely by increasing frequency.

Comparison of tubes — At 7 Mc. and lower frequencies, the signal-to-noise ratio, gain, and selectivity of an r.f.-amplifier stage are sufficiently high with any of the standard receiving tubes. At 14 Mc. and higher, however, this is no longer true, and the choice of a tube must be based on several conflicting considerations.

Gain is highest with high mutual-conductance pentodes, the 6AB7 and 6AC7 being examples of this type. These tubes also develop less noise than any of the others. The input-loading effect is greatest with them, however, so that selectivity is decreased and the tuned-circuit gain is lowered.

Pentodes, such as the 6K7, 6J7 and corresponding types in glass, have lesser input-loading effects at high frequencies, moderate gain, and relatively high inherent noise.

"Acorn" and equivalent miniature pentodes are excellent from the input-loading standpoint; gain is about the same as with standard types, and the inherent noise is somewhat lower.

Where selectivity is paramount the acorns are best, the standard pentodes second, and the 6AB7–6AC7 types worst. On signal-to-noise ratio the latter tubes are first, acorns are second and standard pentodes third. The same order of precedence holds for over-all gain.

At 56 Mc. the standard types are usable, but acorns are capable of better performance because of lesser loading. The 954 and 956 and the corresponding types, 9001 and 9003, are examples of types satisfactory for r.f. amplification at 100 Mc. and higher.

7-7 Tuning and Band-Changing Methods

Band-changing — The resonant circuits which are tuned to the frequency of the incoming signal constitute a special problem in the design of amateur receivers, since the amateur frequency assignments consist of groups or bands of frequencies at widely spaced intervals. The same LC combination cannot be used for, say, 14 Mc. to 3.5 Mc., because of the impracticable maximum-minimum capacity ratio required, and also because the tuning would be excessively critical with such a large frequency range. It is necessary, therefore, to provide a means for 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.

There are two favorite methods of changing inductances. One is to use a switch having an appropriate number of contacts, which connects the desired coil and disconnects the others. The second is to use coils wound on forms with contacts (usually pins) which can be plugged in and removed from a socket.

Bandspreading — The tuning range of a given coil and variable condenser will depend upon the inductance of the coil and the change in tuning capacity. For ease of tuning, it is desirable to adjust the tuning range so that practically the whole dial scale is occupied by the band in use. This is called *bandspreading.* Because of the varying widths of the bands, special tuning methods must be devised to give the correct maximum-minimum capacity ratio on each band. Several of these methods are shown in Fig. 715.

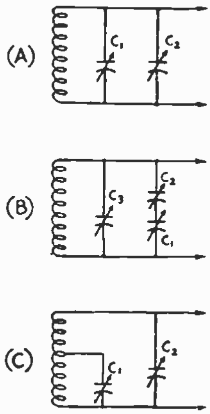


Fig. 715 — Essentials of the three basic band-spread tuning systems.

In A, a small *band-spread condenser,* C_1 (15 to 25 μμfd. maximum capacity), is used in parallel with a condenser, C_2, which is usually large enough (140 to 175 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. In practicable circuits it is almost impossible, because of the non-harmonic relation of the various bands, to get full bandspread on all bands with the same pair of condensers, especially when the coils are wound to give continuous frequency coverage on C_2, which is variously called the *band-setting* or *main-tuning* condenser. C_2 must be reset each time the band is changed.

The method shown at B makes use of condensers in series. The tuning condenser, C_1, may have a maximum capacity of 100 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 pre-adjusted condensers must be switched in.

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

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

True tracking can be obtained only when the inductance, tuning condensers, and circuit 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. 716, 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 which must be tracked. The circuits are identical with those of Fig. 715. 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. 715-B, and C_2 in Fig. 715-C serve as trimmers.

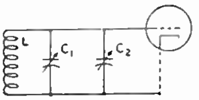


Fig. 716 — Showing the use of a trimmer condenser, to set the minimum circuit capacity in order to obtain true tracking for gang-tuning.

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. 717.

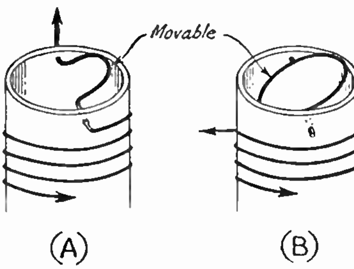


Fig. 717 — Methods of adjusting the inductance for ganging. The half turn in A can be moved so that its magnetic field either aids or opposes the field of the coil. The shorted loop in B is not connected to the coil, but operates by induction. It will have no effect on the coil inductance when the plane of the loop is parallel to the axis of the coil, and will give maximum reduction of the coil inductance when perpendicular to the coil axis.

V.h.f. circuits — Interelectrode capacities are practically constant for a given tube regardless of the operating frequency, and the same is approximately true of stray circuit capacities. Hence, at very-high frequencies these capacities become an increasingly larger part of the usable tuning capacity, and reasonably high L/C ratios (§ 2-10) are more difficult to secure as the frequency is raised. Because of this irreducible minimum capacity, standard types of tubes cannot be tuned to frequencies higher than about 200 Mc., even when the inductance in the circuit is simply that of a straight wire between the tube elements.

Along with these capacity effects, the input loading (§ 7-6) increases rapidly at very-high frequencies, so that ordinary tuned circuits have very low effective Qs when connected to the grid circuit of a tube. The effect is still further aggravated by the fact that losses in the tuned circuit itself are higher, causing a still further reduction in Q. For these reasons, the frequency limit at which an r.f. amplifier will give any gain is in the vicinity of 60 Mc. with standard tubes. At higher frequencies there will be a loss, instead of amplification. This condition can be mitigated somewhat by taking steps to improve the effective Q of the circuit, either by tapping the grid down on the coil, as shown in Fig. 718-A, or by using a lower L C ratio (§ 2-10). The Q of the tuned circuit alone can be greatly improved by using a linear circuit (§ 2-12), which when properly constructed will give Qs much higher than those attainable at lower frequencies with conventional coils and condensers. The concentric type of line, Fig. 718-B, is best both from the standpoint of Q and of adaptability to nonsymmetrical circuits such as are used in receivers. Since the capacity and resistance loading effects of the tube are still present, the Q of such a circuit will be destroyed if the grid-cathode circuit of the tube is connected directly across it. Hence, tapping down on the line, as shown, is necessary.

Very-high-frequency amplifiers employ tubes of the acorn or miniature type, which have the least loading effect as well as low interelectrode capacities. The smaller loading effect means higher input resistance, and, for a given loaded Q of the tuned circuit, a higher voltage is developed between the grid and cathode. Thus the amplification of the stage is higher and the noise level lower.

A concentric circuit may be tuned by varying the length of the inner conductor (usually by using close-fitting tubes, one sliding inside the other) or by connecting an ordinary tuning condenser across the line. Tapping the condenser down, as shown in Fig. 718-B, gives a bandspread effect, which is advantageous. It also helps to keep the Q of the circuit higher than it would be with the condenser connected directly across the open end of the line, since at very-high frequencies most condensers have losses which cannot be neglected.

Ordinary bakelite-based receiving-type tubes will function quite satisfactorily as oscillators and superregenerative detectors at frequencies where r.f. amplification is impossible with standard tubes (as in the 112-Mc. band), since tube losses are compensated for by energy taken from the power supply. Ordinary coil and condenser circuits are practicable with such tubes at 112 Mc. At higher frequencies, however, the special v.h.f. tubes are essential.

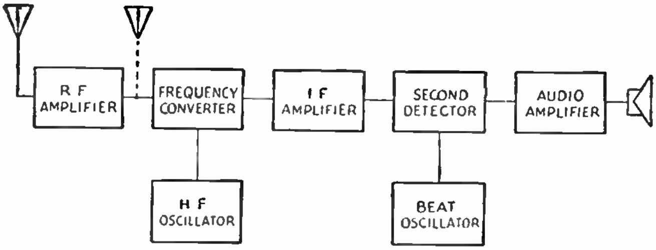


Fig. 719 — Block diagram of the basic elements of the superheterodyne

7-8 The Superheterodyne

Principles — In the *superheterodyne*, or *superhet*, receiver the frequency of the incoming signal is changed to a new radio frequency, the *intermediate frequency* (*i.f.*), then amplified, and finally detected. The frequency is changed by means of the heterodyne process (§ 7-1), the output of an adjustable *local oscillator* (the *h.f. oscillator*) being combined with the incoming signal in a *mixer* or *converter* stage (*first detector*) to produce a beat frequency equal to the intermediate frequency.

Fig. 719 gives the essentials of the superheterodyne in block form. C.w. signals are made audible by heterodyning the signal at the second detector by the *beat-frequency oscillator* (*b.f.o.*) or *beat oscillator*, set to differ from the i.f. by a suitable audio frequency.

As a numerical example, assume that an intermediate frequency of 455 kc. is chosen and that the incoming signal is on 7000 kc. Then the h.f. oscillator frequency may be set to 7455 kc., in order that the beat frequency (7455 minus 7000) will be 455 kc. The h.f. oscillator also could be set to 6545 kc., which will give the same frequency difference. To produce an audible c.w. signal of, say, 1000 cycles at the second detector, the beat oscillator would be set to either 454 kc. or 456 kc.

Characteristics — The frequency-conversion process permits r.f. amplification at a relatively low frequency. Thus high selectivity can be obtained, and this selectivity is constant regardless of the signal frequency. Higher gain also is possible at the lower frequency. The separate oscillators can be designed for

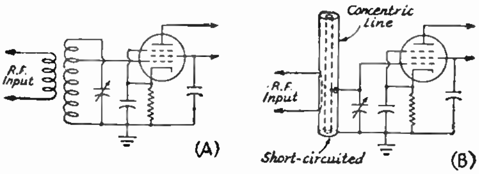


Fig. 718 — Circuits of improved *Q* for very-high frequencies. A, reducing tube loading by tapping down on the resonant circuit; B, use of a concentric-line circuit, with the tube similarly tapped down. The line should be a quarter-wave long, electrically; because of the additional shunt capacity represented by the tube, the physical length will be somewhat less than given by the formula (§ 10-5). In general, this reduction in length will be greater the higher the grid tap on the inner conductor. The coupling turn should be parallel to the axis of the line and must be insulated from the outer conductor.

stability, and, since the h.f. oscillator is working at a frequency considerably removed from the signal frequency, its stability is practically unaffected by the incoming signal.

Images — Each h.f. oscillator frequency will cause i.f. response at two signal frequencies, one higher and one lower than the oscillator frequency. If the oscillator is set to 7455 kc. to respond to a 7000-kc. signal, for example, it will respond also to a signal on 7910 kc., which likewise gives a 455-kc. beat. The undesired signal of the two is called the *image*.

The radio-frequency circuits of the receiver (those used before the frequency is converted to the i.f.) normally are tuned to the desired signal, so that the selectivity of the circuits reduces the response to the image signal. If the desired signal and image have equal strengths at the input terminals of the receiver, 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 farther away from the peak of the resonance curve (§ 2-10) of the signal-frequency input circuits.

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 (§ 2-7) generated by rectification in the second detector may, by stray coupling, be introduced into the r.f. or mixer circuit and converted to the intermediate frequency, to go through the receiver in the same way as an ordinary signal. These "birdies" appear as a heterodyne beat on the desired signal, and are principally bothersome when the frequency of the incoming signal is not greatly different from the intermediate frequency. The cure is proper circuit isolation and shielding.

Harmonics of the beat oscillator also may be converted in similar fashion and amplified through the receiver; these responses can be reduced by shielding the beat oscillator and operating it at low output level.

The double superheterodyne — At high and very-high frequencies it is difficult to secure an adequate image ratio when the intermediate frequency is of the order of 455 kc. To reduce image response the signal frequently is converted first to a rather high (1500, 5000, or even 10,000 kc.) intermediate frequency, and then — sometimes after further amplification — reconverted to a lower i.f. where higher adjacent-channel selectivity can be obtained. Such a receiver is called a *double superheterodyne.*

ℂ 7-9 Frequency Converters

Characteristics — The first detector or mixer resembles an ordinary detector. A circuit tuned to the intermediate frequency is placed in the plate circuit of the mixer, so that the highest possible i.f. voltage will be developed. The signal- and oscillator-frequency voltages appearing in the plate circuit are bypassed to ground, since they are not wanted in the output. The i.f. tuned circuit should have low impedance for these frequencies, a condition easily met if they do not approach the intermediate frequency.

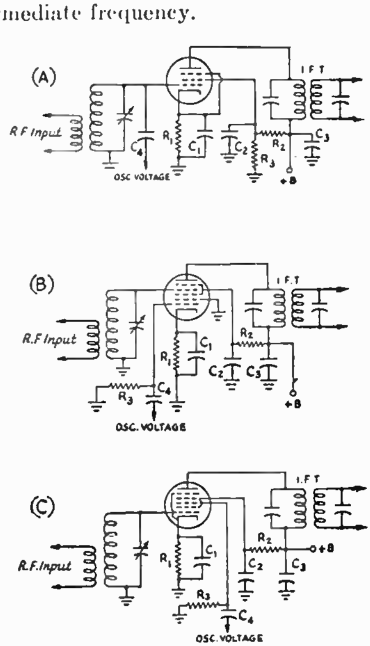


Fig. 720 — Mixer or converter circuits. A, grid injection with a pentode plate detector; B and C, separate injection circuits for converter tubes. Circuit values are:

Component	*Circuit A*	*Circuit B*	*Circuit C*
C_1, C_2, C_3 —	0.01–0.1 μfd.	0.01–0.1 μfd.	0.01–0.1 μfd.
C_4 —	Approx. 1 μμfd.	50–100 μμfd.	50–100 μμfd.
R_1 —	10,000 ohms.	300 ohms.	500 ohms.
R_2 —	0.1 megohm.	50,000 ohms.	15,000 ohms.
R_3 —	50,000 ohms.	50,000 ohms.	50,000 ohms.

Plate voltage should be 250 in all circuits. If a 6AB7 or 6AC7 tube is used in Circuit A, R_1 should be 500 ohms.

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 (§ 3-7). Also, the conversion efficiency should not depend too critically on the oscillator voltage (that is, a small change in oscillator output should not change the gain), since it is difficult to maintain constant output over a wide frequency range.

A change in oscillator frequency caused by tuning of the mixer grid circuit is called *pulling.* If the mixer and oscillator could be completely isolated, mixer tuning would have no effect on the oscillator frequency; but in practice this is a difficult condition to attain. Pulling should be minimized, because the stability of the whole receiver depends critically upon the stability of the h.f. oscillator. Pulling decreases with separation of the signal and h.f. oscillator frequencies, being less with high i.f.s.

Circuits — Typical frequency-conversion circuits are given in Fig. 720. The variations are chiefly in the way in which the oscillator voltage is introduced. In Fig. 720-A, the screen-grid pentode functions as a plate detector; the oscillator is capacity-coupled to the grid of the tube, in parallel with the tuned input circuit. 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.

A pentagrid-converter tube is used in the circuit at B. Although intended for combination oscillator-mixer use, this type of tube usually will give more satisfactory performance when used in conjunction with a separate oscillator, the output of which is coupled in as shown. The circuit gives good conversion efficiency, and, because of the electron coupling, affords desirable isolation between the mixer and oscillator circuits. A small amount of power is required from the oscillator.

Circuit C is for the 6L7 mixer tube. The oscillator voltage can vary over a considerable range without affecting the conversion gain. There are no critical adjustments, and the oscillator-mixer isolation is good. The oscillator must supply somewhat more power than in B.

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

Tubes for frequency conversion — Any sharp cut-off pentode may be used in the circuit of Fig. 720-A. The 6AB7 and 6AC7 give high conversion gain and excellent signal-to-noise ratio — comparable, in fact, to the gain and signal-to-noise ratio obtainable with r.f. amplifiers — and in these respects are far superior to any other tubes used as mixers, particularly between 14 and 100 Mc. However, this type of tube loads the circuit more (§ 7-6) and thus decreases the selectivity.

The 6K8 is a good tube for the circuit at B; its oscillator plate connection may be ignored. The 6SA7 also is excellent in this circuit, although it has no anode grid (No. 2 grid, in the diagram). In addition to these two types, any pentagrid converter tube may be used.

V.h.f. and U.h.f. converters. — At frequencies above the 30-Mc. region the performance of the special mixer and converter tubes employed on the lower frequencies falls off because of greatly reduced input resistance which, by loading the tuned circuit connected to the tube and thus reducing its Q, lowers the signal-to-noise ratio. However, the high-transconductance pentodes such as the 6AC7 and 6AB7 will perform fairly effectively in the circuit of Fig. 720-A up to 100 Mc. or so.

Above about 100 Mc. the loading effect, in addition to the relatively large input capacity which limits the amount of inductance that can be used in the tuned circuit, makes these tubes markedly inferior to the special high-frequency pentodes such as the 9000 and acorn series. The latter perform successfully up to 400 Mc.

At still higher frequencies — or, for that matter, anywhere above 200 Mc. — other types of converters are preferred. At these frequencies triode mixers, when operated as plate-rectifier detectors in suitable circuits, give the least noise and maximum conversion transconductance.

Fig. 721-A shows the elementary circuit for a single triode with cathode oscillator-voltage injection. In such an arrangement the cathode connection usually terminates (with as short a lead as possible) in a small link near the oscillator tank, one end of which is grounded. Alternatively, direct capacity-coupled grid injection may be used in an arrangement similar to that of Fig. 720-A, C_4 being a very small coupling condenser of perhaps 1 or 2 $\mu\mu$fd. — often merely the free end of the coupling lead placed within the field of the oscillator coil or near the oscillator tube plate or grid.

The balanced triode circuit of Fig. 721-B affords the added advantages of symmetry to ground and complete cancellation of both the received-signal and oscillator voltages in the plate circuit. This serves further to improve the signal/noise ratio as well as to stabilize operation. For optimum performance the oscillator-voltage input should be carefully adjusted, by means of the coupling between the two coils, to give maximum converter gain. The balanced converter circuit is most frequently used with miniature dual triodes such as the 6J6, with which it performs effectively up to 600 Mc. or higher. The oscillator may be operated either on its fundamental or a harmonic. At frequencies above 200 Mc. coaxial or "trough"-line circuits are chiefly used.

At still higher frequencies converters employing conventional tubes are inferior to other, basically different types, including highly specialized versions of velocity-modulation tubes of various types. These techniques, however, are beyond the scope of the present treatment; information concerning practical tubes and circuits is largely held confidential by the military services.

For amateur work on these higher frequencies the use of special small u.h.f. diodes with

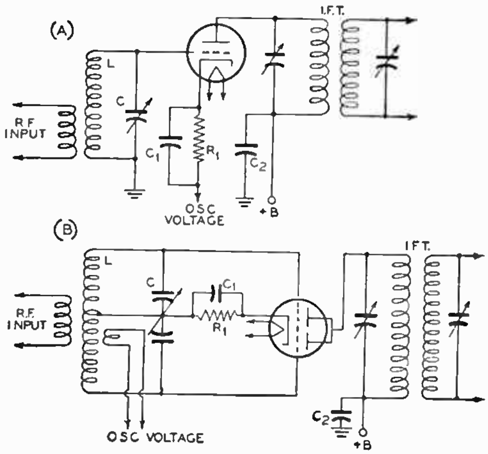


Fig. 721 — V.h.f. frequency converter circuits. A, triode mixer with separate oscillator tube; B, balanced square-law mixer using a dual triode tube with push-pull input circuit. *L* and *C* are tuned to the signal frequency.
C_1 — 100-$\mu\mu$fd. silvered mica.
C_2 — 0.005-$\mu\mu$fd.
R_1 — 10,000-50,000 ohms.

extremely close element spacing as converters is a logical solution. Crystal detectors have also been used extensively because of their ready availability and independence of frequency limitations. Crystal detectors are not susceptible to the transit time limitations of electronic tubes. Silicon is the most popular material for such applications; the crystals are ground to minute dimensions and permanently mounted in fixed miniature holders with tungsten contacts. Fig. 722-A shows a typical crystal mixer circuit with inductive coupling to a triode oscillator (955 or 9002).

Because stability of a crystal detector can be achieved only at the expense of sensitivity, diode detectors are preferred up to the limit of frequency at which they can be made to function. Diodes have the further advantage that they will function as mixers by using a harmonic of the oscillator voltage, making possible the use of conventional triode oscillators for receivers operating up to the 2000-Mc.

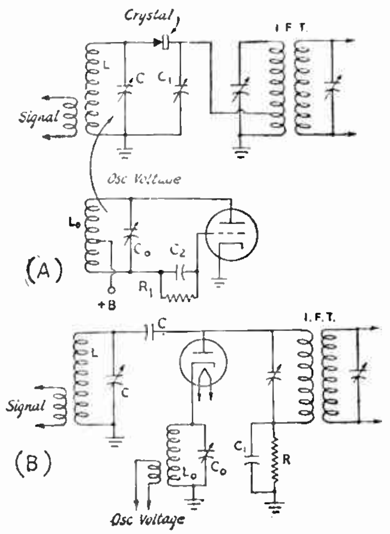


Fig. 722 — U.h.f. frequency converter circuits. A, crystal-detector mixer with an inductively coupled triode oscillator; B, diode mixer with cathode-link coupling to the oscillator circuit. *L* and *C* are tuned to the signal frequency; L_o and C_o to the oscillator frequency.

C_1 — 3–30-$\mu\mu$fd. mica trimmer.
C_2 — 25-$\mu\mu$fd. silvered mica.
C_3 — 10-$\mu\mu$fd. silvered mica.
C_4 — 0.005-$\mu\mu$fd.
R_1 — 50,000 ohms (metallized carbon).
R_2 — 5000–20,000 ohms.

region or higher. While operation of the oscillator on a fundamental is the more efficient method, the loss in conversion efficiency does not exceed 2 to 1 even with third harmonic operation provided the oscillator input is sufficient to establish a diode current of 0.2 to 0.5 ma. Diode mixers are considerably more tolerant as concerns oscillator voltage and other circuit conditions than the crystal type.

In the circuit of Fig. 722-B the cathode tuned circuit, L_oC_o, is tuned to the oscillator fundamental, C_o is being made large enough so that it is effectively a cathode by-pass condenser for the signal frequency.

ℂ 7-10 The High-Frequency Oscillator

Design considerations — Stability of the receiver (§ 7-2) is dependent chiefly upon the stability of the h.f. oscillator, and particular care should be given this part of the receiver. The frequency of oscillation should be insensitive to changes in voltage, loading, and mechanical shock. Thermal effects (slow change in frequency because of tube or circuit heating) should be minimized. These ends can be attained by the use of good insulating materials and circuit components, suitable electrical design, and careful mechanical construction.

In addition, the oscillator must be capable of furnishing sufficient r.f. voltage and power for the particular mixer circuit chosen, at all frequencies within the range of the receiver, and its harmonic output should be as low as possible to reduce spurious response (§ 7-8).

It is desirable to make the *L C* ratio in the oscillator tuned circuit low (high-*C*), since this results in increased stability (§ 3-7). Particular care should be taken to insure that no part of the oscillator circuit can vibrate mechanically. This calls for short leads and "solid" mechanical construction. The chassis and panel material should be heavy and rigid enough so that pressure on the tuning dial will not cause torsion and a shift in the frequency. Care in mechanical construction is well repaid by increased frequency stability.

Circuits — Several oscillator circuits are shown in Fig. 723. 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 6.3-volt heater tubes are used. Hum usually is not bothersome with 2.5-volt tubes, nor, of course, with tubes which are heated by direct current. The circuit of Fig. 723-C overcomes hum, since the cathode is

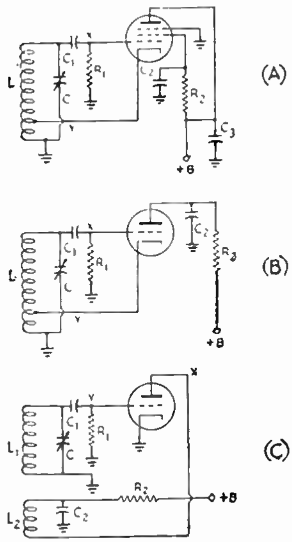


Fig. 723 — High-frequency oscillator circuits. A, screen-grid grounded-plate oscillator; B, triode grounded-plate oscillator; C, triode oscillator with tickler circuit. Coupling to the mixer may be taken from points X and Y. In A and B, coupling from *Y* will reduce pulling effects, but gives less voltage 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 $\mu\mu$fd.	100 $\mu\mu$fd.	100 $\mu\mu$fd.
C_2 —	0.1 μfd.	0.1 μfd.	0.1 μfd.
C_3 —	0.1 μfd.		
R_1 —	50,000 ohms.	50,000 ohms.	50,000 ohms.
R_2 —	50,000 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 (§ 8-10).

grounded. The two-coil arrangement is advantageous in construction, since the feed-back adjustment (altering the number of turns on L_2 or the coupling between L_1 and L_2) is simple mechanically.

Besides the use of a fairly high C/L ratio in the tuned circuit, it is necessary to adjust the feed-back to obtain optimum results. Too much feed-back will cause the oscillator to "squeg," or operate at several frequencies simultaneously (§ 7-4); too little feed-back will cause the output to be low. In the tapped-coil circuits (A, B), the feed-back is increased by moving the tap toward the grid end of the coil; in C, by increasing the number of turns on L_2 or by moving L_2 closer to L_1.

The oscillator plate voltage should be as low as is consistent with adequate output. Low plate voltage will cause reduced tube heating and thereby reduce frequency drift. The oscillator and mixer circuits should be well isolated, preferably by shielding, since coupling other than by the means intended may result in pulling.

To avoid plate-voltage changes which may cause the oscillator frequency to change, it is good practice to use a voltage-regulated plate supply employing a gaseous VR tube (§ 8-8).

Tracking — For ganged tuning, there must be a *constant difference* in frequency between the oscillator and mixer circuits. This difference must be exactly equal to the intermediate frequency (§ 7-8).

Tracking methods for covering a wide frequency range, suitable for general-coverage receivers, are shown in Fig. 724. The tracking capacity, C_5, commonly consists of two condensers in parallel, a fixed one of somewhat less capacity than the value needed and a smaller variable in parallel to allow for adjustment to the exact proper value. In practice, the trimmer, C_4, is first set for the high-frequency end of the tuning range, and then the tracking condenser is set for the low-frequency end. The tracking capacity becomes larger as the percentage difference between the oscillator and signal frequencies becomes smaller (that is, as the signal frequency becomes higher). Typical circuit values are given in the tables under Fig. 724.

In amateur-band receivers, tracking is simplified by choosing a bandspread circuit which 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. 715-C, the tuning will be practically straight-line-frequency if the capacity actually in use at C_2 is not too small; the same is true of 715-A if C_1 is small compared to C_2.

ℂ 7-11 The Intermediate-Frequency Amplifier

Choice of frequency — The selection of an intermediate frequency is a compromise between various conflicting factors. The lower the i.f. the higher the selectivity and gain, but a low i.f. brings the image nearer the desired signal and hence decreases the image ratio (§ 7-8). A low i.f. also increases pulling of the oscillator frequency (§ 7-9). 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

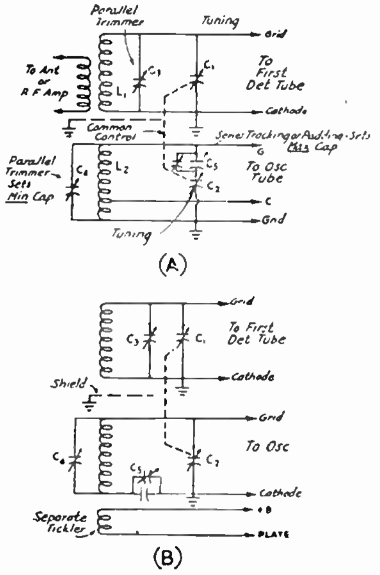


Fig. 724 — Converter-circuit tracking methods. Following are approximate circuit values for 450- to 465-kc. i.f.s, with tuning ranges of approximately 2.15-to-1 and C_2 having 140 μμfd. maximum, and the total minimum capacitance, including C_3 or C_4, being 30 to 35 μμfd.

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-μμfd. maximum, minimum including C_3 and C_4 being 40 to 50 μμfd.

Tuning Range	L_1	L_3	C_5
0.5–1.5 Mc.	240 μh.	130 μh.	425 μμfd.
1.5–4 Mc.	32 μh.	25 μh.	0.00115 μfd.
4–10 Mc.	4.5 μh.	4 μh.	0.0028 μfd.
10–25 Mc.	0.8 μh.	0.75 μh.	None used

tuned r.f. amplifier between antenna and mixer. At 28 Mc. and on the very-high frequencies, the image ratio is very poor unless several r.f. stages are used. Above 14 Mc., pulling is likely to be bad unless very loose coupling can be used between mixer and oscillator.

With an i.f. of about 1600 kc., satisfactory image ratios can be secured on 14, 28 and 56 Mc., and pulling can be reduced to negligible proportions. However, the i.f. selectivity is considerably lower, so that more tuned circuits must be used to increase the selectivity. For very-high frequencies, including 28 Mc., the best solution is to use a double superheterodyne (§ 7-8), choosing one high i.f. for image reduction (5 and 10 Mc. are frequently used) and a lower one for gain and selectivity.

In choosing an i.f. it is wise to avoid frequencies on which there is considerable activity by the various radio services, since such signals may be picked up directly on the i.f. wiring. The frequencies mentioned are fairly free of such interference.

Fidelity, sideband cutting — As described in § 5-2, 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 which contains, for instance, audio frequencies up to 5000 cycles, it must be capable of amplifying equally all frequencies contained in a band extending from 5000 cycles above to 5000 cycles below the carrier frequency. In a superheterodyne, where all carrier frequencies are changed to the fixed intermediate frequency, this means that the i.f. amplifier should amplify equally well all frequencies within that band. In other words, the amplification must be uniform over a band 10 kc. wide, with the i.f. at its center. The signal-frequency circuits usually do not have enough over-all selectivity to affect materially the "adjacent channel" selectivity (§ 7-2), so that only the i.f. amplifier selectivity need be considered.

A 10-kc. band is considered sufficient for reasonably faithful reproduction of music, but much narrower band-widths can be used for communication work where intelligibility rather than fidelity is the primary objective.

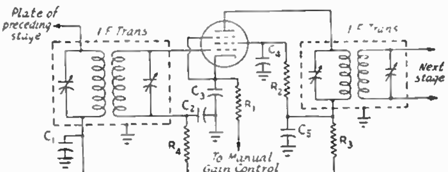


Fig. 725 — Typical intermediate-frequency amplifier circuit for a superheterodyne receiver. Representative values for components are as follows:
C_1 — 0.1 μfd. at 455 kc.; 0.01 μfd. at 1600 kc. and higher.
C_2 — 0.01 μfd.
C_3, C_4, C_5 — 0.1 μfd. at 455 kc.; 0.01 μfd. above 1600 kc.
R_1 — 300 ohms. R_3 — 2000 ohms.
R_2 — 0.1 megohm. R_4 — 0.25 megohm.

If the selectivity is too great to permit uniform amplification over the band of frequencies occupied by the modulated signal, the higher modulating frequencies are attenuated as compared to the lower frequencies; that is, the upper-frequency sidebands are "cut." While sideband cutting reduces fidelity, it is frequently preferable to sacrifice naturalness of reproduction in favor of greater selectivity.

The selectivity of an i.f. amplifier, and hence the tendency to cut sidebands, increases with the number of amplifier stages and also is greater the lower the intermediate frequency. From the standpoint of communication, sideband cutting is not serious with two-stage amplifiers at frequencies as low as 455 kc.

Circuits — I.f. amplifiers usually consist of one or two stages. Two stages at 455 kc. give all the gain usable, in view of the minimum receiver noise level, and also give suitable selectivity for good-quality 'phone reception.

A typical circuit arrangement is shown in Fig. 725. A second stage would simply duplicate the circuit of the first. In principle, the i.f. amplifier is the same as the tuned r.f. amplifier (§ 7-6). However, since a fixed frequency is used, the primary as well as the secondary of the coupling transformer is tuned, giving higher selectivity than is obtainable with a closely coupled untuned primary. The cathode resistor, R_1, is connected to a gain control circuit of the type previously described (§ 7-6); usually both stages, if two are used, are controlled by a single variable resistor. The decoupling resistor, R_3 (§ 2-11), helps isolate the amplifier, and thus prevents stray feed-back. C_2 and R_4 are part of the automatic volume-control circuit (§ 7-13); if no a.v.c. is used, the lower end of the i.f. transformer secondary is simply connected to ground.

In a two-stage amplifier the screen grids of both stages may be fed from a common supply, either through a resistor (R_2) as shown, the screens being connected in parallel, or from a voltage divider (§ 8-10) across the plate supply. Separate screen voltage-dropping resistors are preferable for preventing undesired coupling between stages.

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, is necessary.

I.f. transformers — The tuned circuits of i.f. amplifiers are built up as transformer units consisting of a metal-shield container in which the coils and tuning condensers are mounted. Both air-core and powdered-iron-core universal-wound coils are used, the latter having somewhat higher *Q*s and, hence, greater selectivity and gain per unit. In universal windings the coil is wound in layers with each turn traversing the length of the coil, back and forth, rather than being wound perpendicular to the axis as in ordinary single-layer coils. In a straight multi-layer winding, the turns on adjacent layers at the edges of the coil have a rather large potential difference between them as compared to the difference between any two adjacent turns in the same layer; hence a fairly large capacity current can flow between layers. Universal winding, with its "criss-crossed" turns, tends to avoid building up such potential differences, and hence reduces distributed-capacity effects (§ 2-8).

Variable tuning condensers are of the midget type, air-dielectric condensers being preferable because their capacity is practically unaffected by changes in temperature and humidity. Iron-core transformers may be tuned by varying the inductance (permeability tuning), in which case stability comparable to that of variable air-condenser tuning can be obtained by use of high-stability fixed mica condensers. Such stability is of great importance, since a circuit whose frequency "drifts" with time eventually will be tuned to a different frequency than the other circuits, thereby reducing the gain and selectivity of the amplifier. Typical i.f. transformer construction is shown in Fig. 726.

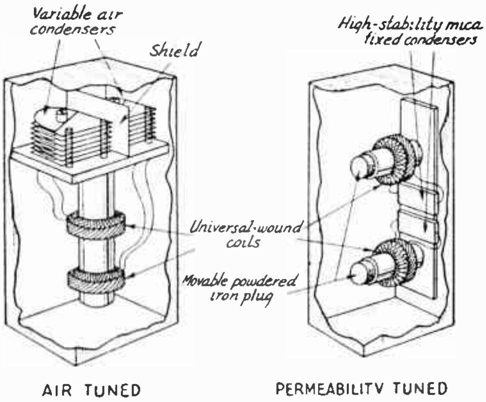


Fig. 726 — Representative i.f. transformer construction. Coils are supported on insulating tubing or (in the air-tuned type) on wax-impregnated wooden dowels. The shield in the air-tuned transformer prevents capacity coupling between the tuning condensers. In the permeability-tuned transformer the cores 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.

Besides the type of i.f. transformer shown in Fig. 726, special units to give desired selectivity characteristics are available. For higher than ordinary adjacent-channel selectivity (§ 7-2) *triple-tuned* transformers, with a third tuned circuit inserted between the input and output windings, are used. The energy is transferred from the input to the output windings via this *tertiary* winding, thus adding its selectivity to the over-all selectivity of the transformer. *Variable-selectivity* transformers also can be obtained. These usually are provided with a third (untuned) winding which can be connected to a resistor, thereby loading the tuned circuits and decreasing the *Q* and selectivity (§ 2-10) to broaden the selectivity curve. The variation in selectivity is brought about by switching the resistor in and out of the circuit. Another method is to vary the coupling between primary and secondary, overcoupling being used to broaden the selectivity curve and undercoupling to sharpen it (§ 2-11).

Selectivity — The over-all selectivity of the i.f. amplifier will depend on the frequency and the number of stages. The following figures are indicative of the band-widths (§ 7-2) to be expected with good-quality transformers in amplifiers so constructed as to keep regeneration to a minimum:

	Band-width in kilocycles		
Intermediate frequency	*2 times down*	*10 times down*	*100 times down*
One stage, 455 kc. (air core) . . .	8.7	17.8	32.3
One stage, 455 kc. (iron core) . .	4.3	10.3	20.4
Two stages, 455 kc. (iron core) .	2.9	6.4	10.8
Two stages, 1600 kc.	11.0	16.6	27.4
Two stages, 5000 kc.	25.8	46.0	100.0

Tubes for i.f. amplifiers — Variable-μ pentodes (§ 3-5) are almost invariably used in i.f. amplifier stages, since grid-bias gain control (§ 7-6) is practically always applied to the i.f. amplifier. Tubes with high plate resistance will have least effect on the selectivity of the amplifier, and those with high mutual conductance will give greatest gain. The choice of i.f. tubes has practically no effect on the signal-to-noise ratio, since this is determined by the preceding mixer and r.f. amplifier (if the latter is used).

When single-ended tubes (§ 3-5) are used, care should be taken to keep the plate and grid leads well separated. With these tubes it is advisable to mount the screen by-pass condenser directly on the bottom of the socket, cross-wise between the plate and grid pins, to provide additional shielding. The outside foil of the condenser should be connected to ground.

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., it will also be heterodyned by the beat oscillator to produce a 1000-cycle beat. This *audio-frequency image* corresponds to the high-frequency images already discussed (§ 7-8). It can be reduced by providing enough i.f. selectivity, since the image signal is off the peak of the i.f. resonance curve.

When this is done, tuning through a given signal will show a strong response at the desired beat note on one side of zero beat only, instead of the two beat notes on either side of zero beat characteristic of less-selective reception; hence the name, "single-signal" reception.

The necessary selectivity is difficult to obtain with non-regenerative amplifiers using ordinary tuned circuits unless a very low intermediate frequency or a large number of circuits is used. In practice it is secured either by regenerative amplification or by a crystal filter.

Regeneration — Regeneration can be used to give a pronounced single-signal effect, particularly when the i.f. is 455 kc. or lower. The resonance curve of an i.f. stage at critical regeneration (just below the oscillating point) is extremely sharp, a band-width of 1 kc. at 10 times down and 5 kc. at 100 times down being obtainable in one stage. The audio-frequency image of a given signal thus can be reduced by a factor of nearly 100 for a 1000-cycle beat note (image 2000 cycles from resonance).

Regeneration is easily introduced into an i.f. amplifier by providing a small amount of capacity coupling between grid and plate. Bringing a short length of wire, connected to the grid, into the vicinity of the plate lead usually will suffice. The feed-back may be controlled by the regular cathode-resistor gain control. When the i.f. is regenerative, it is preferable to operate the tube at reduced gain (high bias) and depend on regeneration to bring up the signal strength. This prevents overloading and increases selectivity.

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

Crystal filters — The most satisfactory method of obtaining high selectivity is by the use of a piezoelectric quartz crystal as a selective filter in the i.f. amplifier (§ 2-10). Compared to a good tuned circuit, the Q of such a crystal is extremely high. The dimensions of the crystal are made such that it is resonant at the desired intermediate frequency. It is then used as a selective coupler between i.f. stages.

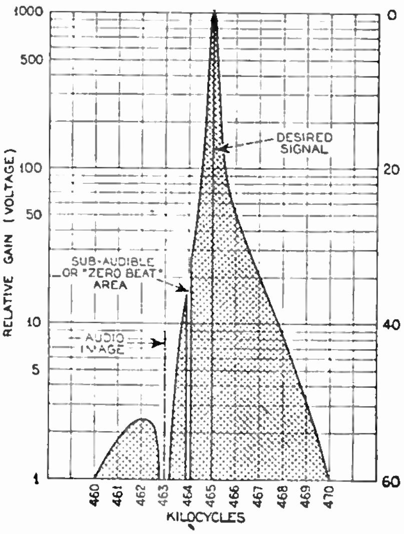


Fig. 727 — Graphical representation of single-signal selectivity. The shaded area indicates the overall band-width, or region in which response is obtainable.

Fig. 727 gives a typical crystal-filter resonance curve. For single-signal reception, the audio-frequency image can be reduced by a factor of 1000 or more. Besides practically eliminating the a.f. image, the high selectivity of the crystal filter provides great discrimination against signals very close to the desired signal in frequency, and, by reducing the band-width, reduces the response of the receiver to noise both from sources external to the receiver and in the r.f. stages of the receiver itself.

Crystal filter circuits: phasing — Several crystal filter circuits are shown in Fig. 728. Those at A and B are practically identical in performance, although differing in details. The crystal is connected in a bridge circuit (§ 2-11), 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, X, and the *phasing condenser*, C_2, which has a maximum capacity somewhat higher than the capacity of the crystal in its holder. When C_2 is set to balance the crystal-holder capacity, the resonance curve of the crystal circuit is practically symmetrical; the crystal acts as a series-resonant circuit of very high Q and thus allows signals of the desired frequency to be fed through C_3 to L_3L_4, the output transformer. Without C_2, the holder capacity (with the crystal acting as a dielectric) would pass signals of undesired frequencies.

The phasing control has an additional function besides neutralization of the crystal-holder capacity. The holder capacity becomes a part of the crystal circuit and causes it to act as a parallel-tuned resonant circuit at a frequency slightly higher than its series-resonant frequency. Signals at the parallel-resonant frequency thus are prevented from reaching the output circuit. The phasing control, by varying the effect of the holder capacity, permits shifting the parallel-resonant frequency over a considerable range, providing adjustable rejection of interfering signals. The effect of rejection is illustrated in Fig. 727, where the audio image is reduced, by proper setting of the phasing control, far below the value that would be expected if the resonance curve were symmetrical.

Variable selectivity — In circuits such as A and B, Fig. 728, variable selectivity is obtained by adjustment of the variable input impedance, which is effectively in series with the crystal resonator. This is accomplished by varying C_1 (the *selectivity control*), which tunes the balanced secondary circuit of T_1. When the secondary is tuned to i.f. resonance the parallel impedance of the L_2C_1 combination is maximum and is purely resistive (§ 2-10). Since the secondary circuit is center-tapped, approximately one-fourth of this resistive impedance is in series with the crystal through C_3 and L_4. This lowers the Q of the crystal circuit and makes its selectivity minimum. At the same time, the voltage applied to the crystal circuit is maximum.

When the input circuit is detuned from the crystal resonant frequency the resistance component of the input impedance decreases, and so does the total parallel impedance. Accordingly, the selectivity of the crystal circuit becomes higher and the applied voltage falls off. At first the resistance decreases faster than the applied voltage, with the result that the c.w. output from the filter *increases* as the selectivity is increased. The output falls off gradually as the input circuit is detuned further from resonance, however, and the selectivity becomes still higher.

In the circuits of A and B in Fig. 728, the minimum selectivity is still much greater than that of a normal two-stage 455-kc. amplifier and it is desirable to provide a wider range of selectivity, particularly for 'phone reception. A circuit which does this is shown at Fig. 728-C. The principle of operation is similar, but a much higher value of resistance can be introduced in the crystal circuit to reduce the selectivity. The output tuned circuit, L_3C_3, must have high Q. A compensated condenser is used at C_2 (phasing) to maintain circuit balance, so that the phasing control does not affect the resonant frequency. The output circuit functions as a voltage divider in such a way that the amplitude of the carrier delivered to the next grid does not vary appreciably with the selectivity setting. The variable resistor, R, may consist of a series of separate fixed resistors selected by a tap switch.

℄ 7-12 The Second Detector and Beat Oscillator

Detector circuits — The second detector of a superheterodyne receiver performs the same function as the detector in the simple receiver, but usually operates at a higher input level because of the relatively great r.f. amplification. 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 (§ 7-13). The basic circuits are as described in § 7-3, although in many cases the diode elements are incorporated in a multi-purpose tube which contains an amplifier section in addition to the diode unit.

The beat oscillator — Any standard oscillator circuit (§ 3-7) may be used for the beat oscillator. Special beat-oscillator transformers are available, usually consisting of a tapped coil with adjustable tuning; these are most conveniently used with circuits such as those shown at Fig. 723-A and -B, with the output taken from Y. A variable condenser of about 25-$\mu\mu$fd. capacity may be connected between cathode and ground to provide fine adjustment. The beat oscillator usually is coupled to the second-detector tuned circuit through a fixed condenser of a few $\mu\mu$fd. capacity.

The beat oscillator should be well shielded, to prevent coupling to any part of the circuit except the second detector and to prevent its harmonics from getting into the front end of the receiver and being amplified like regular signals. To this end, the plate voltage should be as low as is consistent with sufficient audio-frequency output. If the beat oscillator output is too low, strong signals will not give a proportionately strong audio response.

An oscillating second detector may be used to give the audio beat note, but, since the detector must be detuned from the i.f., the selectivity and signal strength will be reduced, while blocking (§ 7-4) will be pronounced because of the high signal level at the second detector.

℄ 7-13 Automatic Volume Control

Principles — Automatic regulation of the gain of the receiver in inverse proportion to the signal strength is a great advantage, especially in 'phone reception, since it tends to keep the output level of the receiver constant regardless of input signal strength. It is readily accomplished in superheterodyne receivers by using the average rectified d.c. voltage, developed by the received signal across a resistance in a detector circuit (§ 7-3), to vary the bias on the r.f. and i.f. amplifier tubes.

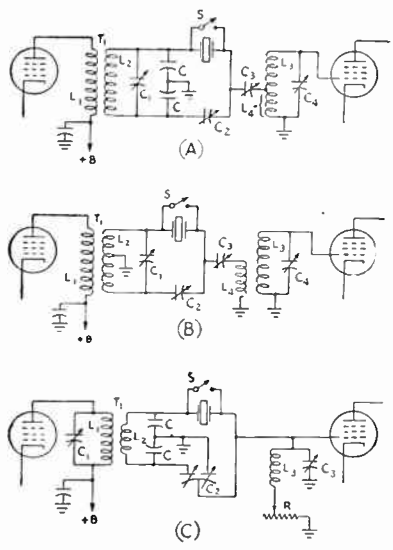


Fig. 728 — Crystal filter circuits of three types. All give variable band-width, with C having the greatest range of selectivity. Their operation is discussed in the text. Suitable circuit values are as follows: Circuit A, T_1, special i.f. input transformer with high-inductance primary, L_1, closely coupled to tuned secondary, L_2; C_1, 50-$\mu\mu$fd. variable; C, each 100-$\mu\mu$fd. fixed (mica); C_2, 10- to 15-$\mu\mu$fd. (max.) variable; C_3, 50-$\mu\mu$fd. trimmer; L_3C_4, i.f. tuned circuit, with L_3 tapped to match crystal-circuit impedance. In circuit B, T_1 is the same as in circuit A except that the secondary is center-tapped; C_1 is 100-$\mu\mu$fd. variable; C_2, C_3 and C_4, same as for circuit A; L_3L_4 is a transformer with primary, L_4, corresponding to tap on L_3 in A. In circuit C, T_1 is a special i.f. input transformer with tuned primary and low-impedance secondary; C, each 100-$\mu\mu$fd. fixed (mica); C_2, opposed stator phasing condenser, approximately 8 $\mu\mu$fd. maximum capacity each side; L_3C_3, high-Q i.f. tuned circuit; R, 0 to 3000 ohms (selectivity control).

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. 729. One plate of the diode section of the tube is used for signal detection and the other for a.v.c. rectification. The a.v.c. diode plate is fed from the detector diode through the small coupling condenser, C_3. A negative bias voltage resulting from the flow of rectified carrier current is developed across R_4, the diode load resistor. This negative bias is applied to the grids of the controlled stages through the filtering resistors (§ 2-11), R_5, R_6, R_7 and R_8. When S_1 is closed the a.v.c. line is grounded, thereby removing the a.v.c. bias from the amplifier without disturbing the detector circuit.

It does not matter which of the two diode plates is selected for audio and which for a.v.c. Frequently the two plates are connected together and used as a combined detector and a.v.c. rectifier. This could be done in Fig. 729. 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. 729 the audio diode return is made directly to the cathode and the a.v.c. diode return to ground. This places negative bias on the a.v.c. diode equal to the d.c. drop through the cathode resistor (a volt or two) and thus delays the application of a.v.c. voltage to the amplifier grids, since no rectification takes place in the a.v.c. diode circuit until the carrier amplitude is large enough to overcome the bias. Without this delay the a.v.c. would start working even with a very small signal. This is undesirable, because the full amplification of the receiver then could not be realized on weak signals. In the audio diode circuit this fixed bias would cause distortion, and must be avoided; hence, the return is made directly to the cathode.

Time constant — The time constant (§ 2-6) of the resistor-condenser combinations in the a.v.c. circuit is an important part of the system. It must be high enough so that the modulation on the signal is completely filtered from the d.c. output, leaving only an average d.c. component which follows the relatively slow carrier variations with fading. Audio-frequency variations in the a.v.c. voltage applied to the amplifier grids would reduce the percentage of modulation on the incoming signal, and in practice would cause frequency distortion. On the other hand, the time constant must not be too great or the a.v.c. would be unable to follow rapid fading. The capacity and resistance values indicated in Fig. 729 will give a time constant which is satisfactory for high-frequency reception.

Signal-strength and tuning indicators — A useful accessory to the receiver is an indicator which will show relative signal strength. Not only is it an aid in giving reports to transmitting stations, but it is helpful also in aligning the receiver circuits, in conjunction with a test oscillator or other steady signal.

Three types of indicators are shown in Fig. 730. That at A uses an electron-ray tube (§ 3-5), several types of which are available. The grid of the triode section usually is connected to the a.v.c. line. The particular type of tube used depends upon the voltage available for its grid; where the a.v.c. voltage is large, a remote cut-off type (6G5 or 6N5) should be used in preference to the more sensitive sharp cut-off type (6E5).

In B, a milliammeter is connected in series with the d.c. plate lead to one or more r.f. and i.f. tubes, the grids of which are controlled by a.v.c. voltage. Since the plate current of such tubes varies with the strength of the incoming signal, the meter will indicate relative signal intensity and may be calibrated in "S" points. The scale range of the meter should be chosen to fit the number of tubes in use; the maximum plate current of the average remote cut-off r.f. pentode is from 7 to 10 milliamperes. The shunt resistor, R, enables setting the plate current to the full-scale value ("zero adjustment"). With this system the ordinary meter reads downwards from full scale with increasing signal strength, which is the reverse of normal pointer movement (clockwise with increasing reading). Special instruments in which the zero-current position of the pointer is on the right-hand side of the scale are used in commercial receivers.

The system at C uses a 0–1 ma. milliammeter in a bridge circuit, arranged so that the

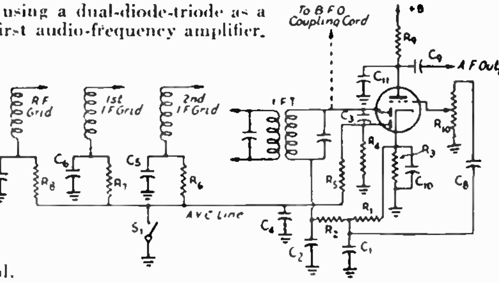


Fig. 729 — Automatic volume control circuit using a dual-diode-triode as a combined a.v.c. rectifier, second detector and first audio-frequency amplifier.

R_1 — 0.25 megohm.
R_2 — 50,000 to 250,000 ohms.
R_3 — 2000 ohms.
R_4 — 2 to 5 megohms.
R_5 — 0.5 to 1 megohm.
R_6, R_7, R_8, R_9 — 0.25 megohm.
R_{10} — 0.5-megohm variable.
C_1, C_2, C_3 — 100 μμfd.
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} — 250 μμfd.

meter reading and the signal strength increase together. The current through the branch containing R_1 should be approximately equal to the current through that containing R_2. In some manufactured receivers this is brought about by draining the screen voltage-divider current and the current to the screens of three r.f. pentodes (r.f. and i.f. stages) through R_2, the sum of these currents being about equal to the maximum plate current of one a.v.c.-controlled tube. Typical values for this type of circuit are given. The sensitivity can be increased by increasing the resistance of R_1, R_2 and R_3. The initial setting is made with the manual gain control set near maximum, when R_3 should be adjusted to make the meter read zero with no signal.

ℂ 7-14 Preselection

Purpose — Preselection is added signal-frequency selectivity incorporated before the mixer stage is reached. An r.f. amplifier preceding the mixer generally is called a *preselector*, its purpose, in part at least, being to discriminate in favor of the signal against the image. The preselector may consist of one or more r.f. amplifier stages. When its tuning control is ganged with those of the mixer and oscillator, its circuits must track with the mixer circuit.

The circuit is the same as discussed earlier (§ 7-6). An external preselector stage may be used with receivers having inadequate image ratios. In this case it is built as a separate unit, often with a tuned output circuit which gives a further improvement in selectivity. The output circuit usually is link-coupled (§ 2-11) to the receiver.

Signal/noise ratio — An r.f. amplifier will have a better signal-to-noise ratio (§ 7-2) than a mixer because the gain is higher and because the mixer-tube electrode arrangement results in higher internal tube noise than does the ordinary pentode structure. Hence, a preselector is advantageous in increasing the signal-to-noise ratio over that obtainable when the mixer is fed directly from the antenna.

Image suppression — The image ratios (§ 7-8) obtainable at frequencies up to and including 7 Mc. with a single preselector stage are high enough, when the intermediate frequency is 455 kc., so that for all practical purposes there is no appreciable image response. Average image ratios on 14 Mc. and 28 Mc. are 50–75 and 10–15, respectively. This is the overall selectivity of the r.f. and mixer tuned circuits. A second preselector stage, adding another tuned circuit, will increase the ratios to several hundred at 14 Mc. and to 30–40 at 28 Mc.

On very-high frequencies, it is impracticable to attempt to secure a good image ratio with a 455-kc. i.f. Good performance can be secured only by using a high i.f. or a double superheterodyne (§ 7-8) with a high-frequency first i.f.

Regeneration — Regeneration may be used in a preselector stage to increase both gain and selectivity. Since its use makes tuning more critical and increases ganging problems, regeneration is seldom employed except at 14 Mc. and above, where adequate image suppression is difficult to obtain with non-regenerative circuits. The same disadvantages exist as in the case of a regenerative i.f. amplifier (§ 7-11). The effect of regeneration is roughly equivalent to adding another non-regenerative preselector stage.

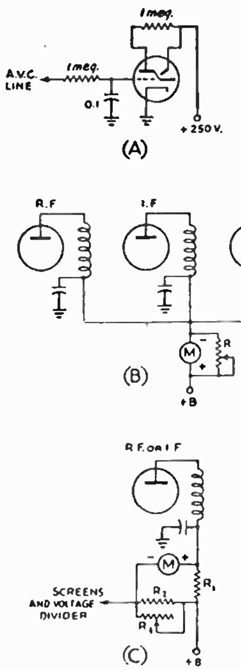


Fig. 730 — Tuning indicator or "S"-meter circuits for superhet receivers. A, electron-ray indicator; B, plate-current meter for tubes on a.v.c.; C, bridge circuit for a.v.c.-controlled tube. In B, resistor R should have a maximum resistance several times that of the milliammeter. In C, representative values for the components are: R_1, 250 ohms; R_2, 350 ohms; R_3, 1000-ohm variable.

Regeneration may be introduced by the same method as used in regenerative i.f. amplifiers (§ 7-11). The manual gain control of the stage will serve as a volume control.

Regeneration in a preselector does not improve the signal-to-noise ratio, since the tube noise is fed back to the grid circuit along with the signal to add to the thermal-agitation noise originally present. This noise also is amplified.

ℂ 7-15 Noise Reduction

Types of noise — In addition to tube and circuit noise (§ 7-6), much of the noise interference experienced in reception of high-frequency signals is caused by domestic electrical equipment and by automobile ignition systems. The interference is of two types in its effects. The first is the "hiss" type, consisting of overlapping pulses similar in nature to the receiver noise. It is largely reduced by high selectivity in the receiver, especially for code reception. The second is the "pistol-shot" or "machine-gun" type, consisting of separated impulses of high amplitude. The "hiss"

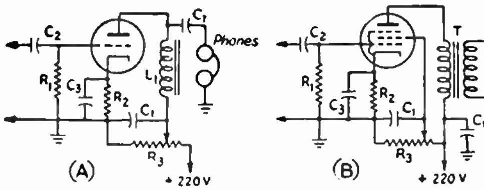


Fig. 731 — Audio output-circuit amplitude-limiting noise-reducing circuits for c.w. reception.
C_1 — 0.25 μfd.
C_2 — 0.01 μfd.
C_3 — 5 μfd.
R_1 — 0.5 megohm.
R_2 — 2000 ohms.
R_3 — 50,000-ohm potentiometer.
T — Output transformer.
L_1 — 15-henry choke.

type of interference usually is caused by commutator sparking in d.c. and series-wound a.c. motors, while the "shot" type results from separated spark discharges (a.c. power leaks, switch and key clicks, ignition sparks, and the like).

Impulse noise — Impulse noise, because of the extremely short duration of the pulses as compared to the time between them, must have high pulse amplitude to contain much average energy. Hence, noise of this type strong enough to cause much interference generally has an instantaneous amplitude much higher than that of the signal being received. The general principle of devices intended to reduce such noise is that of allowing the signal amplitude to pass through the receiver unaffected, but making the receiver inoperative for amplitudes greater than that of the signal. The greater the amplitude of the pulse compared to its time of duration the more successful the noise reduction, since more of the constituent energy can be suppressed.

In passing through selective receiver circuits, the time duration of the impulses is increased, because of the Q or flywheel effect (§ 2-10) of the circuits. Hence, the more selectivity ahead of the noise-reducing device, the more difficult it becomes to secure good noise suppression.

Audio limiting — A considerable degree of noise reduction in code reception can be accomplished by amplitude-limiting arrangements applied to the audio output circuit of a receiver. Such limiters also maintain the signal output nearly constant with fading. Diagrams of typical output-limiter circuits are shown in Fig. 731. Circuit A employs a triode tube operated at reduced plate voltage (approximately 10 volts), so that it saturates at a low signal level. The arrangement of B has better limiting characteristics. A pentode audio tube is operated at reduced screen voltage (35 volts or so), so that the output power remains practically constant over a grid excitation-voltage range of more than 100 to 1. These output-limiter systems are simple, and adaptable to most receivers. However, they cannot prevent noise peaks from overloading previous circuits.

Second-detector circuits — The circuit of Fig. 732 "chops" noise peaks at the second detector of a superhet receiver by means of a biased diode, which becomes non-conducting 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 non-conducting 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 (§ 5-1) 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 732-A, using an infinite-impedance detector (§ 7-3), gives a positive voltage on rectification. When the rectified voltage is negative, as it is from the usual diode detector (§ 7-3), the circuit arrangement shown in Fig. 732-B must be used.

An audio signal of about ten volts is required for good limiting action. When a beat oscillator is used for c.w. reception the b.f.o. voltage should be small, so that incoming noise will not have a strong carrier to beat against and so produce large audio output.

A second-detector noise-limiting circuit which automatically adjusts itself to the received carrier level is shown in Fig. 733. The diode load circuit (§ 7-3) consists of R_6, R_7, R_8 (shunted by the high-resistance audio volume control, R_4) and R_5 in series. The cathode of the 6N7 noise limiter is tapped on the load resistor at a point such that the average rectified carrier voltage (negative) at its grid is approximately twice the negative voltage at the cathode, both measured with reference to ground. A filter network, R_1C_1, is inserted in the grid circuit, so that the audio modulation on the carrier does not reach the grid; hence, the grid potential is maintained at substantially the rectified carrier voltage alone. The cathode, however, is free to follow the modulation, and when the modulation is 100 per cent the peak cathode voltage will just equal the steady grid voltage.

At all modulation percentages below 100 per cent the grid is negative with respect to cathode, and current cannot flow in the 6N7 plate-cathode circuit. A noise pulse exceeding the peak voltage which represents 100 per cent modulation will, however, make the grid positive with respect to cathode. The relatively low plate-cathode resistance of the 6N7 then shunts the high-resistance audio output circuit, effectively short-circuiting it, so that there is practically no response for the duration of the peak over the 100 per cent modulation limit.

R_5 is used to make the noise-limiting tube more sensitive by applying to the plate an audio voltage out of phase with the cathode voltage, so that, at the instant the grid goes positive with respect to cathode, the highest positive potential also is applied to the plate, thus further lowering the effective plate-cathode resistance.

I.f. noise silencer — In the circuit shown in Fig. 734, 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. 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 in a typical instance.

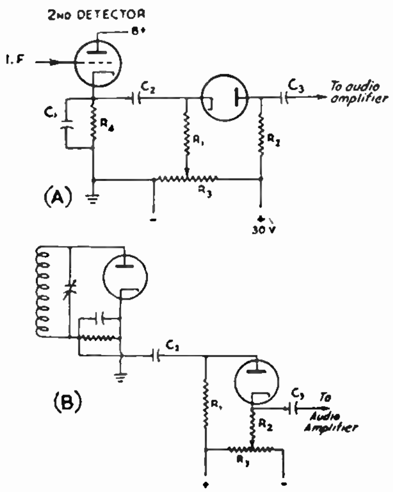


Fig. 732—Series-valve noise-limiter circuits. A, as used with an infinite-impedance detector; B, with a diode detector. Typical values for components are as follows:

R_1 — 0.25 megohm.
R_2 — 50,000 ohms.
R_3 — 10,000-ohms.
R_4 — 20,000 to 50,000 ohms.
C_1 — 250 $\mu\mu$fd.
C_2, C_3 — 0.1 μfd.

All other diode-circuit constants in B are conventional.

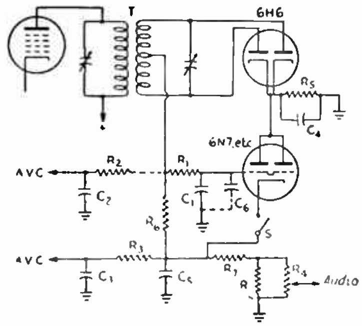


Fig. 733 — Automatic noise-limiter for superheterodynes.

T — I.f. transformer with a balanced secondary for working into a diode rectifier.
R_1, R_2, R_3 — 1 megohm.
R_4 — 1-megohm variable.
R_5 — 250,000 ohms.
R_6, R_8 — 100,000 ohms.
R_7 — 25,000 ohms.
Sw — S.p.s.t. toggle (on-off switch).
C_1 — 0.1-μfd. paper.
C_2, C_3 — 0.05-μfd. paper.
C_4, C_5 — 50-$\mu\mu$fd. mica.
C_6 — 0.001-μfd. mica (for r.f. filtering, if needed).

The switch should be mounted close to the circuit elements and controlled by an extension shaft if necessary.

Circuit values are normal for i.f. amplifiers (§ 7-11), except as indicated. The noise-rectifier transformer, T_1, has an untuned secondary closely coupled to the primary and center-tapped for full-wave rectification. The center-tap rectifier (§ 8-3) is used to reduce the possibility of r.f. feed-back into the i.f. amplifier (noise-silencer) stage. The time constant (§ 2-6) of the noise-rectifier load circuit, $R_1C_1C_2$, must be small, to prevent disabling the noise-silencer stage for a longer period than the duration of the noise pulse. The r.f. choke, *RFC*, must be effective at the intermediate frequency.

Adequate shielding and isolation of the noise-amplifier and rectifier circuits from the noise-silencer stage must be provided to prevent possible self-oscillation and instability. This circuit should be applied to the first i.f. stage of the receiver, before the high-selectivity circuits are reached. On the other hand, it is most effective when the signal and noise levels are fairly high (meaning one or two r.f. stages before the mixer) since several volts must be obtained from the noise rectifier for good silencing.

¶ 7-16 Operating Superheterodyne Receivers

C.w. reception — For making code signals audible, the beat oscillator should be set to a frequency slightly different from the intermediate frequency (§ 7-8). To adjust the beat-oscillator frequency, first tune in a moderately weak but steady carrier with the beat oscillator turned off. Adjust the receiver tuning for maximum signal strength, as indicated by maximum hiss. Then turn on the beat oscillator and adjust its frequency (leaving the receiver tuning unchanged) to give a suitable beat note. The beat oscillator need not subsequently be touched, except for occasional checking to make certain the frequency has not drifted from the initial setting. The b.f.o. may be set on either the high- or low-frequency side of zero beat.

The use of a.v.c. (§ 7-13) is not generally satisfactory in c.w. reception because the receiver gain rises in the spaces between the dots and dashes, giving an increase in noise in the same intervals, and because the rectified beat-oscillator voltage in the second detector circuit also operates the a.v.c. circuit. This gives a constant reduction in gain and prevents utilization of the full sensitivity of the receiver. Hence, the gain preferably should be manually adjusted to give suitable audio-frequency output.

To avoid overloading in the i.f. circuits, it is usually better to control the i.f. and r.f. gain and keep the audio gain at a fixed value than to use the a.f. gain control as a volume control and leave the r.f. gain fixed at its highest level.

Tuning with the crystal filter — If the receiver is equipped with a crystal filter the tuning instructions in the preceding paragraph still apply, but more care must be used both in the initial adjustment of the beat oscillator and in tuning. The beat oscillator is set as described above, but with the crystal filter in operation and adjusted to its sharpest position, if variable selectivity is available. The initial adjustment should be made with the phasing control (§ 7-11) in the intermediate position. After it is completed, the beat oscillator should be left set and the receiver tuned to the other side of zero beat (audio-frequency image) on the same carrier to give a beat note of the same tone. This beat will be considerably weaker than the first, and may be "phased out" almost completely by careful adjustment of the phasing control. This is the adjustment for normal operation; it will be found that one side of zero beat has practically disappeared, leaving maximum response on the desired side.

An interfering signal having a beat note differing from that of the a.f. image can be similarly phased out, provided its carrier frequency is not too near the desired carrier.

Depending upon the filter design, maximum selectivity may cause the dots and dashes to lengthen out so that they seem to "run together." This, plus the fact that tuning is quite critical with extremely high selectivity, may make it desirable to use somewhat less selectivity in ordinary operation. However, it must be emphasized that, to realize the benefits of the crystal filter in reducing interference, it is necessary to do *all* tuning with it in the circuit. Its selectivity is so high that it is almost impossible to find the desired station quickly, should the filter be switched in only when interference is present.

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

A crystal filter will do much toward reducing interference in 'phone reception. Although the high selectivity cuts sidebands (§ 7-11) and thereby reduces the audio output, especially at the higher audio frequencies, it is possible to use quite high selectivity without destroying intelligibility even though the "quality" of the transmission may suffer. As in the case of c.w. reception, it is advisable to do all tuning with the filter in the circuit. Variable-selectivity filters permit a choice of selectivity to suit interference conditions.

An undesired carrier close in frequency to a desired carrier will heterodyne with it to produce a beat note equal to the frequency difference. Such a heterodyne can be reduced by adjustment of the phasing control in the crystal filter. It cannot be prevented in a "straight" superheterodyne having no crystal filter.

A tone control often will be of help in reducing the effects of high-pitched heterodynes, sideband splatter (§ 5-2) and noise, by cutting off the higher audio frequencies. This, like sideband cutting with high selectivity, causes some reduction in naturalness.

Spurious responses — Spurious responses can be recognized without a great deal of difficulty. Often it is possible to identify an image by the nature of the transmitting station, if the frequency assignments applying to the frequency to which the receiver is tuned are known. However, an image also can be recognized by its behavior with tuning. If the signal causes a heterodyne beat note with the

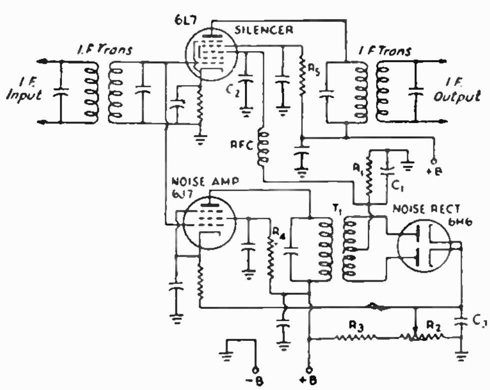


Fig. 734 — I.f. noise-silencing circuit. The plate supply should be 250 volts. Typical values for components are:

C_1 — 50–250 $\mu\mu$fd. (use smallest value possible without r.f. feedback).
C_2 — 50 $\mu\mu$fd.
C_3 — 0.1 μfd.
R_1 — 0.1 megohm.
R_2 — 5000-ohm variable.
R_3 — 20,000 ohms.
R_4, R_5 — 0.1 megohm.
T_1 — Special i.f. transformer for noise rectifier.

desired signal and is actually on the same frequency, the beat note will not change as the receiver is tuned through the signal; but if the interfering signal is an image, the beat will vary in pitch as the receiver is tuned. The beat oscillator in the receiver must be turned off for this test. Using a crystal filter with the beat oscillator on, an image will peak on the side of zero beat opposite that on which the desired signal peaks.

Harmonic response can be recognized by the "tuning rate," or movement of the tuning dial required to give a specified change in beat note. Signals getting into the i.f. via high-frequency oscillator harmonics tune more rapidly (less dial movement) through a given change in beat note than do signals received by normal means.

Harmonics of the beat oscillator can be recognized by the tuning rate of the beat-oscillator pitch control. A smaller movement of the control will suffice for a given change in beat note than is necessary with legitimate signals.

ℂ 7-17 Servicing Superheterodyne Receivers

Troubleshooting — Two basic methods are employed. One is the "point-by-point" system of static analysis, requiring chiefly a multirange volt-ohm-milliammeter. Beginning at the power transformer, the operating voltages at each point in the circuit are measured. Abnormally low or high voltages, or the absence of indication at a given point in the circuit, presumably indicate a defective component at that point. The analysis may then be completed with the aid of the ohmmeter and a little deduction, ending with repair or replacement of unserviceable components.

An alternative method, commonly employed by professional radio servicemen, is that of "dynamic" or "channel" analysis. The principle is that of applying a test signal to the r.f. input and tracing it stage-by-stage through the receiver. The r.f. and i.f. stages are checked by tuned amplifiers feeding a linear detector which operates an indicator such as vacuum-tube voltmeter, electron-ray voltmeter, or cathode-ray tube. A probe on the end of a shielded lead with a very small condenser (1–2 $\mu\mu$fd.) in series is used to pick up the signal in the output of any stage, and the tuned amplifiers are adjusted to the frequency of the stage. Thus the presence or absence of the signal at any point in the receiver may be determined, as well as the relative level.

I.f. alignment — A calibrated signal generator or test oscillator is a practical necessity for initial alignment of an i.f. amplifier. Some means for measuring the output of the receiver also is needed. If the receiver has a tuning meter, its indications will serve for this purpose. Alternatively, if the signal generator is of the modulated type, an a.c. output meter (high-resistance voltmeter with copper-oxide rectifier) can be connected across the primary of the output transformer, or from the plate of the last audio amplifier through a 0.1-μfd. blocking condenser (§ 2-13) to the receiver chassis. The intensity of sound from the loudspeaker can be judged by ear, if no output meter is available, but this method is not as accurate as those using instruments.

The procedure is as follows: The test oscillator is adjusted to the desired intermediate frequency, and the "hot" or ungrounded output lead is clipped on the grid terminal of the last i.f. amplifier tube. The grounded lead is connected to the receiver chassis. The trimmer condensers of the transformer feeding the second detector are then adjusted for maximum signal output. The hot lead from the generator is next clipped on the grid of the next-to-last i.f. tube, and the second from last i.f. transformer is brought into alignment by adjusting its trimmers 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 signal generator as more of the i.f. amplifier is brought into use, because the increased gain otherwise may cause overloading and consequent inaccurate results. It is desirable always to use the minimum signal strength which gives useful output readings.

The i.f. transformer in the plate circuit of the mixer is aligned with the signal-generator output lead connected to the mixer grid. Since the tuned circuit feeding the mixer grid is tuned to a considerably higher frequency, it can effectively short-circuit the signal-generator output, and therefore it may be necessary to disconnect this circuit. With tubes having a top grid-cap connection, this can be done by simply removing the grid clip from the tube cap.

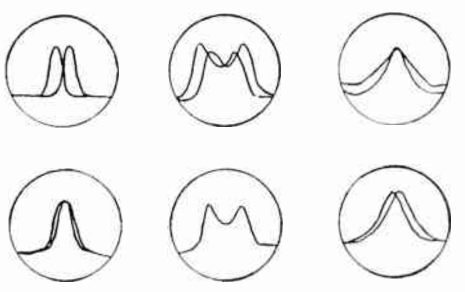
If the tuning indicator is used as an output meter the a.v.c. should be on; if the audio-output method is used, the a.v.c. should be off. The beat oscillator should be off in either case.

If the i.f. amplifier has a crystal filter, the filter should be switched out. Alignment is then carried out as described above, setting the signal generator as closely as possible to the frequency of the crystal. After alignment, 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 its exact frequency, which will be indicated by a sharp rise in output. Leaving the signal generator set on the crystal peak, the i.f. trimmers may be realigned for maximum output. The necessary readjustment should be small. The signal generator 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 of the receiver is used as a criterion of alignment. Lacking an a.v.c. tuning meter the transformers may be 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 transformers for maximum audio output.

An amplifier which is only slightly out of alignment, as a result of normal drift from temperature, humidity or aging effects, can be realigned by using any steady signal, such as a local broadcasting station, in lieu of a test oscillator. Allow the receiver to warm up thoroughly (an hour or so), tune in the signal as usual, and "touch up" the i.f. trimmers.

R.f. alignment — The objective in aligning the r.f. circuits in 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, or even on noise or such signals as may be heard. First set the tuning dial at the high-frequency end of the range in use. Then set the test oscillator to the frequency indicated by the receiver dial. The test-oscillator output may be connected to the antenna terminals of the receiver for this test. Adjust the oscillator trimmer condenser in the receiver to give maximum response on the test-oscillator signal, then reset the receiver dial to the low-frequency end of the range. Set the test-oscillator frequency near the frequency indicated by the receiver dial and carefully tune the test oscillator until its signal is heard in the receiver. If the frequency of the signal as indicated by the test-oscillator calibration is higher than that indicated by the receiver dial, more inductance (or more capacity in the tracking condenser) is needed in the receiver oscillator circuit; if the frequency is lower, less inductance (less tracking capacity) is required in the receiver oscillator. Most commercial receivers provide some means for varying the inductances of the coils



(A) (B) (C)

Fig. 735 Oscilloscope patterns of response characteristics on a visual curve tracer. The upper row illustrates various kinds of misalignment; the lower row shows the same stages properly aligned. A, i.f. curve of a selective communications-type receiver, with all transformers mistuned on one side of resonance (top). Below, the peaks coincide when properly aligned, even though skirts do not precisely match. B, at the top, a broad-band f.m. receiver curve taken after alignment by the fixed-frequency and output-meter method; the lower curve shows the improvement after careful visual alignment. C, the pattern of a medium-selectivity receiver with transformers misaligned symmetrically on either side of resonance (top); below, the same i.f. correctly aligned but with the test oscillator tuned slightly off frequency to displace the return trace for better examination.

or the capacity of the tracking condenser, to permit aligning the receiver tuning with the dial calibration. Set the test oscillator to the frequency indicated by the receiver dial, and then adjust the tracking capacity or inductance of the receiver oscillator coil to obtain maximum response. After making this adjustment, recheck the high-frequency end of the scale as previously described. It may be necessary to go back and forth between the ends of the range several times before the proper combination of inductance and capacity is secured. In many cases, better over-all tracking will result if frequencies near but not actually at the ends of the tuning range are selected, instead of taking the extreme dial settings.

After the oscillator range is properly adjusted, set the receiver and test oscillator to the high-frequency end of the range. Adjust the mixer trimmer condenser for maximum hiss or signal, then the r.f. trimmers. Reset the tuning dial and test oscillator to the low-frequency end of the range, and repeat; if the circuits are properly designed, no change in trimmer settings should be necessary. If it is necessary to increase the trimmer capacity in any circuit, it indicates that more inductance is needed; if less capacity resonates the circuit, less inductance is required.

Tracking seldom is perfect throughout a tuning range, so that a check of alignment at intermediate points in the range may show it to be slightly off. Normally the gain variation from this cause will be small, however, and it will suffice to bring the circuits into line at both ends of the range. If most reception is in a particular part of the range, such as an amateur band, the circuits may be aligned for maximum performance in that region, even though the ends of the frequency range as a whole may be slightly out of alignment.

Visual alignment — More accurate and efficient alignment of receiver circuits may be performed with the aid of a visual curve-tracer or "wobbulator" which traces out the response curve visually on a cathode-ray oscilloscope. This is accomplished by using a special signal generator in which the oscillator frequency is varied over a suitable range at a low audio rate. The horizontal sweep of the oscilloscope is synchronized with the rate of variation of the test frequency, so that the horizontal deflection is a function of frequency. The rectified output of the second detector is connected to the vertical deflection plates of the oscilloscope. The spot on the screen therefore traces a curve proportional to the receiver response in terms of the instantaneous value of the oscillator frequency. This visual response curve, which may be that of the entire receiver or of any stage, is continually visible as a whole. Thus the effect of any adjustment of the circuits may be observed much more rapidly than is possible with an ordinary signal generator and output meter, particularly in the case of wide-band i.f. circuits.

Fig. 736 — A, a typical single-trace response curve of a selective high-fidelity i.f. system. B, pattern of the amplifier in A made highly regenerative, illustrating instability. C, double trace of a single overcoupled i.f. stage with the return trace displaced. A similar "knee" located lower on the skirts would indicate regeneration.

Apparatus and methods for obtaining visual curve traces are described in Chapter Nineteen. The simplest arrangement is that which employs a reactance-tube modulated oscillator operating on 1000 kc., the output of which is combined with that from an unmodulated variable-tuning r.f. oscillator in a mixer tube, to provide a heterodyned signal at the desired center frequency.

Either "double trace" and "single trace" patterns may be used. The double trace pattern is obtained by applying a triangular sweep to the f.m. oscillator at a frequency half that of the sawtooth sweep on the horizontal plates of the cathode-ray tube. The return sweep produces a reversed pattern superimposed on the first, and is useful for checking symmetry and frequency calibration. The single-trace pattern shows the same two opposite-sequence resonance curves, but with the second curve displaced by a half cycle of the audio sweep frequency. It is useful in displaying irregularities in the pattern which might be obscured by superposition of the traces.

The alignment procedure follows that described for the oscillator–output-meter method. Assuming a diode second detector, run a shielded lead to the vertical input terminals of the oscilloscope from the "high" side of the diode load resistor — usually the audio volume control. With a triode biased detector, the bias resistor and by-pass condenser circuit should be opened and the vertical terminal connected to the cathode of the detector tube across a 0.5-megohm leak to ground, by-passed with a 250-μμfd. condenser. The plate load should be shorted out. This will make the resonance patterns appear upside down, but does not change their interpretation.

The r.f. output from the mixer should connect directly to the grid of the last i.f. tube. Add the i.f. frequency to 1000 kc. and set the unmodulated signal generator to this frequency. For example, if the i.f. is 465 kc., set the a.m. signal generator to 1465 kc. At the usual bandwidth of 30 kc., the signal at the grid of the last i.f. stage will swing from 450 kc. to 480 kc. and back. If the signal generator is set to the exact i.f., a double-trace pattern should appear on the screen. Center this pattern with the oscilloscope sweep vernier. Adjust the i.f. trimmers until these peaks coincide. For single-trace analysis, the oscilloscope sweep frequency should be reduced one half.

To align the next i.f. stage, move the r.f. output lead to the grid of the tube and adjust the next i.f. transformer. It may be necessary to readjust the output transformer after this operation. When aligning triple-tuned or high-fidelity i.f. circuits, it is most important that the peaks in the double pattern coincide and have nearly equal amplitude.

To align the r.f. and mixer input circuits, the variable-frequency signal generator should be set to a frequency which, by addition to 1000 kc., produces the desired r.f. signal frequency. As each stage is added, the output level must be reduced to keep the pattern on the screen. To avoid overloading, only enough signal should be used to overcome local interference. Adjust the r.f. trimmers for maximum vertical amplitude of the pattern, as with an output meter. Dial calibration can be checked by setting the test oscillator on frequency and adjusting the h.f. oscillator trimmer in the receiver to center the pattern on the screen.

Fig. 737 — Response curves of a superheterodyne with crystal filter (made at a very low repetition rate). A, crystal in "broad" position, phasing control at center. B, phasing control set to place the rejection slot on low-frequency side. C, with slot on high-frequency side.

Oscillation in r.f. or i.f. amplifiers — Oscillation in high-frequency amplifier and mixer circuits may be evidenced by squeals or "birdies" as the tuning is varied, or by complete lack of audible output if the oscillation is strong enough to cause the a.v.c. system to reduce the receiver gain drastically. Oscillation can be caused by poor connections in the common ground circuits, especially to the tuning-condenser rotors. Inadequate or defective by-pass condensers in cathode, plate and screen-grid circuits also can cause such oscillation. In some cases it may be advisable to provide a shield between the stators of pre-r.f. amplifier and first-detector ganged tuning condensers, in addition to the usual tube and interstage shielding. A metal tube with an ungrounded shell will cause trouble. Improper screen-grid voltage, resulting from a shorted or too-low screen-grid series resistor, also may be responsible for such instability.

Oscillation in the i.f. circuits is independent of high-frequency tuning, and is indicated by a continuous squeal which appears when the gain is advanced with the c.w. beat oscillator on. It can result from similar defects in i.f. amplifier circuits. Inadequate cathode by-pass capacitance is a common cause of such oscillation. An additional by-pass condenser of 0.1 to 0.25 μfd. usually will remedy the trouble. Similar treatment can be applied to the screen-grid and plate by-pass filters of i.f. stages.

Instability — "Birdies" or a mushy hiss occurring with tuning of the high-frequency oscillator may indicate that the oscillator is "squegging" or oscillating simultaneously at high and low frequencies (§ 7-4). 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 (§ 7-10), 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 (§ 7-9) also will cause the beat-note to "chirp" on strong c.w. signals because the oscillator load changes slightly.

In 'phone reception with a.v.c., a peculiar type of instability ("motorboating") may appear if the h.f. oscillator frequency is sensitive to changes in plate voltage. As the a.v.c. voltage rises the electrode currents of the controlled tubes decrease, decreasing the load on the power supply and causing its output voltage to rise. Since this increases the voltage applied to the oscillator, its frequency changes correspondingly, throwing the signal off the peak of the i.f. resonance curve and reducing the a.v.c. voltage, thus tending to restore the original conditions. The process then repeats itself, at a rate determined by the signal strength and the time constant of the power-supply circuits. This effect is most pronounced with high i.f. selectivity, as when a crystal filter is used, and can be cured by making the oscillator relatively insensitive to voltage changes and by regulating the plate voltage supply (§7-10).

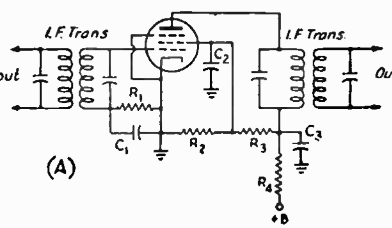
7-18 Reception of Frequency-Modulated Signals

F.m. receivers — A frequency-modulation receiver differs in circuit design from one designed for amplitude modulation chiefly in the arrangement used for detecting the signal. Detectors for amplitude-modulated signals do not respond to frequency modulation. It is also necessary, for full realization of the noise-reducing benefits of the f.m. system, that the signal applied to the detector be completely free from amplitude modulation. In practice, this is attained by preventing the signal from rising above a given amplitude by means of a limiter (§ 3-10, 7-15). Since the weakest signal must be amplitude-limited, high gain must be provided ahead of the limiter; the superheterodyne type of circuit almost invariably is used to provide the necessary gain.

The r.f. and i.f. stages in a superheterodyne for f.m. reception are practically identical in circuit arrangement with those in an a.m. receiver. Since the use of f.m. is confined to the very-high frequencies (above 28 Mc.) a high intermediate frequency is employed, usually between 4 and 5 Mc. This not only reduces image response but also provides the greater band-width necessary to accommodate wide-band frequency-modulated signals.

Receiver requirements — The primary requirements are sufficient r.f. and i.f. gain to "saturate" the limiter even with a weak signal, sufficient band-width (§ 7-2) to accommodate the full frequency deviation either side of the carrier frequency without undue attenuation at the edges of the band, a limiter circuit which functions properly on both rapid and slow variations in amplitude, and a detector which gives a linear relationship between frequency deviation and *amplitude* output. The audio circuits are the same as in other receivers (§ 7-5), except that in communications-type receivers it is desirable to cut off the upper audio range by a low-pass filter (§ 2-11) because higher-frequency noise components have the greatest amplitude in an f.m. receiver.

The limiter — Limiter circuits generally are of the plate-saturation type (§ 7-15), where low plate and screen voltage are used to limit the plate-current flow at high signal amplitudes. Fig. 738-A is a typical circuit. The tube is self-biased (§ 3-6) by a grid leak, R_1, and condenser, C_1. R_2, R_3 and R_4 form a voltage divider



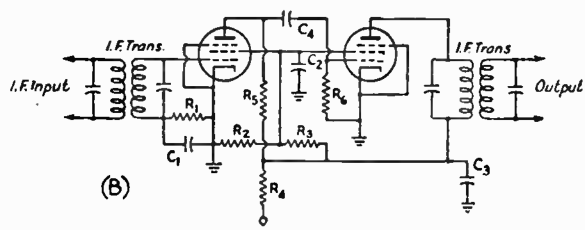


Fig. 738 — F.m. limiter circuits. A, single-tube plate-saturation limiter; B, cascade limiter. Typical values are:

	Circuit A	Circuit B
C_1 —	100 μμfd.	100 μμfd.
C_2, C_3 —	0.1 μfd.	0.1 μfd.
C_4 —		250 μμfd.
R_1 —	0.1 megohm.	50,000 ohms.
R_2 —	2000 ohms.	2000 ohms.
R_3 —	50,000 ohms.	50,000 ohms.
R_4 —	0–50,000 ohms.	0–50,000 ohms.
R_5 —		4000 ohms.
R_6 —		0.2 megohm.

Plate-supply voltage is 250 in both circuits.

(§ 8-10) which puts the desired voltages on the screen and plate. The lower the voltages the lower the signal level at which limiting occurs, but the r.f. output voltage of the limiter also is lower. C_2 and C_3 are the plate and screen by-pass condensers, of conventional value for the intermediate frequency used. The time constant (§ 2-6) of R_1C_1 determines the behavior of the limiter with respect to rapid and slow amplitude variations. For best operation on impulse noise (§ 7-15) the time constant should be small, but a too-small time constant limits the range of signal strengths the limiter can handle without departing from the constant-output condition. A larger time constant is better in this respect but is not so effective for rapid variations. Compromise constants are shown in Fig. 738.

The cascade limiter, Fig. 738-B, overcomes this by making the time constant in the first grid circuit suitable for effective operation on impulse noise, and that in the second grid (C_4R_6) optimum for a wide range of input signal strengths. This results, in addition, in more constant output over a very wide range of input signal amplitudes because the voltage at the grid of the second stage already is partially amplitude-limited. Resistance coupling ($R_5C_4R_6$) is used for simplicity and to prevent unwanted regeneration, additional gain at this point being unnecessary.

The rectified voltage developed across R_1 in either circuit may be applied to the i.f. amplifier for a.v.c. (§ 7-13).

Discriminator circuits and operation — The f.m. detector commonly is called a *discriminator*, because of its ability to discriminate between frequency deviations above and those below the carrier frequency.

A rectifier connected to an ordinary tuned circuit adjusted so that the signal frequency falls on one side of the response curve constitutes an elementary discriminator, because the rectifier output will vary with a change in the carrier frequency. If two such circuits are used with a balanced rectifier, one tuned above and the other below the signal frequency, amplitude variations are balanced out and the combined rectified current is proportional to the frequency deviation.

The circuit most widely used is the "series" or center-tuned discriminator shown in Fig. 739-A. A special i.f. coupling transformer is used between the limiter and detector. Its secondary, L_1, is center-tapped and is connected back to the plate side of the primary circuit, which otherwise is conventional. C_4 is the tuning condenser. The load circuits of the two diode rectifiers ($R_1C_1R_2C_2$) are connected in series; constants are the same as in ordinary diode detector circuits (§7-3). Audio output is taken from across the two load resistances.

The primary and secondary circuits are both adjusted to resonance in the center of the i.f. pass-band. The voltage applied to the rectifiers consists of two components, that induced in the secondary by the inductive coupling and that fed to the center of the secondary through C_2. The phase relations between the two are such that at resonance the rectified load currents are equal in amplitude but flow in opposite directions through R_1 and R_2, hence the net voltage across the terminals marked "audio output" is zero. When the carrier deviates from resonance the induced secondary current either lags or leads, depending upon whether the deviation is to the high- or low-frequency side, and this phase shift causes the induced current to combine with that fed through C_2 in such a way that one diode gets more voltage than the other when the frequency is below resonance, while the second diode gets the larger voltage when the frequency is higher than resonance. The voltage appearing across the output terminals is the difference between the two diode voltages. Thus a characteristic like that of Fig. 740 results, where the net rectified output voltage has opposite polarity for frequencies on either side of resonance, and up to a certain point becomes greater in amplitude as the frequency deviation is greater. The straight-line portion of the curve is the useful detector characteristic. The separation between the peaks which mark the ends of the linear portion of the curve depends upon the Qs of the primary and secondary circuits and the degree of coupling. The separation becomes greater with low Qs and close coupling. The circuit ordinarily is designed so that the peaks fall just outside the limits of the pass-band, thus utilizing most of the straight portion of the curve. Since the audio output is proportional to the change in d.c. voltage with deviation, it is advantageous for maximum output to keep the frequency separation between peaks down to the minimum value necessary for a linear characteristic.

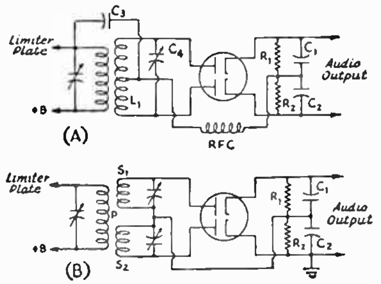


Fig. 739 — F.m. discriminator circuits. In both circuits typical values for C_1 and C_2 are 100 μμfd. each; R_1 and R_2, 0.1 megohm each. C_3 in A is approximately 50 μμfd., depending upon the intermediate frequency; *RFC* should be of a type designed for the i.f. in use (2.5 mh. is satisfactory for i.f.s of 4 to 5 Mc.). In either circuit the ground may be moved from the lower end of C_2 to the junction of C_1 and C_2, for push-pull audio output.

A second type of discriminator is shown in Fig. 739-B. Two secondary circuits are used, one tuned above the center frequency of the i.f. pass-band and the other below. They are coupled equally to the primary, which is tuned to the center frequency. As the carrier fre-

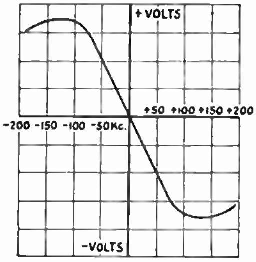


Fig. 740 — Characteristic of a typical f.m. detector. The vertical axis represents the voltage developed across the load resistor as the frequency varies from the exact resonance frequency. This detector would handle f.m. signals up to a band-width of 150 kc. over the linear portion of the curve.

quency deviates the voltages induced in the secondaries will change in amplitude, the larger voltage appearing across the secondary being nearer resonance with the instantaneous frequency. The detection characteristic is similar to that of the center-tuned discriminator. The peak separation is determined by the *Q*s of the circuits, the coefficient of coupling, and the tuning of the secondaries. High *Q*s and loose coupling are required for close peak separation.

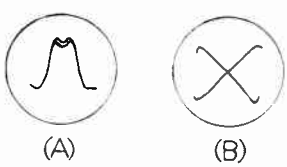
A simple self-quenched superregenerative receiver may be used as a frequency detector if it is tuned so that the carrier frequency falls along the slope of the resonance curve. Two such detectors, off-tuned on either side of the carrier, may be used in push-pull. An alternative arrangement employing a superregenerative stage as a first i.f. amplifier at 75 Mc., following a converter unit, provides high gain and linear response with relatively few stages.

F.m. receiver alignment — Alignment of f.m. receivers up to the limiter is carried out as described in § 7-17. For output measurement, a 0–1 milliammeter or 0–500 microammeter should be connected in series with the limiter grid resistor (R_1 in Fig. 738) at the grounded end; or, if the voltage drop across R_1 is used for a.v.c. and the receiver is provided with a tuning meter (§ 7-13), the tuning meter may be used as an output meter. An accurately calibrated signal generator or test oscillator is desirable, since the i.f. should be aligned to be as symmetrical as possible; that is, the output reading should be the same for any two test oscillator settings the same number of kilocycles above or below resonance. It is not necessary to have uniform response over the whole band to be received, although the output at the edges of the band (limit of deviation (§ 5-11) of the transmitted signals) should not be less than 25 per cent of the voltage at resonance. In communications work, a band-width of 30 kc. or less (15 kc. or less deviation) is commonly used. Output readings should be taken with the oscillator set at intervals of a few kilocycles either side of resonance up to the band limits.

After the i.f. (and front-end) alignment, the limiter operation should be checked. This can be done by temporarily disconnecting C_3, if the discriminator circuit of Fig. 739-A is used, disconnecting R_1 and C_1 on the cathode side, and inserting the milliammeter or microammeter in series with R_2 at the grounded end. This converts the discriminator to an ordinary diode rectifier. Varying the signal-generator frequency over the channel, with the discriminator transformer adjusted to resonance, should show no change in output (at the band-widths used for communications purposes) as indicated by the rectified current read by the meter. At this point various plate and screen voltages can be tried on the limiter tube or tubes, to determine the set of conditions which gives maximum output with adequate limiting (no change in rectified current).

When the limiter has been checked the discriminator connections can be restored, leaving the meter connected in series with R_1. Provision should be made for reversing the connections to the meter terminals, to take care of the reversal in polarity of the net rectified current. Set the signal generator to the center frequency of the band and adjust the discriminator transformer trimmer condensers to resonance, which will be indicated by zero rectified current. Then set the test oscillator at the deviation limit (§ 5-11) on one side of the center frequency, and note the meter reading. Reverse the meter terminals and set the test oscillator at the deviation limit on the other side. The two readings should be the same. If they are not, they can be made so by a slight adjustment of the primary trimmer. This will necessitate rechecking the response at resonance to make sure it is still zero. Generally, the secondary trimmer will chiefly affect the zero-response frequency, while the primary trimmer will have most effect on the symmetry of the discriminator peaks. A detector curve having satisfactory linearity can be obtained by cut-and-try adjustment of both trimmers.

Fig. 741 — Oscilloscope patterns in f.m. i.f. alignment. A — I.f. amplifier response. B — Over-all characteristic through the f.m. detector.



A visual curve tracer is particularly advantageous in aligning the wide-band i.f. amplifiers of f.m. receivers. The i.f. is first aligned with the discriminator circuit converted into an a.m. diode detector, as described above, the pattern appearing as in Fig. 741-A. The over-all characteristic, including the f.m. detector, is shown in Fig. 741-B.

Tuning and operation — An f.m. receiver gives greatest noise reduction when the carrier is tuned exactly to the center of the receiver pass-band and to the point of zero response in the discriminator. Because of the decrease in noise, this point is readily recognized.

When an amplitude-modulated signal is tuned in its modulation practically disappears at exact resonance, only those nonsymmetrical modulation components which may be present being detected. If the signal is to one side or the other of resonance, however, it is capable of causing interference to an f.m. signal.

Chapter Eight

Power Supply

8-1 Power-Supply Requirements

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 (§ 2-9) 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 (§ 2-6) across it. The filament or heater transformer generally is center-tapped, to provide a balanced circuit for eliminating hum (§ 3-6).

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.

Plate supply — Direct current must be used for the plates of tubes, since any variation in plate current arising from power-supply causes will be superimposed on the signal being received or transmitted, giving an undesirable type of modulation (§ 5-1) if the variations occur at an audio-frequency (§ 2-7) rate. Unvarying direct current is called *pure d.c.*, to distinguish it from current which may be unidirectional but of pulsating character. The use of pure d.c. on the plates of transmitting tubes is required by FCC regulations on all frequencies below 60 Mc.

Sources of plate power — D.c. plate power is usually obtained from rectified and filtered alternating current, but in low-power and portable installations may be secured from batteries. Dry batteries may be used for very low-power portable equipment, but in many cases a storage battery is used as the primary power source, in conjunction with an interrupter giving pulsating d.c. which is applied to the primary of a step-up transformer (§8-10).

Rectified-a.c. supplies — Since the power-line voltage ordinarily is 115 or 230 volts, a step-up transformer (§ 2-9) is used to obtain the desired voltage for the plates of the tubes in the equipment. The alternating secondary current is changed to unidirectional current by means of diode rectifier tubes (§ 3-1), and then passed through an inductance-capacity filter (§ 2-11) to the load circuit. The *load resistance* in ohms is equal to the d.c. output voltage of the power supply divided by the current in amperes (Ohm's Law, § 2-6).

Voltage regulation — Since there is always some resistance in power-supply circuits, and since the filter normally depends to a considerable extent upon the energy storage of inductance and capacity (§ 2-3, 2-5), the output voltage will depend upon the current drain on the supply. The change in output voltage with change in load current is called the *voltage regulation.* It is expressed as a percentage:

$$\% \text{ Regulation} = \frac{100\,(E_1 - E_2)}{E_2}$$

where E_1 is the no-load voltage (no current in the load circuit) and E_2 the full-load voltage (rated current in load circuit).

8-2 Rectifiers

Purpose and ratings — A rectifier is a device which will conduct current only in one direction. The diode tube (§ 3-1) is used almost exclusively for rectification in d.c. power supplies used with radio equipment. The important characteristics of tubes used as power-supply rectifiers are the voltage drop between plate and cathode at rated current, the maximum permissible inverse peak voltage, and the permissible peak plate current.

Voltage drop — Tube voltage drop depends upon the type of tube. In vacuum-type rectifiers it increases with the current flowing because of space-charge effect (§ 3-1), but can be minimized by using very small spacing between plate and cathode as is done in some rectifiers for receiver power supplies. Mercury-vapor rectifiers (§ 3-5) have a constant drop of about 15 volts, regardless of current. This is much smaller than the voltage drops encountered in vacuum-type rectifiers.

Inverse peak voltage — This is the maximum voltage developed between the plate and cathode of the rectifier when the tube is not conducting; i.e., when the plate is negative with respect to the cathode.

Peak plate current — This is the maximum *instantaneous* current through the rectifier. It can never be smaller than the load current in ordinary circuits, and may be several times higher.

Operation of mercury-vapor rectifiers — Because of its constant voltage drop, the mercury-vapor rectifier is more susceptible to damage than the vacuum type. With the latter, the increase in voltage drop tends to limit current flow on heavy overloads, but the mercury-vapor rectifier does not have this limiting action and the cathode may be damaged under similar conditions.

In mercury-vapor rectifiers a phenomenon known as "arc-back," or breakdown of the mercury vapor and conduction in the opposite direction to normal, occurs at high inverse peak voltages, hence such tubes always should be operated within their inverse-peak voltage ratings. Arc-back also may occur if the cathode temperature is below normal; therefore the heater or filament voltage should be checked to make sure that the rated voltage is applied. This check should be made at the tube socket, to avoid errors caused by voltage drop in the leads. For the same reason, the cathode should be allowed to come up to its final temperature before plate voltage is applied; the time required for this is of the order of 15 to 30 seconds. When a tube is first installed, or is put into service after a long period of idleness, the cathode should be heated for a period of 10 minutes or so before application of plate voltage.

8-3 Rectifier Circuits

Half-wave rectifiers — The simple diode rectifier (§ 3-1) is called a *half-wave rectifier*, because it can pass only half of each cycle of alternating current. Its circuit is shown in Fig. 801-A. At the top of the figure is a representation of the applied a.c. voltage, with positive and negative alternations (§ 2-7) marked.

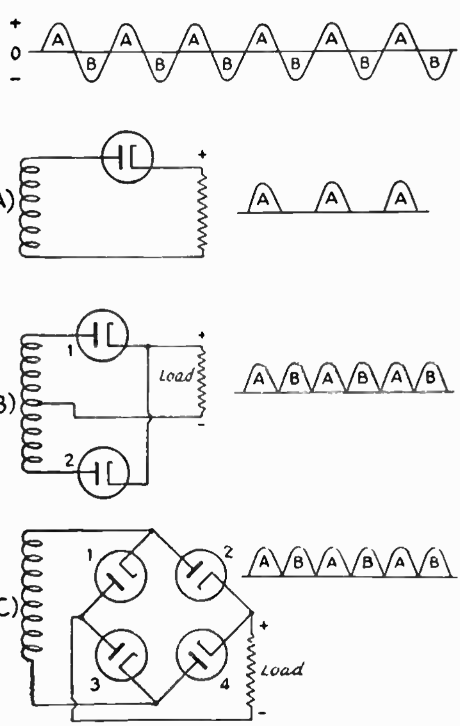


Fig. 801 — Fundamental vacuum-tube rectifier circuits.

When the plate is positive with respect to cathode, plate current flows through the load as indicated in the drawing at the right, but when the plate is negative with respect to cathode no current flows. This is indicated by the gaps in the output drawing. The output current is unidirectional but pulsating.

In this circuit the inverse peak voltage is equal to the maximum transformer voltage, which in the case of a sine wave is 1.41 times the r.m.s. voltage (§ 2-7).

Full-wave center-tap rectifier — Fig. 801-B shows the "full-wave center-tap" rectifier circuit, so called because both halves of the a.c. cycle are rectified and because the transformer secondary winding must consist of two equal parts with a connection brought out from the center. When the upper end of the winding is positive, current can flow through rectifier No. 1 to the load; this current cannot pass through rectifier No. 2 because its cathode is positive with respect to its plate. The circuit is completed through the transformer center-tap. When the polarity reverses the upperend of the winding is negative and no current can flow through No. 1, but the lower end is positive and therefore No. 2 passes current to the load, the return connection again being the center-tap. The resulting waveshape is shown at the right.

Since the two rectifiers are working alternately in this circuit, each half of the transformer secondary must be wound to deliver the full-load voltage; hence the total voltage across the transformer terminals is twice that required with the half-wave rectifier. Assuming negligible voltage drop in the particular rectifier which may be conducting at any instant, the inverse peak voltage on the other rectifier is equal to the maximum voltage between the outside terminals of the transformer. In the case of a sine wave, this is 1.41 times the total secondary r.m.s. voltage (§ 2-7).

Because energy is delivered to the load at twice the average rate as in the case of a half-wave rectifier, each tube carries only half the load current.

The bridge rectifier — The "bridge" type of full-wave rectifier is shown in Fig. 801-C. Its operation is as follows: When the upper end of the winding is positive, current can flow through No. 2 to the load but not through No. 1. On the return circuit, current flows through No. 3 by way of the lower end of the transformer winding. When the polarity reverses and the lower end of the winding becomes positive, current flows through No. 4 and the load and through No. 1 by way of the upper side of the transformer. The output waveshape is shown at the right.

The inverse peak voltage is equal to the maximum transformer voltage, or 1.41 times the r.m.s. secondary voltage in the case of a sine wave (§ 2-7). Energy is delivered to the load at the same average rate as in the case of the full-wave center-tap rectifier, each *pair* of tubes in series carrying half the load current.

8-4 Filters

Purpose of filter — As shown in Fig. 801, the output of a rectifier is pulsating d.c., which would be unsuitable for most vacuum-tube applications (§ 8-1). A *filter* is used to smooth out the pulsations so that practically unvarying direct current flows through the load circuit. The filter utilizes the energy-storage properties of inductance and capacity (§ 2-3, 2-5), by virtue of which energy stored in electromagnetic and electrostatic fields when the voltage and current are rising is restored to the circuit when the voltage and current fall, thus filling in the "gaps" or "valleys" in the rectified output.

Ripple voltage and frequency — The pulsations in the output of the rectifier can be considered to be caused by an alternating current superimposed on a steady direct current (§ 2-13). Viewed from this standpoint, the filter may be considered to consist of by-pass condensers which short-circuit the a.c. while not interfering with the flow of d.c., and *chokes* or inductances which permit d.c. to flow through them but which have high reactance for the a.c. (§ 2-13). The alternating component is called the *ripple*. The effectiveness of the filter may be measured by the *per cent ripple*, which is the r.m.s. value of the a.c. ripple voltage expressed as a percentage of the d.c. output voltage. With an effective filter, the ripple percentage will be low. Five per cent ripple is considered satisfactory for c.w. transmitters, but lower values (of the order of 0.25 per cent) are necessary for hum-free speech transmission and for receiver plate supplies.

The ripple frequency depends upon the line frequency and the type of rectifier. In general, it consists of a fundamental plus a series of harmonics (§ 2-7), the latter being relatively unimportant since the fundamental is hardest to smooth out. With a half-wave rectifier, the fundamental is equal to the line frequency; with a full-wave rectifier, the fundamental is equal to twice the line frequency, or 120 cycles in the case of a 60-cycle supply.

Types of filters — Inductance-capacity filters are of the low-pass type (§ 2-11), using series inductances and shunt capacitances. Practical filters are identified as *condenser-input* and *choke-input*, depending upon whether a capacity or inductance is used as the first element in the filter. Resistance-capacity filters (§ 2-11) are used in applications where the current is very low and the voltage drop in the resistor can be tolerated.

Bleeder resistance — Since the condensers in a filter will retain their charge for a considerable time after power is removed (provided the load circuit is open at the time), it is good practice to connect a resistor across the output of the filter to discharge the condensers when the power supply is not in use. The resistance usually is high enough so that only a relatively small percentage of the total output current is consumed in it during normal operation.

Components — Filter condensers are made in several different types. Electrolytic condensers, which are available for voltages up to about 800, combine high capacity with small size, since the dielectric is an extremely thin film of oxide on aluminum foil. Condensers for higher voltages usually are made with a dielectric of thin paper impregnated with oil. The *working voltage* of a condenser is the voltage which it will withstand continuously.

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 (§ 2-5) 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 load value.

8-5 Condenser-Input Filters

Ripple voltage — The conventional condenser-input filter is shown in Fig. 802-A. No simple formulas are available for computing

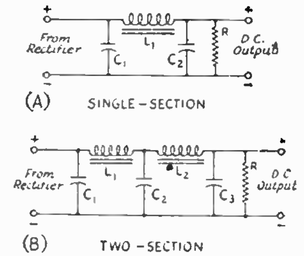


Fig. 802 — Condenser-input filter circuits.

the ripple voltage, but it will be smaller as both capacity and inductance are made larger. Adequate smoothing for transmitting purposes can be secured by using 4 to 8 μfd. at C_1 and C_2 and 20 to 30 henrys at L_1, for full-wave rectifiers with 120-cycle ripple (§ 8-4). A higher ratio of inductance to capacity may be used at higher load resistances (§ 8-1).

For receivers, as shown in Fig. 802-B, an additional choke, L_2, and condenser, C_3, of the same approximate values, are used to give additional smoothing. In such supplies the three condensers generally are 8 μfd. each, although the input condenser, C_1, sometimes is reduced to 4 μfd. Inductances of 10 to 20 henrys each will give satisfactory filtering with these capacity values.

For ripple frequencies other than 120 cycles, the inductance and capacity values should be multiplied by the ratio $120/F$, where F is the actual ripple frequency.

The bleeder resistance, R, should be chosen to draw 10 per cent or less of the rated output current of the supply. Its value is equal to $1000E/I$, where E is the output voltage and I the bleeder current in milliamperes.

Rectifier peak current — The ratio of rectifier peak current to average load current is high with a condenser-input filter. Small rectifier tubes designed for low-voltage supplies (type 80, etc.) generally carry load-current ratings based on the use of condenser-input filters. With rectifiers for higher power, such as the 866, 866-A, the load current should not exceed 25 per cent of the rated peak plate current for one tube when a full-wave rectifier is used, or one-eighth the half-wave rating.

Output voltage — The d.c. output voltage from a condenser-input supply will, with light loads or no load, approach the peak transformer voltage. This is 1.41 times the r.m.s. voltage (§ 2-7) of the transformer secondary, in the case of Figs. 801-A and C, or 1.41 times the voltage from the center-tap to one end of the secondary in Fig. 801-B. At heavy loads, it may decrease to the *average* value of secondary voltage or about 90 per cent of the r.m.s. voltage, or even less. Because of this wide range of output voltage with load current, the voltage regulation (§ 8-1) is inherently poor.

The output voltage obtainable from a given supply cannot readily be calculated, since it depends critically upon the load current and filter constants. Under average conditions it will be approximately equal to or somewhat less than the r.m.s. voltage between the center-tap and one end of the secondary in the full-wave center-tap rectifier circuit (§ 8-3).

Ratings of components — Because the output voltage may rise to the peak transformer voltage at light loads, the condensers should have a working-voltage rating (§ 8-4) at least as high and preferably somewhat higher, as a safety factor. Thus, in the case of a center-tap rectifier having a transformer delivering 550 volts each side of the center-tap, the minimum safe condenser voltage rating will be 550 × 1.41 or 775 volts. An 800-volt, or preferably a 1000-volt, condenser should be used. Filter chokes should have the inductance specified at full-load current, and must have insulation between the winding and the core adequate to withstand the maximum output voltage.

§ 8-6 Choke-Input Filters

Ripple voltage — The circuit of a single-section choke-input filter is shown in Fig. 803-A. For 120-cycle ripple, a close approximation of the ripple to be expected at the output of the filter is given by the formula:

$$\left.\begin{array}{l}\text{Single}\\\text{Section}\\\text{Filter}\end{array}\right\}\ \%\ \text{Ripple} = \frac{100}{LC}$$

where L is in henrys and C in μfd. The product, LC, must be equal to or greater than 20 to reduce the ripple to 5 per cent or less. This figure represents, in most cases, the economical limit for the single-section filter. Smaller percentages of ripple usually are more economically obtained with the two-section filter of Fig. 803-B. The ripple percentage (120-cycle ripple) with this arrangement is given by the formula:

$$\left.\begin{array}{l}\text{Two}\\\text{Section}\\\text{Filter}\end{array}\right\}\ \%\ \text{Ripple} = \frac{650}{L_1L_2\ (C_1 + C_2)^2}$$

For a ripple of 0.25 per cent or less, the denominator should be 2600 or greater.

These formulas can be used for other ripple frequencies by multiplying each inductance and capacity value in the filter by the ratio 120/F, where F is the actual ripple frequency.

The distribution of inductance and capacity in the filter will be determined by the value of input-choke inductance required (next paragraph), and the permissible a.c. output impedance. If the supply is intended for use with an audio-frequency amplifier, the reactance (§ 2-8) of the last filter condenser should be small (20 per cent or less) compared to the other a.f. resistance or impedance in the circuit, usually the tube plate resistance and load resistance (§3-2, 3-3). On the basis of a lower a.f. limit of 100 cycles for speech amplification (§ 5-9), this condition is usually satisfied when the output capacity (last filter capacity) of the filter is 4 to 8 μfd., the higher value being used for the lower tube and load resistances.

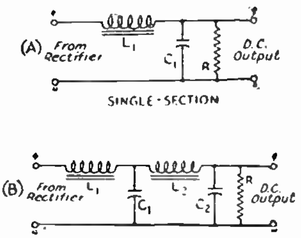


Fig. 803 — Choke-input filter circuits.

The input choke — The rectifier peak current and the power-supply voltage regulation depend almost entirely upon the inductance of the input choke in relation to the load resistance (§ 8-1). The function of the choke is to raise the ratio of average to peak current (by its energy storage), and to prevent the d.c. output voltage from rising above the average value (§ 2-7) of the a.c. voltage applied to the rectifier. For both purposes, its impedance (§ 2-8) to the flow of the a.c. component (§ 8-4) must be high.

The value of input-choke inductance which prevents the d.c. output voltage from rising above the average of the rectified a.c. wave is the *critical inductance*. For 120-cycle ripple, it is given by the approximate formula:

$$L_{crit.} = \frac{\text{Load resistance (ohms)}}{1000}$$

For other ripple frequencies, the inductance required will be the above value multiplied by the ratio of 120 to the actual ripple frequency.

With inductance values less than critical, the d.c. output voltage will rise because the filter tends to act as a condenser-input filter (§ 8-5). With critical inductance, the peak plate current of one tube in a center-tap rectifier will be approximately 10 per cent higher than the d.c. load current taken from the supply.

An inductance of twice the critical value is called the *optimum* value. This value gives a further reduction in the ratio of peak to average plate current, and represents the point at which further increase in inductance does not give correspondingly improved operating characteristics.

Swinging chokes — The formula for critical inductance indicates that the inductance required varies widely with the load resistance. In the case where there is no load except the bleeder (§ 8-4) on the power supply, the critical inductance required is highest; much lower values are satisfactory when the full-load current is being delivered. Since the inductance of a choke tends to rise as the direct current flowing through it is decreased (§ 8-4), it is possible to effect an economy in materials by designing the choke to have a "swinging" characteristic such that it has the required critical inductance value with the bleeder load only, and about the optimum inductance value at full load. If the bleeder resistance is 20,000 ohms and the full-load resistance (including the bleeder) is 2500 ohms, a choke which swings from 20 henrys to 5 henrys over the full output-current range will fulfill the requirements.

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 (§ 2-10). 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 (§ 8-4), and resonance will occur when the product of choke inductance in henrys times condenser capacity in microfarads is equal to 1.77. The corresponding figure for 50-cycle supply (100-cycle ripple frequency) is 2.53, and for 25-cycle supply (50-cycle ripple frequency), 13.5. At least twice these products should be used to ensure against resonance effects.

Output voltage — Provided the input-choke inductance is at least the critical value, the output voltage may be calculated quite closely by the 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 (§ 8-2). These voltage drops are shown in Fig. 804.

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 (§ 8-1).

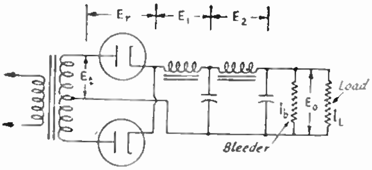


Fig. 804 — Voltage drops in the power-supply circuit.

Ratings of components — Because of better voltage regulation, filter condensers are subjected to smaller variations in d.c. voltage than in the condenser-input filter (§ 8-5). However, it is advisable to use condensers rated for the peak transformer voltage in case the bleeder resistor should burn out when there is no external load on the power supply, since the voltage then will rise to the same maximum value as with a condenser-input filter.

The input choke may be of the swinging type, the required no-load and full-load inductance values being calculated as described above. The second choke (*smoothing choke*) should have constant inductance with varying d.c. load currents. Values of 10 to 20 henrys ordinarily are used. Since chokes usually are placed in the positive leads, the negative being grounded, the windings should be insulated from the core to withstand the full d.c. output voltage of the supply.

8-7 The Plate Transformer

Output voltage — The output voltage of the plate transformer depends upon the required d.c. load voltage and the type of rectifier circuit. With condenser-input filters, the r.m.s. secondary voltage usually is made equal to or slightly more than the d.c. output voltage, allowing for voltage drops in the rectifier tubes and filter chokes as well as in the transformer itself. The full-wave center-tap rectifier requires a transformer giving this voltage each side of the secondary center-tap (§ 8-3).

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 resistances of the filter chokes, and E_r is the voltage drop in the rectifier. E_t is the full-load r.m.s. (§ 2-7) secondary voltage; the open-circuit voltage usually will be 5 to 10 per cent higher.

Volt-ampere rating — The volt-ampere rating (§ 2-8) 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 (§ 8-6), the secondary volt-amperes can be calculated quite closely by the equation:

$$\text{Sec. V.A.} = 0.00075\ EI$$

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.

8-8 Voltage Stabilization

Gaseous regulator tubes — There is frequent need for maintaining the voltage applied to a low-voltage low-current circuit (such as the oscillator in a superhet receiver or the frequency-controlling oscillator in a transmitter) 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. The first number in the tube designation indicates the terminal voltage, the second the maximum permissible tube current.

The fundamental circuit for a gaseous regulator is shown in Fig. 805-A. The tube is connected in series with a *limiting resistor*, R_1, across a source of voltage which must be higher than the *starting* voltage, or voltage required for ionization of the gas in the tube. 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 30 ma.; consequently, the load current cannot exceed 20 to 25 ma. if the voltage is to be stabilized over a range from zero to maximum load current.

The value of the limiting resistor must lie between that which just permits minimum tube current to flow and that which just passes the maximum permissible tube current when there is no load current. The latter value is generally used. It is given by the equation:

$$R = \frac{1000\ (E_s - E_r)}{I}$$

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 30 ma.).

Fig. 805-B shows how two tubes may be

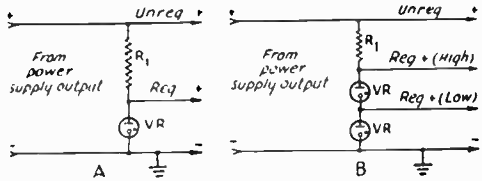


Fig. 805 Voltage-stabilizing circuits using VR tubes.

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 20 to 25 milliamperes.

Voltage regulation of the order of 1 per cent can be obtained with circuits of this type.

Electronic voltage regulation — A voltage regulator circuit suitable for higher voltages and currents than the gaseous tubes, and also having the feature that the output voltage can be varied over a rather wide range, is shown in Fig. 806. A high-gain voltage amplifier tube (§ 3-3), usually a sharp cut-off pentode (§ 3-5) is connected in such a way that a small change in the output voltage of the power supply causes a change in grid bias, and thereby a corresponding change in plate current. Its plate current flows through a resistor (R_5), the voltage drop across which is used to bias a second tube — the "regulator" tube — whose plate-cathode circuit is connected in series with the load circuit. The regulator tube therefore functions as an automatically variable series resistor. Should the output voltage increase slightly the bias on the control tube will become more positive, causing the plate current of the control tube to increase and the drop across R_5 to increase correspondingly. The bias on the regulator tube therefore becomes more negative and the effective resistance of the regulator tube increases, causing the terminal voltage to drop. A decrease in output voltage causes the reverse action. The time lag in the action of the system is negligible, and with proper circuit constants the output voltage can be held within a fraction of a per cent throughout the useful range of load currents and over a wide range of supply voltages.

An essential in this system is the use of a constant-voltage bias source for the control tube. The voltage change which appears at the grid of the tube is the *difference* between a fixed negative bias and a positive voltage which is taken from the voltage divider across the output. To get the most effective control, the negative bias must not vary with plate current. The most satisfactory type of bias is a dry battery of 45 to 90 volts, but a gaseous regulator tube (VR75-30) or a neon bulb of the type without a resistor in the base may be used instead. If the gas tube or neon bulb is used, a negative-resistance type of oscillation (§ 3-7) may take place at audio frequencies or higher, in which case a condenser of 0.1 μfd. or more should be connected across the tube. A similar condenser between the control-tube grid and cathode also is frequently helpful in this respect.

The variable resistor, R_3, is used to adjust the bias on the control tube to the proper operating value. It also serves as an output voltage control, setting the value of regulated voltage within the existing operating limits.

The maximum output voltage obtainable is equal to the power-supply voltage minus the minimum drop through the regulator tube. This drop is of the order of 50 volts with the tubes ordinarily used. The maximum current also is limited by the regulator tube; 100 milliamperes is a safe value for the 2A3. Two or more regulator tubes may be connected in parallel to increase the current-carrying capacity, with no change in the circuit.

8-9 Bias Supplies

Requirements — A bias supply is not called upon to deliver current to a load circuit, but simply to furnish a fixed grid voltage to set the operating point of a tube (§ 3-3). However, in most applications it is nevertheless true that current flows through the bias supply, because such supplies are used chiefly in connection with power amplifiers of the Class-B and Class-C type, where grid-current flow is a feature of operation (§ 3-4). In circuit design a bias supply resembles the rectified-a.c. plate supply (§ 8-1), having a transformer-rectifier-filter system employing similar circuits. Bias supplies may be classified in two types, those furnishing only *protective* bias, intended to prevent excessive plate current flow in a power tube in case of loss of grid leak bias (§ 3-6) from excitation failure, and those which furnish the actual *operating* bias for the tubes. In the former type, voltage regulation (§ 8-1) is relatively unimportant; in the latter it may be of considerable importance.

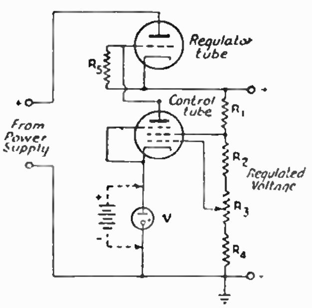


Fig. 806 — Electronic voltage regulator. The regulator tube is ordinarily a 2A3 or a number of them in parallel, the control tube a 6SJ7 or similar type. The filament transformer for the regulator tube must be insulated for the plate voltage, and cannot supply current to other tubes when a filament-type regulator tube is used. Typical values: R_1, 10,000 ohms; R_2, 25,000 ohms; R_3, 10,000-ohm potentiometer; R_4, 5000 ohms; R_5, 0.5 megohm.

In general, a bias supply should have well-filtered d.c. output, especially if it furnishes the operating bias for the stage, since ripple voltage may modulate the signal on the grid of the amplifier tube (§ 5-1). Condenser-input filters are generally used, since the regulation of the supply is not a function of the filter. The constants given in § 8-5 are applicable.

Voltage regulation — A bias supply must always have a bleeder resistance (§ 8-4) connected across its output terminals, to provide a d.c. path from grid to cathode of the tube being biased. Although the grid circuit takes no current from the supply, grid current flows through the bleeder resistor and the voltage across the resistor therefore varies with grid current. This variation in voltage is practically independent of the bias-supply design unless special voltage-regulating means are used.

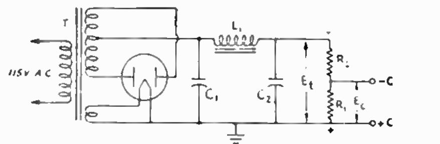


Fig. 807 — Supply for furnishing protective bias to a power amplifier. The transformer, *T*, should furnish peak voltage at least equal to the protective bias required.

Protective bias — This type of bias supply is designed to give an output voltage sufficient to bias the tube to which it is applied at or near the plate-current cut-off point (§ 3-2). A typical circuit is given in Fig. 807. The resistance, R_1, is the grid-leak resistor (§ 3-6) for the amplifier tube with which the supply is used, and the normal operating bias is developed by the flow of grid current through this resistor. R_2 is connected in series with R_1 across the output of the supply, to reduce the voltage across R_1, when there is no grid-current flow, to the cut-off value for the tube being biased. The value of R_2 is given by the formula:

$$R_2 = \frac{E_t - E_c}{E_c} \times R_1$$

where E_t is the output voltage of the supply with R_2 and R_1 in series as a load, E_c is the cut-off bias, and R_1 is as described above.

When such a supply is used with a Class-C amplifier, the voltage across R_1 from grid-current flow will normally be higher than that from the bias supply itself, since the latter is adjusted to cut-off while the operating bias will be twice cut-off or higher (§ 3-4). In some cases the grid-leak voltage may even exceed the peak output voltage of the transformer (1.41 times half the total secondary voltage, in the circuit shown). The filter condensers in such a bias supply must, therefore, be rated to stand the maximum operating bias voltage on the Class-C amplifier, if this voltage exceeds the nominal output voltage of the supply.

Voltage stabilization — When the bias supply furnishes operating rather than simply protective bias, the value of bias voltage should be as constant as possible even when the grid current of the biased tube varies. A simple method of improving bias voltage regulation is to make the bleeder resistance low enough so that the current through it from the supply is several times the maximum grid current to be expected. By this means, the percentage variation in current is reduced. This method requires, however, that a considerable amount of power be dissipated in the bleeder, which in turn calls for a relatively large power transformer and filter choke.

Bias-voltage variation may also be reduced by means of a regulator tube, as shown in Fig. 808. The regulator tube usually is a triode having a plate-current rating adequate to carry the expected grid current. It is cathode-biased

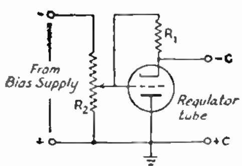


Fig. 808 — Automatic voltage regulator for bias supplies. For best operation the tube used should be one having high mutual conductance (§ 3-2).

(§ 3-6) by the resistor, R_1, which is of the order of several hundred thousand ohms or a few megohms, so that with no grid current the tube is biased practically to cut-off. Because of this high resistance, the grid current will flow through the plate resistance of the regulator tube, which is comparatively low, rather than through R_1 and R_2; hence the voltage from the supply, across R_1 and the cathode-plate circuit of the regulator tube in series, can be considered constant. The bias voltage is equal to the voltage across the tube alone. When grid current flows, the voltage across the tube will tend to increase; hence the drop across R_1 decreases, lowering the bias on the regulator and reducing its plate resistance. This, in turn, reduces the tube voltage drop, and the bias voltage tends to remain constant over a fairly wide range of grid current values.

At low bias voltages it may be necessary to use a number of tubes in parallel to get sufficient variation of plate resistance for good regulating action. The bias supply must furnish the required bias voltage plus the voltage required to bias the regulator tube to cut-off, considering the output bias voltage as the plate voltage applied to the regulator. The current taken from the bias supply is negligible. R_2 may be tapped to provide a range of bias voltages to meet different tube requirements.

Multistage bias supplies — Where several power amplifier tubes are to be biased from a single supply, the various bias circuits must be isolated by some means. If the grid currents of all stages should flow through a single bleeder resistor, a variation in grid current in one stage would change the bias on all, a condition which would interfere with effective adjustment and operation of the transmitter.

When protective bias is to be furnished several stages, the circuit arrangement of Fig.

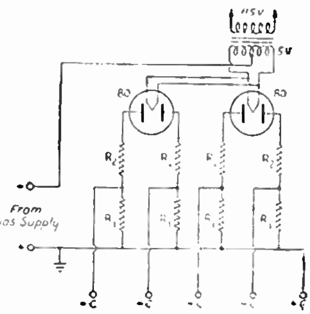


Fig. 809 — Isolating circuit for multiple bias supply.

809, using rectifier tubes to isolate the individual grid-leaks of the various stages, may be employed. In the diagram, two type 80 rectifiers are used to furnish bias to four stages. Each pair of resistors (R_1R_2) constitutes a separate bleeder across the bias supply. R_1 is the grid-leak for the biased stage; R_2 is a dropping resistor to adjust the voltage across R_1 to the cut-off value (without grid-current flow) for the biased tube. The values of R_1 and R_2 may be calculated as described in the paragraph on protective bias. In this case, the bias supply should be designed to have inherently good voltage regulation; i.e., a choke-input filter with appropriate filter and bleeder constants (§ 8-6) should be used, the bleeder being separate from those associated with the rectifier tubes. When the voltage across R_1R_2 rises because of grid-current flow through R_1, the load on the supply will vary (hence the necessity for good voltage regulation in the supply), but there is no interaction of grid currents in the separate bleeders because the rectifiers can pass current only in one direction.

When a single supply is to furnish operating bias for several stages, a separate regulator-tube circuit (Fig. 808) may be used for each one. Individual voltages for the various stages can be obtained by appropriate taps on R_2.

Well-regulated bias for several stages may be obtained by the use of gaseous regulator tubes, when the voltage and current ratings of the tubes permit their use. This is shown in Fig. 810. A single tube or two or more in series can be used to give the desired bias-voltage drop; the bias supply voltage must be high enough to provide starting voltage for the tubes in series. R_1 is the protective resistance (§ 8-8); its value should be calculated for minimum stable tube current. The maximum grid current that can be handled is 20 to 25 milliamperes with available regulator tubes.

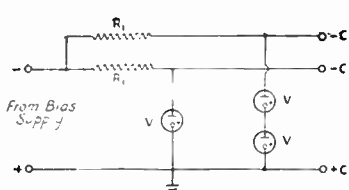


Fig. 810 — Use of VR tubes to stabilize bias voltage.

8-10 Miscellaneous Power-Supply Circuits

Voltage dividers — A voltage divider is a resistor connected across a source of voltage and tapped at appropriate points (§ 2-6). Since the voltage at any tap depends upon the current drawn from the tap, the voltage regulation (§ 8-1) is inherently poor. Hence, a voltage divider is best suited to applications where the currents drawn are constant, or where separate voltage-regulating circuits (§ 8-8) are used to compensate for voltage variations at the taps.

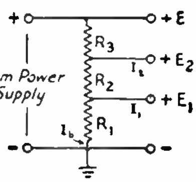
A typical voltage-divider arrangement is shown in Fig. 811. The terminal voltage is E, and two taps are provided to give lower voltages, E_1 and E_2, at currents I_1 and I_2 respectively. The smaller the resistance between taps in proportion to the total resistance, the smaller the voltage between the taps. For convenience, the voltage divider in the figure is considered to be made up of separate resistances, R_1, R_2, R_3, between taps. R_1 carries only the bleeder current, I_b. R_2 carries I_1 in addition to I_b; R_3 carries I_2, I_1 and I_b. To calculate the resistances required, a bleeder current,

Fig. 811 — Typical voltage-divider circuit.

$$R_1 = \frac{E_1}{I_b}$$

$$R_2 = \frac{E_2 - E_1}{I_b + I_1}$$

$$R_3 = \frac{E - E_2}{I_b + I_1 + I_2}$$



I_b, must be assumed; generally it is low compared to the total load current (10 per cent or so). Then the required values can be calculated as shown below, I being in amperes.

The method may be extended to any desired number of taps, each resistance section being calculated by Ohm's Law (§ 2-6) using the voltage drop across it and the total current through it. The power dissipated by each section may be calculated by multiplying I and E.

Transformerless plate supplies — The line voltage is rectified directly, without a step-up power transformer, for certain applications (such as some types of receivers) where the low voltage so obtained is satisfactory. A simple power supply of this variety, often called the "a.c.-d.c." type, is shown in Fig. 812. Rectifier tubes for this purpose have heaters operating at relatively high voltages (12.6, 25, 35, 45, 50, 70 or 115 volts), which can be connected across the a.c. line in series with other tube filaments and/or a resistor, R, of suitable value to limit the current to the rated value for the tubes.

The half-wave circuit shown has a fundamental ripple frequency equal to the line frequency (§ 8-4) and hence requires more inductance and capacity in the filter for a given ripple percentage (§ 8-5) than the full-wave rectifier. A condenser-input filter generally is used. The input condenser should be at least 16 μfd. and preferably 32 or 40 μfd., to keep the output voltage high and to improve voltage regulation. Frequently a second filter section (§ 8-5) is sufficient to provide smoothing.

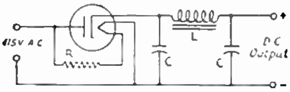


Fig. 812 — Transformerless plate supply with half-wave rectifier. Other filaments are connected in series with R.

No ground connection can be used on the power supply unless the grounded side of the power line is connected to the grounded side of the supply. Receivers using an a.c.-d.c. supply usually are grounded through a low capacity (0.05 μfd.) condenser, to avoid short-circuiting the line should the line plug be inserted in the socket the wrong way.

Voltage multiplier circuits — Transformerless voltage multiplier circuits make it possible to obtain d.c. voltages higher than the line voltage without using step-up transformers. By alternately charging two or more condensers to the peak line voltage and allowing them to discharge in series, the total output voltage becomes the sum of the voltages appearing across the individual condensers. The required switching operation is performed automatically by diode rectifier tubes associated with the condensers.

A half-wave voltage doubler is shown in Fig. 813-A. In this circuit when the plate of the lower diode is positive the tube passes current, charging C_1 to a voltage equal to the peak line voltage less the tube drop. When the line polarity reverses at the end of the half cycle the voltage resulting from the charge in C_1 is added to the line voltage, the upper diode meanwhile similarly charging C_2. C_2, however, does not receive its full charge because it be-

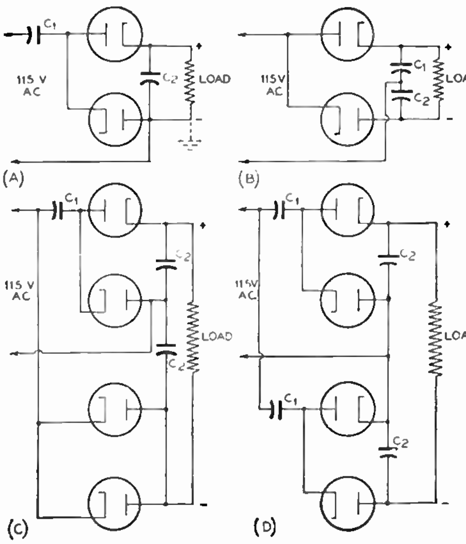


Fig. 813 — Voltage multiplier circuits. A, half-wave voltage doubler. B, full-wave doubler. C, tripler. D, quadrupler. Dual diode rectifier tubes may be used.

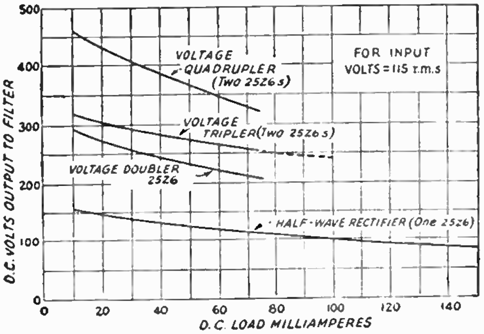


Fig. 814 — Curves showing the d.c. output voltage and the regulation under load for voltage-multiplier circuits.

gins discharging into the load resistance as soon as the upper diode becomes conductive. For this reason, the output is somewhat less than twice the line peak voltage. As with any half-wave rectifier, the ripple frequency corresponds to the line frequency.

The full-wave voltage doubler at B is more popular than the half-wave type. One diode charges C_1 when the polarity between its plate and cathode is positive while the other section charges C_2 when the line polarity reverses. Thus each condenser is charged separately to the same d.c. voltage, and the two discharge in series into the load circuit. The ripple frequency with the full-wave doubler is twice the line frequency (§ 8-1). The voltage regulation is inherently poor and depends critically upon the capacities of C_1 and C_2, being better as these capacities are made larger. A typical supply with 16 μfd. at C_1 and C_2 will have an output voltage of approximately 300 at light loads, as shown in Fig. 814.

The voltage tripler in Fig. 813-C comprises four diodes in a full-wave doubler and half-wave rectifier combination. The ripple frequency is that of the line as in a half-wave circuit, because of the unbalanced arrangement, but the output voltage of the combination is very nearly three times the line voltage, and the regulation is better than in other voltage multiplier arrangements, as shown in Fig. 814.

Fig. 813-D is a voltage quadrupler with two half-wave doublers connected in series, discharging the sum of the accumulated voltages in the associated condensers into the filter input. The quadrupler is by no means the ultimate limit in voltage multiplication. Practical power supplies have been built using up to twelve doubler stages in series.

In the circuits of Fig. 813, C_2 should have a working voltage rating of 350 volts and C_1 of 250 volts for a 115-volt line. Their capacities should be at least 16 μfd. each. Subsequent filter condensers must, however, withstand the *peak* total output voltage — 450 volts in the case of the tripler and 600 for the quadrupler.

No direct ground can be used on any of these supplies or on associated equipment. If an r.f. ground is made through a condenser the capacity should be small (0.05 μfd.), since it is in shunt from plate to cathode of one rectifier.

Duplex plate supplies — In some cases it may be advantageous economically to obtain two plate-supply voltages from a single power supply, making one or more of the components serve a double purpose. Circuits of this type are shown in Figs. 815 and 816.

In Fig. 815, a bridge rectifier is used to obtain the full transformer voltage, while a connection is also brought out from the center-tap to obtain a second voltage corresponding to half the total transformer secondary voltage. The sum of the currents drawn from the two taps should not exceed the d.c. ratings of the rectifier tubes and transformer. Filter values for each tap are computed separately (§ 8-6).

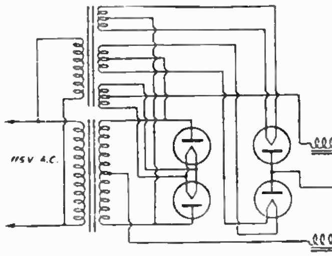


Fig. 815 — Duplex plate supply, delivering two output voltages, both with good regulation.

Fig. 816 shows how a transformer with multiple secondary taps may be used to obtain both high and low voltages simultaneously. A separate full-wave rectifier is used at each tap. The filter chokes are placed in the common negative lead, but separate filter condensers are required. The sum of the currents drawn from each tap must not exceed the transformer rating, and the chokes must be rated to carry the total load current. Each bleeder resistance should have a value in ohms 1000 times the maximum rated inductance in henrys of the swinging choke, L_1, for best regulation (§ 8-6).

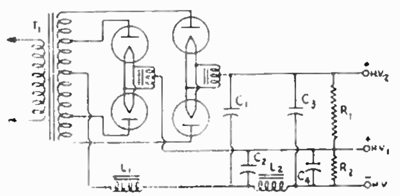


Fig. 816 — Power supply in which a single transformer and set of chokes serve for two different output voltages.

Rectifiers in parallel — Vacuum-type rectifiers may be connected in parallel (plate to plate and cathode to cathode) for higher current-carrying capacity with no circuit changes.

When mercury-vapor rectifiers are connected in parallel, slight differences in tube characteristics may make one ionize at a slightly lower voltage than the other. Since the ignition voltage is higher than the operating voltage the first tube to ionize carries the whole load, as the voltage drop is then too low to ignite the second tube. This can be prevented by connecting 50- to 100-ohm resistors in series with each plate, thereby insuring that a high-enough voltage for ignition will be available.

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 (§ 2-5). 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 (§ 8-5). The smoothing filter can be a single-section affair, but the filter output capacity should be fairly large — 16 to 32 μfd.

Fig. 817 shows the two types of circuits. At A is shown the *nonsynchronous* type of vibrator. When the battery is disconnected the reed is midway between the two contacts, touching neither. On closing the battery circuit the magnet coil pulls the reed into contact with one contact point, causing current to flow through the lower half of the transformer primary winding. Simultaneously, the magnet coil is short-circuited, deënergizing it, and the reed swings back. Inertia carries the reed into contact with the upper point, causing current to flow through the upper half of the transformer primary. The magnet coil again is energized, and the cycle repeats itself.

The synchronous circuit of Fig. 817-B 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 which occur on breaking the current, when the magnetic field collapses practically instantaneously and hence causes very high voltages to be induced in the secondary (§ 2-5). 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 capacity 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 capacity is changed. When the system is operating properly there should be practically no sparking at the vibrator contacts. A 5000-ohm resistor in series with C_2 will limit the secondary current to a safe value should the condenser fail.

A more exact check on the operation can be secured with an oscilloscope having a linear sweep circuit which can be synchronized with the vibrator. The vertical plates should be connected across the outside ends of the transformer primary winding to show the input voltage waveshape. Fig. 818-C shows an idealized trace of the optimum waveform when the buffer capacity is adjusted to give proper operation throughout the life of the vibrator. The horizontal lines in the trace represent the voltage during the time the vibrator contacts are closed, which should be approximately 90 per cent of the total time. When the contacts are open the trace should be partly tilted and partly vertical, the tilted part being 60 per cent of the total connecting trace. The oscilloscope will show readily the effect of the buffer capacity on the percentage of tilt. In actual patterns the horizontal sections are likely to droop somewhat because of the resistance drop in the battery leads as the current builds up through the primary inductance (Fig. 818-D).

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. C_1 is usually from 0.5 to 1 μfd., a 50-volt rating being adequate. RFC_1 consists of about 50 turns of No. 12 or No. 14 wound to about half-inch diameter, large wire being required to carry the rather heavy battery current without undue loss of voltage. A choke of these specifications should

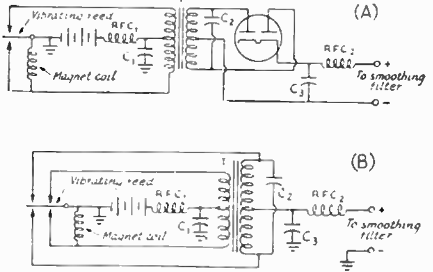


Fig. 817 — Basic types of vibrator power-supply circuits.

be adequate, but if there is persistent trouble with hash it may be beneficial to experiment with other sizes. Bank-wound chokes are more compact and give higher inductance for a given resistance. In the secondary filter, C_3 may be of the order of 0.01 to 0.1 μfd., and RFC_2 a 2.5-millihenry r.f. choke of ordinary design.

A 100-μμfd. mica condenser, connected from the positive output lead to the "hot" side

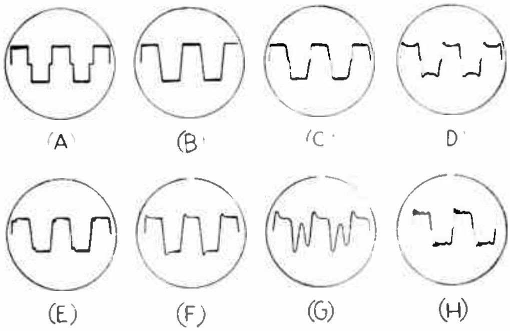


Fig. 818 — Characteristic vibrator waveforms as viewed on the oscilloscope. A, ideal theoretical trace for resistive load; current flow stops instantly when vibrator contacts open and resumes approximately 1 microsecond later (for standard 115-cycle vibration frequency) after interrupter arm moves across for the next half-cycle. B, ideal practical waveform for inductive load (transformer primary) with correct buffer capacity. C, practical approximation of B for loaded nonsynchronous vibrator. D, satisfactory practical trace for synchronous (self-rectifying) vibrator under load; the peaks result from voltage drop in the primary when the secondary load is connected, not from faulty operation.

Faulty operation is indicated in traces E through H: E, effect of insufficient buffering capacity (not to be mistaken for "bouncing" of contacts). The opposite condition — excessive buffering capacity — is indicated by slow build-up with rounded corners, especially on "open." F, overclosure caused by too-small buffer condenser (same condition as in E) with vibrator unloaded. G, "skipping" of worn-out or misadjusted vibrator, with interrupter making poor contact on one side. H, "bouncing" resulting from worn-out contacts or sluggish reed. G and H usually call for replacement of the vibrator.

of the "A" battery, may be helpful in reducing hash in certain power supplies. A trial is necessary to see whether or not it is required. It should be mounted right at the output socket.

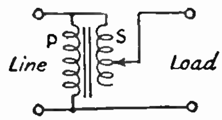
Equally as important as the hash filter is thorough shielding of the power supply and its connecting leads, since even a small piece of wire or metal will radiate enough r.f. to cause interference in a sensitive receiver.

Testing in connection with hash elimination should be carried out with the supply operating a receiver. Since the interference usually is picked up on the receiving antenna leads by radiation from the supply itself and from the battery leads, it is advisable to keep the supply and battery as far from the receiver as the connecting cables will permit. Three or four feet should be ample. The microphone cord likewise should be kept away from the supply and leads.

The power supply should be built on a metal chassis, with all unshielded parts underneath. A bottom plate to complete the shielding is advisable. The transformer case, vibrator cover and the metal shell of the tube all should be grounded to the chassis. If a glass tube is used it should be enclosed in a tube shield. The battery leads should be evenly twisted, since these leads are more likely to radiate hash than any other part of a well-shielded supply. Experimenting with different values in the hash filters should come *after* radiation from the battery leads has been reduced to a minimum. Shielding the leads is not particularly helpful.

Line-voltage adjustment — In some localities the line voltage may vary considerably from the nominal 115 volts as the load on the power system changes. Since it is desirable to operate tube equipment, particularly filaments and heaters, at constant voltage for maximum life, a means of adjusting the line voltage to the rated value is desirable. This can be accomplished by the circuit shown in Fig. 819, utilizing a step-down transformer with a tapped secondary connected as an autotransformer (§ 2-9). The secondary preferably should be tapped in steps of two or three volts, and should have sufficient total voltage to compensate for the widest variations encountered. Depending upon the end of the secondary to which the line is connected, the voltage to the load can be made either higher or lower than the line voltage. A secondary winding capable of carrying five amperes will serve for loads up to 500 volt-amperes on a 115-volt line.

Fig. 819 — Line-voltage compensation by a tapped step-down autotransformer.



ℂ 8-11 — Emergency Power Supply

Dry batteries — Dry-cell batteries are ideal for emergency receiver and low-power transmitter supplies because they provide steady, pure, direct current. Their disadvantages are weight, high cost, and limited current capability. In addition, they will lose their power even when not in use if allowed to stand idle for periods of a year or more. This makes them uneconomical if not used more or less continuously.

Table I in Chapter Eighteen gives service life of representative types of batteries for various current drains, based on intermittent service simulating typical operation. The continuous-service life will be somewhat greater at very low current drains and from one-half to two-thirds the intermittent life at higher drains.

The secret of long battery life at normal current drains lies in intermittent operation. The duration of "on" periods should be reduced to a minimum. The more frequent the rests given a dry-cell battery, the longer it will last. As an example, one standard type will last 50 per cent longer if it is operated for periods of one minute, with five-minute rest intervals, in 24-hour intermittent operation than if it is operated continuously for four hours per day, although the actual energy consumption in the 24-hour period is the same in both cases.

Storage batteries — The most universally acceptable self-contained power source is the storage battery. It has high initial capacity and can be recharged, so that its effective life is practically indefinite. It can be used to provide filament or heater power directly, and plate power through associated devices such as vibrator-transformers, dynamotors and genemotors, and a.c. converters. For emergency work a storage battery is a particularly convenient power source, since such batteries are universally available. In a serious emergency it is possible to obtain 6-volt storage batteries so long as there are automobiles to borrow them from, and for this reason the 6-volt storage battery makes an excellent unit around which to design a low-powered emergency station.

For maximum efficiency and usefulness the power drain on the storage battery should not exceed 15 or 20 amperes from the ordinary 100- or 120-ampere-hour 6-volt battery. Heavy connecting leads should be used to minimize the voltage drop; similarly, heavy-duty low-resistance switches are required.

Vibrator power supplies — For portable or mobile work, the most common source of power for both filaments and plates is the 6-volt automobile-type storage battery. Filaments may be heated directly from the battery, while plate power is obtained by passing current from the battery through the primary of a suitable transformer, interrupting it at regular intervals and rectifying the secondary output (§ 2-5) providing outputs as high as 400 volts at 200 ma. The high-voltage filter circuit usually is identical with that of an equivalent power source operating from the a.c. line (§ 8-5). Noise suppression filters, serving to minimize r.f. interference caused by the vibrator, are incorporated in manufactured units.

Although vibrator supplies are ordinarily used with 6-volt tubes, their use with 2-volt tubes is quite possible provided additional filament filtration is incorporated. This filter may consist of a small low-resistance iron-core filter choke or the voice-coil winding of a speaker transformer. The field coil of a loudspeaker designed to operate on 4 volts at the total filament current of the receiver may be used. The filaments are then connected in parallel, as usual, and placed in series with this winding across the 6-volt battery. In both 6- and 2-volt receivers, "hash" can be reduced by heavily by-passing the battery at the vibrator supply terminals, using fixed condensers of 0.25 to 1 μfd. capacity or more, and by including an r.f. choke of heavy wire in the battery lead near the condenser. Noise will be minimized if a single ground, consisting of a short, heavy copper strap, is used. Thorough shielding of the vibrator also will contribute to the noise reduction.

Table II in Chapter Eighteen lists standard commercial vibrator supplies suitable for use as emergency or portable power sources. Those units which include a hum filter are indicated. The vibrator supplies used with automobile receivers are satisfactory for receiver applications and for use with transmitters where the power requirements are small.

The efficiency of vibrator packs runs between about 60 to 75 per cent.

Dynamotors and genemotors — A dynamotor is a double-armature high-voltage generator, the additional winding serving as a driving motor. Dynamotors usually are operated from 6-, 12- or 32-volt storage batteries, and deliver from 300 to 1000 volts or more.

The genemotor is a refinement of the dynamotor, designed especially for automobile receiver, sound truck and similar applications. It has good regulation and efficiency, combined with economy of operation. Standard models of genemotors have ratings ranging from 135 volts at 30 ma. to 300 volts at 200 ma. or 500 volts at 200 ma. (See Table III in Chapter Eighteen.) 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.

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 by-passed 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.

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 through a "brute force" smoothing filter using 4- to 8-μfd. electrolytic condensers with a 15- or 30-henry choke having low d.c. resistance.

A.c.-d.c. converters — In some instances it is desirable to utilize existing equipment built for 115-volt a.c. operation. To operate such equipment with any of the power sources outlined above would require a considerable amount of rebuilding. This can be obviated by using a rotary converter capable of changing the d.c. from 6-, 12- or 32-volt batteries to 110-volt 60-cycle a.c. Such converter units are built to deliver output ranging from 40 to 300 watts.

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 as well as plate power.

Chapter Nine

Wave Propagation

9-1 Characteristics of Radio Waves

Relation to other forms of radiation — Radio waves differ from other forms of electromagnetic radiation principally in the order of their wavelength, which ranges from approximately 30,000 meters to a small fraction of a centimeter; i.e., their frequency ranges between about 10 kc. and 1,000,000 Mc. They travel at the same velocity as light waves (about 300,000,000 meters per second in free space) and can be similarly reflected, refracted and diffracted.

The total energy in a radio wave is evenly divided between traveling electrostatic and electromagnetic fields. The lines of force of these fields are at right angles to each other in a plane perpendicular to the direction of travel, as shown in Fig. 901.

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

Reflection — Radio waves may be reflected from any sharply defined discontinuity of suitable characteristics and dimensions encountered in the medium in which they are traveling. Any conductor (or any insulator having a dielectric constant differing from that of the medium) offers such a discontinuity if its dimensions are at least comparable to the wavelength. The surface of the earth and the boundaries between ionospheric layers are examples of such discontinuities. Objects as small as an airplane, a tree or even a man's body will readily reflect the shorter waves.

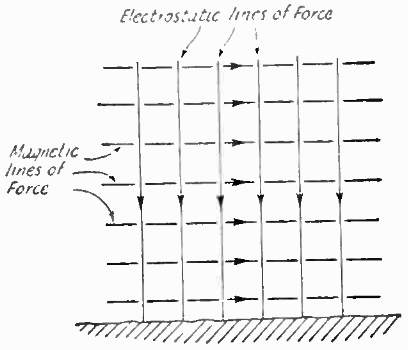


Fig. 901 — Representation of electrostatic and electromagnetic lines of force in a radio wave. Arrows indicate instantaneous directions of the fields for a wave traveling toward the reader. Reversing the direction of one set of lines would reverse the direction of travel.

Refraction — As in the case of light, a radio wave is bent when it moves obliquely into any medium having a different refractive index from that of the medium which it leaves. Since the velocity of propagation or travel differs in the two mediums, that part of the wave front which enters first travels faster or slower than the part which enters last, and so the wave front is turned or refracted (usually downward in the vertical plane). Refraction may take place in either the ionosphere (ionized upper atmosphere) or the troposphere (lower atmosphere).

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

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

The ionospheric wave (sometimes called the "sky wave,") is that part of the total radiation which is directed toward the ionosphere. Depending upon variable conditions in that region, as well as upon wavelength (or frequency), 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 which 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 which is directly affected by the presence of the earth and its surface features. The ground wave has two components. One is the *surface wave*, which is an earth-guided wave, and the other is the *space wave* (not to be confused with the ionospheric or "sky wave"). The space wave is itself the resultant of two components — the *direct wave* and the *ground-reflected wave*, as shown in Fig. 902.

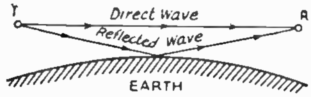


Fig. 902 — Showing how both direct and reflected waves may be received simultaneously in v.h.f. transmission.

9-2 Ionospheric Propagation

The ionosphere — Communication between distant points by means of radio waves of frequencies ranging between 3 and 30 Mc. depends principally upon the ionospheric wave. Upon leaving the transmitting antenna, this wave travels upward from the earth's surface at such an angle that it would continue out into space were its path not bent sufficiently to bring it back to earth. The medium which causes such bending is the ionosphere, a region in the upper atmosphere, above a height of about 60 miles, where free ions and electrons exist in sufficient quantity to cause a change in the refractive index. This condition is believed to be the effect of ultraviolet radiation from the sun. The ionosphere is not a single region but is composed of a series of layers of varying densities of ionization occurring at different heights. Each layer consists of a central region of relatively dense ionization which tapers off in intensity both above and below.

Refraction, absorption and reflection — For a given density of ionization, the degree of refraction becomes less as the wavelength becomes shorter (or as the frequency increases). The bending therefore is less at high than at low frequencies, and if the frequency is raised to a sufficiently high value, a point is finally reached where the refractive bending becomes too slight to bring the wave back to earth, even though it may enter the ionized layer along a path which makes a very small angle with the boundary of the ionosphere.

The greater the density of ionization, the greater the bending at any given frequency. Thus, with an increase in ionization, the minimum wavelength which can be bent sufficiently for long-distance communication is lessened and the maximum usable frequency is increased.

The wave necessarily loses some of its energy in traveling through the ionosphere, this absorption loss increasing with wavelength and also with ionization density. Unusually high ionization, especially in the lower strata of the ionosphere, may cause complete absorption of the wave energy.

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

Critical frequency — When the frequency is sufficiently low, a wave sent vertically upward to the ionosphere will be bent sharply enough to cause it to return to the transmitting point. The highest frequency at which such reflection can occur, for a given state of the ionosphere, is called the *critical frequency*. Although the critical frequency may serve as an index of transmission conditions, it is not the highest useful frequency, since other waves of the same frequency which enter the ionosphere at angles smaller than 90 degrees (less than vertical) will be bent sufficiently to return to earth. The *maximum usable frequency*, for waves leaving the earth at very small angles to the horizontal, is in the vicinity of three times the critical frequency.

Besides being directly observable, the critical frequency is of more practical interest than the ionization density because it includes the effects of absorption as well as refraction.

Virtual height — Although an ionospheric layer is a region of considerable depth it is convenient to assign to it a definite height, called the *virtual height*. This is the height from which a simple reflection would give the same effect as the gradual refraction which actually takes place, as illustrated in Fig. 903. The wave traveling upward is bent back over a path having an appreciable radius of turning, and a measurable interval of time is consumed in the turning process. The virtual height is the height of a triangle formed as shown, having equal sides of a total length proportional to the time taken for the wave to travel from T to R.

Normal structure of the ionosphere — The lowest normally useful layer is called the E layer. The average height of the region of maximum ionization is about 70 miles. The ionization density is greatest around local noon; the layer is only weakly ionized at night, when it is not exposed to the sun's radiation. The air at this height is sufficiently dense so that free ions and electrons very quickly meet and recombine.

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

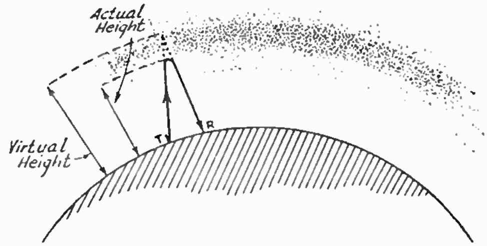


Fig. 903 — Showing bending in the ionosphere and the echo or reflection method of determining virtual height.

the F layer splits into two parts, the F_1 and F_2 layers, with average virtual heights of, respectively, 140 miles and 200 miles. These layers are most highly ionized at about local noon, and merge again at sunset into the F layer.

Cyclic variations in the ionosphere — Since ionization depends upon ultraviolet radiation, conditions in the ionosphere vary with changes in the sun's radiation. In addition to the daily variation, seasonal changes result in higher critical frequencies in the E layer in summer, averaging about 4 Mc. as against a winter average of 3 Mc. The F layer shows little variation, the critical frequency being of the order of 4 to 5 Mc. in the evening. The F_1 layer, which has a critical frequency near 5 Mc. in summer, usually disappears entirely in winter. The critical frequencies for the F_2 are highest in winter (11 to 12 Mc.) and lowest in summer (around 7 Mc.). The virtual height of the F_2 layer, which is about 185 miles in winter, averages 250 miles in summer.

Seasonal transition periods occur in spring and fall, when ionospheric conditions are found highly variable.

There are at least two other regular cycles in ionization. One such cyclic period covers 28 days, which corresponds with the period of the sun's rotation. For a short time in each 28-day cycle, transmission conditions reach a peak. Usually this peak is followed by a fairly rapid drop to a lower level, and then a slow building up to the next peak. The 28-day cycle is particularly evident in the 14- and 28-Mc. amateur bands.

The longest cycle yet observed covers about 11 years, corresponding to a similar cycle of sunspot activity. The effect of this cycle is to shift upward or downward the values of the critical frequencies for F- and F_2-layer transmission. The critical frequencies are highest during sunspot maxima and lowest during sunspot minima. It is during the period of minimum sunspot activity when long-distance transmissions occur on the lower frequencies. At such times the 28-Mc. band is seldom useful for DX work, while the 14-Mc. band performs well in the daytime but is not ordinarily useful at night. The most recent sunspot maximum is considered to have occurred in 1938.

Magnetic storms and other disturbances — Unusual disturbances in the earth's magnetic field (magnetic storms) usually are accompanied by disturbances in the ionosphere, when the layers apparently break up and expand. There is usually also an increase in absorption during such a period. Radio transmission is poor and there is a drop in critical frequencies so that lower frequencies must be used for communication. A storm may last for several days.

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

Magnetic storms frequently are accompanied by unusual auroral displays, creating an ionized "curtain" in the polar regions which can act as a reflector of radio waves. Auroral reflection is occasionally observed at frequencies as high as 60 Mc.

Sporadic E-layer ionization — Occasionally scattered patches or clouds of relatively dense ionization appear at heights approximately the same as that of the E layer. The effect is to raise the critical frequency to a value perhaps twice that which is returned from any of the regular layers by normal refraction. Distances of about 500 to 1250 miles may be covered at 56 Mc. if the ionized cloud is situated midway between transmitter and receiver, or is of any very considerable extent. This effect, while infrequently observed in winter, is prevalent during the late spring and early summer, with no apparent correlation of the condition with the time of day.

The presence of sporadic-E refraction on the 14- and 28-Mc. bands is indicated by an abnormally short distance between the transmitter and the point where the wave first is returned to earth as when, for example, 14-Mc. signals from a transmitter only 100 miles distant may arrive with an intensity usually associated with distances of this order on 7 and 3.5 Mc.

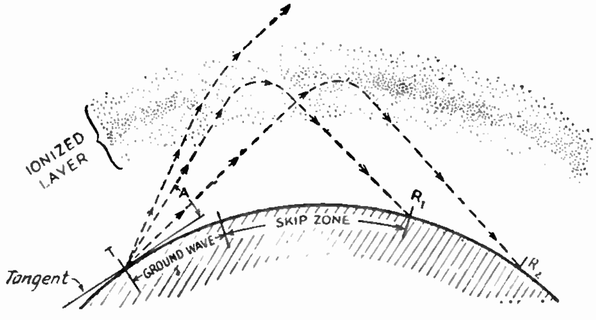


Fig. 904 — Refraction of sky waves, showing the critical wave angle and the skip zone. Waves leaving the transmitter at angles above the critical (greater than A) are not bent enough to be returned to earth. As the angle is increased, the waves return to earth at increasingly greater distances.

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. 901. The vertical angle which the wave makes with a tangent to the earth is called the *wave angle* or *angle of radiation*.

Skip distance — Since greater bending is required to return the wave to earth when the wave angle is high, at the higher frequencies the refraction frequently is not enough to give the required bending unless the wave angle is smaller than a certain angle called the *critical angle*. This is illustrated in Fig. 901, where waves at angles of A or less give useful signals while waves sent at higher angles penetrate the layer and are not returned. The distance between T and R_1 is, therefore, the shortest possible distance over which communication by normal ionospheric refraction can be accomplished.

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

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

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

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

It frequently occurs that transmission conditions are different for waves of slightly different frequencies, so that in the case of voice-modulated transmission, involving side-bands differing slightly from the carrier in frequency, the carrier and various side-band 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.

ℂ 9-3 Tropospheric Propagation

Air masses and fronts — In the lower atmosphere wave propagation is affected by the changes in refractive index between differing air masses. A mass of air hundreds of miles in area may remain at rest over one region until it becomes affected by the surface temperature and humidity characteristic of that region. Eventually being moved on by the forces of atmospheric circulation, the mass may travel over regions quite different from its origin and retain for some time its original characteristics. When it meets a dissimilar air mass, the lighter, warmer and drier mass overruns the heavier, cold, moist mass creating a boundary between the two called a *front*. This front, which represents a discontinuity in the dielectric constant of the troposphere, serves to refract and reflect the higher-frequency radio waves in much the same manner as the ionospheric layers, but at lesser heights and more restricted angles. As a result frequencies above 50 Mc. are returned to earth at distances considerably beyond the range of ground-wave propagation, sometimes up to 400 miles.

Temperature inversions — The temperature of the lower atmosphere normally decreases at a constant rate with increasing height. When for any reason the normal variation or *lapse rate* of approximately 3° F. per 1000 feet of elevation is altered, a *temperature inversion* is said to take place. The resulting change in the dielectric constants of the air masses affected causes reflection and refraction similar to that in the ionosphere.

Types of inversion other than the *dynamic* type described in the preceding paragraph include the *subsidence* inversion, caused by the sinking of an air mass which has been heated by compression; the *nocturnal* inversion, brought about by the rapid cooling of surface air after sunset; and the *cloud-layer* inversion, caused by the heating of air above a cloud layer by reflection of the sun's rays from the upper surface of the clouds. Refraction and reflection of v.h.f. waves are brought about also, although to a lesser degree, by the presence of sharp transitions in the water-vapor content of the atmosphere. Fig. 905 illustrates the conditions existing when the air is "normal" and when a temperature inversion is present.

ℂ 9-4 Ground-Wave Propagation

Surface wave — The surface wave is continuously in contact with the surface of the earth and, in cases where the distance of transmission makes the curvature of the earth a factor, extends its range by diffraction. The surface wave is practically independent of seasonal and day and night effects at frequencies above 1500 kc.

The surface wave must be vertically polarized because the electrostatic field of a horizontally polarized wave would be short-circuited by the ground, which acts as a conductor at the frequencies for which the surface wave is of most interest.

The wave induces a current in the ground in traveling along its surface. If the ground were a perfect conductor there would be no loss of energy, but actual ground has appreciable resistance, so that the current flow causes some energy dissipation. This loss must be supplied by the wave which is correspondingly weakened. Hence, the transmitting range depends upon the ground characteristics. Because sea water is a good conductor, the range will be greater over the ocean than over land. The losses increase with frequency, so that the surface wave is rapidly attenuated at high frequencies and above about 2 Mc. is of little importance, except in purely local communication. The range at frequencies in the vicinity of 2 Mc. is of the order of 200 miles over average land and perhaps two or three times as far over sea water, for a medium-power transmitter (500 watts or so) using a good antenna. At higher frequencies the range drops off rapidly.

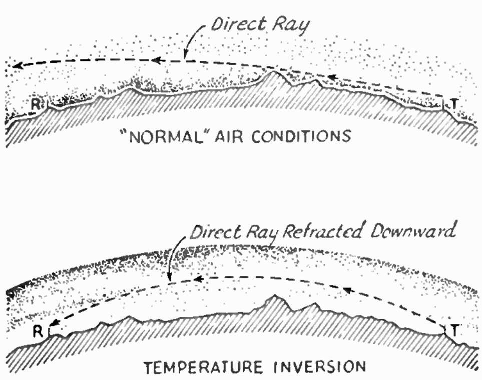


Fig. 905 — Illustrating the effect of a temperature inversion in extending the range of v.h.f. signals.

Space wave — In the v.h.f. portion of the spectrum (above 30 Mc.) the bending of the waves in the normal ionosphere is so slight that the ionospheric wave (§9-2) is not ordinarily useful for communication. The range of the surface wave also is extremely limited, as stated above. Hence, normal v.h.f. transmission is by means of the space wave in which the *direct-wave* component travels directly from the transmitter to the receiver through the atmosphere along a line-of-sight path.

Part of the space wave strikes the ground between the transmitter and receiver and is reflected upward at a slight angle, as was shown in Fig. 902. The effect of this *ground-reflected wave*, which is out of phase with the direct wave, is to reduce the net field strength at the receiving point. The degree of cancellation depends upon the heights of the transmitting and receiving antennas above the point of reflection, the ground losses when reflection takes place, and the frequency — the cancellation decreasing with an increase in any of these.

The energy lost in ground absorption by a wave traveling close to the ground decreases very rapidly with its height in terms of wave-lengths above the ground. A v.h.f. direct wave, therefore, can be relatively close (in physical height) to the ground without suffering the absorption effects which would occur at the same physical heights with longer wave-lengths.

Normal refraction — There is normally some change in the refractive index of the air with height above ground, its nature being such as to cause the wave to bend slightly towards the ground. Where curvature of the earth must be considered, this has the effect of lengthening the distance over which it is possible to transmit a direct wave. It is convenient to consider the effect of this "normal refraction" as equivalent to an increase in the earth's radius, in determining the antenna heights necessary to provide a clear path for the wave. The equivalent radius, taking refraction into account, is 4/3 the actual radius.

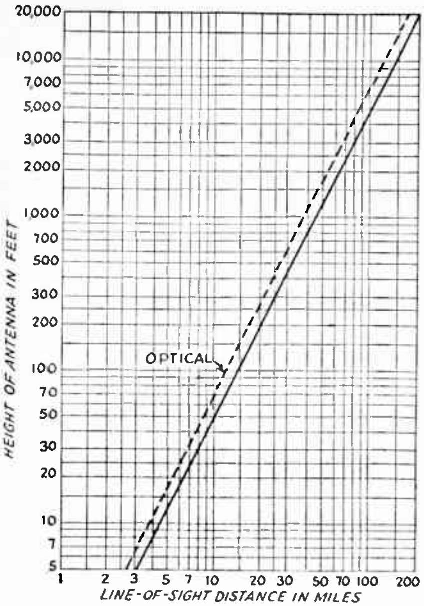


Fig. 906 — Chart for determining line-of-sight distance for v.h.f. transmission. The solid line includes effect of refraction, while the dotted line is the optical distance.

Range vs. height — Since the direct wave travels in practically a straight line, the maximum signal strength can be obtained only when there is an unobstructed atmospheric path between the transmitter and receiver. This means that antennas should be sufficiently elevated to provide such a path. On long paths the curvature of the earth, as well as the intervening terrain, must be taken into account.

The height required to provide a clear line-of-sight path over level terrain from an elevated transmitting point to a receiving point on the surface, not including the effect of refraction, is

$$h = \frac{d^2}{1.51}$$

where h is the height of the transmitting antenna in feet and d the distance in miles. Conversely, the line-of-sight distance in miles for a given height in feet is determined by

$$d = 1.23\sqrt{h}.$$

Taking refraction into account, this equation becomes

$$d = 1.41\sqrt{h}.$$

Fig. 906 gives the answer directly when either value is known.

When transmitter and receiver both are elevated, the maximum direct-wave distance to ground level can be determined separately for each. Adding the two distances thus obtained will give the maximum distance by which they can be separated for direct-wave communication. This is shown in Fig. 907.

9-5 Optimum Wave Angles

One of the requirements in high-frequency radio transmission is to send a wave to the ionosphere in such a way that it will have the best chance of being returned to earth. This is chiefly a matter of the angle at which the wave enters the layer, although in some cases polarization may be of importance. Furthermore, the desirable conditions may change considerably with frequency.

The desirable conditions for waves of different frequencies can be summarized as follows, in terms of the various amateur bands:

1.75 Mc. — Low-angle radiation is indicated for the longer distances. High-angle radiation may cause fading toward the limit of the ground-wave signal, because the downcoming waves add in random phase to the ground wave. Vertical polarization is to be preferred.

3.5 Mc. — Waves at all angles of radiation usually will be reflected, so that no energy is lost by high-angle radiation. However, the lower-angle waves will, in general, give the greatest distances. Polarization on this band is not of great importance.

7 Mc. — Under most conditions, angles of radiation up to about 45 degrees will be returned to earth; during the sunspot maximum still higher angles are useful. It is best to concentrate the radiation below 45 degrees. Polarization is not important, except that losses probably will be higher with vertical polarization.

14 Mc. — For long-distance transmission, most of the energy should be concentrated at angles below about 20 degrees. Higher angles are useful for comparatively short distances (300–400 miles), although 30 degrees is about the maximum useful angle. Aside from the probable higher losses with vertical polarization, the polarization may be of any type.

28 Mc. — Angles of 10 degrees or less are most useful. As in the case of 14 Mc., polarization is not important.

56 Mc. — The lowest possible angle of radiation is most useful for all types of transmission. Vertical polarization has been chiefly used for line-of-sight and lower atmosphere transmission, although horizontal polarization may be slightly better for long distances. In any event, the *same* polarization should be used at both transmitter and receiver.

Higher frequencies — As in the case of 56 Mc. either horizontal or vertical polarization may be used, so long as the same type is employed at both ends of the circuit.

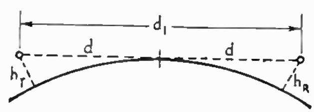


Fig. 907 — Method of determining total line-of-sight distance when both transmitter and receiver are elevated, based on Fig. 906. Since only earth curvature is taken into account in Fig. 906, irregularities in the ground between the transmitting and receiving points must be considered when computing each actual path.

Chapter Ten

Antenna Systems

ℂ 10-1 Antenna Properties

Wave propagation and antenna design — For most effective transmission, the propagation characteristics of the frequency under consideration must be given due consideration in selecting the type of antenna to use. These have been discussed in Chapter Nine. On some frequencies the angle of radiation and polarization may be of relatively little importance; on others they may be all-important. On a given frequency, the particular type of antenna best suited for long-distance transmission may not be as good for shorter-range work as would a different type.

The important properties of an antenna or antenna system are its polarization, angle of radiation, impedance, and directivity.

Polarization — The polarization of a straight-wire antenna is its position with respect to the earth. That is, a vertical wire transmits vertically polarized waves and a horizontal antenna generates horizontally polarized waves (§ 9-1). The wave from an antenna in a slanting position contains both vertical and horizontal components.

Angle of radiation — The wave angle (§ 9-4) at which an antenna radiates best is determined by its polarization, height above ground, and the nature of the ground. Radiation is not all at one well-defined angle, but rather is dispersed over a more or less large angular region, depending upon the type of antenna. The angle is measured in a vertical plane with respect to a tangent to the earth at the transmitting point.

Impedance — The impedance (§ 2-8) of the antenna at any point is the ratio of voltage to current at that point. It is important in connection with feeding power to the antenna, since it constitutes the load represented by the antenna.

Directivity — All antennas radiate more power in certain directions than in others. This characteristic, called *directivity*, must be considered in three dimensions, since directivity exists in the vertical plane as well as in the horizontal plane. Thus, the directivity of the antenna will affect the wave angle as well as the actual compass directions in which maximum transmission takes place.

Current — The field strength produced by an antenna is proportional to the current flowing in it. When there are standing waves on an antenna, the parts of the wire carrying the higher current have the greatest radiating effect.

Power gain — The ratio of power required to produce a given field strength, with a "comparison" antenna, to the power required to produce the same field strength with a specified type of antenna is called the *power gain* of the latter antenna. The field is measured in the optimum direction of the antenna under test. The comparison antenna almost always is 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 (§ 3-3).

ℂ 10-2 The Half-Wave Antenna

Physical and electrical length — 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 wave in space is:

$$\text{Length (feet)} = \frac{492}{\text{Freq. (Mc.)}} \qquad (1)$$

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. 1001, 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*). Under average conditions the following formula is sufficiently accurate for wire antennas at frequencies up to 30 Mc.:

$$\text{Length of half-wave antenna (feet)} = \frac{492 \times 0.95}{\text{Freq. (Mc.)}} = \frac{468}{\text{Freq. (Mc.)}} \qquad (2)$$

Above 30 Mc. the formulas below should be used, particularly for antennas constructed from rod or tubing. The factor K is taken from Fig. 1001.

$$\text{Length of half-wave antenna (feet)} = \frac{492 \times K}{\text{Freq. (Mc.)}} \qquad (3)$$

$$\text{or length (inches)} = \frac{5905 \times K}{\text{Freq. (Mc.)}} \qquad (4)$$

Current and voltage distribution — When power is fed to such an antenna the current and voltage vary along its length (§ 2-12-A). 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* (§ 2-12-A), because of the end effect; similarly, the voltage is not zero at its node because of the resistance of the antenna, which consists of both the r.f. resistance of the wire (*ohmic resistance*) and the radiation resistance (§ 2-12-A). Usually the ohmic resistance of a half-wave antenna is small enough, in comparison with the radiation resistance, to be neglected for all practical purposes.

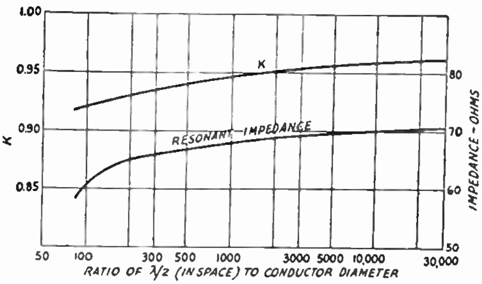


Fig. 1001 — Effect of antenna diameter on length for half-wave resonance, shown as a multiplying factor, K, to be applied to the free-space half wavelength (Equation 1). The effect of conductor diameter on the impedance measured at the center also is shown.

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, 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. 1001. If the diameter of the conductor is made large, the capacitance per unit length increases and the inductance per unit length decreases. Since the radiation resistance is affected relatively little, the decreased L/C ratio causes the Q of the antenna to decrease, so that the resonance curve becomes less sharp. Hence, the antenna is capable of working over a wide frequency range. This effect is greater as the diameter is increased, and is a property of some importance at the very-high frequencies where the wavelength is small.

Radiation characteristics — The radiation from a half-wave antenna is not uniform in all directions but varies with the angle with respect to the axis of the wire. It is most intense in directions at right-angles to the wire and zero along the direction of the wire itself, with intermediate values at intermediate angles. This is shown by the sketch of Fig. 1002, 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 (§ 9-1) 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.

ℂ 10-3 Ground Effects

Reflection — When the antenna is near the ground the free-space pattern of Fig. 1002 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 reflected waves may be in such phase relationship to the directly-radiated waves that the two completely reinforce each other, or the phase relationship may be such that complete cancellation takes place. All intermediate values also are possible. Thus, 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. 1003 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.

When the half-wave antenna is vertical the maximum and minimum points in the curves of Fig. 1003 exchange positions, so that the

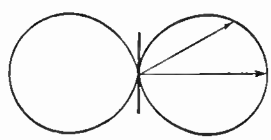


Fig. 1002 — The free-space radiation pattern of a half-wave antenna. The antenna is shown in the vertical position. This is a cross-section of the solid pattern described by the figure when rotated on its vertical axis. The "doughnut" form of the solid pattern can be more easily visualized by imagining 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.

nulls become maxima, and vice versa. In this case, the height is taken as the distance from ground to the center of the antenna.

Radiation angle — The vertical angle, or angle of radiation, is of primary importance, especially at the higher frequencies (§ 9-2, 9-4). It is advantageous, therefore, to erect the an-

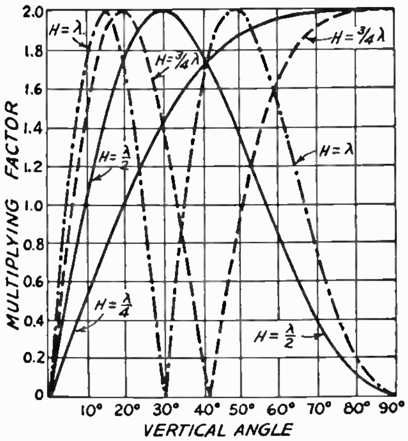


Fig. 1003 — Effect of ground on radiation of horizontal antennas at vertical angles for four antenna heights. This chart is based on perfectly-conducting ground.

tenna at a height which will take advantage of ground reflection in such a way as to reinforce the space radiation at the most desirable angle. Since low radiation angles usually are desirable, this generally means that the antenna should be high — at least ½ wavelength at 14 Mc., and preferably ¾ or 1 wavelength; at least 1 wavelength, and preferably higher, at 28 Mc. and the very-high frequencies. The physical height required for a given height in wavelengths decreases as the frequency is increased, so that good heights are not impracticable; a half wavelength at 14 Mc. is only 35 feet, approximately, while the same height represents a full wavelength at 28 Mc. At 7 Mc. and lower frequencies the higher radiation angles are effective, so that again a reasonable antenna height is not difficult of attainment. Heights between 35 and 70 feet are suitable for all bands, the higher figures generally being preferable where circumstances permit their use.

Imperfect ground — Fig. 1003 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 at heights of less than several wavelengths. Above 15 degrees, however, the curves are accurate enough for all practical purposes, and may be taken as indicative of the sort of result to be expected at angles between 5 and 15 degrees.

The effective ground plane — that is, the plane from which ground reflections can be considered to take place — seldom is the actual surface of the ground but is a few feet below it, depending upon the character of the soil.

Impedance — Waves which 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. 1004. 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 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. This can be readily seen by imagining that Fig. 1002 is lying on the ground, and that the pattern is looked at from above.

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. In practice, however, at high frequencies both types work about alike at low angles.

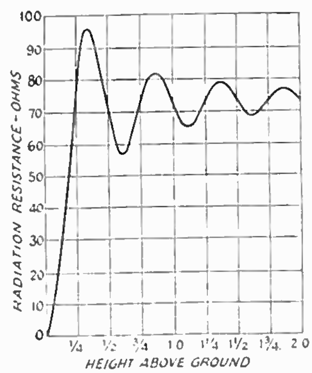


Fig. 1004 — Theoretical curve of variation of radiation resistance for a half-wave horizontal antenna, as a function of height in wavelength above perfectly-reflecting ground.

Effective radiation patterns — In determining the radiation pattern it is necessary to consider radiation in both the horizontal and vertical planes. When the half-wave antenna is vertical, the vertical angle of radiation chosen does not affect the *shape* of the horizontal pattern, but only its relative amplitude. When the antenna is horizontal, however, both the shape and amplitude are dependent upon the angle of radiation chosen.

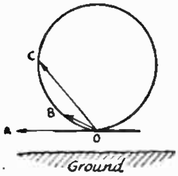


Fig. 1005 — Illustrating the importance of vertical angle of radiation in determining antenna directional effects. Ground reflection is neglected in this drawing of the free-space field pattern of a horizontal antenna.

Fig. 1005 illustrates this point. The "free-space" pattern of the horizontal antenna shown is a section cut vertically through the solid pattern. In the direction *OA*, horizontally along the wire axis, the radiation is zero. At some vertical angle, however, represented by the line *OB*, the radiation is appreciable, despite the fact that this line runs *in the same geographical direction* as *OA*. At some higher angle, *OC*, the radiation, *still in the same geographical direction*, is still more intense. The effective radiation pattern therefore depends upon which angle of radiation is most useful, and for long-distance transmission is dependent upon the conditions existing in the ionosphere. These conditions may vary not only from day to day and hour to hour, but even from minute to minute. Obviously, then, the effective directivity of the antenna will change along with transmission conditions

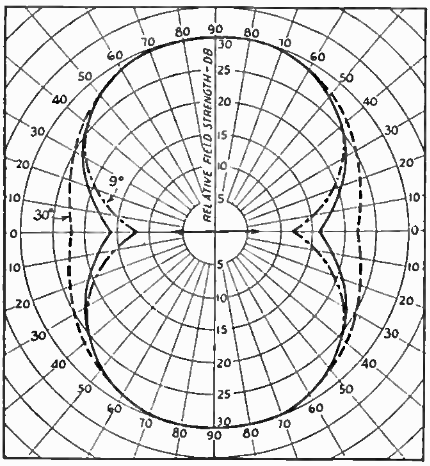


Fig. 1006 — Horizontal pattern of a horizontal half-wave antenna at three vertical radiation angles. The solid line is relative radiation at 15 degrees. Dotted lines show deviation from the 15-degree pattern for angles of 9 and 30 degrees. The patterns are useful for shape only, since the amplitude will 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.

At very-high frequencies, where only extremely low angles are useful for any but sporadic-*E* transmission (§ 9-2), the effective radiation pattern of the antenna approaches the free-space pattern. A horizontal antenna therefore shows more marked directive effects than it does at lower frequencies, on which high radiation angles are effective.

Theoretical horizontal-directivity patterns for half-wave horizontal antennas at vertical angles of 9, 15, and 30 degrees (representing average useful angles at 28, 14 and 7 Mc. respectively) are given in Fig. 1006. At intermediate angles the values in the affected regions also will be intermediate. Relative field strengths are plotted on a decibel scale (§ 3-3), so that they represent as nearly as possible the actual aural effect at the receiving station.

10-4 Applying Power to the Antenna

Direct excitation — When power is transferred directly from the source to the radiating antenna, the antenna is said to be directly excited. While almost any coupling method (§ 2-11) may be used, those most commonly employed are shown in Fig. 1007. Power usually is fed to the antenna at either a current or voltage loop (§ 10-2). If power is fed at a current loop, the coupling method is called *current feed;* if at a voltage loop, the method is called *voltage feed.*

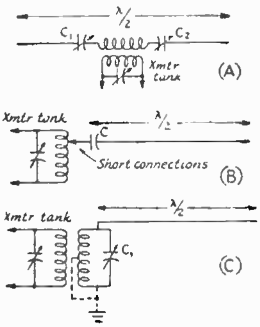


Fig. 1007 — Methods of directly exciting the half-wave antenna. A, current feed, series tuning; B, voltage feed, capacity coupling; C, voltage feed, with an inductively-coupled antenna tank. In A, the coupling circuit is not included in the effective electrical length of the antenna system proper.

Current feed — This method is shown in Fig 1007-A. The antenna is cut at the center and a small coil coupled to the output tank circuit of the transmitter, with adjustable coupling so that the transmitter loading can be controlled. Since the addition of the coil "loads" the antenna, or increases its effective length because of the additional inductance, the series condensers, C_1 and C_2, are used to provide electrical means for reducing the length to its original unloaded value; in other words, their capacitive reactance serves to cancel the effect of the inductive reactance of the coil (§ 2-10).

Voltage feed — In Fig. 1007, at B and C the power is introduced into the antenna at a point of high voltage. In B, the end of the antenna is coupled to the output tank circuit through a small condenser, C; in C, a separate tank circuit, connected directly to the antenna, is used. This tank is tuned to the transmitter frequency, and should be grounded at one end or at the center of the coil, as shown.

Adjustment of coupling — Methods of tuning and adjustment of direct-feed systems correspond to those used with transmission lines, which are discussed in § 10-6.

Disadvantages of direct excitation — Direct excitation seldom is used except on the lowest amateur frequencies, because it involves bringing the antenna proper into the operating room and hence into close relationship with the house and electric wiring. This usually means that some of the power is wasted in heating poor conductors in the field of the antenna. Also, it often means that the shape of the antenna must be distorted, so that the expected directional effects are not realized, and likewise that the height will be limited. For these reasons, in high-frequency work practically all amateurs use transmission lines or feeder systems, which permit placing the antenna in a desirable location.

ℂ 10-5 Transmission Lines

Requirements — A transmission line (§ 2-12-A) is used to transfer power, with a minimum of loss, from the transmitter to the antenna from which the power is to be radiated. At radio frequencies, where every wire carrying r.f. current tends to radiate energy in the form of electromagnetic waves, special design is necessary to minimize radiation and thus cause as much of the power as possible to be delivered to the receiving end of the line.

Radiation can be minimized by using a line in which the current is low, and by using two conductors carrying currents of equal magnitudes but opposite phase so that the fields about the conductors cancel each other. For good cancellation of radiation, the two conductors should be kept parallel and quite close to each other.

Types — The most common form of transmission line consists of two parallel wires, maintained at a fixed spacing of two to six inches by insulating spacers or spreaders placed at suitable intervals (*open-wire line*). A second type consists of insulated wires twisted together to form a flexible line, without spacers (*twisted-pair line*). A third has the parallel wires maintained at a fixed spacing of a half inch or less by molding them in a flexible tape of low-loss insulating material. Another type of line has a wire inside of and coaxial with a tubing outer conductor, separated from the outer conductor by insulating spacers or "beads" at regular intervals (*coaxial* or *concentric line*). A variation of this type uses solid but flexible insulating material to fill the space between the inner and outer conductors, the latter usually being made of metal braid rather than of solid tubing, so that the line will be flexible. Still another type of line uses only a single wire, without a second conductor (*single-wire feeder*); in this type, radiation is minimized by keeping the line current low.

Spacing of open-wire lines — The spacing between the wires of an open-wire line should be small in comparison to the operating wavelength, to prevent appreciable radiation. It is impracticable to make the spacing of an open-wire line very small, however, because when the wires swing with respect to each other in a wind the line constants (§ 2-12-A) will vary, and thus cause a variation in tuning or loading on the transmitter. It is also desirable to use as few insulating spacers as possible, to keep the weight of the line to a minimum. In practice, a spacing of about six inches is used for 14 Mc. and lower bands, with four- and two-inch spacings being common on very-high frequencies.

Electrical length — The electrical length of a transmission line may be quite different from its physical length, because waves travel more slowly along a transmission line than they do in space. The difference is small in the case of air-insulated lines, but is considerable in lines having solid dielectrics. The ratio of the physical length of a line one electrical wavelength long to a wavelength in space is called the *velocity factor* of the line. A line with a velocity factor of 0.65, for example, will have an *electrical* length of 10 meters (space wavelength) when it is 6.5 meters long.

Table I gives velocity factors for various types of lines in common use. This factor must always be used in calculating the length of a solid-dielectric line used, for instance, as a quarter-wave matching section as described later in this chapter. The physical length of a quarter-wave line is

$$\text{Length of quarter-wave line in feet} = \frac{246 \times V}{\text{Freq. (Mc.)}} \quad (5)$$

or

$$\text{Length of quarter-wave line in inches} = \frac{2950 \times V}{\text{Freq. (Mc.)}} \quad (6)$$

where V is the velocity factor given in Table I.

Balance to ground — For maximum cancellation of the fields about the two wires, it is necessary that the currents be equal in amplitude and opposite in phase. Should the capacity or inductance per unit length in one wire differ from that in the other, this condition cannot be fulfilled. Insofar as the line itself is concerned, the two wires will have identical characteristics only when the two have exactly the same physical relationships to ground and to other objects in the vicinity. Thus, the line should be symmetrically constructed and the two wires should be at the same height. Line unbalance can be minimized by keeping the line as far above the ground and as far from other objects as possible.

To overcome unbalance the line sometimes is transposed, which means that the positions of the wires are interchanged at regular intervals. This procedure is more helpful on long

TABLE I

Transmission-Line Velocity Factors and Attenuation

Type of Line	*Velocity Factor V*	** *Attenuation, db./100 ft.; Mc.* 3.5	7	14	28	50	144	*Capacitance per foot μμfd.*
Open-wire, 400 to 600 ohms	0.975*	0.03	0.05	0.07	0.1	0.13	0.25	
Parallel-tubing	0.95*	***						
Coaxial, air-insulated	0.85*	0.2	0.28	0.42	0.55	0.7	1.4	
RG-8/U (53 ohms)	0.66	0.28	0.42	0.61	1.0	1.4	2.6	29.5
RG-58/U (53 ohms)	0.66	0.53	0.8	1.2	1.9	2.7	5.1	28.5
RG-11/U (75 ohms)	0.66	0.27	0.41	0.61	0.92	1.3	2.4	20.5
Twin-Lead, 300 ohms	0.82	0.18	0.3	0.5	0.84	1.3	2.8	5.8
Twin-Lead, 150 ohms	0.77	0.2	0.35	0.6	1.0	1.6	3.5	10
Twin-Lead, 75 ohms	0.68	0.37	0.64	1.1	1.9	3.0	6.8	19
Transmitting Twin-Lead, 75 ohms	0.71	0.29	0.49	0.82	1.4	2.1	4.8	
Rubber-insulated twisted-pair or coaxial ****	0.56 to 0.65	0.96	1.6	2.5	4.2	6.2	13	

* Average figures for air-insulated lines taking into account effect of insulating spacers.
** For lines terminated in characteristic impedance.
*** Losses between open-wire line and air-insulated coaxial cable. Actual loss with both open-wire and parallel-tubing lines is higher than listed because of radiation, especially at higher frequencies.
**** Approximate figures for good-quality rubber insulation.

than on short lines, and need not be resorted to for lines less than a wavelength long.

Resonant and nonresonant lines — Lines are classified as *resonant* or *nonresonant*, depending upon the standing-wave ratio. If the ratio is near 1, the line is said to be nonresonant. Reactive effects will be small, and consequently no special tuning provisions need ordinarily be made for canceling them even when the line length is not an exact multiple of a quarter wavelength. Such a line must be terminated in its characteristic impedance (§ 2-12-A). If the standing-wave ratio is fairly

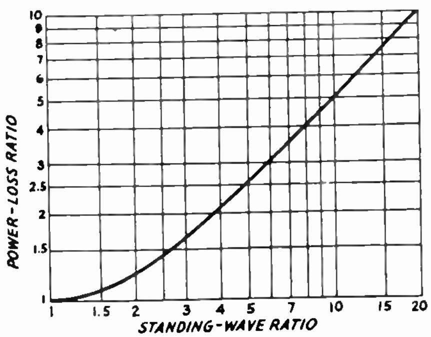


Fig. 1008 — Effect of standing-wave ratio on line loss. The power-loss ratio given by the curve, multiplied by the power that would be lost in the same line if perfectly matched, gives the actual power lost in the line when standing waves are present.

large, the input reactance must be canceled or "tuned out" unless the line is a multiple of a quarter wavelength and resonant.

Losses — There are three ways by which power may be lost in a transmission line: by radiation, by heating of the conductors (I^2R loss), and by heating of the dielectric, if any. Loss by radiation will occur if the line is unbalanced and, particularly with open-wire lines, may greatly exceed the heat losses. It can be reduced to a minimum by terminating the line in a balanced load and by symmetrical, uniform construction.

Heat losses in both the conductor and the dielectric increase with frequency. Conductor losses also are greater the lower the characteristic impedance of the line, because a higher current flows in a low-impedance line for a given power input. The converse is true of dielectric losses because these increase with the voltage, which is greater on high-impedance lines. The dielectric loss in air-insulated lines is negligible (the only loss is in the insulating spacers) and such lines operate at high efficiency when radiation losses are low. In solid-dielectric lines most of the loss is in the dielectric, the conductor losses being small.

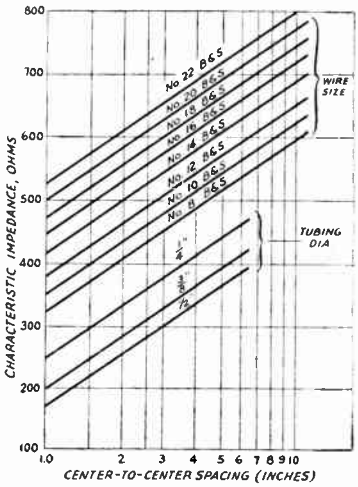


Fig. 1009 — Chart showing the characteristic impedance of typical spaced-conductor parallel transmission lines. Tubing sizes given are for outside diameters.

It is convenient to express the loss in a transmission line in decibels per unit length, since the loss in db. is directly proportional to the line length. Losses in various types of lines operated without standing waves (that is, terminated in a resistive load equal to the characteristic impedance of the line) are given in Table I. In these figures the radiation loss is assumed to be negligible. When there are standing waves on the line the power loss increases as shown in Fig. 1008.

The losses in air-insulated lines may increase considerably when the line is wet or the spacers become dirty. Moisture may also cause a change in the characteristic impedance of parallel-wire lines with solid dielectric.

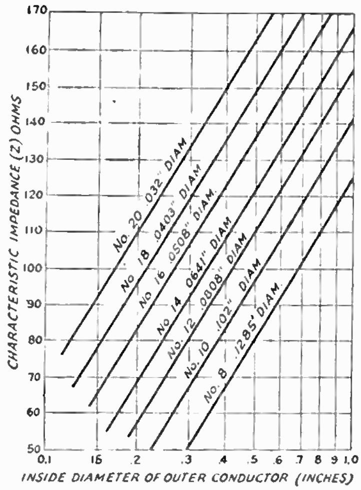


Fig. 1010 — Chart showing characteristic impedance obtained with various air-insulated concentric lines.

ℂ 10-6 Coupling to Transmission Lines

Requirements — The coupling system between a transmitter and the input end of a transmission line must provide means for adjusting the load on the transmitter to the proper value (impedance matching), and for tuning out any reactive component that may be present (§ 2-9, 2-10, 2-11). The resistance and reactance considered are those present *at the input end of the line*, and hence have nothing to do with the antenna itself except insofar as the antenna load may affect the operation of the line.

Untuned coil — One of the simplest systems, shown in Fig. 1011-A, uses a coil of a few turns tightly coupled to the plate tank coil. Since no provision is made for tuning, this system is suitable only for nonresonant lines which show practically no reactance at the input end. Loading on the transmitter may be varied by varying the coupling between the tank inductance and the *pick-up coil*, as it is frequently called, or by changing the number of turns on the pick-up coil. A slight amount of reactance is coupled into the tank circuit by the pick-up coil, since the flux leakage (§ 2-11) is high, so that some slight retuning of the plate tank condenser may be necessary when the load is connected.

Taps on tank circuit — A method suitable for use with open-wire lines is shown in Fig. 1011-B, where the line is tapped on a balanced tank circuit with taps equidistant from the center or ground point. This symmetry is necessary to maintain line balance to ground (§ 10-5). Loading is increased by moving the taps outward from the center. Any reactance present may be tuned out by readjustment of the plate tank condenser, but this method is not suitable for large values of reactance and therefore direct tapping is best confined to use with nonresonant lines.

Adjustment of untuned systems — Adjustment of either of the above systems is quite simple. Starting with loose coupling, apply power to the transmitter, and adjust the plate tank condenser for minimum plate current. If the current is less than the desired load value, increase the coupling and again resonate the plate condenser. Continue until the desired plate current is obtained, always keeping the plate tank condenser at the setting which gives minimum current.

Pi-section coupling — A coupling system which is electrically equivalent to tapping on the tank circuit, but using a capacitance voltage divider in the plate tank circuit for the purpose, is shown in Fig. 1011-C. Since one side of the condenser across which the line is connected is grounded, some unbalance will be introduced into the transmission line. This method is used chiefly with low-power portable sets, because it is readily adjustable to meet a fairly wide range of impedance values. A single-ended amplifier, using either a screen-grid tube or a grid-neutralized triode (§ 4-7), is required, since the plate tank circuit is not balanced. Coupling is adjusted by varying C_1, reresonating the circuit each time by means of C_2 until the desired amplifier plate current is obtained. In general, the coupling will increase as C_1 is made smaller with respect to C_2. Relatively large-capacity condensers are required to give a suitable impedance-matching range while maintaining resonance.

Pi-section filter — The coupling circuit shown in Fig. 1011-D is a low-pass filter capable of coupling between a fairly wide range of impedances. The method of adjustment is as follows: First, with the filter disconnected from the transmitter tank, tune the transmitter tank to resonance, as evidenced by minimum plate current. Then, with trial settings of the clips on L_1 and L_2 (few turns for high frequencies, more for lower), tap the input clips on the final tank coil at points equidistant from the center, so that about half the coil is included between them. A balanced tank circuit must be used. Set C_2 at about half scale, apply power, and rapidly rotate C_1 until the plate current drops to minimum. If this minimum is not the desired full-load plate current, try a new setting of C_2 and repeat. If, for all settings of C_2, the plate current is too high or too low, try new settings of the taps on L_1 and L_2, and also of the taps on the transmitter tank. Do not touch the tank condenser during these adjustments.

With some lengths of resonant lines, particularly those which are not exact multiples of a quarter wavelength, it may be difficult to get proper loading with the pi-section coupler. Usually antennas of these lengths also will be difficult to feed with other systems of coupling. In such cases, the proper output loading often

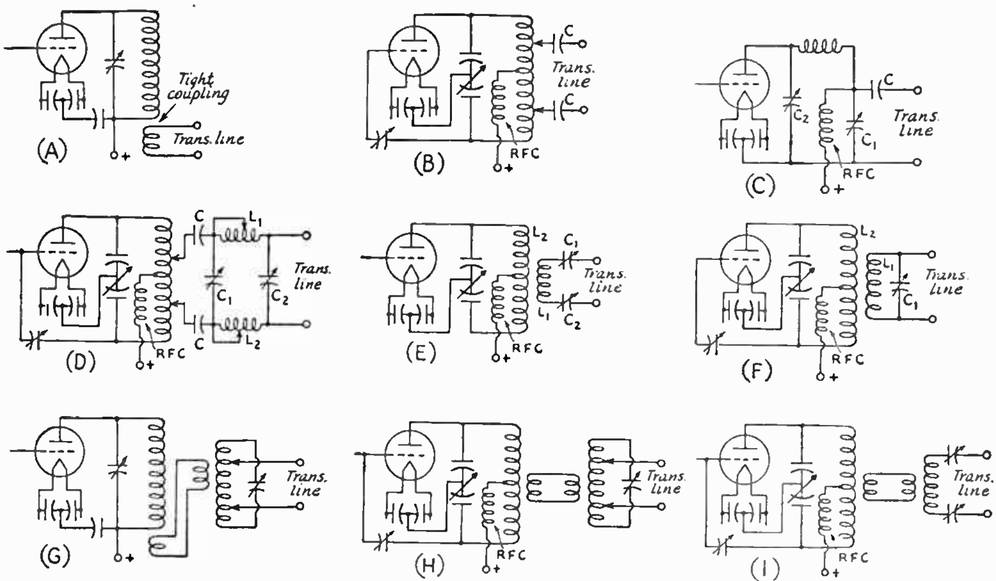


Fig. 1011— Methods of coupling the transmitter output to the transmission line. Application, circuit values and adjustment are discussed in the text. The coupling condensers, *C*, are fixed blocking condensers used to isolate the transmitter plate voltage from the antenna. Their capacity is not critical, 500 μμfd. to 0.002 μfd. being satisfactory values, but their voltage rating should at least equal the plate voltage on the final stage.

can be obtained by varying the L/C ratio of the filter over a considerably wider range than is necessary for normal loads.

Series tuning — When the input impedance of the line is low, the coupling method shown in Fig. 1011-E may be used. This system, known as *series tuning*, places the coupling coil, tuning condensers and load all in series, and is particularly suitable for use with resonant lines when a current loop appears at the input end. As shown, two tuning condensers are used, to keep the line balanced to ground. However, one will suffice, the other end of the line being connected directly to the end of L_1.

The tuning procedure with series tuning is as follows: With C_1 and C_2 at minimum capacitance, couple the antenna coil, L_1, loosely to the transmitter output tank coil, and observe the plate current. Then increase C_1 and C_2 simultaneously until a setting is reached which gives maximum plate current, indicating that the antenna system is in resonance with the transmitting frequency. Readjust the plate tank condenser to minimum plate current. This is necessary because tuning the antenna circuit will have some effect on the tuning of the plate tank. The new minimum plate current will be higher than with the antenna system detuned, but should still be well below the rated value for the tube or tubes. Increase the coupling between L_1 and L_2 by a small amount, readjust C_1 and C_2 for maximum plate current, and again set the plate tank condenser to minimum. Continue this process until the minimum plate current is equal to the rated plate current for the amplifier. Always use the degree of coupling between L_1 and L_2 which will just bring the amplifier plate current to rated value when C_1 and C_2 pass through resonance.

Parallel tuning — When the line has high input impedance, the use of parallel tuning, as shown in Fig. 1011-F, is required. Here the coupling coil, tuning condenser and line all are in parallel, the load represented by the line being directly across the tuned coupling circuit.

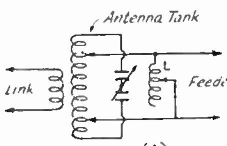
If the line is nonreactive, the coupling circuit will be tuned independently to the transmitter frequency; line reactance can be compensated for by tuning of C_1 and, if necessary, adjustment of L_1 by means of taps. Parallel tuning is suited to resonant lines when a voltage loop appears at the input end.

The tuning procedure is quite similar to that with series tuning. Find the value of coupling between L_1 and L_2 which will bring the plate current to the desired value as C_1 is tuned through resonance. Again, a slight readjustment of the amplifier tank condenser may be necessary to compensate for the effect of coupled reactance.

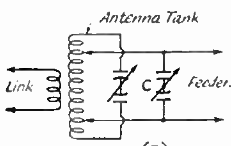
Link coupling — Where tuning of the circuit connected to the line is necessary or desirable, it is possible to separate physically the line-tuning apparatus and the plate tank circuit by means of link coupling (§ 2-11). This is often convenient from a constructional standpoint, and has the advantage that there will be somewhat less harmonic transfer to the antenna, since stray capacity coupling is lessened with the smaller link coils.

Figs. 1011-G and H show a method which can be considered to be a variation of Fig. 1011-B. The first (G) is suitable for use with a single-ended plate tank, the second (H) for a balanced tank. The auxiliary tank on which the transmission line is tapped may have adjustable inductance as well as capacitance, to provide a wide range of reactance variation for compensating for line reactance. The center of the auxiliary tank inductance may be grounded, if desired. The link windings should be placed at the grounded parts of the coils, to reduce capacitance coupling and consequent harmonic transfer. With this inductively-coupled system, the loading on the auxiliary tank circuit increases as the taps are moved outward from the center, but, since this decreases the Q of the circuit, the coupling to the plate tank simultaneously decreases (§ 2-11). Hence, a compromise adjustment giving proper loading must be found in practice. Loading also may be varied by changing the coupling between one link winding and its associated tank coil; either tank may be used for this purpose. When the auxiliary tank is properly tuned to compensate for line reactance, the plate-tank tuning will be practically the same as with no load; hence, the plate tank condenser need be readjusted only slightly to compensate for the small reactance introduced by the link.

With some antenna systems and line lengths it may be difficult to make these perform simultaneously the functions of compensating for the input reactance of the line and matching the input resistance of the line to the transmitter. In such cases it will be hard to find a definite resonance point when tuning the antenna tank circuit, and it may also be impossible to load the amplifier to normal plate current. This condition frequently is accompanied by excessive heating of parts of the antenna tank coil. It may be overcome by separately tuning out the line reactance as shown in Fig. 1012. The tuning procedure is as follows: First, with the feeder taps disconnected and with very loose coupling between the two tank circuits, tune the antenna tank to resonance as indicated by a rise in plate current. Then attach the feeder taps, keeping them quite close together, and note whether the antenna tank condenser capacitance has to be increased or decreased to reresonate the circuit. If the capacitance has to be decreased,



(A)



(B)

Fig. 1012 — Use of auxiliary coil (L) or condenser (C) to tune out line input reactance with the link-coupled circuits of Figs. 1011-G and H.

use Fig. 1012-A; if increased, use circuit B. Adjust the auxiliary inductance (L) or capacitance (C) to the value which permits tapping the line on the antenna tank coil without changing the tuning of this circuit. The spread between the taps may then be adjusted as described above to give normal loading. Values of auxiliary inductance and capacitance required must be determined experimentally.

Link coupling also may be used with series tuning, as shown in Fig. 1011-I. The coupling between one link and its associated coil may be made variable, to give the same effect as changing the coupling between the plate tank and antenna coils in the ordinary system. The tuning procedure is the same as described above for series tuning. In the case of single-ended tank circuits the input link is coupled to the grounded end of the tank coil, as in Fig. 1011-G.

Circuit values — The values of inductance and capacity to use in the antenna coupling system will depend upon the transmitting frequency, but are not particularly critical. With series tuning (Figs. 1011-E, I), the coil may consist of a few turns of the same construction as is used in the final tank; average values will run from one or two turns at very-high frequencies to perhaps 10 or 12 at 3.5 Mc. The number of turns preferably should be adjustable so that the inductance can be changed should it not be possible to reach resonance with the condensers used. The series condensers should have a maximum capacitance of 250 or 350 $\mu\mu$fd. at the lower frequencies; the same values will serve even at 28 Mc., although 100 $\mu\mu$fd. will be ample for this and the 14-Mc. band. Still smaller condensers can be used at very-high frequencies. Since series tuning is used at a low-voltage point in the feeder system, the plate spacing of the condensers does not have to be large. Ordinary receiving-type condensers are large enough for plate voltages up to 1000, and the smaller transmitting condensers have high-enough voltage ratings for higher-power applications. In high-power radiotelephone transmitters it may be necessary to use condensers having a plate spacing of approximately 0.15 to 0.2 inch.

In parallel-tuned circuits (F, G, H) the antenna coil and condenser should be approximately the same as those used in the final tank circuit. The antenna tank circuit must be capable of being tuned independently to the transmitting frequency, and, if possible, provision should be made for tapping the coil, so that the L/C ratio can be varied to the optimum value (§ 2-11) as determined experimentally.

In Fig. 1011-D, C_1 and C_2 may be 100 to 250 $\mu\mu$fd. each, the higher-capacitance values being used for lower-frequency operation (3.5 Mc. and lower). Plate spacing should be, in general, at least half that of the final-amplifier tank condenser. For operation up to 14 Mc., L_1 and L_2 each may consist of 12 turns, 2½ inches in diameter, spaced to occupy 3 inches length, and tapped every three turns. Approximate settings are 9 turns for 3.5 Mc., 6 turns for 7 Mc., and 3 turns for 14 Mc. The coils may be wound with No. 14 or No. 12 wire. This method of coupling is very seldom used at very-high frequencies.

Harmonic reduction — It is important to prevent harmonics in the output of the transmitter from being transferred to the antenna system. Harmonics are readily fed to the antenna system by coupling methods which require a connection to the plate tank circuit, either direct or through condensers, as in B, C and D, Fig. 1011. Harmonic transfer is much less likely with inductively-coupled systems, particularly when a separate tuning system is provided at the input end of the line as in E, F, G, H, and I.

In inductively-coupled systems, care must be taken to prevent stray capacitance coupling between coils. Link coils always should be coupled at a point of ground potential (§ 2-13) on the plate tank coil, as also should series- and parallel-tuned coils (E and F), when possible. The effect of stray capacitance can be reduced by grounding (to the amplifier chassis) the center of the coupling coil in Fig. 1011-E and F, and by similarly grounding one side of the coupling coil at the amplifier end in G, H, and I. Capacitance coupling can be practically eliminated by the use of a Faraday shield (§ 4-9) between the plate-tank and antenna coils.

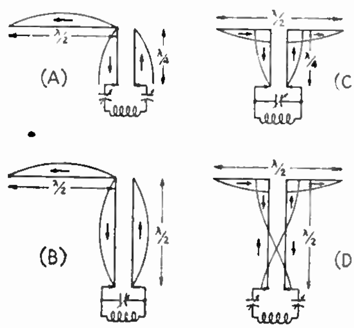


Fig. 1013 — **Half-wave antennas fed from resonant lines. A and B are end-feed systems for use with quarter- and half-wave lines; C and D are center-feed systems. The current distribution is shown for all four cases, arrows indicating the instantaneous direction of current flow.**

ℂ 10-7 Resonant Lines

Two-wire lines — Because of its simplicity of adjustment and flexibility with respect to the frequency range over which an antenna system will operate, the resonant line is widely used with simple antenna systems. Because resonant lines operate with relatively high standing-wave ratios, lines with air dielectric are to be preferred for this purpose in view of their low losses (§ 10-5). However, if the line is short — say less than 100 feet — lines having low-loss solid dielectric (polyethylene) such as 300-ohm "Twin-Lead" can be used without undue loss at frequencies below 30 Mc.

Connection to antenna — A resonant line is usually — in fact, practically always — connected to the antenna at either a current or voltage loop. This is advantageous, especially when the antenna is to be operated at harmonic frequencies, since it simplifies the problem of determining the coupling system to be used at the input end of the line.

Half-wave antenna with resonant line — It is often helpful to look upon the resonant line simply as an antenna folded back on itself. Such a line may be any whole-number multiple of a quarter wave in length; in other words, any total wire length which will accommodate a whole number of standing waves. (The "length" of a two-wire line is, however, always taken as the length of one of the wires.)

Quarter- and half-wave resonant lines feeding half-wave antennas are shown in Fig. 1013. The current distribution on both antenna and line is indicated. It will be noted that the quarter-wave line has maximum current at one end and minimum current at the other, determined by the point of connection to the antenna. The half-wave line, however, has the same current (and voltage) values at both ends.

If a quarter-wave line is connected to the end of an antenna, as shown in Fig. 1013-A, then at the transmitter end of the line the current is high and the voltage low (low impedance), so that series tuning (§ 10-6) can be used. Should the line be a half-wave long, as at 1013-B, current will be minimum and voltage maximum (high impedance) at the transmitter end of the line, just as it is at the end of the antenna. Parallel tuning therefore is required (§ 10-6). The line could be coupled to a balanced final tank through small condensers, as in Fig. 1011-B, but the inductively-coupled circuit is preferable. An end-fed antenna with resonant feeders, as in 1013-A and B, is known as the "Zeppelin" or "Zepp" antenna.

The line also may be inserted at the center of the antenna at the maximum-current point. Quarter- and half-wave lines used in this way are shown at Fig. 1013-C and D. In C, the antenna end of the line is at a high-current low-voltage point (§ 10-2); hence, at the transmitter end the current is low and the voltage high. Parallel tuning therefore is used. The half-wave line at D has high current and low voltage at both ends, so that series tuning is used at the transmitter end.

The four arrangements shown in Fig. 1013 are thoroughly-useful antenna systems, and are shown in more practical form in Fig. 1014. In each case the antenna is a half wavelength long, the exact length being calculated from Equations 2, 3 or 4 (§ 10-2). The line length should be an integral multiple of a quarter wavelength and may be calculated from Equations 5 and 6 (§ 10-5), the result being multiplied by any whole number which gives a total length convenient for reaching from the antenna to the transmitter. If there is an *odd* number of quar-

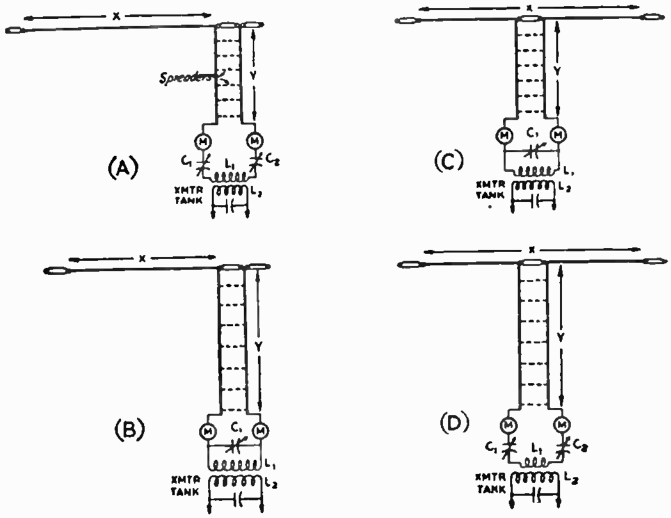


Fig. 1014 — Practical half-wave antenna systems using resonant-line feed. In the center-feed systems, the antenna length, X, does not include the length of the insulator at the center. Line length is measured from the antenna to the tuning apparatus; leads in the latter should be kept short enough so their effect can be neglected. The use of two r.f. ammeters, *M*, as shown is helpful for balancing feeder currents; however, one meter is sufficient to enable tuning for maximum output, and may be transferred from one feeder to the other, if desired. The systems at (A) and (C) are for feeders an odd number of quarter waves in length; (B) and (D) are for feeders a multiple of a half wavelength. The detailed drawings shown here correspond electrically to the elementary-schematic half-wave antenna systems shown in Fig. 1013.

ter waves on the line in the case of the end-fed antenna, series tuning should be used at the transmitter end; if an *even* number of quarter waves, then parallel tuning should be used. With the center-fed antenna the reverse is true.

Practical line lengths — In general, it is best to use line lengths that are integral multiples of a quarter wavelength. Intermediate lengths will give intermediate impedance values and will show reactance (§ 2-12-A) as well. The tuning apparatus is capable of compensating for reactance, but it may be difficult to get suitable transmitter loading because simple series and parallel tuning are suitable for only low and high impedances, respectively, and neither will perform well with impedances of the order of a few hundred ohms. Such values of impedance may reduce the *Q* of the coupling circuit to a point where adequate coupling cannot be obtained (§ 2-11). However, some departure from the ideal length is possible — even as much as 25 per cent of a quarter wave in many cases — without undue difficulty in tuning and coupling. In such cases the type of tuning to use, whether series or parallel, will depend on whether the feeder length is nearer an odd number of quarter waves or nearer an even number, as well as on the point at which the feeder is connected to the antenna — at the end or in the center.

Line current — The feeder current as read by the r.f. ammeters is useful for tuning purposes only; the absolute value is of little importance. When series tuning is used the current will be high, but very little current will be indicated in a parallel-tuned system. This is because of the current distribution on the feeders, as shown by Fig. 1013. With a given antenna and tuning system, of course, the greatest power will be delivered to the antenna when the readings are highest. However, should the feeder length be changed no useful conclusions can be drawn from comparison between the new and old readings. For this reason, any indicator which registers the relative intensity of r.f. current can be used for tuning purposes. Many amateurs, in fact, use

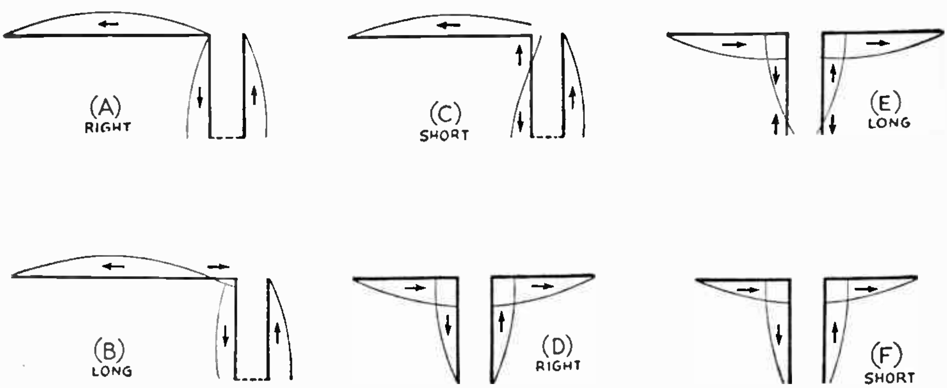


Fig. 1015 — Illustrating the effect on feeder balance of incorrect antenna length for various types of antenna systems. In end-feed systems, the current minimum shifts above or below the feeder junction, unbalancing the line. With center feed, incorrect antenna length does not unbalance the transmission line as it does with end feed.

flashlight or dial lamps for this purpose instead of meters. Such lamps are inexpensive indicators, and, when shunted by short lengths of wire so that considerable current can be passed without danger of burn-out, will serve very well even with high-power transmitters.

Antenna length and line operation — Insofar as the operation of the antenna itself is concerned, departures of a few per cent from the exact length for resonance are of negligible consequence. However such inaccuracies may influence the behavior of the feeder system, and as a result may have an adverse effect on the operation of the system as a whole. This is true particularly of end-fed antennas, such as are shown in Figs. 1014-A and B.

For example, Fig. 1015-A shows the current distribution on the half-wave antenna and quarter-wave feeder when the antenna length is correct. At the junction of the "live" feeder and the antenna the current is minimum, so that the currents in the two feeder wires are equal at all corresponding points along their length. When the antenna is too long, as in B, the current minimum occurs at a point on the antenna proper, so that at the top of the live feeder there is already appreciable current flowing, whereas at the top of the "dead" feeder the current must be zero. As a result the feeder currents are not balanced, and some power will be radiated from the line. In C, the antenna is too short, bringing the current minimum to a point on the live feeder, so that again the currents are unbalanced. The more serious the unbalance, the greater the radiation from the line.

If the antenna is fed at the center the undesirable effects of incorrect antenna length balance out, so that the line operates properly under all conditions. This is shown in Fig. 1015 at D, E and F. So long as the two halves of the antenna are of equal length the distribution of current on the feeders will be symmetrical, so that no unbalance exists even for antenna lengths considerably removed from the correct value.

In the interests of reducing radiation from the transmission line, therefore, feeding at the center of the antenna system is preferable to feeding at the end. Strictly speaking, end-fed systems of the type shown in Fig. 1014 at A and B cannot be truly balanced because the current at the end of the wire connected to the antenna is finite, though small, while the current at the end of the open wire is zero.

10-8 Nonresonant Lines

Requirements — The advantages of nonresonant transmission lines — minimum losses, and elimination of the necessity for tuning — make the use of this type of line attractive. The chief disadvantage of the nonresonant line, aside from the necessity for more care in initial adjustment, is that when "matched" to the ordinary antenna the match is perfect only for one frequency, or at most for a small band of frequencies on either side of the frequency for which the matching is done. Except for a few special systems, such an antenna is unsuitable for work on more than one amateur band.

Adjustment of a nonresonant line is simply a process of adjusting the terminating resistance to match the characteristic impedance of the line. To accomplish this the antenna itself must be resonant at the selected frequency, and the line must then be connected to it in such a way that the antenna impedance as looked at by the line is the right value. The matching may be done by connecting the line at the proper spot along the antenna, by inserting an impedance-transforming device between the antenna and line, or by using a line having an impedance equal to the center impedance of the antenna.

An impedance mismatch of several per cent is of little consequence so far as power transfer to the antenna is concerned. It is relatively easy to get the standing-wave ratio down to 2 or 3 to 1, a perfectly satisfactory condition in practice. Of considerably greater importance is the necessity for getting the currents in the two wires balanced, both as to amplitude and phase. If the currents are not the same at corresponding points on adjacent wires and the loops and nodes do not also occur at corresponding points, there will be considerable radiation loss. Perfect balance can be brought about only by perfect symmetry in the line, particularly with respect to ground. This symmetry should extend to the coupling apparatus at the transmitter. An electrostatic shield between the line and the transmitter coupling coils often will be of value in preventing capacitance unbalance, and at the same time will reduce harmonic radiation.

In the following discussion of ways in which different types of lines may be matched to the antenna, a half-wave antenna is used as an example. Other types of antennas may be treated by the same methods, making due allowance for the order of impedance that appears at the end of the line when more elaborate systems are used.

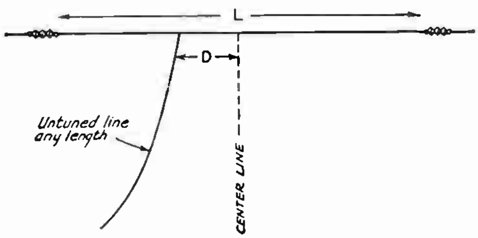


Fig. 1016 — Single-wire feed system.

Single-wire feed — In the single-wire feed system, the return circuit is through the ground. There will be no standing waves on the feeder when its characteristic impedance is matched by the impedance of the antenna at the connection point. The principal dimensions (Fig. 1016) are the length of the antenna, *L*, and the distance, *D*, from the exact center of the antenna to the point at which the feeder is attached. The antenna length may be calculated from Equation 2 (§ 10-2). The distance *D* depends upon the diameter of the feeder wire, since this diameter determines its characteristic impedance. For No. 14 wire *D* is equal to the antenna length multiplied by 0.139; for No. 12 wire the factor is 0.133.

In constructing an antenna system of this type, the feeder must run straight away from the antenna (at a right angle) for a distance of at least one-third the length of the antenna. Otherwise the field of the antenna will affect the feeder and cause faulty operation. There should be no sharp bends in the feeder wire at any point.

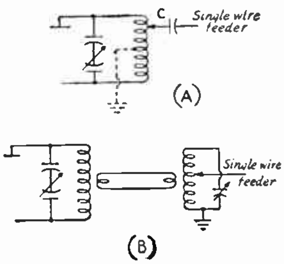


Fig. 1017 **— Methods of coupling the feeder to the transmitter in a single-wire feed system. Circuits are shown for both single-ended and balanced tank circuits.**

With the coupling system shown in Fig. 1017-A, the process of adjustment is as follows: Starting at the ground point on the tank coil, the tap is moved toward the plate end until the amplifier draws the rated plate current. The plate tank condenser should be readjusted each time the tap is changed, to bring the plate current back to minimum. The amplifier is loaded properly when this "minimum" value is equal to the rated current. The condenser, *C*, in the feeder is for the purpose of insulating the antenna system from the high-voltage plate supply when series plate feed is used. It should have a voltage rating somewhat higher than that of the plate supply. Almost any capacitance greater than 500 μμfd. will be satisfactory. The condenser is unnecessary, of course, if parallel plate feed is used.

Inductive coupling to the output circuit is shown in Fig. 1017-B. The antenna tank circuit should tune to resonance at the operating frequency, and the loading is adjusted by varying the coupling between the two tanks, both being kept tuned to resonance.

Regardless of the type of coupling employed, a good ground connection is essential with this system. Single-wire feed works best over moist ground, and comparatively poorly over rock and sand.

Twisted-pair feed — A two-wire line composed of twisted rubber-covered wires or close-spaced parallel wires with polyethylene insulation can be constructed to have a surge impedance approximately equal to the 70-ohm impedance at the center of the antenna itself, thus permitting connecting the line to the antenna as shown in Fig. 1018. Any discrepancy which may exist between line and antenna impedance can be compensated for by a slight fanning of the line where it connects to the two halves of the antenna, as indicated at B in Fig. 1018. The twisted-pair line is a convenient type to use, since it is easy to install and the r.f. voltage on it is low because of the low impedance.

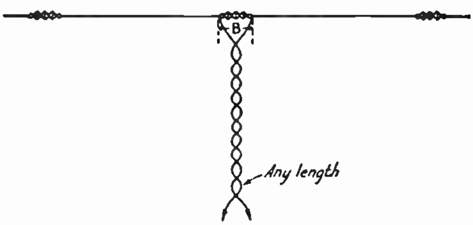


Fig. 1018 **— Half-wave antenna center fed by a twisted-pair line. Fanning (*B*) compensates for line impedance.**

The antenna should be one-half wavelength long for the frequency of operation, as determined by the formulas (§10-2). The amount of "fanning" (dimension *B*) will depend upon the kind of cable used; the required spacing usually will be between 6 and 18 inches. It may be checked by inserting ammeters in each antenna leg at the junction of the feeder and antenna; the value of *B* which gives the largest current is correct. Alternatively, the system may be operated continuously for a time with fairly high r.f. power input, after which the feeder may be inspected (by touch) for hot spots. These indicate the presence of standing waves, and the fanning should be adjusted until they are eliminated or minimized. Each leg of the feeder forming the triangle at the antenna should be equal in length to dimension *B*.

Coupling between the transmitter and the transmission line is ordinarily accomplished by the untuned coil method shown in Fig. 1011-A (§ 10-6).

Concentric-line feed — A concentric transmission line can be constructed to have a surge impedance equal to the 70-ohm impedance at the center of a half-wave antenna. Such a line can be connected directly to the center of the antenna, therefore, forming the system shown in Fig. 1019.

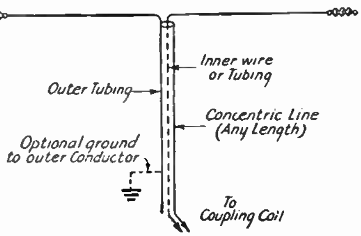


Fig. 1019 **— Half-wave antenna centerfed by a concentric transmission line of 70 ohms surge impedance.**

An air-insulated concentric line will have a surge impedance of 70 ohms when the inside diameter of the outer conductor is approximately 3.2 times the outside diameter of the inner conductor. This condition can be fulfilled by using standard 5/16-inch (outside diameter) copper tubing for the outer conductor and No. 14 wire for the inner. Ceramic insulating spacers are available commercially for this combination. Flexible solid coaxial cable having the requisite impedance for connection to the center of the antenna also is available.

The operation of such an antenna system is similar to that of the twisted-pair system just described, and the same transmitter coupling arrangements may be used (§ 10-6).

The outer conductor of the line may be grounded, if desired. The feeder system is slightly unbalanced, because the inner and outer conductors do not have the same capacitance to ground. Although the line itself, being shielded, cannot radiate, an "antenna" current can flow on the outside of the shield (outer conductor) and the cable therefore may become part of the radiating system. The magnitude of this current will depend upon the length of the cable and will be greatest when the length is such as to be resonant, in conjunction with the antenna itself, at the operating frequency. The current can be reduced by grounding the shield (with a very short lead) at any point an odd number of quarter wavelengths from the point of connection to the antenna.

Delta matching transformer — Because of the extremely close spacing required, it is impracticable to construct an open-wire transmission line which will have a surge impedance low enough to work directly into the center of a half-wave antenna. Such wire lines usually have impedances between 400 and 700 ohms, 600 ohms being a widely-used value. It is necessary, therefore, to use other means for matching the line to the antenna.

One method of matching is illustrated by the system shown in Fig. 1020. The matching section, E, is "fanned" to have a gradually increasing impedance so that its impedance at the antenna end will be equal to the impedance of the antenna section, C, while the impedance at the lower end matches that of a practicable transmission line.

The antenna length, L, the feeder clearance, E, the spacing between centers of the feeder wires, D, and the coupling length, C, are the important dimensions of this system. The system must be designed for exact impedance values as well as frequency values, and the dimensions therefore are fairly critical.

The length of the antenna is figured from Equation 2 (§ 10-2). The length of section C is computed by the formula:

$$C\ (feet) = \frac{118}{Freq.\ (Mc.)}$$

The feeder clearance, E, is found from the equation:

$$E\ (feet) = \frac{148}{Freq.\ (Mc.)}$$

The above equations are for wire antennas and for feeders having a characteristic impedance of 600 ohms and will not apply to feeders of any other impedance. The proper feeder spacing for a 600-ohm transmission line is computed to a sufficiently close approximation by the following formula:

$$D = 75 \times d$$

where D is the distance between the centers of the feeder wires and d is the diameter of the wire. If the wire diameter is in inches the spacing also will be in inches, and if the wire diameter is in millimeters the spacing also will be in millimeters.

Methods of coupling to the transmitter are discussed in § 10-6, those shown in Figs. 1011-C, D, G and H being suitable.

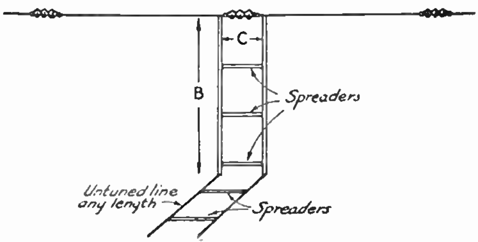


Fig. 1021 — The "Q" antenna, using a quarter-wave impedance-matching section with close-spaced conductors.

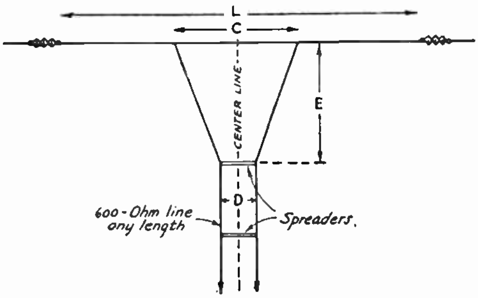


Fig. 1020 — Delta-matched antenna system. The dimensions C, D, and E are found by formulas given in the text. It is important that the matching section, E, come straight away from the antenna without any bends.

"Q"-section transformer — The impedance of a two-wire line of ordinary construction (400 to 600 ohms) can be matched to the impedance of the center of a half-wave antenna by utilizing the impedance-transforming properties of a quarter-wave line (§ 10-5). The matching section must have low surge impedance and therefore is commonly constructed of large-diameter conductors such as aluminum or copper tubing, with fairly close spacing. This system is known as the "Q" antenna. It is shown in Fig. 1021. The important dimensions are the length of the antenna, the length of the matching section, B, the spacing between the two conductors of the matching section, C, and the impedance of the untuned transmission line connected to the lower end of the matching section.

The required surge impedance for the matching section is

$$Z_s = \sqrt{Z_1 Z_2}$$

where Z_1 is the input impedance and Z_2 the output impedance. Thus a quarter-wave section matching a 600-ohm line to the center of a half-wave antenna (72 ohms) should have a surge impedance of 208 ohms. The spacings between conductors of various sizes of tubing and wire for different surge impedances are given in graphical form in Fig. 1009. With ½-inch tubing, the spacing should be 1.5 inches for an impedance of 208 ohms.

The length of the matching section, B, should be equal to a quarter wavelength, and is given by Equation 5 (§ 10-5). The length of the antenna can be calculated from Equation 2 (§ 10-2).

This system has the advantage of the simplicity of adjustment of the twisted-pair feeder system and at the same time the superior insulation of an open-wire system. Figs. 1011-B, D, G and H (§ 10-6) represent suitable methods of coupling to the transmitter.

Linear transformers — Fig. 1022 shows two methods of coupling a nonresonant line to a half-wave antenna through a quarter-wave linear transformer or matching section. In the case of the center-fed antenna, the free end of the matching section, B, is open (high impedance) since the other end is connected to a low-impedance point on the antenna. With the end-fed antenna, the free end of the matching section is closed through a shorting bar or link; this end of the section has low impedance, since the other end is connected to a high-impedance point on the antenna.

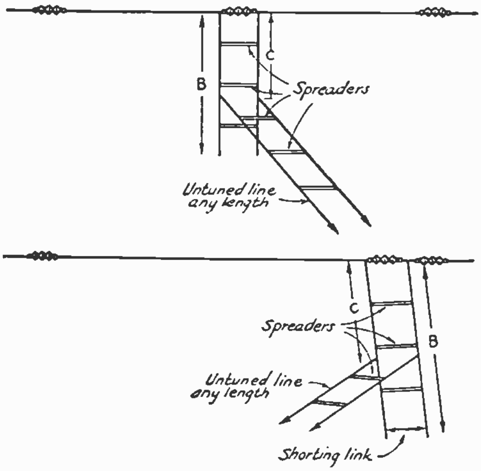


Fig. 1022 — Half-wave antenna systems with quarter-wave open-wire linear impedance-matching transformers.

When the connection between the matching section and the antenna is unbalanced, as in the end-fed system, it is important that the antenna be the right length for the operating frequency if a good match is to be obtained (§ 10-7). The balanced center-fed system is less critical in this respect. The shorting-bar method of tuning the center-fed system to resonance may be used if the matching section is extended to a half wavelength, bringing a current loop at the free end.

In the center-fed system, the antenna and matching section should be cut to lengths found from the equations in § 10-2 and § 10-5. Any necessary on-the-ground adjustment can be made by adding to or clipping off the open ends of the matching section. In the end-fed system the matching section can be adjusted by making the line a little longer than necessary and adjusting the system to resonance by moving the shorting link up and down. Resonance can be determined by exciting the antenna at the proper frequency from a temporary antenna near by and measuring the current in the shorting bar by a low-range r.f. ammeter or galvanometer using one of the devices of this type described in the chapter on measurements. The position of the bar should be adjusted for maximum current reading. This should be done before the transmission line is attached to the matching section.

The position of the line taps will depend upon the impedance of the line as well as on the antenna impedance at the point of connection. The procedure is to take a trial point, apply power to the transmitter, and then check the transmission line for standing waves. This can be done by measuring the current in, or voltage along, the wires. At any one position along the line the currents in the two wires should be identical. Readings taken at intervals of a quarter wavelength will indicate whether or not standing waves are present.

It will not usually be possible to obtain complete elimination of standing waves when the matching stub is exactly resonant, but the line taps should be adjusted for the smallest obtainable standing-wave ratio. Then a further "touching up" of the matching-stub tuning will eliminate the remaining standing wave, provided the adjustments are carefully made. The stub must be readjusted, because when resonant it exhibits some reactance as well as resistance at all points except at the ends, and a slight lengthening or shortening of the stub is necessary to tune out this reactance.

Since the line impedance is ordinarily between 500 and 600 ohms, the same methods of coupling may be used between the transmitter and the line as are recommended for the delta-matching system and the "Q" matching transformer.

Matching stubs — The operation of the quarter-wave matching transformer of Fig. 1022 may be considered from another — and more general — viewpoint. Suppose that section C is looked upon simply as a continuation of the transmission line. Then the "free" end of the transformer becomes a "stub" line, shunting a section of the main transmission line. From this viewpoint, matching the line to the antenna becomes a matter of selecting the right type and length of stub and attaching it to the proper spot along the line.

Referring to Fig. 1023, at any distance (X) from the antenna, the line will have an impedance which may be considered to be made up of reactance (either inductive or capacitive) and resistance, in parallel. The reactive component can be eliminated by shunting the line at distance X from the antenna with another reactance equal in value but opposite in sign to the reactance presented by the line at that point. If distance X is such that the line presents an inductive reactance, a corresponding shunting capacitive reactance will be required.

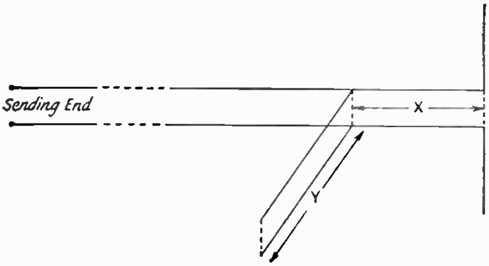


Fig. 1023 — When antenna and transmission line differ in impedance, they may be matched by a short length of transmission line. Y, called a stub. Determination of the critical dimensions, X and Y, for proper matching depends on whether the stub is open or closed at the end.

The required compensating reactance may be supplied by shunting the line with a stub cut to proper length, Y. With the reactances canceled only a pure resistance remains as a termination for the remainder of the line between the sending end and the stub, and this resistance can be adjusted to match the characteristic impedance of the line by adjusting the distance X.

Distances X and Y may be determined experimentally, but since their values are interdependent the cut-and-try method is somewhat laborious. If the standing-wave ratio and the positions of the current loops and nodes can be measured, the length and position of the stub can be found from Figs. 1024 and 1025.

Although the standing-wave ratio can be measured in terms of either current or voltage, measurement of current usually is more convenient. (If the measurements are made with a current-squared galvanometer an appropriate correction must be made, since scale readings with this type of meter are proportional to power.) With the antenna connected to the line but with the stub disconnected, the r.f. meter should be moved along the line from the antenna toward the sending end until a current loop or node is found. Its location should be marked and the value of the current

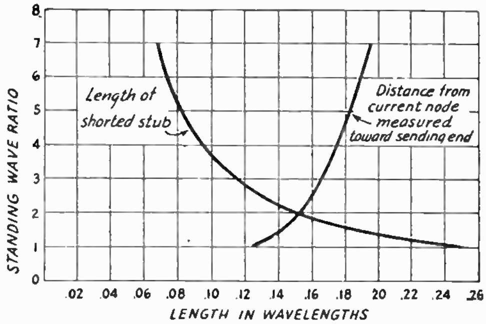


Fig. 1024 — Graph for determining position and length of a *shorted* stub. Dimensions may be converted to linear units after values have been taken from the graph.

recorded. Then the meter should be moved along toward the sending end until the next loop or node is located (if the first was a loop the second will be a node, and vice versa), and the current at this point recorded. As a cross-check for wavelength, the distance between a loop and node should be ¼ wavelength. The standing-wave ratio is the ratio of current at a loop to current at a node.

Once the standing-wave ratio is known, the length and position of the stub, in terms of wavelength, can be found directly from Figs. 1024 and 1025. The wavelength in feet for any frequency can be found from Equation 1 (§ 10-2).

Methods of coupling to the line shown in Figs. 1011-B, D, G and H (§ 10-6) can be used.

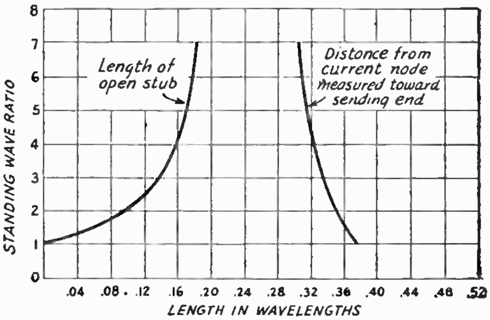


Fig. 1025 — Graph for determining position and length of an *open* stub. Dimensions may be converted to linear units after values have been taken from the graph.

Measuring standing waves — Equipment for measuring the standing-wave ratio along the transmission line is described in the chapter on measurements. At frequencies below 30 megacycles the thermomilliammeter probably is the most reliable instrument and the easiest to use. The *absolute* value of the current in the line is not important; the *ratio* between the maximum and minimum currents is what is required.

When the standing-wave ratio is low it may be difficult to determine the exact location of a node or loop since the current changes rather slowly at these points. In such a case the following procedure may be adopted: Measure the minimum current, then choose a somewhat higher value and locate two points on either side of the minimum at which the current equals the chosen value. For example, if the minimum current is 0.1 ampere, a value of 0.15 ampere might be chosen and the meter moved first to one side and then the other of the minimum point until two spots are found where the reading is 0.15 ampere. Then the node will be just half-way between these two points and may be determined very easily by measuring the distance. The same method may be used to locate a current loop with more exactness than by trying to locate the actual point of maximum current. In this case, of course, a value of current slightly lower than the maximum value should be chosen.

A crystal-detector probe pick-up measures maximum and minimum voltage rather than current. The standing-wave ratio may be measured in terms of voltage equally as well as in terms of current. However, in using the charts for the matching stub system it must be kept in mind that a voltage loop occurs at the same point as a current node, and vice versa.

ℂ 10-9 Long-Wire Antennas

Definition — 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. 1026 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.

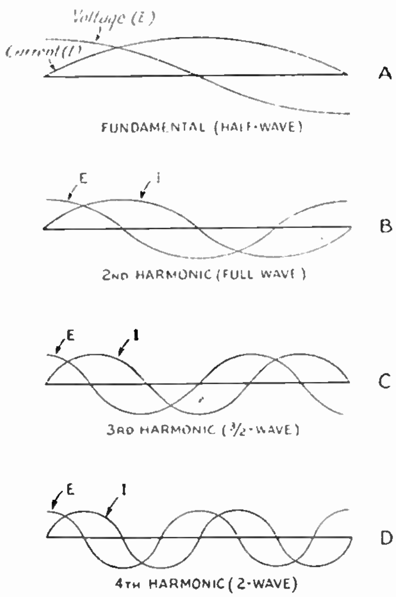


Fig. 1026 — **Standing-wave current and voltage distribution along an antenna when it is operated at various harmonics of its fundamental resonant frequency.**

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 (§ 10-2) operate only on the end sections of the antenna; in other parts of the wire these effects are absent, and the wire length is approximately that of an equivalent portion of the wave in space. The formula for the length of a long-wire antenna, therefore, is

$$Length\ (feet) = \frac{492\ (N-0.05)}{Freq.\ (Mc.)} \qquad (7)$$

where N is the number of *half* waves on the antenna. From this, 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 (on the second harmonic) because of the different behavior of end effects when there is more than one standing wave on the antenna. The effect is not very important except for a possible unbalance in the feeder system (§ 10-7), which may result in some radiation from the feeder in end-fed systems.

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. 1027 shows how the radiation resistance and the power in the lobe of maximum radiation vary with the antenna length.

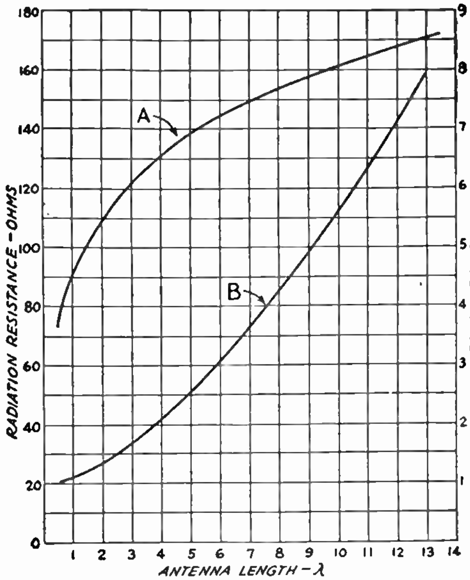


Fig. 1027 — Curve *A* shows variation in radiation resistance with antenna length. Curve *B* shows power in lobes of maximum radiation for long-wire antennas as a ratio to the maximum radiation for a half-wave antenna.

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.

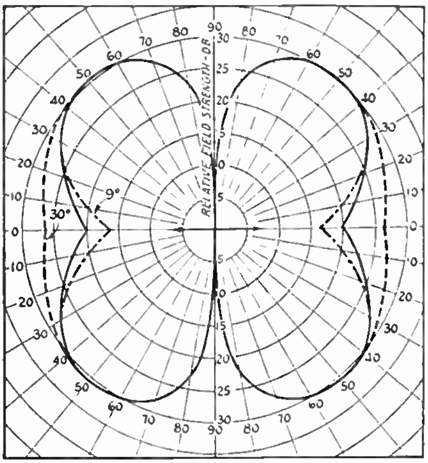


Fig. 1028 — Horizontal patterns of radiation from a *full-wave* antenna. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. All three patterns are drawn to the same relative scale: actual amplitudes will depend upon the height of the antenna.

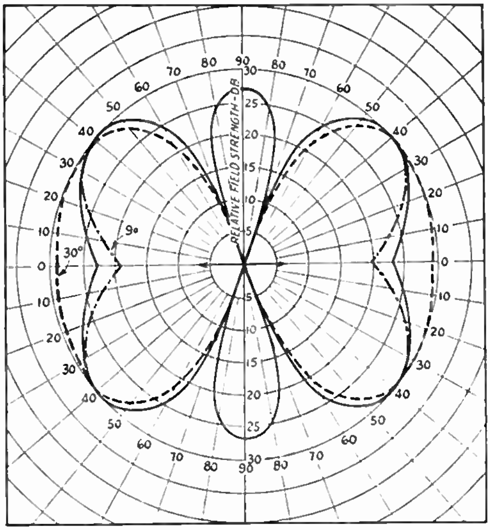


Fig. 1029 — Horizontal patterns of radiation from an antenna *three half-waves* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. Minor lobes coincide for all three angles.

Directional characteristics for antennas one wavelength, three half-wavelengths, and two wavelengths long are given in Figs. 1028, 1029 and 1030, for three vertical angles of radiation. Note that, as the wire length increases, the radiation along the line of the antenna becomes more pronounced. Still longer antennas can be considered to have practically "end-on" directional characteristics, even at the lower radiation angles.

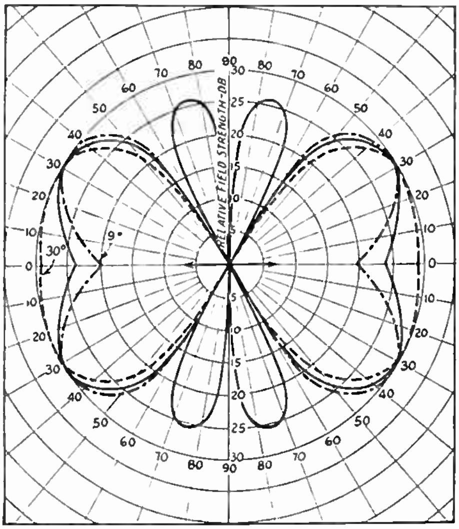


Fig. 1030 — Horizontal patterns of radiation from an antenna *two wavelengths* long. The solid line shows the pattern for a vertical angle of 15 degrees; dotted lines show deviation from the 15-degree pattern at 9 and 30 degrees. The minor lobes coincide for all three angles.

Methods of feeding — In a long-wire antenna, the currents in adjacent half-wave sections must be out of phase, as shown in Fig. 1026 and Fig. 1031. 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.

Either resonant or nonresonant feeders may be used. With the latter, the systems employing a matching section (§ 10-8) are best. The nonresonant line may be tapped on the matching section, as in Fig. 1022, or a "Q"-type section, Fig. 1021, may be employed. In such case, Fig. 1032 gives the required surge impedance for the matching section. It can also be calculated as described in § 10-8 from the radiation resistance data in Fig. 1027.

Methods of coupling the line to the transmitter are the same as described in § 10-6 for the particular type of line used.

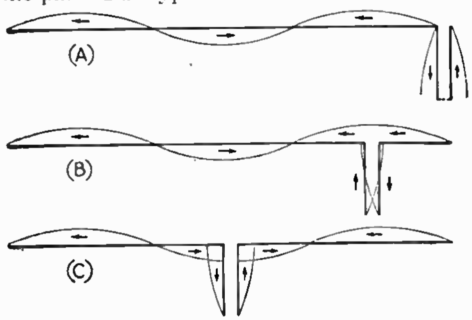


Fig. 1031 — Current distribution and feed points for long-wire antennas. A 3/2-wave antenna is used as an illustration. With two-wire feed, the line may be connected at the end of the antenna or at any current *loop* (but not at a current node) for harmonic operation.

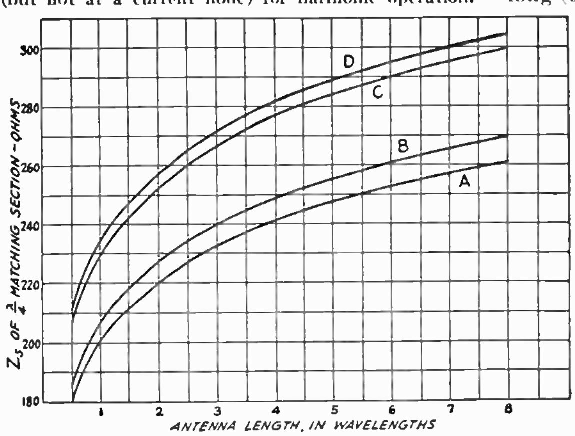


Fig. 1032 — Required surge impedance of quarter-wave matching sections for radiators of various lengths. Curve *A* is for a transmission-line impedance of 440 ohms, curve *B* is for 470 ohms, curve *C* for 580 ohms and curve *D* for 600 ohms. Dimensions for matching sections of the required impedance are obtained from Fig. 1009.

10-10 Multiband Antennas

Principles — 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. A matching section which is only a quarter wavelength long at one frequency will be a half wavelength long at twice that frequency, and so on; and changing the length of the wires, even by switching, is so inconvenient as to be impracticable.

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 which is center fed by a rubber-insulated line is practically useless for harmonic operation; on all even harmonics there is a voltage maximum occurring right at the feed point, and the resultant impedance mismatch is so bad that there is a large standing-wave ratio and consequently high losses arise in the rubber dielectric. It is also wise not to attempt to use a half-wave antenna center fed with coaxial cable, even the type using polyethylene dielectric, on its harmonics. Higher-impedance solid-dielectric lines such as 300-ohm Twin-Lead may be used, however, provided the power does not exceed a few hundred watts.

When the same antenna is used for work in several bands, it must be realized that the directional characteristic will vary with the band in use.

Simple systems — Any of the antenna arrangements shown in § 10-7 may be used for multiband operation by making the antenna a half wave long at the lowest frequency to be used. The feeders should be a quarter wave long (electrical length), or some multiple of a quarter wave, at the same frequency. Typical examples, together with the type of tuning to be used, are given in Table II. The figures given represent a compromise designed to give satisfactory operation on all the bands considered, taking into account the change in required length as the order of the harmonic goes up.

A center-fed half-wave antenna will not operate as a long wire on harmonics, because of the phase reversal at the feeders previously mentioned (§ 10-9). On the second harmonic the two antenna sections are each a half wave long, and, since the currents are in phase, the directional characteristic is different from that of a full-wave antenna even though the over-all length is the same. On the fourth harmonic each section is a full

TABLE II

MULTIBAND RESONANT-LINE FED ANTENNAS

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
With end feed: 120	60	4-Mc. 'phone	series
136	67	3.5-Mc. c.w. 7 Mc. 14 Mc. 28 Mc.	series parallel parallel parallel
134	67	3.5-Mc. c.w. 7 Mc.	series parallel
67	33	7 Mc. 14 Mc. 28 Mc.	series parallel parallel
With center feed: 137	67	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel parallel
67.5	34	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel

The antenna lengths given represent compromises for harmonic operation because of different end effects on different bands. The 136-foot end-fed antenna is slightly long for 3.5 Mc., but will work well in the region which quadruples into the 14-Mc. band (3500–3600 kc.). Bands not listed are not recommended for the particular antenna. The center-fed systems are less critical as to length.

On harmonics, the end-fed and center-fed antennas will not have the same directional characteristics, as explained in the text.

wave long, and, again, because of the direction of current flow, the system will not operate as a two-wavelength antenna. It should not be assumed that these systems are not effective radiators: it simply means that the directional characteristic will not be that of a long wire having the same over-all length. Rather, it will resemble the characteristic of one side of the antenna, although not necessarily having the same exact form.

Antennas with a few other types of feed systems may be operated on harmonics for the higher-frequency bands, although their performance is somewhat impaired. The single-wire fed antenna (§ 10-8) may be used in this way; the feeder and antenna will not be matched exactly on harmonics, with the result that standing waves will appear on the feeder, but the system as a whole will radiate. A better match will be obtained if the point of connection of the feeder to the antenna is made exactly one-third the over-all antenna length from one end. While this disagrees slightly with the figures given for a half-wave antenna, it has been found to work better on the harmonic frequencies.

The "Q" antenna system (§ 10-8) also can be operated on harmonics, but the line cannot operate as a nonresonant line except at the fundamental frequency of the antenna. For harmonic operation the line must be tuned, and therefore the feeder length is important. The tuning system will depend upon the number of quarter waves on the line, including the "Q" bars. The concentric-line fed antenna (§ 10-8) may be used on harmonics, if the concentric line is air-insulated. Its operation on harmonics is similar to that of the "Q." This antenna is not recommended for multiband operation with a solid-dielectric line, however.

The delta-match system (§ 10-8) can be used on harmonics, although some standing waves will appear on the line. For that matter, *any* antenna system can be used on harmonic frequencies by tying the feeders together at the transmitter end and feeding the system as a single wire by means of a tuned circuit coupled to the transmitter.

A simple antenna system without feeders, useful for operation on five bands, is shown in Fig. 1033. On all bands from 3.5 Mc. upward it operates as an end-fed antenna — half wave on 3.5 Mc., long wire on the other bands. On 1.75 Mc. it is only a quarter wave in length, and must be worked against ground. On this band, since it is fed at a high-current point, series tuning (§ 10-6) must be used.

Antennas for restricted space — If the space available for the antenna is not large enough to accommodate the length necessary for a half wave at the lowest frequency to be used, quite satisfactory operation can be secured by using a shorter antenna and making up the missing length in the feeder system. The antenna itself may be as short as a quarter wavelength and still radiate fairly well, although of course it will not be as effective as one a half wave long. Nevertheless, such a system is useful where operation on the desired band otherwise would be impossible.

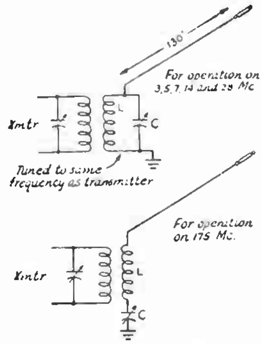


Fig. 1033 — A simple antenna system for five amateur bands. The antenna is voltage fed on 3.5, 7, 14 and 28 Mc., working on the fundamental, second, fourth and eighth harmonics, respectively. For 1.75 Mc. the system is a quarter-wave grounded antenna, in which case series tuning must be used. The antenna wire should be kept well in the clear and should be as high as possible. If the length of the antenna is increased to approximately 260 feet, voltage feed can be used on all five bands.

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 and, since lengths midway between those requiring series or parallel tuning ordinarily must be used to bring the entire system to resonance, coupling to the transmitter often becomes difficult.

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. Typical cases are shown in Fig. 1034, one for

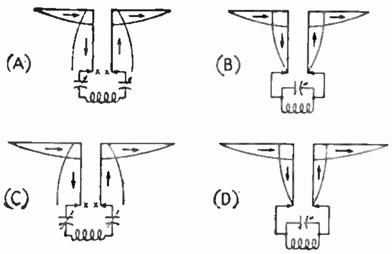


Fig. 1034 — Current distribution on short antennas. Those at the left are too short for fundamental operation, one (A) having an over-all length of one quarter-wave; the other (C) being longer but not a half wave long. These systems may be used wherever space to erect a full half-wave antenna is not available. The current distribution for second-harmonic operation is shown at the right of each figure (B and D). In A and C, the total length around the system is a half wave at the fundamental. In B and D, the over-all length is a full wave. Arrows show the instantaneous direction of current flow.

an antenna having a length of one quarter-wave (A) and the other for an antenna somewhat longer (C) but still not a half wave long. Current distribution is shown for both fundamental and second harmonic. From the points marked X, resonant feeders any convenient number of quarter waves in length may be extended to the operating room. The sum of the distances on each wire from X to the antenna end must equal a half wave. It is sufficiently accurate to use Equation 2 (§ 10-2) in calculating this length. Note that X-X is a high-current point on these shortened antennas, corresponding to the center of a half-wave antenna. It is also apparent that the antenna at A is a half-wave antenna on the next higher-frequency band (B).

A practical antenna of this type can be made as shown in Fig. 1035. Table III gives a few recommended lengths. Remembering the preceding discussion, however, the antenna can be made any convenient length, provided the feeder is considered to "begin" at X-X and the line length is adjusted accordingly.

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 (§ 10-1). Advantage can be taken of this fact when the space available does not permit erecting an antenna a half wave long. In this case the ends may be bent, either horizontally or vertically, so that the total length equals a half wave, even though the straightaway horizontal length may be as short as a quarter wave.

TABLE III

Antenna and Feeder Lengths for Short Multiband Antennas, Center Fed

Antenna Length (ft.)	*Feeder Length (ft.)*	*Band*	*Type of Tuning*
100	38	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	parallel series series series or parallel
67.5	34	3.5 Mc. 7 Mc. 14 Mc. 28 Mc.	series parallel parallel parallel
50	43	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel
33	51	7 Mc. 14 Mc. 28 Mc.	parallel parallel parallel
33	31	7 Mc. 14 Mc. 28 Mc.	parallel series parallel

The operation is illustrated in Fig. 1036. Such an antenna will be a somewhat better radiator than the arrangement of Fig. 1034-A 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,

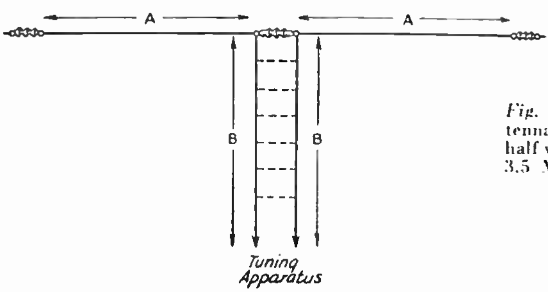


Fig. 1035 — Practical arrangement of a shortened antenna. The total length, A + B + B + A, should be a half wavelength for the lowest-frequency band, usually 3.5 Mc. See Table III for lengths and tuning data.

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.

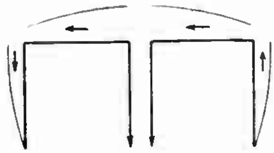
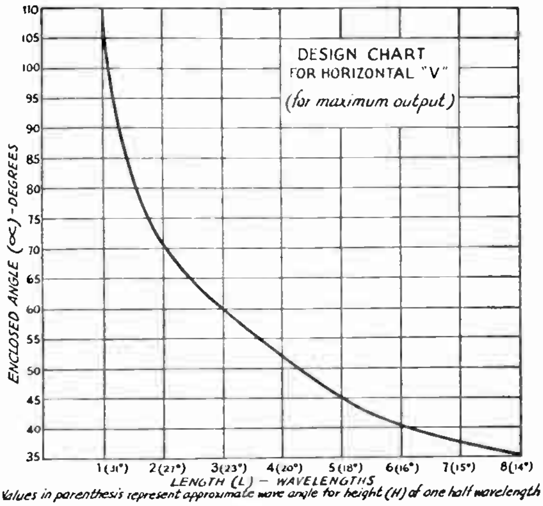


Fig. 1036 — Folded arrangement for shortened antennas. The total length is a half wave, not including the feeders. The horizontal part is made as long as convenient and the ends dropped down to make up the required length. The ends may be bent back on themselves like feeders to cancel radiation partially. The horizontal section should be at least a quarter wave long.



Values in parenthesis represent approximate wave angle for height (H) of one half wavelength

Fig. 1038 — Design chart for horizontal "V" antennas, giving the enclosed angle between sides *vs.* the length of the wires.

ℂ 10-11 Long-Wire Directive Arrays

The "V" antenna — It has been emphasized that, as the antenna length is increased, the lobe of maximum radiation makes a more acute angle with the wire (§ 10-9). Two such wires may be combined in the form of a horizontal "V" so that the main lobes from each wire will reinforce along a line bisecting the angle between the wires. This increases both gain and directivity, since the lobes in directions other than along the bisector cancel to a greater or lesser extent. The horizontal "V" antenna therefore transmits best in either direction (is bidirectional) along a line bisecting the "V" made by the two wires. The power gain depends upon the length of the wires. Provided the necessary space is available, the "V" is a simple antenna to build and operate. It can also be used on harmonics, so that it is suitable for multiband work. The "V" antenna is shown in Fig. 1037.

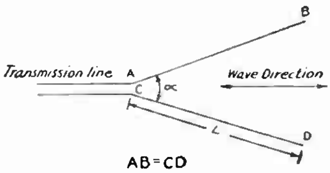


Fig. 1037 — The "V" antenna, made by combining two long wires in such a way that each reinforces the radiation from the other. The important quantities are the length of each leg and the angle between the legs.

Fig. 1038 shows the dimensions that should be followed for an optimum design to obtain maximum power gain for different-sized "V" antennas. The longer systems give good performance in multiband operation. Angle ∝ is approximately equal to twice the angle of maximum radiation for a single wire equal in length to one side of the "V."

The wave angle referred to in Fig. 1038 is the vertical angle of maximum radiation (§ 10-1). 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. 1027. 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 8-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. 1037. Alternatively, a quarter-wave matching section may be employed and the antenna fed through a nonresonant line (§ 10-8). If the antenna wires are made multiples of a half wave in length (use Equation 7 in § 10-9, for computing the length), the matching section will be closed at the free end.

The rhombic antenna — The horizontal rhombic or "diamond" antenna is shown in Fig. 1039. Like the "V," it requires a good 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. 1039, 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. 1039. 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. 1040 gives design information based on a given length and wave angle to determine the

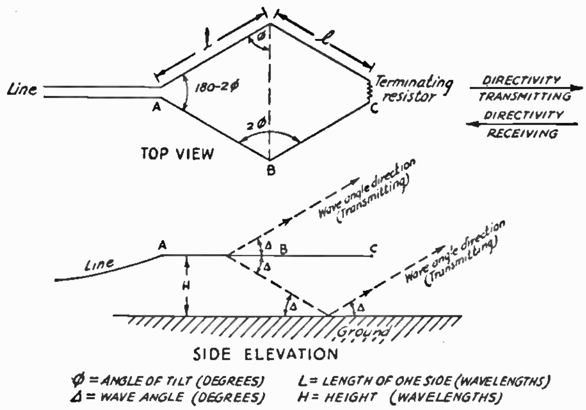


Fig. 1039 — **The horizontal rhombic or diamond antenna, terminated. Important design dimensions are indicated; details in text.**

The same design details apply to the unterminated rhombic as to the terminated type. When used without a terminating resistor, the system is bidirectional. Resonant feeders are preferable for the unterminated rhombic. A nonresonant line may be used by incorporating a matching section at the antenna, but is not readily adaptable to 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.

remaining optimum dimensions for best operation. Curves for values of length of 2, 3 and 4 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 the 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.

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.

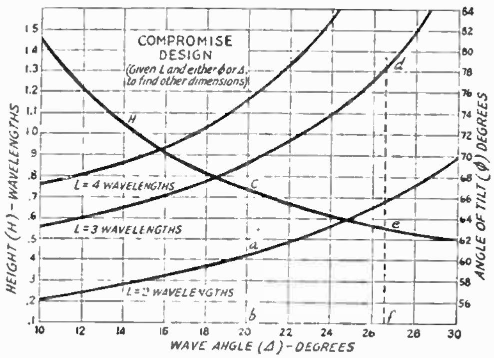


Fig. 1040 — **Compromise-method design chart for rhombic antennas of various leg lengths and wave angles. The following examples illustrate the use of the chart:**

(1) Given:

Length (L) = 2 wavelengths.
Desired wave angle (Δ) = 20°.

To Find: H, Φ.

Method:

Draw vertical line through point *a* (L = 2 wavelengths) and point *b* on abscissa (Δ = 20°). Read angle of tilt (Φ) for point *a* and height (H) from intersection of line *ab* at point *c* on curve H.

Result:

Φ = 60.5°.
H = 0.73 wavelength.

(2) Given:

Length (L) = 3 wavelengths.
Angle of tilt (Φ) = 78°.

To Find: H, Δ.

Method:

Draw a vertical line from point *d* on curve L = 3 wavelengths at Φ = 78°. Read intersection of this line on curve H (point *e*) for height, and intersection at point *f* on the abscissa for Δ.

Result:

H = 0.56 wavelength.
Δ = 26.6°.

10-12 Directive Arrays with Driven Elements

Principles — 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 which 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 which 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. 1041. 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. 1041-B. Note that quarter-wave transmission lines are used between each element; 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. Another way of looking at it is to consider that the whole system is a long wire, with alternate half-wave sections folded so that they do not radiate. Any phase-reversing section may be used as a quarter-wave matching section for attaching a nonresonant feeder (§ 10-8), or a resonant transmission line may be substituted for any of the quarter-wave sections. Also, the antenna may be end fed by any of the systems previously described (§ 10-7, 10-8), or any element may be center fed. It is best to feed at the center of the array, so that the energy will be distributed as uniformly as possible among the elements.

TABLE IV

Theoretical Gain of Collinear Half-Wave Antennas

Spacing between centers of adjacent half waves	*Number of half waves in array vs. gain in db.*				
	2	3	4	5	6
½ Wave	1.8	3.3	4.5	5.3	6.2
¾ Wave	3.2	4.8	6.0	7.0	7.8

The gain and directivity depend upon the number of elements and their spacing, center-to-center. This is shown by Table IV. Although ¾-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 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. 1042 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. 1043. Half-wave spacing generally is used, since it simplifies the problem of feeding the system when the array has more than two elements. Table V gives theoretical gain as a function of the number of elements with half-wave spacing.

Broadside arrays may be suspended either with the elements all vertical or with them horizontal and one above the other (*stacked*). In the former case the horizontal pattern becomes quite sharp, while the vertical pattern is the same as that of one element alone. If the array is suspended horizontally, the horizontal pattern is equivalent to that of one element

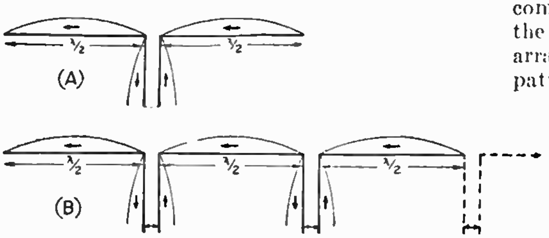


Fig. 1041 — Collinear half-wave antennas in phase. The system at A is generally known as "two half waves in phase." B is an extension of the system; in theory the number of elements may be carried on indefinitely, but practical considerations usually limit the elements to four.

TABLE V	
THEORETICAL GAIN *vs.* NUMBER OF BROADSIDE ELEMENTS (HALF-WAVE SPACING)	
No. of elements	*Gain*
2	4 db.
3	5.5 db.
4	7 db.
5	8 db.
6	9 db.

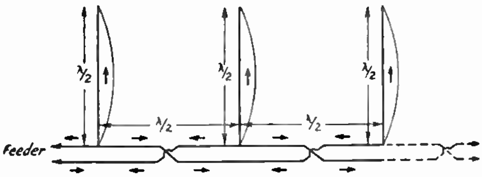


Fig. 1042 — Broadside array using parallel half-wave elements. Arrows indicate the direction of current flow. Transposition of the feeders is necessary to bring the antenna currents in phase. Any reasonable number of elements may be used. The array is bidirectional, with maximum radiation "broadside" or perpendicular to the antenna plane (perpendicularly through this page).

while the vertical pattern is sharpened, giving low-angle radiation.

Broadside arrays may be fed either by resonant transmission lines (§ 10-7) or through quarter-wave matching sections and non-resonant lines (§ 10-8). In Fig. 1042, note the "crossing over" of the feeders, which is necessary to bring the elements in proper phase relationship.

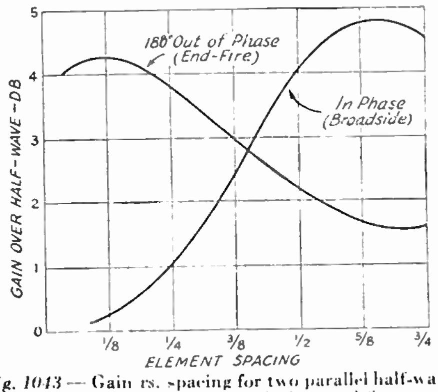


Fig. 1043 — Gain *vs.* spacing for two parallel half-wave elements combined as either broadside or end-fire arrays.

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. 1044. 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.

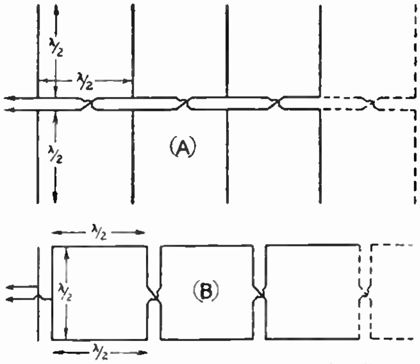


Fig. 1044 — Combination broadside and collinear arrays. A, with vertical elements; B, with horizontal elements. Both arrays give low-angle radiation. Two or more sections may be used. The gain in db. will be equal, approximately, to the sum of the gain for one set of broadside elements (Table V) plus the gain of one set of collinear elements (Table IV). For example, in A each broadside set has four elements (gain 7 db.) and each collinear set two elements (gain 1.8 db.), giving a total gain of 8.8 db. In B, each broadside set has two elements (gain 4 db.) and each collinear set three elements (gain 3.3 db.), making the total gain 7.3 db. The result is not strictly accurate, because of mutual coupling between the elements, but is good enough for practical purposes.

The arrays in Fig. 1044 are shown fed from one end, but this is not especially desirable in the case of large arrays. Better distribution of energy between elements, and hence better all-around performance, will result when the feeders are attached as nearly as possible to the center of the array. Thus, in the 8-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. 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. 1044-B, known as the "lazy H" antenna, has been quite frequently used. This arrangement is shown, with the feed point indicated, in Fig. 1045.

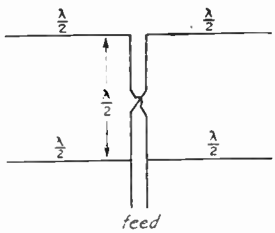


Fig. 1045 — A four-element combination broadside-collinear array, popularly known as the "lazy H" antenna. A closed quarter-wave stub may be used at the feed point to match into a 600-ohm transmission line, or resonant feeders may be attached at the point indicated. The gain over a half-wave antenna is 5 to 6 db.

End-fire arrays — Fig. 1046 shows a pair of parallel half-wave elements with currents out of phase. This is known as an *end-fire* array, because it radiates best along the line of the antennas, as shown.

The end-fire array may be used either 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. 1043 shows how the gain varies with spacing. End-fire elements may be combined with additional collinear and broadside elements to give a further increase in gain and directivity.

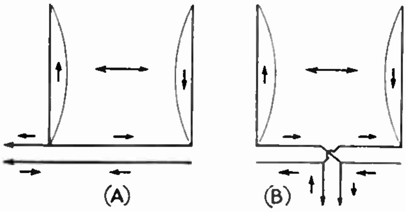


Fig. 1046 — End-fire arrays using parallel half-wave elements. The elements are shown with half-wave spacing to illustrate feeder connections. In practice, closer spacings are desirable, as shown by Fig. 1043. Direction of maximum radiation is shown by the large arrows.

Either resonant or nonresonant lines may be used with this type of array. Nonresonant lines preferably are matched to the antenna through a quarter-wave matching section (§ 10-8).

Checking phasing — Figs. 1044 and 1046 illustrate a point in connection with feeding a phased antenna system which sometimes is confusing. In Fig. 1046, when the transmission line is connected as at A there is no cross-over in the line connecting the two antennas, but when the transmission line is connected to the center of the connecting line the cross-over 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. 1044. Note that, under these conditions, the antenna elements are in phase when the line is not transposed, and out of phase when the transposition is made. The opposite is the case when the half-wave line simply joins two antenna elements and does not have the feed line connected to its center, as in Fig. 1042.

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 the equation for a half-wave antenna in § 10-2, while the half-wave phasing lines between the parallel elements can be calculated from the formula:

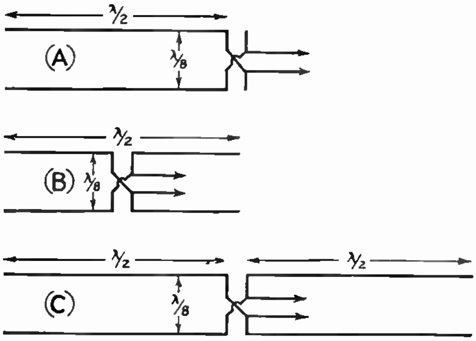
$$\text{Length of half-wave line (feet)} = \frac{492 \times 0.975}{\text{Freq. (Mc.)}} = \frac{480}{\text{Freq. (Mc.)}}$$

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 carefully.

With collinear arrays of the type shown in Fig. 1041-B, 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:

$$\text{Length of quarter-wave line (feet)} = \frac{240}{\text{Freq. (Mc.)}}$$

If the array is fed at its 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 on each phasing section and moving the link back and forth to find the maximum current position. This refinement is hardly necessary in practice, however, so long as all elements are the same length and the system is symmetrical.



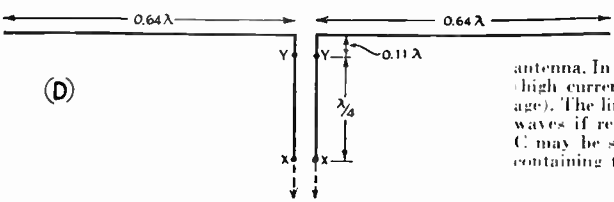


Fig. 1047 — Simple directive-antenna systems. A is a two-element end-fire array; B is the same array with center feed, which permits use of the array on the second harmonic, where it becomes a four-element array with quarter-wave spacing. C is a four-element end-fire array with 1/8-wave spacing. D is a simple two-element broadside array using extended in-phase antennas ("extended double-Zepp"). The gain of A and B is slightly over 4 db. On the second harmonic, B will give about 5 db. gain. With C, the gain is approximately 6 db., and with D, approximately 3 db. In A, B and C, the phasing line contributes about 1/16 wavelength to the transmission line; when B is used on the second harmonic, this contribution is 1/8 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. 1047. Tuned feeders are assumed in all cases; however, a matching section (§ 10-8) 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 in § 10-2 for the antenna and from the equation above for the resonant transmission line or matching section. In cases where the transmission-line proper connects to the midpoint of a phasing line, only *half* the length of the latter should be added to the line to find the quarter-wave point.

At A and B are two-element end-fire arrangements using close spacing. They are electrically equivalent; the only difference is in the method of connecting the feeders. B may also be used as a four-element array on the second harmonic, although the spacing is not quite optimum (Fig. 1043) 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 quite 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.

ℭ 10-13 Directive Arrays with Parasitic Elements

Parasitic excitation — The antenna arrays described in § 10-12 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 (for instance, north only, instead of north-south), 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.

The parasitic element is called a *director* when it reinforces radiation on a line pointing to it from the antenna, and a *reflector* when the reverse is the case. Whether the parasitic element is a director or reflector depends upon the parasitic element tuning (which usually is adjusted by changing its length) and, particularly when the element is self-resonant, upon the spacing between it and the antenna.

Gain vs. spacing — The gain of an antenna-reflector or an antenna-director combination varies chiefly with the spacing between the elements. The way in which gain varies with spacing is shown in Fig. 1048, for the special case of self-resonant parasitic elements. This chart also shows how the attenuation to the "rear" varies with spacing. The same spacing does not necessarily give both maximum forward gain and maximum backward attenuation. Backward attenuation is desirable when the antenna is used for receiving, since it greatly reduces interference coming from the opposite direction to the desired signal.

Element lengths — The antenna length is given by the formulas in § 10-2. The director and reflector lengths must be determined experimentally for maximum performance. The preferable method is to aim the antenna at a receiver a mile or more distant and have an observer check the signal strength (on the receiver "S" meter) while the reflector or director is adjusted a few inches at a time, until the length which gives maximum signal is found. The attenuation may be similarly checked, the length being adjusted for minimum signal. In general, for best front-to-back ratio the length of a director will be about 4 per cent less than that of the antenna. The reflector will be about 5 per cent longer than the antenna.

Simple systems: the rotary beam — Four practical combinations of antenna, reflector and director elements are shown in Fig. 1049. Spacings which give maximum gain or maximum front-to-back ratio (ratio of power radiated in the desired direction to power radiated in the opposite direction) may be taken from Fig. 1048. In the chart, the front-to-back ratio in db. will be the sum of gain and attenuation at the same spacing.

Systems of this type are popular for rotary-beam antennas, where the entire antenna system is rotated, to permit its gain and directivity to be utilized for any compass direction. They may be mounted either horizontally (with the plane containing the elements parallel to the earth) or vertically.

Arrays using more than one parasitic element, such as those shown at C and D in Fig. 1049, will give more gain and directivity than is indicated for a single reflector and director by the curves of Fig. 1048. The gain with a properly-adjusted three-element array (antenna, director and reflector) will be 5 to 7 db. over a half-wave antenna. Somewhat higher gain still can be secured by adding a second director to the system, making a four-element array. The front-to-back ratio is correspond-

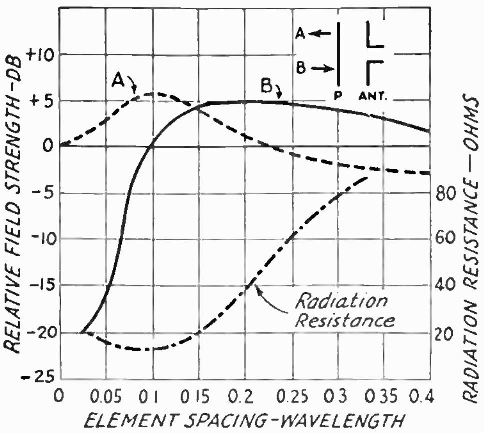


Fig. 1048 — Gain *vs.* element spacing for an antenna and one parasitic element. The reference point, 0 db., is the field strength from a half-wave antenna alone. The greatest gain is in direction *A* at spacings of less than 0.14 wavelength, and in direction *B* at greater spacings. The front-to-back 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.

ingly improved as the number of elements is increased.

The elements in close-spaced (less than one-quarter wavelength element spacing) arrays preferably should be made of tubing of one-half to one-inch diameter. A conductor of large diameter not only has less ohmic re-

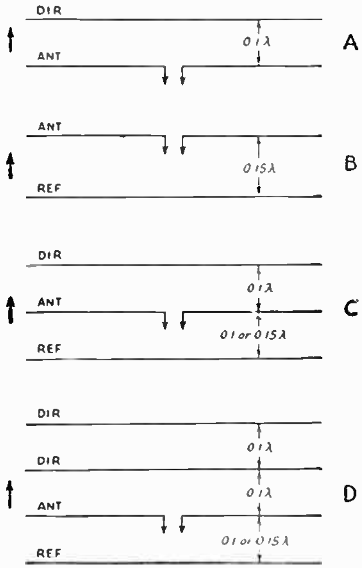


Fig. 1049 — Half-wave antennas with parasitic elements. A, with director; B, with reflector; C, with both director and reflector; D, two directors and one reflector. Gain is approximately as shown by Fig. 1048, in the first two cases, and depends upon the spacing and length of the parasitic element. In the three- and four-element arrays a reflector spacing of 0.15 wavelength will give slightly more gain than 0.1-wavelength spacing. Arrows show the direction of maximum radiation.

sistance but also has lower *Q;* both these factors are important in close-spaced arrays because the impedance of the driven element usually is quite low compared to that of a single half-wave dipole. With 3- and 4-element arrays the radiation resistance of the driven element may be as low as 6 or 8 ohms, so that ohmic losses in the conductor can consume an appreciable fraction of the power. Low radiation resistance means that the antenna will work over only a small frequency range without retuning unless large-diameter conductors are used. In addition, the antenna elements should be rigid because if they are free to move with respect to each other, the array will tend to show detuning effects under windy conditions.

Feeding close-spaced arrays — While any of the usual methods of feed may be applied to the driven element of a parasitic array, the fact that, with close spacing, the radiation resistance as measured at the center of the driven element drops to a very low value makes some systems more desirable than others. The preferred methods are shown in Fig. 1050. Resonant feeders are not recommended for lengths greater than a half wavelength.

The quarter- or half-wave matching stubs shown at A and B in Fig. 1050 preferably should be constructed of tubing with rather close spacing, in the manner of the "Q" section. This lowers the impedance of the matching section and makes the position of the line taps somewhat less difficult to determine accurately. The line adjustment should be made only with the parasitic elements in place, and after the correct element lengths have been determined, it should be checked to compensate for changes likely to occur because of element tuning. The procedure is the same as that described in § 10-8.

The concentric-line matching section at C will work with fair accuracy into a close-spaced parasitic array of 2, 3 or 4 elements without necessity for adjustment. The line is used as an impedance-inverting transformer, and, if its characteristic impedance is 70 ohms, it will give an exact match to a 600-ohm line when the resistance at the termination is about 8.5 ohms. Over a range of 5 to 15 ohms the mismatch, and therefore the standing-wave ratio, will be less than 2 to 1. The length of the quarter-wave section may be calculated from Equation 5 (§ 10-5).

The delta matching transformer shown at D is probably easier to install, mechanically, than any of the others. The positions of the taps (dimension *a*) must be determined experimentally, along with the length, *b*, by checking the standing-wave ratio on the line as adjustments are made. Dimension *b* should be about 15 per cent longer than *a*.

The system shown at E ("T" match) resembles the delta match in principles of operation. It has the advantage that, with close spacing between the two parallel conductors,

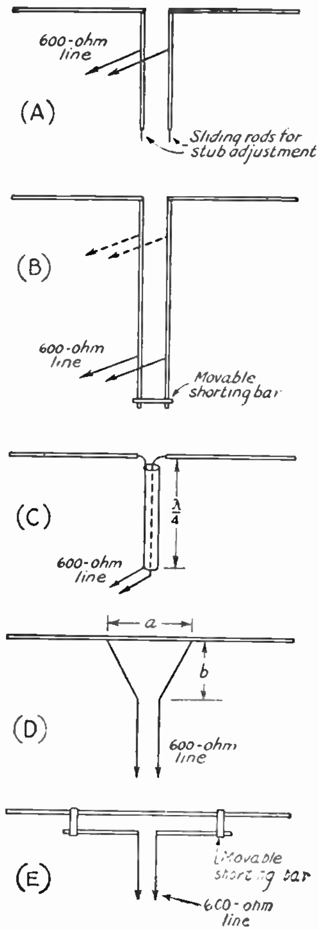


Fig. 1050 — Recommended methods of feeding the driven antenna element in close-spaced parasitic arrays. The parasitic elements are not shown. A, quarter-wave open stub; B, half-wave closed stub; C, concentric-line quarter-wave matching section; D, delta matching transformer; E, "T" matching transformer. Adjustment details are discussed in the text.

line radiation from the matching section is negligible whereas radiation from a delta may be considerable. It is adjusted by moving the shorting bars, keeping them equidistant from the center, until there are no standing waves on the line. The matching section may be made of the same type of conductor used for the driven element and spaced a few inches from it.

The "folded dipole" shown in Fig. 1051 may be used as the driven element of a close-spaced parasitic array to secure an impedance step-up to the transmission line and also to broaden the resonance curve of the antenna. The folded dipole consists of two or more half-wave antennas connected together at the ends with the feeder connected to the center of only one of the antennas. The spacing between the parallel antennas should be small — of the order of the spacing used between wires of a transmission line. The current in the system divides in approximate proportion to the areas

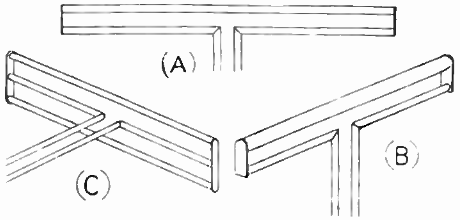


Fig. 1051 — Various forms of folded dipole. In calculating the element lengths, the total length around any loop starting with the transmission-line terminals should equal one wavelength (twice the length given by the appropriate formula, in view of conductor diameter, in § 10-2) so that the lengths of the connecting bars at the ends are included.

of the conductors, resulting in an impedance step-up at the input terminals. With two similar conductors (equal areas) the impedance step-up is 4 to 1; if there are three similar conductors (or if the one not connected to the transmission line has twice the diameter of the other) the step-up is 9 to 1; if the ratio of the areas is 3 to 1 the step-up is 16 to 1, and so on. Thus if a 3-conductor dipole (all conductors the same diameter) is used as the driven element of a four-element parasitic array the center impedance of approximately 8 ohms is multiplied by 9 and appears as approximately 72 ohms at the input terminals. Such a system therefore can be fed directly from a 70-ohm line with no additional means for matching.

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 (§ 10-2). 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 IV, the gain of four collinear elements is 4.5 db. with half-wave spacing; from Fig. 1043 or Table V, the gain of two broadside elements at half-wave spacing is 4.0 db.; from Fig. 1048, the gain of a parasitic reflector at quarter-wave spacing is 4.5 db. The total gain is then the sum, or 13 db. for the sixteen elements. Note that using two *sets* of elements in broadside is equivalent to using two elements, so far as gain is concerned; similarly with sets of reflectors, as against one antenna and one reflector. The actual gain of the combination array will depend, in practice, upon the way in which the power is distributed between the various elements and upon the effect which mutual coupling between elements has upon the radiation resistance of the array, and may be somewhat higher or lower than the estimate.

A great many directive antenna combinations can be worked out by combining elements according to these principles.

Receiving Antennas

Nearly all of the properties possessed by an antenna as a radiator also apply when it is used for reception. Current and voltage distribution, impedance, resistance and directional characteristics are the same in a receiving antenna as if it were used as a transmitting antenna. This reciprocal behavior makes possible the design of a receiving antenna of optimum performance based on the same considerations that have been discussed for transmitting antennas.

The simplest receiving antenna is a wire of random length. The longer the wire, the more energy it abstracts from the wave. Because of the high sensitivity of modern receivers, a large antenna is not necessary for picking up signals at good strength. An indoor wire only 15 to 20 feet long will serve at frequencies below the v.h.f. range, although a longer wire outdoors is better.

The use of a tuned antenna improves the operation of the receiver, however, because the signal strength is raised more in proportion to the stray noises picked up than is the case with wires of random length. Since the transmitting antenna usually is given the best 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. A change-over switch or relay, connected in the antenna leads, can be used to transfer the connections from the receiver to the transmitter.

In selecting a directional receiving antenna it is preferable to choose a type which 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.

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 of ordinary soft-drawn No. 14 or No. 12 enameled copper wire, since hard-drawn or copper-clad steel wire is difficult to handle unless it is under considerable tension at all times. The wires should be all in one piece; where a joint cannot be avoided, it should be carefully soldered.

In building a resonant two-wire feeder, the spacer insulation should be of as good quality as in the antenna insulators proper. For this reason, good ceramic spacers are advisable. Wooden dowels boiled in paraffin may be used with untuned lines, but their use is not recommended for tuned lines. The wooden dowels can be attached to the feeder wires by drilling small holes and binding them to the feeders with wire.

At points of maximum voltage insulation is most important, and Pyrex glass, Isolantite or steatite insulators with long leakage paths are recommended for the antenna. Glazed porcelain also is satisfactory. Insulators should be cleaned once or twice a year, especially if they are subjected to much smoke and soot.

In most cases poles or masts are desirable to lift the antenna clear of surrounding buildings, although in some locations the antenna will be sufficiently in the clear when strung from one chimney to another or from a chimney to a tree. Small trees usually are not satisfactory as points of suspension for the antenna because of their movement in windy weather. If the antenna is strung from a point near the center of the trunk of a large tree, this difficulty is not so serious. Where the antenna wire must be strung from one of the smaller branches, it is best to tie a pulley firmly to the branch and run a rope through the pulley to the antenna, with the other end of the rope attached to a counterweight near the ground.

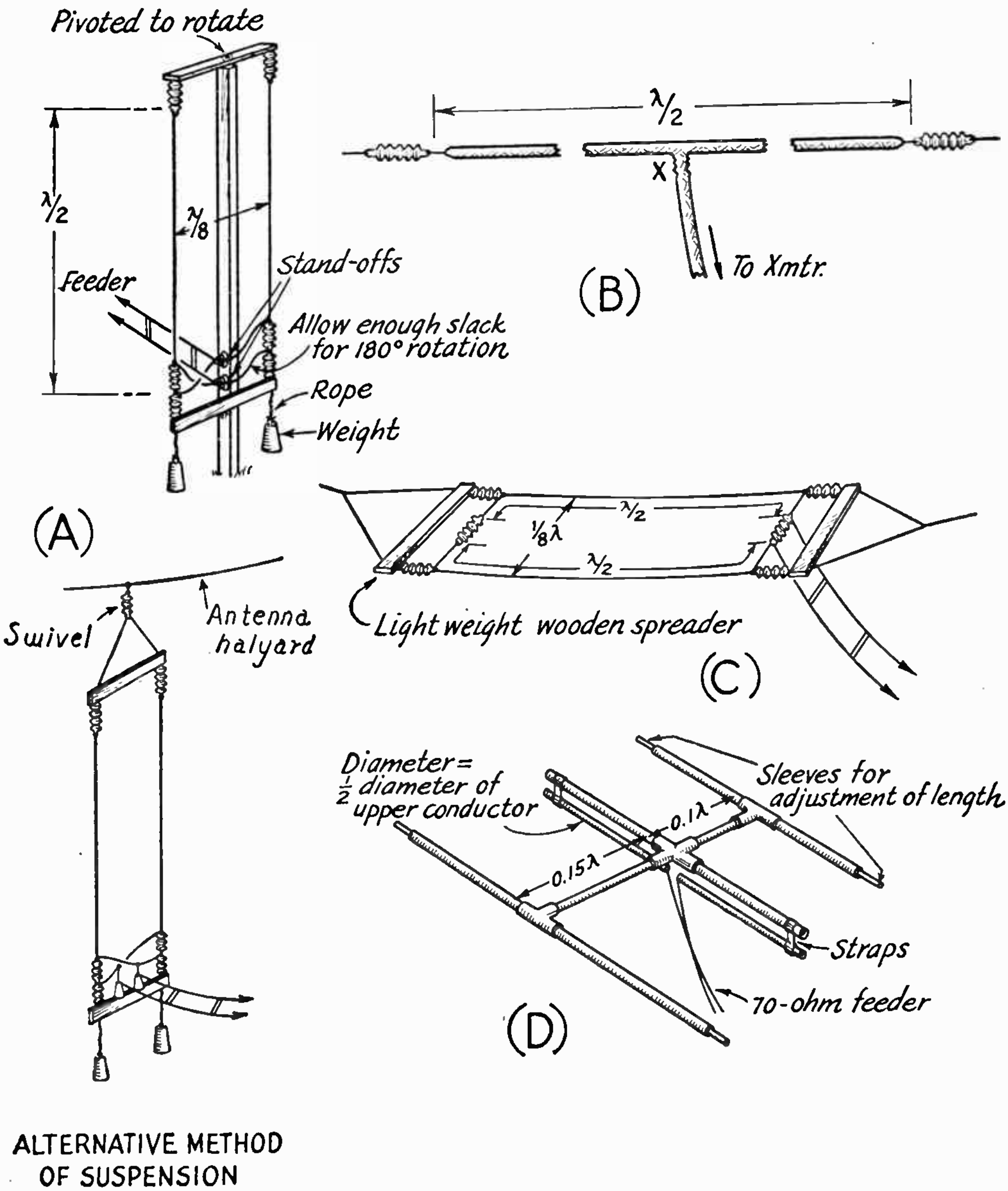
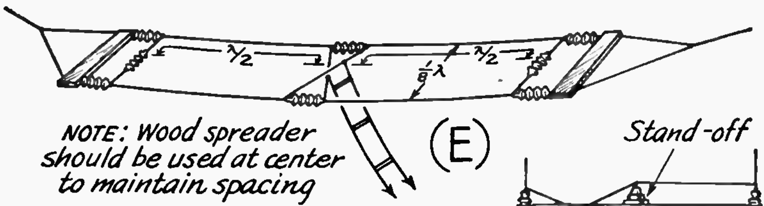
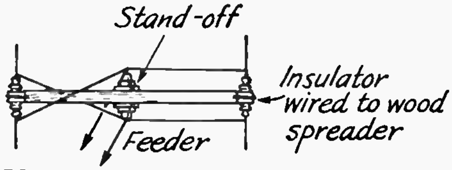


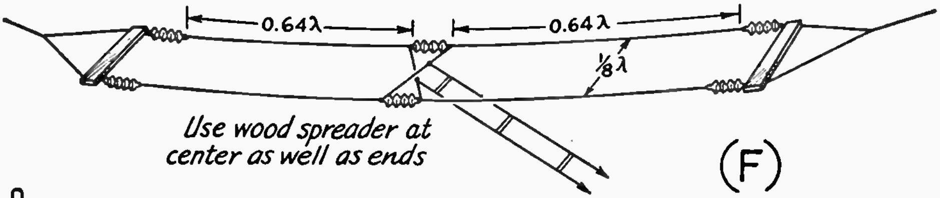
Fig. 1052 — Some suggested antenna systems. A — Simple bidirectional rotatable end-fire array using 1/8-wave spacing between out-of-phase elements. Suitable for either 11 or 28 Mc. and can be rotated by hand. It can also be suspended from the halyard holding another antenna, as suggested in the lower drawing. B — Folded dipole using 300-ohm Twin-Lead for both antenna and feeder. The junction X at the center is made by opening one conductor of the antenna section and soldering to the feeder leads. The joint may be made mechanically firm by heating the dielectric with a soldering iron, using extra bits of dielectric for a good bond. C — An end-fire array for use where space is limited. The ends of the two half-wave elements are folded to meet at an insulator in the center. The antenna may be made still shorter by increasing the spacing: spacings up to 1/4 wavelength may be used. D — Pipe-assembly three-element beam ("plumber's delight") with folded-dipole driven element. Because all three elements are at the same r.f. potential at their centers it is possible to join them electrically as well as mechanically with no effect on the performance. Provision is made for adjusting the element lengths for optimum performance at a given frequency (§10–13). E — An extension of the folding principle shown in C. The collinear in-phase elements give additional gain and directivity. F — End-fire array with extended double Zepps. This antenna should give a gain of about 7 db. in the direction perpendicular to the line of the antenna. G — An 8-element array combining broadside, end-fire and collinear elements. The gain of an antenna of this type is about 10 db. This antenna also can be used at half the frequency for which it is designed. H — Using two half-wave antennas at right angles to change direction. With the three feeders indicated, either antenna alone can be fed as a Zepp and will radiate best perpendicular to its



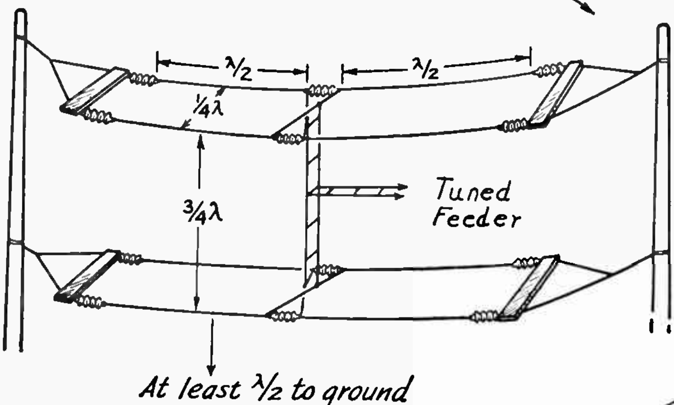
(E)



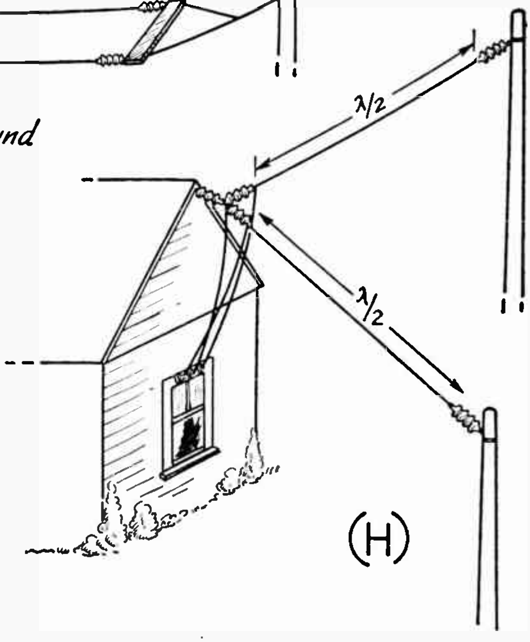
SUGGESTED CONSTRUCTION AT CENTER



(F)



(G)



(H)

direction. By feeding the two together, leaving the third feeder wire idle, the optimum direction is the bisector of the angle between the wires. This system is most useful at high frequencies such as 14 Mc. and above.

In these drawings, wavelength dimensions on conductors refer to lengths calculated for the conductor size as described in §10–2. Dimensions between elements are free-space dimensions.

The feeders to the various directive systems in A, C, E, F and G must be tuned if used as shown. For one-band operation, matching stubs (§10–8) may be attached to the feeders if a matched line is desired.

The counterweight will keep the tension on the antenna wire reasonably constant even when the branches sway or the rope tightens and stretches with varying climatic conditions.

"A"-Frame Mast

The simple and inexpensive mast shown in Fig. 1053 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.

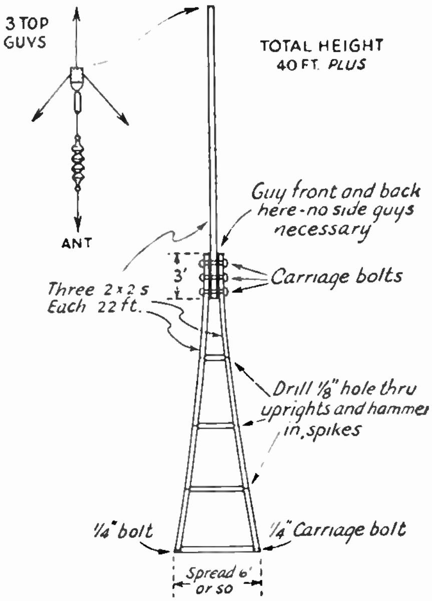


Fig. 1053 — Details of a simple 40-foot "A"-frame mast suitable for erection in locations where space is limited.

Simple 40-Foot Mast

The mast shown in Fig. 1054 is relatively strong, easy to construct, readily dismantled, and costs very little. Like the "A" frame, it is suitable for heights of the order of 40 feet.

The top section is a single 2 × 3, bolted at the bottom between a pair of 2 × 3s with an overlap of about two feet. The lower section thus has two legs spaced the width of the narrow side of a 2 × 3. At the bottom the two legs are bolted to a length of 2 × 4 which is set in the ground. A short length of 2 × 3 is placed between the two legs about half way up the bottom section, to maintain the spacing.

The two back guys at the top pull against the antenna, while the three lower guys prevent buckling at the center of the pole.

The 2 × 4 section should be set in the ground so that it faces the proper direction, and then made vertical by lining it up with a plumb bob. The holes for the bolts should be drilled beforehand. With the lower section laid on the ground, bolt *A* should be slipped in place through the three pieces of wood and tightened just enough so that the section can turn freely on the bolt. Then the top section may be bolted

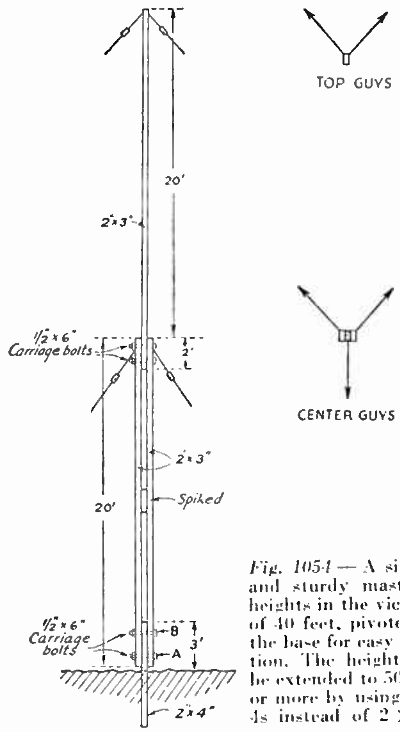


Fig. 1054 — A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 × 4s instead of 2 × 3s.

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.

"T"-Section Mast

A type of mast suitable for heights up to about 80 feet is shown in Fig. 1055. The mast is built up by butting 2 × 4 or 2 × 6 timbers edgewise against a second 2 × 4, as shown at A, with alternating joints in the edgewise and

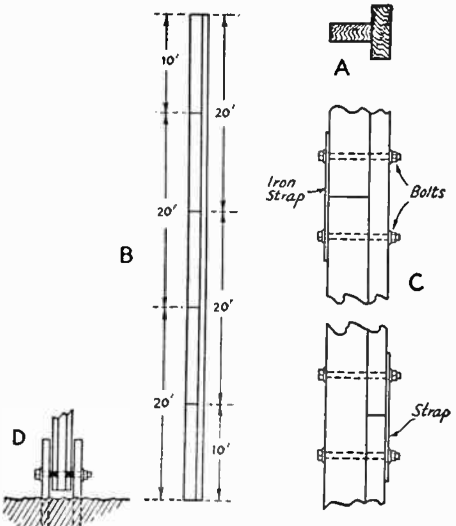


Fig. 1055 — "T"-section mast made of overlapping 2 × 4s or 2 × 6s.

flatwise sections. The construction can be carried out to greater lengths simply by continuing the 20-foot sections. Longer or shorter sections may be used, if more convenient.

The method of making the joints is shown at C. Quarter-inch or 3/16-inch iron, 1½ to 2 inches wide, is recommended for the straps, with ½-inch bolts to hold the pieces together. One bolt should be run through the pieces midway between joints, to provide additional rigidity.

Although there are many ways in which such a mast can be secured at the base, the "cradle" illustrated at D has many advantages. Heavy timbers set firmly in the ground, spaced far enough apart so the base of the mast will pass between them, hold a large carriage bolt or steel bar which serves as a bearing. This bolt goes through a hole in the mast so that it is pivoted at the bottom.

Half of the guys can be put in place and tightened up before the mast leaves the ground. Four sets of guys should be used, one in front, one directly in the rear, and two on each side at right angles to the direction in which the mast will face. A set of guys should be used at each of the joints in the edgewise sections, the guy wires being wrapped around the pole for added strength.

For heights up to 50 feet, 2 × 4-inch members may be used throughout. For greater heights, use 2 × 6s for the edgewise sections; 2 × 4-inch pieces will do for the flat sections.

Pole and Tower Supports

Poles, which often may be purchased at a reasonable price from the local telephone or power company, have the advantage that they do not require guying unless they are called upon to carry a very heavy load. The life of a pole can be extended many years by proper precautions before erecting, and regular maintenance.

Before setting the pole, it should be given four or five coats of creosote, applying it liberally so it can soak into and preserve the wood. The bottom of the pole and the part which will be buried in the ground should have a generous coating of hot pitch, poured on while the pole is warm. This will keep termites out and prevent rotting.

The pole should be set in the ground four to eight feet depending upon the height. It is a good idea to pour concrete around the bottom three feet of the base, packing the rest of the excavation with soil. The concrete will help hold the pole against strong winds. After filling the hole with dirt, a stream from a hose should be played on the dirt slowly for several hours. This will help to settle the soil quickly.

If desired, the pole may be extended by the arrangement shown in Fig. 1056. Three 2 × 4s are required for the top section, two being 18 feet long and one 10 feet long. The 10-foot section is placed between the other two and bolted in place. A half-inch hole should be bored through the pole about 2 feet from its top and through both 18-foot 2 × 4s about 5 feet from their bottom ends, which are spread apart to fit the top of the pole. The bottom end of the extension is then hauled up to the top of the pole and bolted loosely so that the section can be swung up into place by the leverage of another 2 × 4 temporarily fastened to the section, as shown in Fig. 1056.

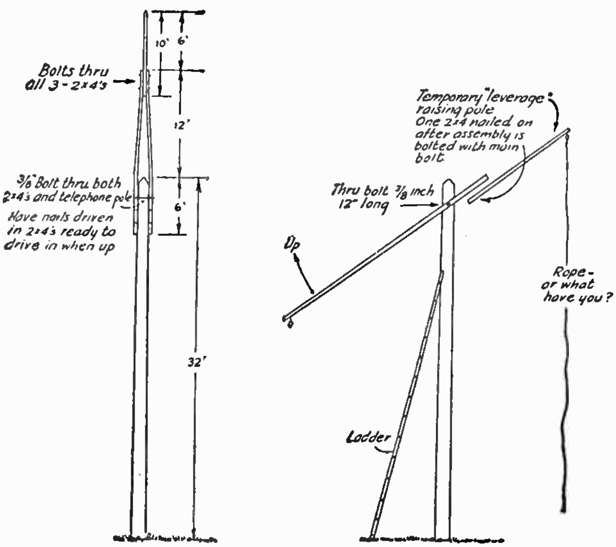


Fig. 1056 — This type of mast may be carried to a height of fifty feet or more. No guy wires are required.

Lattice towers built of wood should be assembled with brass screws and casein glue, rather than with nails which work loose in a short time. A tower constructed in this manner will give trouble-free service if treated with a coat of paint every year.

In painting outside structures, use pure white lead, thinned with three parts of pure linseed oil to one part of turpentine, for the first coat on new wood. The use of a drier is not recommended if the paint will possibly dry without it, since it may cause the paint to peel after a short time. For the second and third coats pure white lead thinned only with pure linseed oil is recommended. Plenty of time for drying should be allowed between coats. White paint will last fifty per cent longer than any colored paint.

Guys and Guy Anchors

For masts or poles up to about 50 feet, No. 12 iron wire is a satisfactory guy-wire material. Heavier wire or stranded cable may be used for taller poles or poles installed in locations where the wind velocity is high.

More than three guy wires in any one set usually are unnecessary. If a horizontal antenna is to be supported, two guy wires in the top set will be sufficient in most cases. These should run to the rear of the mast about 100 degrees apart to offset the pull of the antenna. 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 which 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. 1057 can be made

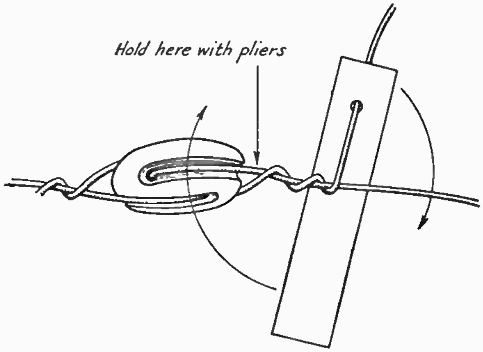


Fig. 1057 — Using a lever for twisting heavy guy wires.

from a piece of heavy iron or steel by drilling a hole about twice the diameter of the guy wire about a half inch from one end of the piece. The wire is passed through the insulator, given a single turn by hand, and then held with a pair of pliers at the point shown in the sketch. By passing the wire through the hole in the iron and rotating the iron as shown, the wire may be quickly and neatly twisted.

Guy wires may be anchored to a tree or building when they happen to be in convenient spots. For small poles, a 6-foot length of 1-inch pipe driven into the ground at an angle will suffice. Additional bracing will be provided by using two pipes, as shown in Fig. 1058.

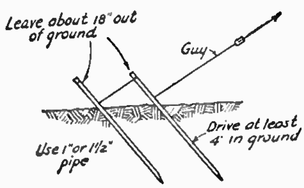


Fig. 1058 — Pipe guy anchors. One pipe is sufficient for small masts, but two installed as shown will provide the additional strength required for the larger poles.

Halyards and Pulleys

Halyards or ropes and pulleys are important items in the antenna-supporting system. Particular attention should be directed toward the choice of a pulley and halyards for a high mast since replacement, once the mast is in position, may be a major undertaking if not entirely impossible.

Galvanized-iron pulleys will have a life of only a year or so. Especially for coastal-area installations, marine-type pulleys with hardwood blocks and bronze wheels and bearings should be used.

An arrangement which has certain advantages over a pulley when a mast is used is shown in Fig. 1059. In case the rope breaks, it may be possible to replace it by heaving a line over the brass rod, making it unnecessary to climb or lower the pole.

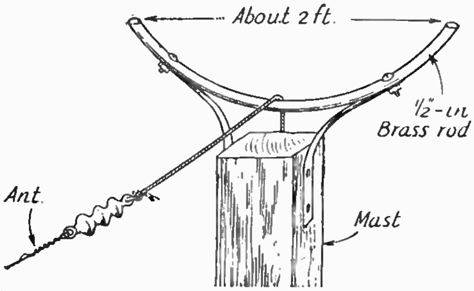


Fig. 1059 — This device is much easier than a pulley to "rethread" when the rope breaks.

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.

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.

ℂ Bringing the Antenna or Transmission Line into the Station

The antenna or transmission line should be anchored to the outside wall of the building, as shown in Fig. 1060, 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 which 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. 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. 1060-B may be used.

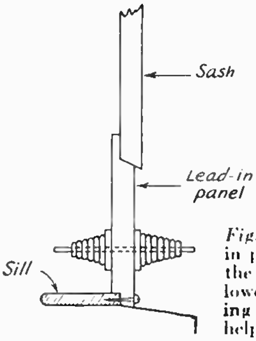


Fig. 1061 — An antenna lead-in panel may be placed over the top sash or under the lower sash of a window. Sealing the overlapping joint will help make it weatherproof.

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. 1061, and covering the opening between sashes with a sheet of soft rubber from a discarded inner tube.

ℂ Lightning Protection

An ungrounded radio antenna, particularly if large and well elevated, is a lightning hazard. When grounded, it provides a measure of protection. Therefore, grounding switches or lightning arresters should be provided. Examples of construction of low-loss arresters are shown in Fig. 1062. 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

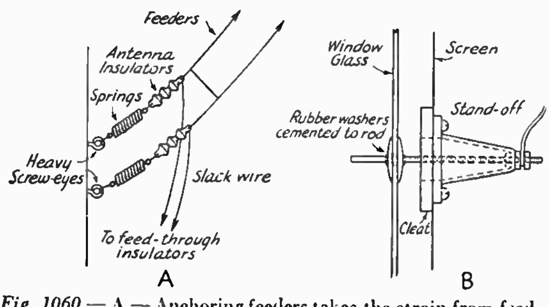


Fig. 1060 — A — Anchoring feeders takes the strain from feed-through insulators or window glass. B — Going through a full-length screen, a cleat is fastened to the frame of the screen on the inside. Clearance holes are cut in the cleat and also in the screen.

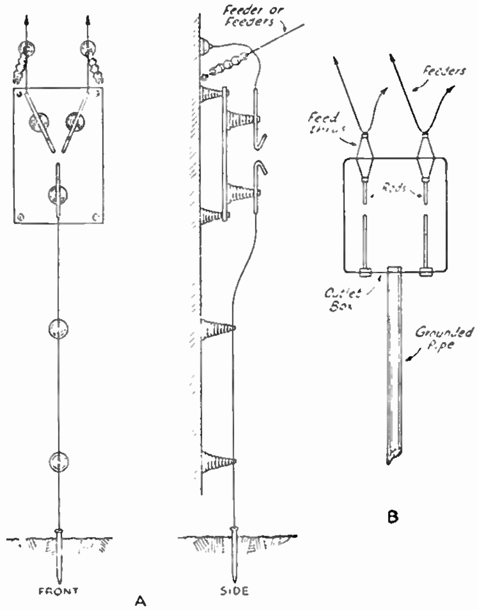


Fig. 1062 — Low-loss lightning arresters for transmitters.

the feeder wires may be used. The ground lead should be short and run, if possible, directly to a driven pipe or water pipe where it enters the ground outside the building.

Antenna Switching

It is often desirable, particularly in DX work, to use the same antenna for transmitting and receiving. This requires switching of antenna from transmitter to receiver. One of two general systems may be employed. In the first, the transmitter and receiver each are provided with an antenna tuner, and the antenna transmission line is switched from one to the other. In the second system, one antenna tuner is provided for each antenna and the switch is in the low-impedance coupling line. Several typical arrangements are shown in Fig. 1063. Frequently relays with low-capacity contacts are substituted for switches.

Rotary-Beam Construction

It is a distinct advantage to be able to shift the direction of a beam antenna at will, thus securing the benefits of power gain and directivity in any desired compass direction. A favorite method of doing this is to construct the antenna so that it can be rotated in the horizontal plane. Obviously, the use of such rotatable antennas is limited to the higher frequencies — 14 Mc. and above — and to the simpler antenna element combinations if the structure size is to be kept within practicable bounds. For the 14- and 28-Mc. bands such antennas usually consist of two to four elements and are of the parasitic-array type described earlier in this chapter. At 50 Mc. and higher it becomes possible to use more elaborate arrays because of the shorter wavelength and thus obtain still higher gain. Antennas for these bands are described in Chapter Seventeen.

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 di-

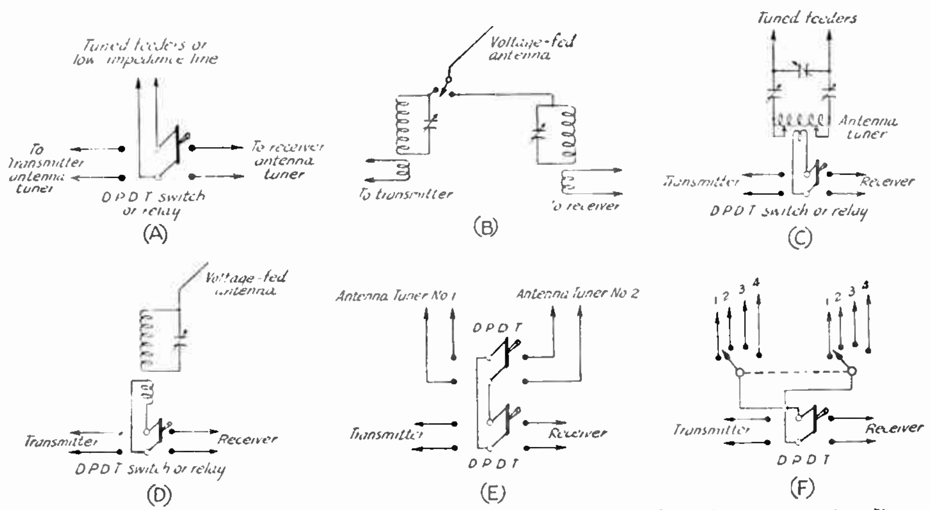


Fig. 1063 — Antenna-switching arrangements for various types of antennas and coupling systems. A — For tuned lines with separate antenna tuners or low-impedance lines. B — For a voltage-fed antenna. C — For a tuned line with a single antenna tuner. D — For a voltage-fed antenna with a single tuner. E — 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.

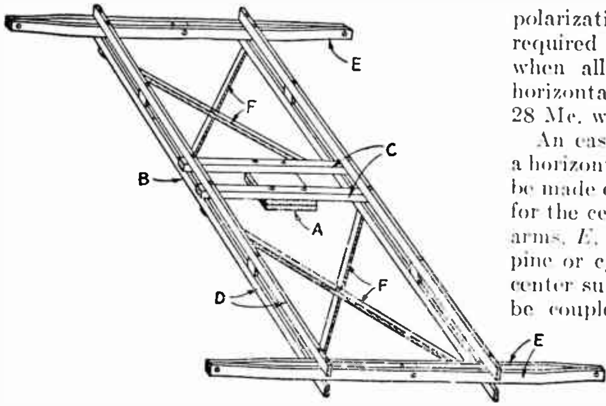


Fig. 1064 — Easily-built supporting structure for horizontal rotary beams. Made chiefly of 1 × 2″ wood strip, it is strong yet lightweight. Antenna elements are supported on stand-off insulators on the arms, *E*. The length of the *D* sections will depend upon the element spacing, while the length of the *E* sections and the spacing between the *D* sections should be 1/4 to 1/2 the length of the antenna elements.

ameter 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. Electricians' 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, using as lightweight construction as is consistent with the required strength. Generally, the frame is not required to hold much weight, but it must be extensive enough so that the antenna elements can be supported near enough to their ends to prevent excessive sag, and it must have sufficient strength to stand up under the maximum wind in the locality. The design of the frame will depend chiefly on the size of the antenna elements, whether they are mounted horizontally or vertically, and the method to be used for rotating the antenna.

The general preference is for horizontal polarization, primarily because less height is required to clear surrounding obstructions when all the antenna elements are in the horizontal plane. This is important at 14 and 28 Mc. where the elements are fairly long.

An easily-constructed supporting frame for a horizontal array is shown in Fig. 1064. It may be made of 1 × 2-inch lumber, preferably oak, for the center sections *B*, *C*, and *D*. The outer arms, *E*, and crossbraces, *F*, may be of white pine or cypress. The square block, *A*, at the center supports the whole structure and may be coupled to the pole by any convenient means that permits rotation. The bearing shown in Fig. 1068, for example, may easily be modified for the purpose. Alternatively, the block may be firmly fastened to the pole and the latter rotated in bearings affixed to the side of the house.

Another type of construction is shown in Fig. 1065, with details in Figs. 1066 and 1067. This method, suitable for 28-Mc. beams, uses a section of ordinary ladder as the main support, with crosspieces to hold the tubing antenna elements. Fig. 1066 also indicates a method of adjusting the lengths of the parasitic elements and bringing the transmission line down through the supporting pole from a delta match. The latter is especially adapted to construction in which the pole rather than the framework alone is rotated.

The problem of feeding a parasitic array is somewhat simplified if the elements are mounted vertically, since in such a case it is not necessary to rotate the driven element but only to rotate the parasitic elements around it. Thus no special provision need be made for maintaining contact to the feeders through a complete rotation. A suitable method of construction is shown in Fig. 1068.

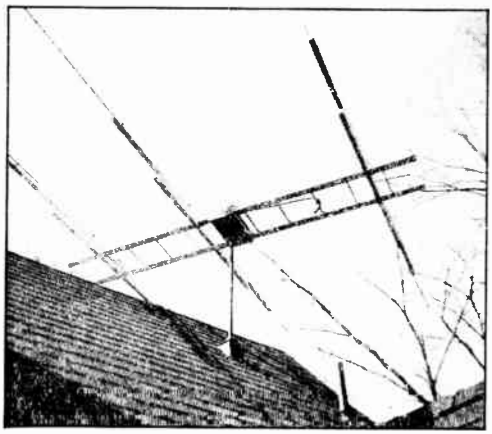


Fig. 1065 — A ladder-supported 3-element 28-Mc. beam. It is mounted on a pipe mast that projects through a bearing in the roof and is turned from the attic operating room. (W1MRK in August, 1946, *QST*.)

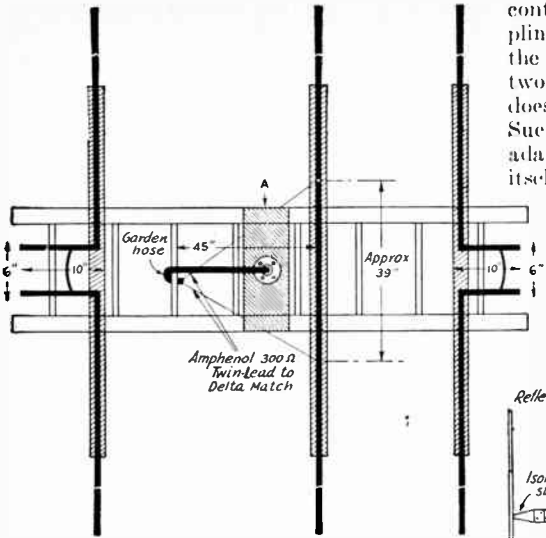


Fig. 1066 — Top-view drawing of the ladder support and mounted elements. Lengths of director and reflector are adjusted by means of the shorting bars on the small stubs at the center. The drawing also shows a method for pulling off the wires of a delta match and feeding 300-ohm Twin-Lead transmission line through the pipe support.

Feeder connections — For beams which rotate only 180 degrees, it is relatively simple to bring off feeders by making a short section of the feeder, just where it leaves the rotating member, of flexible wire. Enough slack should be left so that there is no danger of breaking or twisting. Stops should be placed on the rotating shaft of the antenna so that the feeders cannot "wind up." This method also can be used with antennas which rotate the full 360 degrees, but again a stop is necessary to avoid jamming the feeders.

For continuous rotation, the sliding contact is simple and, when properly built, quite practicable. Fig. 1069 shows two methods of making sliding contacts. The chief points to keep in mind are that the contact surfaces should be wide enough to take care of wobble in the rotating shaft, and that the contact surfaces should be kept clean. Spring contacts are essential, and an "umbrella" or other scheme for keeping rain off the contacts is a desirable addition. Sliding contacts preferably should be used with nonresonant open lines where the impedance is of the order of 500 to 600 ohms so that the current is low.

The possibility of poor connections in sliding contacts can be avoided by using inductive coupling at the antenna, with one coil rotating on the antenna and the other fixed in position, the two coils being arranged so that the coupling does not change when the antenna is rotated. Such an arrangement is shown in Fig. 1070, adapted to an antenna system in which the pole itself rotates. A quarter-wave feeder system is connected to a tuned pick-up circuit whose inductance is coupled to a link. In the drawing, the link coil connects to a twisted-pair transmission line, but any type of line such as flexible coaxial cable can be used. The circuit would be adjusted in the same way as any link-

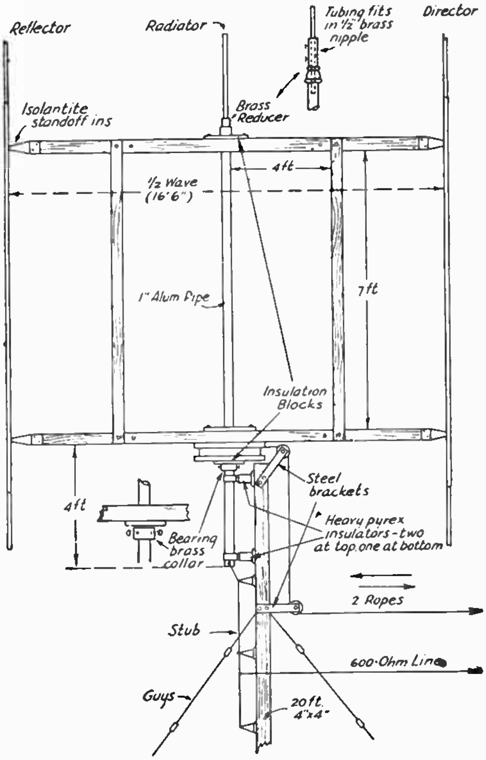


Fig. 1068 — A practical vertical-element rotatable array for 28 Mc. The driven antenna is fixed and the reflector and director elements, parasitically excited, rotate around it. Close-spaced elements may be used if desired.

coupled circuit, and the number of turns in the link should be varied to give proper loading on the transmitter. The rotating coupling circuit of course tunes to the transmitting frequency. The whole thing is equivalent to a link-coupled antenna tuner mounted on the pole, using a parallel-tuned tank at the end of a quarter-wave line to center feed the antenna. To maintain constant coupling, the two coils should be quite rigid and the pole should rotate without wobble. The

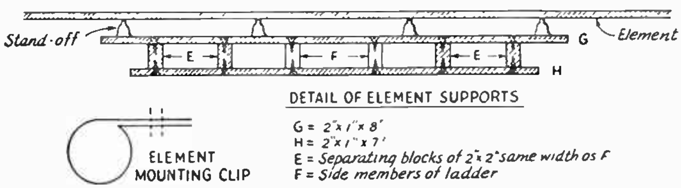


Fig. 1067 — Detail of element supports for the ladder beam.

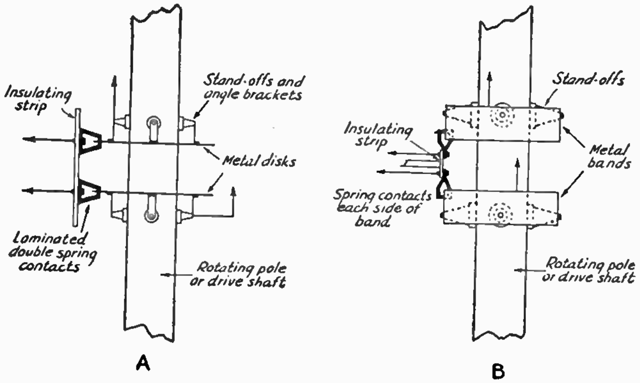


Fig. 1069 — Ideas in sliding contacts for rotatable antenna feeder connection to permit continuous rotation. The broad bearing surfaces take care of any wobble in the rotating mast or driving shaft.

two coils might be made a part of the upper bearing assembly holding the rotating pole in position.

Other variations of the inductive-coupled system can be worked out. The tuned circuit might, for instance, be placed at the end of a 600-ohm line, and a one-turn link used to couple directly to the center of the antenna, if the construction of the rotary member permits. In this case the coupling can be varied by changing the L/C ratio in the tuned circuit. For mechanical strength the coils preferably should be made of copper tubing, well braced with insulating strips to keep them rigid.

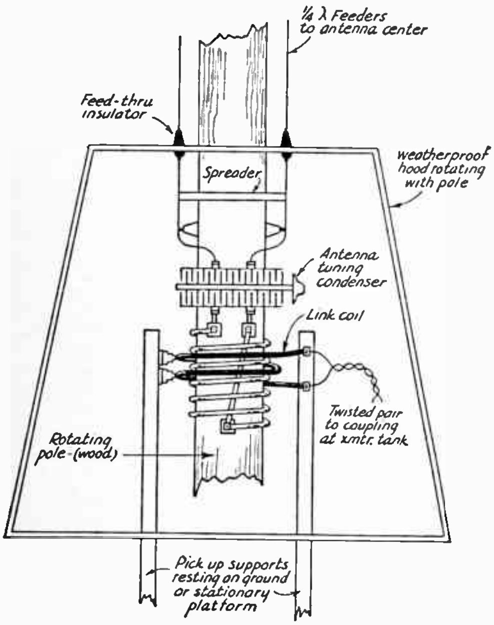


Fig. 1070 — One method of transmission line–antenna system coupling which eliminates sliding contacts. The low-impedance line is link-coupled to a tuned line.

Rotation — It is convenient to use a motor to rotate the beam, but it is not always necessary, especially if a rope and pulley arrangement such as that shown in Fig. 1068 can be brought into the operating room. If the pole can be mounted near a window in the operating room, hand rotation of the beam will work out quite well. If the use of a rope and pulleys is impracticable, motor drive is about the only alternative. The speed of rotation should not be too great — 1 or 2 r.p.m. is about right. This requires a considerable gear reduction from the usual 1750-r.p.m. speed of small induction motors; a large reduction is advantageous because the gear train will prevent the beam from turning in weathervane fashion in a wind. The ordinary structure does not require a great deal of power for rotation at slow speed, and a ⅛-h.p. motor will be ample. Even small series motors of the sewing-machine type will develop enough power to turn a 28-Mc. beam at slow speed. If possible, a reversible motor should be used so 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 top of the pole, but if the pole rotates the motor can usually be installed in a more accessible location.

Parts from junked automobiles often provide gear trains and bearings for rotating the antenna. Rear axles, in particular, can readily be adapted to the purpose. Driving motors and gear housings will stand the weather better if given a coat of aluminum paint followed by two coats of enamel and a coat of glyptal varnish. Even commercial units will last longer if treated with glyptal varnish.

Lead-sheathed twin-conductor cable is recommended for power wiring to the motor to prevent r.f. pick-up. It will also reduce "hash" if a series-wound motor is used. With such motors it is wise to install r.f. filters at the motor terminals as an additional precaution against interference to reception, since it is usual practice to determine the proper direction for the beam by rotating it while listening to the station it is desired to work and setting it at the point that gives maximum signal strength.

Chapter Eleven

Workshop Practice

Tools

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. A few additional tools will make certain operations easier, so it is a good idea for the amateur who does constructional work to add to his supply of tools from time to time. The following list will be found helpful in making a selection:

Bench vise, 4-inch jaws.
Tin shears, 10-inch, for cutting thin sheet metal.
Taper reamer, 1/2-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, cross-cut.
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, 1/2-inch.
Cold chisel, 1/2-inch.
Wing dividers, 8-inch, for scribing circles.
Set of machine-screw taps and dies.
Folding rule, 6-foot.
Dusting brush.

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.

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 well-kept sharp-edged tools. A few minutes spent now and then with the oil stone or emery wheel will maintain the fine cutting edges of knives, drills, chisels, etc.

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 oil-stoning of the cutting edges of a drill or reamer will extend the time between grindings. Stoned cutting edges also will stand more feed and speed.

The soldering iron can be kept in good condition by keeping the tip well tinned with solder and not allowing it to run at full voltage for long periods when it is not being used. After each period of use, the tip should be removed and cleaned of any scale which may have accumulated. An oxidized tip may be cleaned by dipping it in sal ammoniac while hot and then wiping it clean with a rag. If the tip becomes pitted, it should be filed until smooth and bright, and then tinned by dipping it in solder.

All tools should be wiped occasionally with an oily cloth to prevent rust.

INDISPENSABLE TOOLS

Long-nose pliers, 6-inch.
Diagonal cutting pliers, 6-inch.
Screwdriver, 6 to 7-inch, 1/4-inch blade.
Screwdriver, 4 to 5-inch, 1/8-inch blade.
Scratch awl or scriber for marking lines.
Combination square, 12-inch, for laying out work.
Hand drill, 1/4-inch chuck or larger, 2-speed type preferable.
Electric soldering iron, 100 watts.
Hacksaw, 12-inch blades.
Center punch for marking hole centers.
Hammer, ball peen, 1-lb. head.
Heavy knife.
Yardstick or other straight-edge.
Carpenter's brace with adjustable hole cutter or socket-hole punches (see text).
Pair of small C-clamps for holding work.
Large, coarse, flat file.
Large round or rat-tail file, 1/4-inch diameter.
Three or four small and medium files—flat, round, half-round, triangular.
Drills, particularly 1/4-inch and Nos. 18, 28, 33, 42 and 50.
Combination oil stone for sharpening tools.
Solder and soldering paste (noncorroding).
Medium-weight machine oil.

Useful Materials

Small stocks of various miscellaneous materials will be required in constructing radio apparatus, most of which are available from hardware or radio supply stores. A representative list follows:

½ × 1/16-inch brass strip for brackets, etc. (half-hard for bending).
¼-inch square brass rod or ½ × ½ × 1/16-inch angle brass for corner joints.
¼-inch diameter round brass rod for shaft extensions.
Machine screws: Round-head and flat-head, with nuts to fit. Most useful sizes: 4-36, 6-32 and 8-32, in lengths from ¼ inch to 1½ inches. (Nickel-plated iron will be found satisfactory except in strong r.f. fields, where brass should be used.)
Bakelite and hard-rubber scraps.
Soldering lugs, panel bearings, rubber grommets, terminal-lug wiring strips, varnished-cambric insulating tubing.

Machine screws, nuts, washers, soldering lugs, etc., are most reasonably purchased in quantities of a gross.

Chassis Construction

With a few essential tools and proper procedure, it will be found that building radio gear on a metal chassis is no more of a chore than building with wood, and a more satisfactory job results.

The placing of components on the chassis is shown quite clearly in the photographs in this *Handbook*. Aside from certain essential dimensions, which usually are given in the text, exact duplication is not necessary.

Much trouble and energy can be saved by spending sufficient time in planning the job. When all details are worked out beforehand the actual construction is greatly simplified.

Cover the top of the chassis with a piece of wrapping paper or, preferably, cross-section paper, folding the edges down over the sides of the chassis and fastening with adhesive tape. Then assemble the parts to be mounted on top of the chassis and move them about until a satisfactory arrangement has been found, keeping in mind any parts which are to be mounted underneath, so that interferences in mounting may be avoided. Place condensers and other parts with shafts extending through the panel first, and arrange them so that the controls will form the desired pattern on the panel. Be sure to line up the shafts squarely with the chassis front. Locate any partition shields and panel brackets next, and then the tube sockets and any other parts, marking the mounting-hole centers of each accurately on the paper. Watch out for condensers whose shafts are off center and do not line up with the mounting holes. Do not forget to mark the centers of socket holes and holes for leads under i.f. transformers, etc., as well as holes for wiring leads.

By means of the square, lines indicating accurately the centers of shafts should be extended to the front of the chassis and marked on the panel at the chassis line, the panel being fastened on temporarily. The hole centers may then be punched in the chassis with the center punch. After drilling, the parts which require mounting underneath may be located and the mounting holes drilled, making sure by trial that no interferences exist with parts mounted on top. Mounting holes along the front edge of the chassis should be transferred to the panel, by once again fastening the panel to the chassis and marking it from the rear.

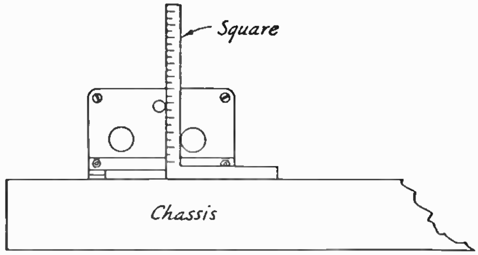


Fig. 1101 — Method of measuring the heights of condenser shafts, etc. If the square is adjustable, the end of the scale should be set flush with the face of the head.

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. 1101. 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.

Cutting and Bending Sheet Metal

If a sheet of metal is too large to be cut conveniently with a hacksaw, 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 so far that 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.

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. Care should be taken not to use too much pressure with small drills, which bend or break easily. When the drill starts to break through, special care must be used. Often it is an advantage to shift a two-speed drill to low gear at this point. Holes more than ¼-inch in diameter may be started with a smaller drill and reamed out with the larger drill.

The chuck on the usual type of hand drill is limited to ¼-inch drills. Although it is rather tedious, the ¼-inch hole may be filed out to larger diameters with round files. Another method possible with limited tools is to drill a series of small holes with the hand drill along the inside of the diameter of the large hole, placing the holes as close together as possible. The center may then be knocked out with a cold chisel and the edges smoothed up with a file. Taper reamers which fit into the carpenter's brace will make the job easier. A large rat-tail file clamped in the brace makes a very good reamer for holes up to the diameter of the file, if the file is revolved counterclockwise.

For socket holes and other large round holes, an adjustable cutter designed for the purpose may be used in the brace. The cutter should be kept well-sharpened. Occasional application of machine oil in the cutting groove will help. The cutter first should be tried out on a block of wood, to make sure that it is set for the correct diameter. Probably the most convenient device for cutting socket holes is the socket-hole punch. The best type is that which works by turning a take-up screw with a wrench.

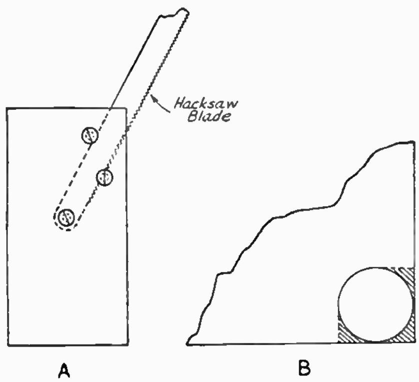


Fig. 1102 — To cut rectangular holes in a chassis, corner holes may be filed out as shown in the shaded portion of B, making it possible to start the hacksaw blade along the cutting line. A shows how a single-ended handle may be constructed for a hacksaw blade.

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. 1102, and using these holes for starting and turning the hacksaw. The socket-hole punch 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.

ℂ 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 below

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	—	2–46
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.

will be the drills most commonly used in construction of amateur radio equipment. It is usually desirable to purchase several of each of the commonly-used sizes rather than a quantity of odd sizes, most of which will be used infrequently, if at all.

Cutting Threads

Brass rod may be threaded, or the damaged threads of a screw repaired, by the use of *dies*. Holes of suitable size (see drill chart) may be threaded for screws by means of *taps*. Taps and dies are obtainable in all standard machine-screw sizes. A set usually consists of taps and dies for 4–36, 6–32, 8–32, 10–32 and 14–20 sizes, with a holder suitable for use with either tap or die. The die may be started easily by first filing a sharp taper or bevel on the end of the rod. In tapping a hole, extreme care should be used to prevent breaking the tap. The tap should be kept at right angles to the surface of the material, and rotation should be reversed a revolution or two whenever the tap begins to turn hard. With care, holes can be tapped rapidly by clamping the tap in the chuck of the hand drill and using slow speed. Machine oil applied to the tap usually makes cutting easier and sticking less troublesome.

Crackle Finish

Wood or metal parts can be given a crackle finish by applying one coat of clear Duco or Tri-Seal and allowing it to dry over night. A coat of Kem-Art Metal Finish is then sprayed or applied thickly with a brush, taking care that the brush marks do not show. This should be allowed to dry for two or three hours and the part should then be baked in the kitchen oven at 215 degrees for one-and-one-half hours. This will produce a regular commercial job. This finish, which comes in several different colors, is made by Sherwin-Williams Paint Co.

Cleaning and Finishing Metal

Parts made of aluminum can be cleaned up and given a satin finish, after all holes have been drilled, by placing them in a solution of lye for one-half to three-quarters of an hour. Three or four tablespoonfuls of lye should be used to each gallon of water. If more than one piece is treated in the same bath, each piece should be separated from the others so as to expose all surfaces to the solution. Overlapping of pieces may result in spots or stains.

Wiring

A popular type of wire for receivers and low-power transmitters is that known as "push-back" wire. It comes in sizes No. 16, 18, 20, etc., which are sufficiently large for all power circuits except filament. The insulating covering, which is sufficient for circuits where voltages do not exceed 400 or 500, can be pushed back a few inches at the end, making cutting of the insulation unnecessary when making a connection. Filament wiring should be done with sufficiently large conductors to carry the required current without appreciable voltage drop (see Copper Wire Table, Chapter Twenty). Rubber-covered house-wire sizes No. 14 to No. 10 are suitable for heavy-current transmitting tubes, while No. 18 to No. 14 flexible wire is satisfactory for receivers and low-drain transmitting tubes where the total length of the leads is not excessive.

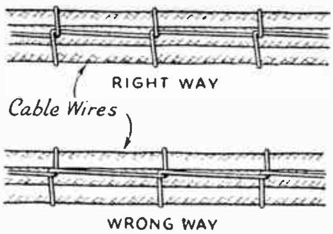


Fig. 1103 — Right and wrong methods of lacing cable. With the right way the leading line is pinched under each turn and will not loosen if a break occurs in the lacing.

Stiff bare wire, sometimes called *bus wire* or *bus bar*, is most favored for the high r.f.-potential wiring of transmitters and, where practicable, in receivers. It comes in sizes No. 14 and No. 12 and is usually tin-dipped. Soft-drawn antenna wire also may be used. Kinks or bends can be removed by stretching 10 or 15 feet of the wire and then cutting it into small usable lengths.

The insulation covering power wiring which is to carry high transmitter voltages should be appropriate for the voltage involved. Wire with rubber and varnished cambric covering, similar to ignition cable, is available from radio parts dealers. The smaller sizes have sufficient insulation to be safe at 1000 to 1500 volts, while the more heavily insulated types should be used for voltages above 1500.

It is usually advisable to do the power-supply wiring first. The leads should be bunched together as much as possible and kept down close to the surface of the chassis. The lacing of power wiring in cable form not only improves its appearance but also strengthens the wiring. Fig. 1103 shows the correct way of lacing cabled wires. When done correctly the leading line is held tightly pinched in place after tension has been removed, and therefore does not loosen readily. When the wrong method is used the turns will loosen up as soon as tension is removed.

Chassis holes for wires should be lined with *rubber grommets* which fit the hole, to prevent chafing of the insulation. In cases where power-supply leads have several branches, it is often convenient to use fiber *terminal strips* as anchorages. These strips also form handy mountings for wire-terminal resistors, etc. When any particular unit is provided with a nut or thumb-screw terminal, soldering-lug wire terminals to fit are useful.

High-voltage wiring should have exposed points kept at a minimum and those which cannot be avoided rendered as inaccessible as possible to accidental contact.

Soldering

The secret of good soldering is in allowing time for the *joint*, as well as the solder, to attain sufficient temperature. Enough heat should be applied so that the solder will melt when it comes in contact with the wires being joined, without touching the solder to the iron.

Wartime solder, which is still with us, has a much smaller ratio of tin to lead, requires considerably more heat, and its use makes it especially important that the iron be kept clean at all times. More care must be exercised in making the joint because this solder does not flow as readily, and also has a tendency to crystallize.

Soldering paste, if of the noncorroding type, is extremely helpful when used correctly. In general, it should not be used for radio work except when necessary. The joint should first be warmed slightly and the soldering paste applied with a piece of wire. Only the bit of paste which melts from the warmth of the joint should be used. If the soldering iron is clean it will be possible with one hand to pick up a drop of solder on the tip of the iron which can be applied to the joint, while the other hand is used to hold the connecting wires together. The use of excessive soldering paste causes the paste to spread over the surface of adjacent insulation, causing leakage or breakdown of the insulation. Except where absolutely necessary, solder should never be depended upon for the mechanical strength of the joint; the wire should be wrapped around the terminals or clamped with soldering terminals.

Do not attempt to make ground connections to a cadmium-plated chassis by soldering to the surface of the chassis, since the plating may be loosened by the heat and later fall off, breaking the connection. Drill a hole in the chassis and solder the wire in the hole.

Construction Notes

Lockwashers should be used under nuts to prevent loosening with use, particularly when mounting tube sockets or plug-in coil receptacles subject to frequent strain.

If a control shaft must be extended or insulated, a flexible shaft coupling with adequate insulation should be used. Satisfactory support for the shaft extension can be provided by means of a *metal* panel bearing made for the purpose. Never use panel bearings of the nonmetal type unless the condenser shaft is grounded. *The metal bearing should be connected to the chassis with a wire or grounding strip.* This prevents any possible danger of shock.

The standard way of mounting toggle switches is with the switch "On" when the lever is in the upward position.

Variable condensers and resistors, having one-hole mountings, should be firmly fastened using the special lockwashers provided for shaft nuts.

The use of fiber washers between ceramic insulation and metal brackets, screws or nuts will prevent the ceramic parts from breaking.

Coil Winding

Dimensions for coils for the various units described in the constructional chapters are given under the circuit diagrams. Where no wire size is given, the power is sufficiently low to permit use of any available size within reason.

Unless a close-wound winding is definitely specified, the number of turns indicated should be spaced out to fill the specified length on the form. The length should be marked on the form and holes drilled opposite the pins to which the ends of the winding are to connect. Scrape one end of the wire and pass it through the lower hole in the form to the pin to which the bottom end of the winding is to connect, and solder this end fast. Unroll a length of wire approximately sufficient for the winding, and clamp the spool in a vise so it will not turn. The wire should be pulled out straight and the winding started by turning the form in the hands and walking toward the vise. A fair tension should be kept on the wire at all times. The spacing can be judged by eye. If, as the winding progresses, it becomes evident that the spacing is going to be incorrect to fill the required length, the winding can be started over again with a different spacing. If the spacing is only slightly off, the winding may be finished, the top end fastened, and the spacing corrected by pushing each turn. When complete, the turns should be fastened in place with coil cement. After a little practice, the job of determining the correct spacing will not be difficult.

Sometimes it is necessary to adjust the number of turns on a coil experimentally. The easiest way to do this is to bring a wire up from one of the pins, extending it through a hole in the form for a half inch or so. The end of the winding may then be soldered to this extension rather than to the pin itself, and the nuisance of repeatedly fishing the wire through the pin avoided until the correct size of the winding has been determined.

Coil Cement

Duco cement, obtainable universally at hardware, stationery or 5-and-10-cent stores, is satisfactory for fastening coil turns. For small coils, a better-looking job will result if it is thinned out with acetone (amyl acetate), sometimes referred to as banana oil. If desired, the solution may be made thin enough to permit application with a brush.

Special low-loss coil "dopes" are available, including some with a polystyrene base.

Receiver Construction

ℂ A Two-Tube Superheterodyne Receiver

Although all the advantages of the superheterodyne-type receiver cannot be secured without going to rather elaborate multitube circuits, it is possible to use the superhet principle to overcome most of the disadvantages of the simple regenerative receiver. These are chiefly the necessity for critical adjustment of the regeneration control with tuning, antenna "dead spots," lack of stability (both in the detector circuit itself and because of slight changes in frequency when the antenna swings with the wind), and blocking, or the tendency for strong signals to pull the detector into zero-beat. These effects can be largely eliminated by making the regenerative detector operate on a fixed low frequency and designing it for maximum stability. The incoming signal is then converted to the fixed detector frequency before being detected.

A two-tube receiver operating on this principle is shown in Figs. 1201 to 1205.

The circuit diagram is given in Fig. 1202. A 6K8 is used to convert the frequency of the incoming signal to the fixed or intermediate frequency, and the two triode sections of a 6SN7 serve as the regenerative detector and audio amplifier respectively. L_1C_1 is the r.f. circuit, tuned to the signal, and L_2 is the antenna coupling coil. C_7 is a by-pass condenser across the 1.5-volt battery used to bias the signal grid of the 6K8. The high-frequency

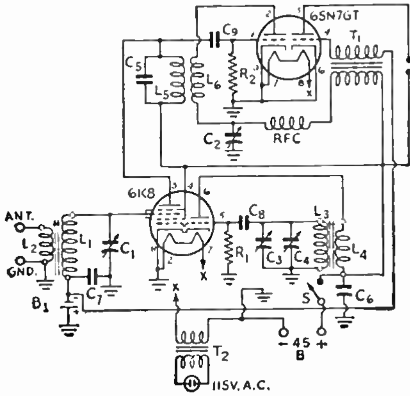


Fig. 1202 — Circuit diagram of the two-tube superheterodyne receiver.

C_1, C_2, C_3 — 100-μμfd. variable (Millen 19100).
C_4 — 15-μμfd. variable (Millen 20015).
C_5 — 240-μμfd. silvered mica.
C_6 — 0.01-μfd. paper.
C_7 — 0.005-μfd. mica.
C_8, C_9 — 100-μμfd. mica.
R_1 — 47,000 ohms, ½ watt.
R_2 — 1 megohm, ½ watt.
L_1, L_2, L_3, L_4 — See coil table.
L_5 — 55 turns No. 30 d.s.c., close-wound on ¾-inch diam. form (National PRF-2); inductance 40 uh.
L_6 — 18 turns No. 30 d.s.c., close-wound on same form as L_5; see Fig. 1203.
B_1 — 1.5-volt bias battery.
J_1 — Open-circuit jack.
RFC — 2.5-mh. r.f. choke.
S — S.p.s.t. toggle switch.
T_1 — Interstage audio transformer (Stancor A-4205).
T_2 — 6.3-volt filament transformer.

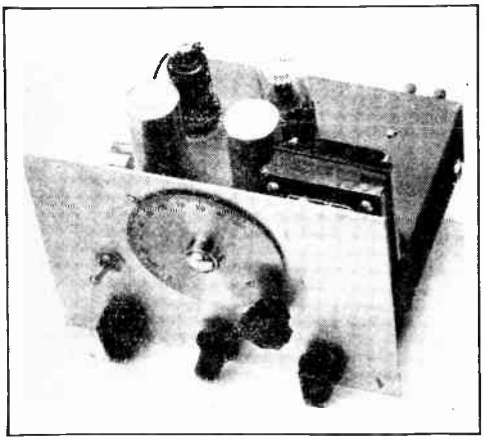


Fig. 1201 — Panel view of the two-tube superheterodyne receiver. The panel is cut from a sheet of ⅛-inch aluminum. It is 6 inches high and 8 inches wide. The controls along the bottom, from left to right, are mixer tuning, oscillator padder and i.f. regeneration. The "B" switch is to the left of the tuning dial.

oscillator tank circuit is $L_3C_3C_4$, with C_3 for band-setting and C_4 for bandspread.

The i.f. tuned circuit (or regenerative detector circuit) is L_5C_5. This must be a high-C circuit if stability better than that of an ordinary regenerative detector is to be secured. The frequency to which it is tuned should be in the vicinity of 1600 kc. L_5 and its tickler coil, L_6, are wound on a small form, and L_5 is tuned by a fixed mica condenser of the low-drift type. Since these condensers are rated with a capacity tolerance of 5 per cent, it is sufficient to wind L_5 as specified under Fig. 1203. The resulting resonant frequency will be in the correct region. No manual tuning is necessary, and therefore the frequency of this circuit need not be adjusted. C_2 is the regeneration-control condenser, isolated from the d.c. supply by the choke, RFC. Only enough turns need be used on L_6 to make the detector oscillate readily when C_2 is at half capacity or more.

The second section of the 6SN7 is trans-

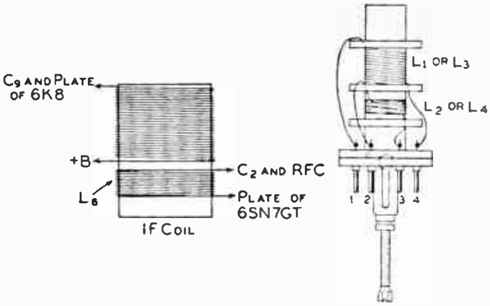


Fig. 1203 — How the coils for the two-tube superheterodyne receiver are wound. In both cases both windings are in the same direction. In the case of the i.f. coil at the left, the top end of the upper winding, L_5, is connected to C_9 and Pin 3 of the 6K8 socket, the lower end of L_5 is connected to Pin 4 of the 6K8, the upper end of the lower winding, L_6, is connected to the stator of C_2 and the lower end of L_6 goes to Pin 2 on the 6SN7GT socket.

In the case of the plug-in coils, the coil sockets and plug-in form bases are wired so that the upper end of L_3 connects to the stator of C_3, the lower end of this winding to the chassis, the upper end of the lower winding, L_4, to C_6 and the lower end of L_4 goes to Pin 6 on the 6K8 socket. When the coil is plugged into the mixer stage, the upper end of the top winding should go to the stator of C_1, the lower end to C_7 and the biasing battery, the upper end of the lower winding to the chassis and the lower end of the bottom winding to the antenna terminal.

former-coupled to the detector. The grid is biased by the same battery that furnishes bias for the 6K8.

Looking at the top of the chassis from in front, the r.f. or input circuit is at the left, with C_1 below the chassis and L_1L_2 just behind it. The 6K8 is directly to the rear of the coil. The h.f. oscillator padding condenser, C_3, underneath, the socket for L_3L_4 and the 6SN7 are in line at the center of the chassis. At the right, underneath the audio transformer, T_1, is the i.f. regeneration-control condenser, C_2. The bandspread tuning condenser, C_4, is mounted on the panel with its shaft 3⅞ inches from the bottom edge of the panel. The audio transformer should be set back far enough so that there will be sufficient space for the bearing for the vernier knob of the National Type G dial. The "B" switch, S, is to the left of the dial.

A pair of terminals set in the left-hand edge of the chassis provides connections for antenna and ground, while another pair at the rear are for the "B"-battery connections. The antenna and B+ terminals must be insulated from the chassis. A jack in the right-hand side is provided for headphones and 115 volts a.c. for the heater transformer, T_2, is plugged in at the rear. The jack is insulated from the chassis by means of fiber washers. T_2 is placed under the chassis near the headphone jack.

Referring to the bottom view of Fig. 1205, the biasing battery is to the left below C_1. It is a pen-light flashlight cell soldered between the coil-socket terminal and ground. Immediately below it is the by-pass condenser, C_7. C_6 is soldered between the socket terminal for L_4

TWO-TUBE SUPERHET COIL DATA

	L_1 or L_3	L_2 or L_4
A.	90 turns No. 30 d.s.c., close-wound	20 turns No. 30 d.s.c.
B.	65 turns No. 26 d.s.c., close-wound	15 turns No. 26 d.s.c.
C.	45 turns No. 22 d.s.c., close-wound	15 turns No. 26 d.s.c.
D.	24 turns No. 22 enam., 1⅛ in. long	15 turns No. 26 d.s.c.
E.	20 turns No. 22 enam., 1⅛ in. long	15 turns No. 26 d.s.c.

Frequency Range	*Coil at L_1-L_2*	*Coil at L_3-L_4*
1700 to 3200 kc.	A	B
3000 to 5700 kc.	B	C
5400 to 10,000 kc.	C	D
9500 to 14,500 kc.	E	D

and ground. The r.f. choke is supported at one end by a small fiber lug strip and soldered to C_2 at the other. The i.f. transformer, L_5L_6, is between the two tube sockets. L_5 is connected between the proper tube-socket terminals and C_5 is soldered across these same terminals. C_9 is fastened directly between the two tube sockets and C_8 between the 6K8 socket and the proper terminal of the socket for L_3. Clearance holes are drilled in the chassis for wiring to the switch, to the stator terminal of C_4 and the grid cap of the 6K8. The rotor terminal of C_4 is grounded to the panel by a lug fastened under one of the mounting pillars. Two holes also are provided for the leads to T_1.

Coils for the receiver are wound on Millen shielded ½-inch diameter forms, Type 74001, which are provided with slug-type inductance trimmers.

The method of winding is indicated in Fig. 1203; if the connections to the circuit are made as shown, there will be no trouble in obtaining the necessary oscillation. Both coils on each form should be wound in the same direction.

Adjustment — To test the receiver, first

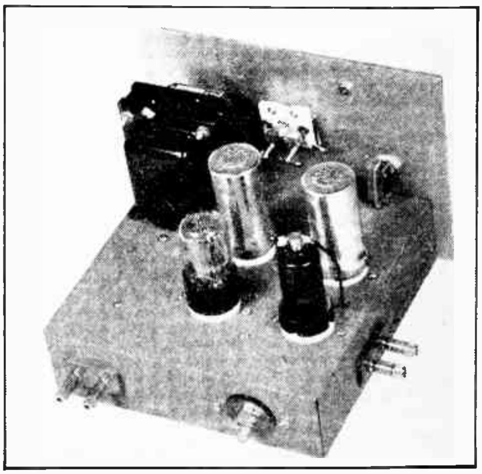


Fig. 1204 — A back-of-panel view of the two-tube superheterodyne receiver. The chassis is 7 × 7 × 2 inches.

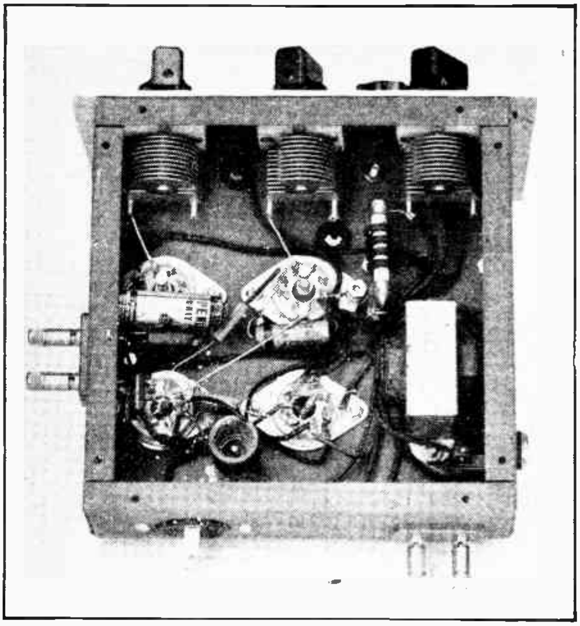


Fig. 1205 — Bottom view of the two-tube superheterodyne receiver. The i.f. coil is between the two tube sockets near the rear of the chassis. The transformer to the right is the filament transformer.

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try out the i.f. circuit. Connect the filament and "B" supplies and place both tubes in their sockets. Put a high-frequency coil in the r.f. socket, but do not insert a coil in the oscillator socket. The only test which need be made is to see if the detector oscillates properly. Advance C_2 from minimum capacity until the detector goes into oscillation, which will be indicated by a soft hiss. This should occur at around half scale on the condenser. If it does not occur, check the coil (L_5L_6) connections and winding direction and, if these seem right, add a few turns to the tickler, L_6. If the detector oscillates with very low capacity at C_2, it will be advisable to take a few turns off L_6 until oscillation starts at about midscale.

After the i.f. has been checked, plug in an oscillator coil for a range on which signals are likely to be heard at the time. The 5400–10,000-kc. range is usually a good one. The coils are arranged so that a minimum number is needed, even though two are used at a time. With Coil C in the r.f. socket and D in the oscillator circuit, set C_1 at about half scale and turn C_3 slowly around midscale until a signal is heard. Then tune C_1 for maximum volume. Should no signals be heard, the probability is that the oscillator section of the 6K8 converter tube is not working, in which case the same method of testing is used as described above for the i.f. detector — check wiring, direction of windings of coils, and finally, add turns to the tickler, L_4, if necessary.

The same oscillator coil, D, is used for two frequency ranges. This is possible because the oscillator frequency is placed on the low-frequency side of the signal on the higher range. This gives somewhat greater stability at the highest-frequency range. Some pulling — a change in beat-note as the r.f. tuning is varied by means of C_1 — will be observed on the highest-frequency range, but it is not serious in the region of resonance with the incoming signal frequency.

The receiver will respond to signals either 1600 kc. lower or 1600 kc. higher than the oscillator frequency. The unwanted response is discriminated against by the selectivity of the r.f. circuit. On the three lower-frequency ranges, when it is possible to find two tuning spots on C_1 at which incoming noise peaks up, the lower-frequency peak is the right one. The oscillator frequency is 1600 kc. higher than that of the incoming signal on these three ranges and 1600 kc. lower on the fourth range. The inductance of the coils to hit the desired ranges can be adjusted by means of the trimming slug in the coil forms.

The regeneration control may be set to give desired sensitivity and left alone while tuning; only when an exceptionally strong signal is encountered is it necessary to advance it more to keep the detector in oscillation. It should be set just on the edge of oscillation for 'phone reception.

The "B"-battery current is between 4 and 5 ma., so that a standard 45-volt block will last hundreds of hours.

ℂ A Three-Tube General-Coverage and Bandspread Superheterodyne

A superhet receiver of simple construction, having a wide frequency range for general listening-in as well as full bandspread for amateur-band reception, is shown in Figs. 1206 to 1210. The circuit uses only three tubes and gives continuous frequency coverage from about 75 kc. (4000 meters) to 60 Mc. (5 meters). The receiver is intended for operation from either a 6.3-volt transformer or 6-volt battery for heater supply, and a 90-volt "B" battery delivering 15 ma. for plate supply.

The circuit diagram is given in Fig. 1207. A 6K8 is used as a combined oscillator-mixer followed by a 6SK7 i.f. amplifier. The intermediate frequency is 1600 kc., a frequency which reduces image response on the higher frequencies and simplifies the design for low-frequency operation in the region below the broadcast band. One section of the 6C8G double triode is used as a second detector and the other section as a beat-frequency oscillator.

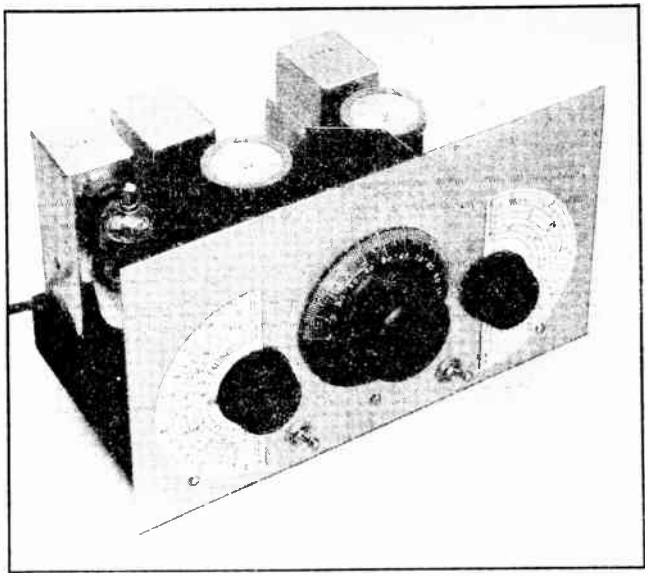


Fig. 1206 — A three-tube superheterodyne receiver, designed for either a.c. or d.c. heater operation and for 90-volt "B"-battery plate supply.

To simplify construction, the antenna and oscillator circuits are separately tuned. The antenna tuning control, C_1, may be used as a volume control by detuning from resonance. The oscillator circuit, $L_3C_2C_3$, is tuned 1600 kc. higher than the signal on frequencies up to 5 Mc.; above 5 Mc. the oscillator is 1600 kc. lower than the signal.

The parts arrangement is shown in the photographs of Figs. 1208 and 1209. The mixer tuning condenser, C_1, is at the right. The bandspread oscillator tuning condenser, C_3, is in the center, controlled by the National Type A 3½-inch dial, and the bandset condenser, C_2, is at the left.

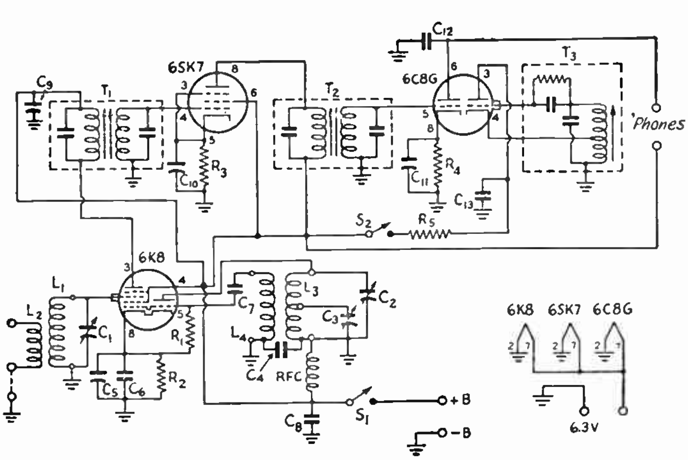
Referring to the top view, Fig. 1208, the i.f. section is along the rear edge, with T_1 at the right. Next is the socket for the 6SK7, then T_2, and finally T_3 at the extreme left. The socket for the 6C8G is just in front of T_3. The triode section in which the grid is brought out to the top cap is the one which is used for the beat oscillator.

The r.f. section has been arranged for short leads to favor high-frequency operation. The three sockets grouped closely together in the center are, from left to right, the oscillator-coil socket, socket for the 6K8, and the mixer-coil socket. All are mounted above the chassis by means of mounting pillars, so that practically all r.f. leads are above deck. The oscillator grid leak, R_1, and the high-frequency cathode by-pass condenser, C_6, should be mounted directly on the socket before it is installed. So also should the oscillator grid condenser, C_7, which can be seen extending to the left toward the oscillator-coil socket in Fig. 1208. Power-supply connections should be soldered to the 6K8 socket prongs before the socket is mounted.

The general-coverage condensers, C_1 and C_2, are mounted directly on the chassis. C_3 is held from the panel by means of a small bracket made from metal strip, bent so that the condenser shaft lines up with the dial coupling. A baffle shield made of aluminum separates the oscillator and mixer sections.

The first step in putting the receiver into operation is to align the i.f. amplifier. This should preferably be done with the aid of a test oscillator, but if one is not available the circuits may be aligned on hiss or noise. The beat oscillator can also be used to furnish a signal for alignment. Further information on alignment

Fig. 1207 — Wiring diagram for the three-tube superheterodyne.



C_1 — 100-μμfd. variable (Hammarlund MC-100-M).
C_2 — 140-μμfd. variable (Hammarlund MC-140-M).
C_3 — 35-μμfd. variable (Hammarlund HF-35).
C_4 — Oscillator padder; see coil table.
C_5 — 0.1-μfd. paper.
C_6, C_8, C_{12}, C_{13} — 0.002-μfd. mica.
C_7 — 270-μμfd. mica.
C_9, C_{10} — 0.01-μfd. paper.
C_{11} — 5-μfd. electrolytic, 50 volts.
R_1, R_5 — 47,000 ohms, ½ watt.
R_2, R_3 — 270 ohms, ½ watt.
R_4 — 12,000 ohms, ½ watt.
L_1, L_2, L_3, L_4 — See coil table.
T_1, T_2 — 1600-kc. i.f. transformer (Millen 64161).
T_3 — 1600-kc. oscillator transformer (Millen 65163).
S_1, S_2 — S.p.s.t. toggle switch.
RFC — 2.5-mh. r.f. choke.

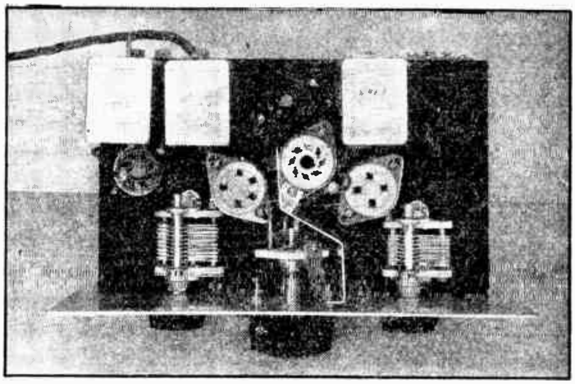


Fig. 1208 — A plan view of the three-tube superheterodyne with the coils and tubes removed. The chassis measures $5\frac{1}{2} \times 9\frac{1}{2} \times 1\frac{1}{2}$ inches and the panel size is $10\frac{1}{2} \times 6$ inches.

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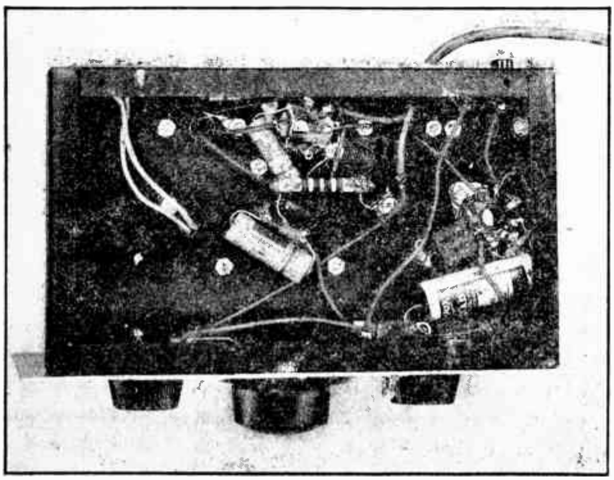
may be found in Chapter Seven.

The coils are wound as shown in Fig. 1210. A complete set of specifications is given in the coil table. Ordinary windings are used for all oscillator coils, and for all mixer coils for frequencies above 1600 kc. Below 1600 kc., readily available r.f. chokes are used for the tuned circuits. For the broadcast band and the 600–750-meter ship-to-shore channels, the mixer coil is a Hammarlund 2.5-mh. r.f. choke, with the pies tapped as shown in Fig. 1210. The grid end and the intermediate tap are connected to machine screws mounted near the top of the coil form, and a flexible lead is brought out from the grid pin in the coil form to be fastened to either lead as desired. Mixer coils for the two lowest-frequency ranges are constructed as shown. The antenna winding in each case is a coil taken from an old 465-kc. i.f. transformer, having an inductance of about 1 millihenry. The inductance is not critical, and a pie from a 2.5-mh. choke may be used instead.

COIL DATA FOR THE THREE-TUBE SUPERHETERODYNE

Range	*Turns* L_1	L_2	L_3	L_4	L_3 *Tap*	C_4
A — 76–154 kc.	30 mh.	1 mh.				
166–360 kc.	8 mh.	1 mh.	65	12	Top	300 μμfd.
400–1500 kc.	2.5 mh.*	*				
B — 1.6 to 3.2 Mc. (160 meters)	56	10	42	11	Top	75 μμfd.
C — 3.0 to 5.7 Mc. (80 meters)	32	8	27	9	Top	100 μμfd.
D — 5.4 to 10.0 Mc. (40 meters)	18	8	22	9	12	0.002 μfd.
E — 9.5 to 18.0 Mc. (20 meters)	10	8	12	3½	6	400 μμfd.
F — 15.0 to 30 Mc. (10 meters)	6	4	6	2½	2½	400 μμfd.
G — 30 to 60 Mc. (5 meters)	3	3	3½	1	1	300 μμfd.

* See Fig. 1210 and text for details. C_4 is mounted inside oscillator coil form; see Fig. 1210. Bandspread taps on L_3 measured from bottom ("B"+ end) of coil. L_3-A and L_1-B coils close-wound with No. 22 enameled wire; L_3-B close-wound with No. 20 enameled; all other L_1 and L_3 coils wound with No. 18 enameled, spaced to give a length of 1½ inches on a 1½-inch diameter form (Hammarlund SWF) except the G coils, which are spaced to a length of 1 inch on 1-inch diameter forms (Millen 45004 and 45005). Antenna and tickler coils, L_2 and L_4, are close-wound with No. 24 enameled, spaced about ⅛ inch from bottoms of grid coils, except for L_4-G, which is interwound with L_3.



With the i.f. aligned, the mixer grid and oscillator coils for a band can be plugged in. C_3 should be set near minimum capacity and C_2 tuned from minimum capacity until a signal is heard. Then C_1 is adjusted for maximum signal strength. If C_2 is set at the high-

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Fig. 1209 — Below the chassis of the three-tube receiver. The r.f. choke is mounted near the oscillator coil socket to keep the r.f. leads short. In the i.f. stage, care should be taken to keep the plate and grid leads from the i.f. transformer short and well separated. A four-wire cable is used for power-supply connections. The headphone tip jacks may be seen near the upper right-hand corner.

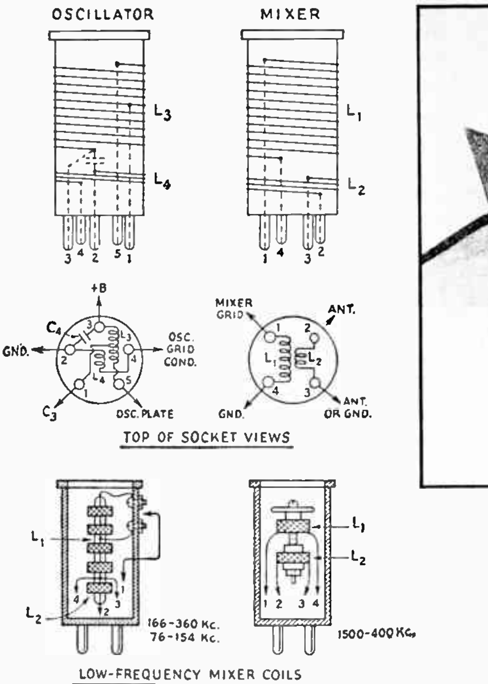


Fig. 1210 — How the coils for the three-tube superheterodyne are constructed. On the hand-wound oscillator and mixer coils, all windings are in the same direction.

frequency end of an amateur band, further tuning should be done with C_3, and the band should be found to cover about seventy-five per cent of the dial. C_3 can of course be used for bandspread tuning outside as well as inside the amateur bands. It is convenient to calibrate the receiver, using a homemade paper scale for the purpose as shown in Fig. 1206. Calibration points may be taken from incoming signals whose frequencies are known, from a calibrated test oscillator, or from the harmonics of a 100-kc. oscillator, as described in Chapter Nineteen. The mixer calibration need be only approximate, since tuning of the mixer circuit has little effect on the oscillator frequency. It is sufficient to make a calibration which ensures that the mixer is tuned to the desired signal rather than to an image.

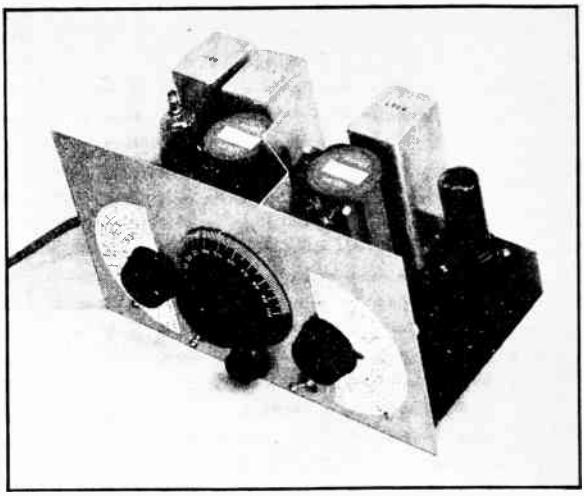


Fig. 1211 — The modified three-tube superheterodyne receiver with the audio-amplifier stage added for loudspeaker operation.

On the broadcast band, the tuning range is such that, with C_2 set at 1500 kc., the entire band will be covered on C_3. It is necessary, however, to change the tap on the mixer coil to make the antenna circuit cover the entire band. Only one oscillator coil is needed for the range from 75 to 1500 kc., but a series of coils is needed to cover the same range in the mixer circuit.

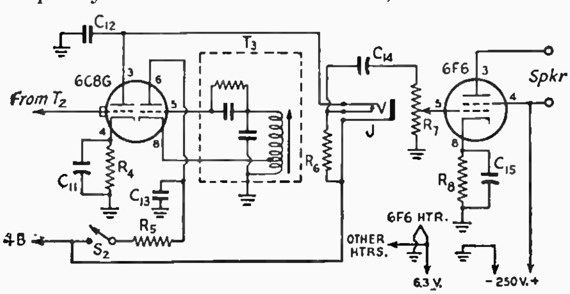


Fig. 1212 — Circuit diagram of the single-tube pentode audio-amplifier stage which may be added for loudspeaker operation of the three-tube superheterodyne. Except as noted below, the values for components correspond to those bearing the same designations in Fig. 1207.

C_{14} — 0.1-μfd. paper.
C_{15} — 25-μfd. electrolytic, 50 volts.
R_6 — 0.12 megohm, ½ watt.
R_7 — 0.5-megohm volume control.
R_8 — 400 ohms, 1 watt.
J — Closed-circuit jack.

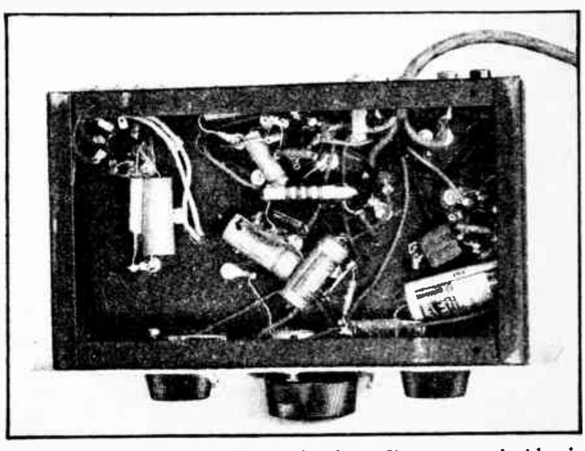


Fig. 1213 — The additional parts for the audio stage can be identified in this subchassis view of the modified three-tube receiver.

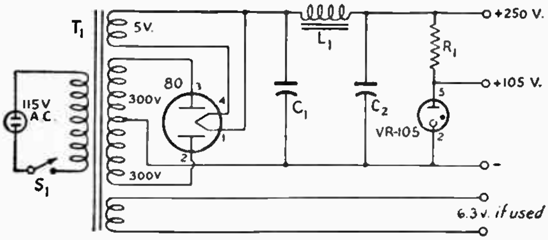


Fig 1214 — Circuit diagram of a power supply suitable for small receivers.

C_1, C_2 — 8- or 16-μfd. electrolytic, 450 volts.
R_1 — 5000 ohms, 10 watts, wire-wound.
L_1 — Standard replacement-type filter choke, 15 to 30 henries at 70 ma.
S_1 — S.p.s.t. toggle switch.
T_1 — Standard replacement-type power transformer with 6.3-volt, 5-volt, and 600-volt center-tapped windings, 70 ma. d.c. output rating.

Adding an audio stage to the three-tube superheterodyne — The three-tube receiver just described is designed for headphone operation, but readily can be converted to a four-tube set for use with a 'speaker. For this purpose a 6F6 pentode can be added to the circuit diagram, as shown in Fig. 1212. Figs. 1211 and 1213 show the receiver when completed.

For the purpose of driving the audio stage, resistance coupling is used from the plate of the second detector to the grid of the 6F6. A volume control is used for the grid resistor of the 6F6, and a jack is installed in the second-detector plate circuit so that a headphone plug may be inserted. The volume control, R_7, should be of the midget type so that it will fit in the chassis; it is installed with its shaft projecting under the tuning dial. In the bottom view, Fig. 1213, the 6F6 socket is in the upper left corner, along with the cathode resistor and by-pass condenser, R_8 and C_{15}. The coupling condenser, C_{14}, and the plate resistor, R_6, are mounted on an insulated lug strip near the volume control.

The 6F6 will require a plate supply of 250 volts at about 40 milliamperes. This may be taken from a regular power pack, and a five-wire connection cable is used to provide an extra lead for the purpose. The first three tubes may be operated from a "B" battery, as before. Alternatively, the power supply may be constructed with a tap giving 90 or 100 volts for these tubes, the tap being connected to the proper wire in the connection cable. For best performance, the output voltage should be regulated by a VR-105 regulator tube. A suitable power supply is shown in Fig. 1214.

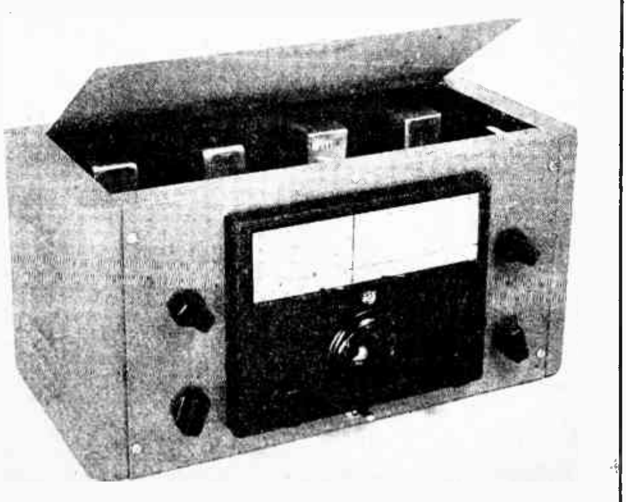
The primary winding of the 'speaker output transformer always should be connected in the plate circuit of the 6F6. Operation without the plate circuit closed is likely to damage the screen grid. Any 'speaker having a transformer with a primary impedance of 7000 ohms will be satisfactory; a permanent-magnet dynamic is convenient, since no field supply for the 'speaker is necessary.

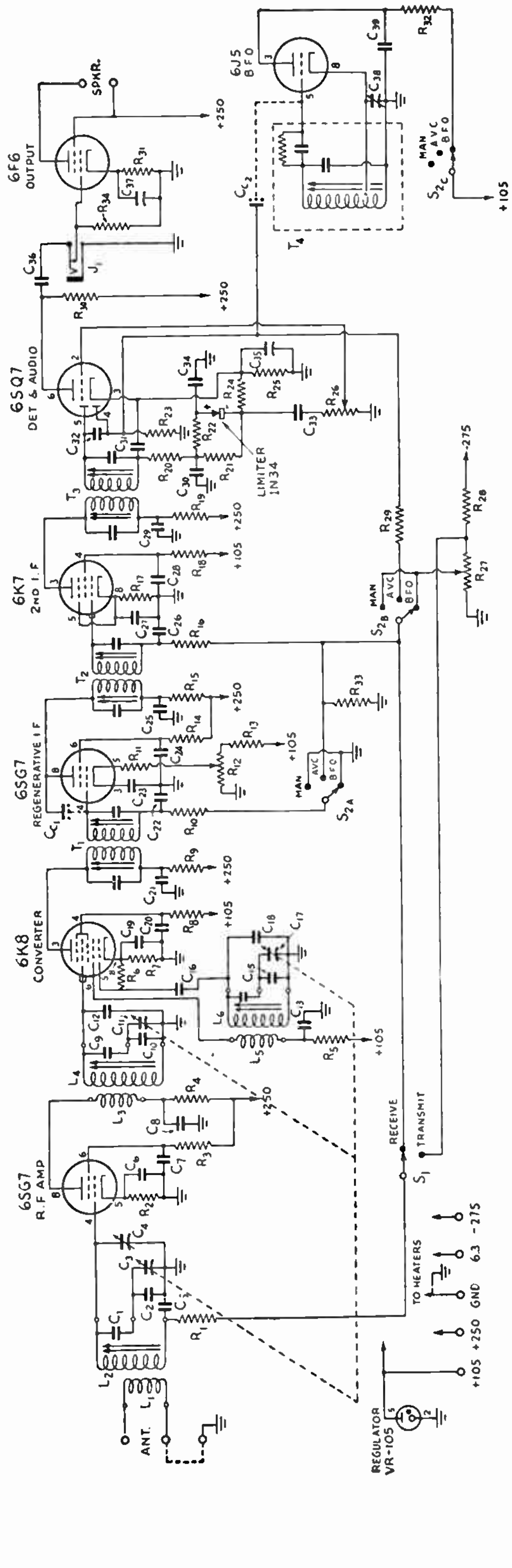
Power supply — Components for the a.c. power supply of Fig. 1214 may be mounted on a 7 × 7 × 2-inch steel chassis or a baseboard made of wood. The placement of parts is not important. If the steel chassis is used, the smaller components may be mounted underneath. The voltage of the filament winding should, of course, correspond to the rated heater voltage of the tubes used, unless a separate heater transformer is used.

An Amateur-Band Eight-Tube Receiver

A receiver with good mechanical and electrical stability, variable selectivity through the use of a regenerative i.f. amplifier, good a.v.c. and gain-control characteristics, an audio noise limiter, and adequate audio for loudspeaker reception is shown in Figs. 1215, 1217

Fig. 1215 — An amateur-band eight-tube receiver. The knobs on the left control audio volume (upper) and b.f.o. pitch, and the two on the right handle r.f. and i.f. gain (upper) and i.f. regeneration. The knob to the left of the large tuning knob is fastened to the *MAN.-A.V.C.-B.F.O* switch, and the one on the right is for the antenna trimmer. The toggle switch under the dial throws high negative bias on the r.f. stage during transmission periods.





and 1218. As can be seen from the circuit in Fig. 1216, a 6SG7 pentode is used for the tuned r.f. stage ahead of the 6K8 converter. An antenna compensator, C_4, controlled from the panel, allows one to trim up the r.f. stage when using different antennas that might modify the tracking. The cathode bias resistor of the r.f. stage is made as low as possible consistent with the tube ratings, to keep the gain and hence the signal-to-noise ratio of the stage high. The oscillator portion of the 6K8 mixer is tuned to the high-frequency side of the signal except on the 28-Mc. band, the usual custom nowadays in communications receivers. The oscillator tuning condenser, C_{17}, is of higher capacity than the r.f. and mixer tuning condensers, in the interest of better oscillator stability.

The i.f. amplifier is tuned to 455 kc., and the first stage is made regenerative by soldering a short length of wire to the plate terminal of the socket and running it near the grid terminal, as indicated by C_{c1} in the diagram. Regeneration is controlled by reducing the gain of the tube, and R_{12}, a variable cathode-bias control, serves this function. The second i.f. stage uses a 6K7, selected because high gain is not necessary at this point.

Manual gain-control voltage is applied to the r.f. and second i.f. stages. It is not applied to

Fig. 1216 — Circuit diagram of the eight-tube receiver.

C_1, C_9, C_{14} — See table.
C_2, C_{10}, C_{12}, C_{18} — 10-μμfd. ceramic.
C_3, C_{11} — 15-μμfd. midget variable (National UM-15).
C_4 — 15-μμfd. midget variable (Hammarlund HF-15).
C_5, C_6, C_7, C_8, C_{13}, C_{19}, C_{20}, C_{21}, C_{22}, C_{23}, C_{24}, C_{25}, C_{26}, C_{27}, C_{28}, C_{29}, C_{39} — 0.01-μfd. mica.
C_{15} — 37-μμfd. ceramic (10 and 27 in parallel).
C_{16}, C_{30}, C_{32} — 100-μμfd. mica.
C_{17} — 35-μμfd. midget variable (National UM-35).
C_{31} — 250-μμfd. mica.
C_{33} — 0.05-μfd. paper, 200 volts.
C_{34} — 0.1-μfd. paper, 200 volts.
C_{35}, C_{37} — 10-μfd. 25-volt electrolytic.
C_{36} — 0.1-μfd. paper, 400 volts.
C_{38} — 35-μμfd. midget variable (Hammarlund HF-35).
C_{c1}, C_{c2} — See text.
R_1, R_{10}, R_{16}, R_{30} — 0.1 megohm.
R_2 — 68 ohms.
R_3, R_{14} — 33,000 ohms.
R_4, R_5, R_6, R_8, R_9, R_{13}, R_{15}, R_{18}, R_{19}, R_{20}, R_{21} — 47,000 ohms.
R_7 — 220 ohms.
R_{11} — 180 ohms.
R_{12} — 2000-ohm wire-wound potentiometer.
R_{17} — 330 ohms.
R_{22}, R_{23}, R_{29}, R_{33} — 1.0 megohm.
R_{24}, R_{28} — 0.15 megohm.
R_{25} — 2700 ohms.
R_{26} — 1.0-megohm carbon potentiometer.
R_{27} — 25,000-ohm carbon potentiometer.
R_{31} — 470 ohms, 1 watt.
R_{32} — 27,000 ohms.
R_{34} — 0.2 megohm.

All resistors ½ watt unless otherwise noted.

L_1 through L_6 — See table.
J_1 — Closed-circuit telephone jack.
S_1 — S.p.d.t. toggle switch.
S_{2A-B-C} — Three-pole 3-position wafer switch (Centralab 2507).
T_1, T_2 — 456-kc. interstage i.f. transformer, permeability tuned (Millen 64456).
T_3 — 456-kc. diode transformer, permeability tuned (Millen 64454).
T_4 — 456-kc. b.f.o. assembly, permeability tuned (Millen 65456).

the mixer because it might pull the oscillator frequency, and it is not tied in with the first i.f. amplifier because it would interlock with the regeneration control used for controlling the selectivity. However, the a.v.c. voltage is applied to the r.f. and both i.f. stages, with the result that the selectivity of the regenerative stage decreases with loud signals and gives a measure of automatic selectivity control. Using a negative-voltage power supply for the manual gain control is more expensive than the familiar cathode control, but it allows a wide range of control with less dissipation in the components. The a.v.c. is of the delayed type, the a.v.c. diode being biased about 1½ volts by the cathode resistor of the diode-triode detector-audio stage.

The second-detector-and-first-audio is the usual diode-triode combination and uses a 6SQ7. A 1N34 crystal diode is used as a noise limiter, and is left in the circuit all of the time. As is common with this type of circuit, it has little or no effect when the b.f.o. is on, but it is of considerable help to 'phone reception on the bands where automobile ignition is a factor. The constructor can satisfy himself on its operation when first building the receiver and working on it out of the case. By leaving one end of the 1N34 floating and touching it to the proper point in the circuit, a marked drop in ignition noise will be noted.

The b.f.o. is capacity-coupled to the detector by soldering one end of an insulated wire to the a.v.c. diode plate and wrapping several turns of the wire around the b.f.o. grid lead. This capacity is designated C_{42} in the diagram. The wire was connected to the a.v.c. diode plate lead for wiring convenience — the a.v.c. coupling condenser, C_{32}, passing the b.f.o. voltage without appreciable attenuation.

Headphone output is obtained from the plate circuit of the 6SQ7 at J_1, and loudspeaker output is available from the 6F6 audio-amplifier stage. High-impedance or crystal headphones are recommended for maximum headphone output.

The receiver is built on an aluminum chassis mounted in a Par-Metal CA-202 cabinet and a Millen 10035 dial is used for tuning. The chassis is made of 1/16-inch-thick stock, bent into a "U"-channel, and measures 13 inches wide and 7¼ inches deep on the top. It is 3¾ inches deep at the rear and ⅛ inch less at the front. The rear edge is reinforced with a piece of ⅜-inch square dural rod that is tapped for screws through the bottom of the cabinet, further to add to the strength of the structure when finally assembled. The various components that are common to the front lip of the chassis and the panel are used to tie the two together.

The shield panel used to mount the antenna compensator condenser is also made of 1/16-inch aluminum with a ⅝-inch lip on the side for mounting. Part of the lip must be cut away to clear wires and mounting plates on some sockets, so it is advisable to put in the panel after most of the assembly and wiring have been completed. Flexible couplings and bakelite rod couple the condenser to the panel bushing.

The three tuning condensers are mounted on individual brackets of 1/16-inch aluminum. The brackets measure 2½ inches wide and 1 9/16 high, with ½-inch lips. A cover of thin aluminum — not shown in the photographs — slides over the condenser assembly to dress up the top view a bit. The dust cover is not necessary for the satisfactory operation of the receiver.

Ceramic sockets are used for the plug-in coils and the r.f. amplifier, converter and b.f.o. tubes. Mica condensers were used throughout the receiver for by-passing wherever feasible, because they lend themselves well to compact

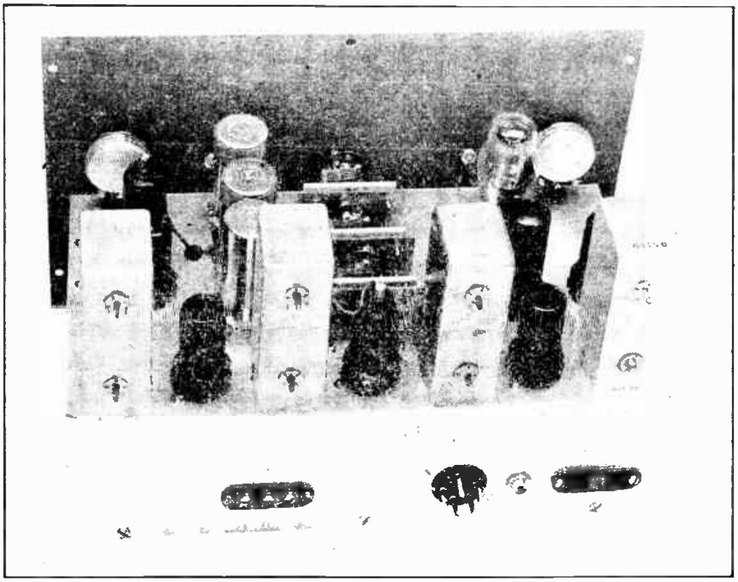


Fig. 1217 — This view of the eight-tube receiver chassis shows the mounting of the tuning condensers and the placement of most of the large components. The three shielded plug-in coil assemblies can be seen to the left of the tuning gang. The 6K8 converter is the tube on the left nearest the panel.

The antenna terminal strip, power-supply plug, headphone jack and 'speaker terminals are mounted on the rear (foreground in this view) of the chassis.

construction. Paper condensers could be used in the i.f. amplifier but they would crowd things a bit more.

In wiring the receiver, small tie-points were used wherever necessary to support the odd ends of resistors and condensers, and rubber grommets were used wherever wires run through the chassis, with the exception of the tuning-condenser leads. The latter leads, being of No. 14 wire, are self-supporting through the 5/16-inch clearance holes and do not require grommets. The same heavy wire was used for the grid and plate leads of the r.f. stage and the plate lead of the oscillator, to reduce the inductance in these leads. The tuning condensers are grounded back at the coil sockets and not above the chassis as might be the tendency. Screen, cathode and plate by-pass condensers are grounded at a single point for any tube wherever possible, although C_2 is grounded at the r.f.-coil socket, C_8 is grounded at the converter-coil socket, and C_{13} is returned at the oscillator-coil socket. The plate and B+ leads from T_1 are brought back to the converter socket through shield braid, and C_{21} is returned to ground at the converter socket.

The b.f.o. pitch condenser, C_{38}, is insulated from the chassis and panel by fiber washers, and the rotor is connected back to the tube socket by braid that shields the stator lead. This is done to reduce radiation from the b.f.o. which might get in at the front end of the i.f. amplifier.

The coils are wound on Millen 74001 permeability-tuned coil forms, according to the coil table. Series condensers are mounted inside the forms on all bands except the 80-meter range, where no condenser is required and the tuning condenser is jumped directly to the grid end of the coils. In building the coils, the washers are first drilled for the leads and then cemented to the form with Duco or other cement. The bottom washer is cemented close to the terminal pins, leaving just enough room to get the soldering iron in to fasten the coil ends and to leave room for the series condenser. The large coils, L_2, L_4 and L_6, were wound first in every case, and then a layer of polystyrene Scotch Tape wrapped over the coil, after which the smaller winding was put on and the ends of the windings soldered in place. Since for maximum range of adjustment it is desirable to allow the powdered-iron slug to be fully withdrawn from the coil, keeping the coils at the base end of the form allows the iron slug to travel out at the other end, under which condition the adjusting screw on the slug projects the least. To secure the wires after winding, drops of cement should be placed on them where they feed through the polystyrene washers.

If a signal generator is available, it can be used to align the i.f. amplifier on 455 kc. in the usual manner. If one is not available, the coupling at C_{C1} can be increased to the point where the i.f. stage oscillates readily and the b.f.o. transformer is then tuned until a beat

COIL DATA FOR THE EIGHT-TUBE SUPERHETERODYNE

Coil	*3.5 Mc.*	*7 Mc.*	*14 Mc.*	*28 Mc.*
L_1	15 t	9 t	6 t	4 t
L_2, L_4	76 t	33 t	19 t	8 t
C_1, C_9	short	27 μμfd.	15 μμfd.	20 μμfd.
L_3	25 t	11 t	7 t	4 t
L_5	10 t	8 t	4 t	2 t
L_6	47 t	32 t	14 t	6 t
C_{14}	short	42 μμfd.	27 μμfd.	51 μμfd.

All coils wound on Millen 74001 forms, close-wound. 3.5-Mc. coils wound with No. 30 enam.; 7-Mc. coils wound with No. 30 d.s.c.; 14- and 28-Mc. coils wound with No. 30 d.s.c. on primaries and ticklers and No. 24 enam. on secondaries. C_{14} for 7-Mc. range made by connecting 27- and 15-μμfd. condensers in parallel. C_1, C_9 and C_{14} Erie Ceramicons mounted in coil form.

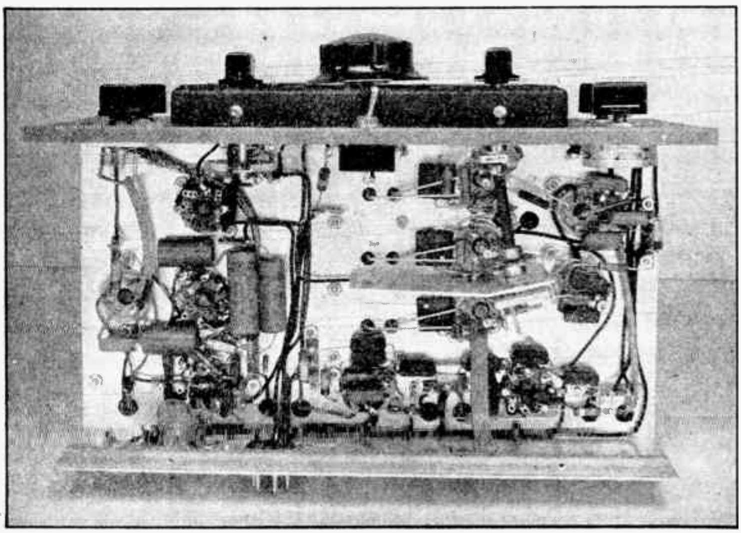


Fig. 1218 — The mica by-pass condensers used throughout the r.f. and i.f. stages are grouped around the sockets of their respective tubes. Tie-points are used wherever necessary to support small resistors and condensers. The antenna trimmer condenser is mounted on a bracket which also serves as shielding between the mixer- and r.f.-coil sockets, and it is offset to allow access to the trimmer screws on the coil forms. The plate and B+ leads from the first i.f. transformer, T_1, are run in shielded braid, as are the leads from the b.f.o. pitch-control condenser and the volume control.

note is heard. The other transformers can then be aligned until the signal is loudest, after which C_{c1} should be decreased until the i.f. oscillates with the regeneration control, R_{12}, about 5 degrees from maximum. The trimmers on T_1 then should be tuned to require maximum advancing of the regeneration control for oscillation, with a set value of C_{c1}. When properly tuned, the oscillation frequency of the i.f. stage and the frequency for maximum gain in the regenerative condition, will be the same.

With a set of coils in the front end, set the tuning dial near the high-frequency end and tune in a strong signal or marker with the adjustment screw on the oscillator coil. The converter and r.f. coils can then be peaked, with the antenna compensator set at about half capacitance. Then tune to the other end of the band and see if you have enough bandspread. If the bandspread is inadequate, it means that C_{14} is too large, and it should be reduced by using a smaller size of condenser or a combination that gives slightly less capacitance. The tracking of the converter and r.f. coils can be checked by repeaking the position of the slugs in the coils at the low-frequency end. If the converter or r.f. coil tuning slug has to be advanced farther into the coil (to increase the inductance) it indicates that C_9 or C_1 should be larger. Tracking by the method described is at best a compromise, although to all intents and purposes the loss from some slight misalignment is completely unimportant. Another method would be to tap the tuning condensers on the coil in the familiar bandspreading manner, but this requires considerable time and patience. However, with the series condensers as used in this receiver, the tuning curve is more crowded at the high-frequency end of a range than at the low, and this would be reduced somewhat by the tapped-coil method of bandspread.

The adjustment of L_5 can be made, if deemed necessary, by lifting the cathode end of R_6 and inserting a 0-1 milliammeter. If the tickler coil has the right number of turns, the current will be from 0.15 to 0.2 ma., and it won't change appreciably over the band. Although such a grid-current check is a fine point and not really necessary, it is a simple way to determine that the oscillator portion is working, since the cold ends of L_5 and L_6 are at the same end of the form — the plug end — and this necessitates winding the two coils in opposite directions.

Some trouble may be experienced with oscillation in the r.f. stage at 28 Mc. However, a grounding strap of spring brass mounted under one of the screws holding the mixer-coil socket to ground the shield when the coil is plugged in will normally clear up the trouble. Inadequate coupling to the antenna will also let the r.f. stage oscillate under some tuning conditions, and close coupling is highly recommended for stability in this stage and also for best signal response. A 10-ohm resistor from L_2 to the grid of the 6SG7 will also do the trick.

It will be found that the over-all gain of the receiver is quite high on the lower-frequency bands, requiring that the r.f. gain be cut down to prevent overloading on strong signals. For c.w. reception, the regeneration control is advanced to the point just below oscillation and the b.f.o. is detuned slightly to give the familiar single-signal effect. For 'phone reception, S_2 is switched to *A.V.C.* and volume-control adjustments made with the audio control, R_{26}. If desired, the regeneration control can be advanced until the i.f. is oscillating weakly, and then a heterodyne will be obtained on weak carriers, making them easy to spot. Strong carriers will pull the i.f. out of oscillation because the developed a.v.c. voltage reduces the gain, and hence a simple form of automatic selectivity control is obtained. If it is considered desirable to reduce the i.f. gain when switched to the *A.V.C.* position, the regeneration control can be used for this purpose. The *MAN.* position permits manual gain-control operation with the b.f.o. off.

The switch S_1 is used for receive-transmit and throws about 40 volts negative on the grid of the first r.f. stage.

Power supply — A power supply suitable for the eight-tube receiver is shown in Figs. 1219 and 1220. An idea of the parts arrangement can be obtained from Fig. 1219, although there is nothing critical about this portion of the receiver. If one wants a neat-looking station with no loose power supplies in sight, the power supply can be built into one corner of the loudspeaker cabinet.

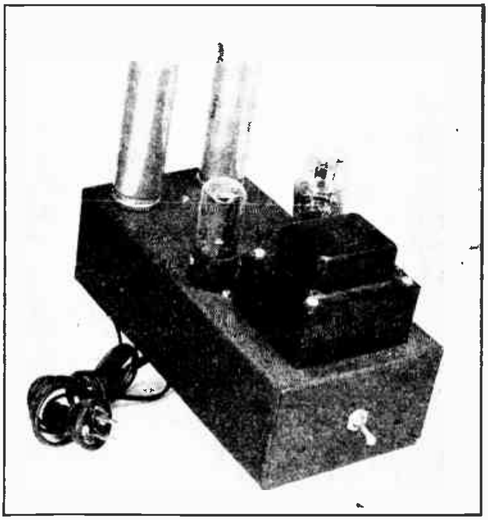
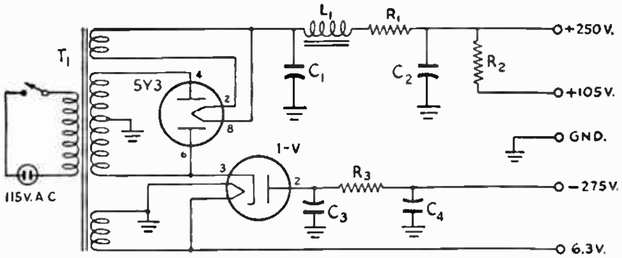


Fig. 1219 — Power supply for the eight-tube receiver. Two rectifiers are required because a separate supply is incorporated for gain-control purposes. The filter choke and the negative-supply filter condensers are mounted under the chassis. At the rear of the chassis is the socket for the power cable.

Fig. 1220 — Power-supply wiring diagram.

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



A Band-Pass Converter for 14, 28 and 50 Mc.

To extend the frequency range of a communications receiver, a converter can be used, to convert from the signal frequency to that of the receiver. Such a converter is shown in Figs. 1221, 1223, 1224 and 1225, which will give reception in the 14-, 28- and 50-Mc. bands with any receiver capable of tuning to 7.3 Mc. To simplify construction, the r.f. stages are fixed-tuned and only the local oscillator is tuned when running across a band. The band-width of the r.f. stages is sufficient to accept any signal over an amateur band without noticeable attenuation. The broad-banding is obtained by loading the circuits with resistors to reduce the *Q*, using a minimum of capacity for the same reason, and then "staggering" the circuits: i.e., tuning them to slightly different frequencies so that the resultant pass band is broad and nearly flat within the required range. The input circuit, from the antenna, must be broad, and this can only be obtained by heavy coupling to the antenna. This condition coincides with the condition for best signal transfer.

As can be seen from the wiring diagram in Fig. 1222, the only tuning controls in the r.f. stages are the powdered-iron slugs of the coils. These are used to resonate the coils with the circuit capacities to the signal frequency. The loading resistors, R_3 and R_6, are used to broaden the circuits. The plate and screen voltages are the same on each r.f. amplifier tube, to reduce the number of by-pass condensers, and filter resistors are used to prevent over-all feed-back through the common power lead. Another possible source of over-all feed-back is the heater circuit, and in this converter the "hot" heater lead to the input stage was run in shield braid to reduce the possibility of feed-back.

The oscillator is a straight plate-tickler type using a 6C4, and it is coupled to the mixer through a capacity shown as dotted lines in the diagram. Actually the coupling capacitor consists of a short length of wire near the grid of the mixer tube.

The output frequency is 7.3 Mc. approximately, and this is the frequency to which $C_{10}L_5$ is tuned. If a frequency slightly below 7.0 Mc. is used, there is a possibility that the fourth harmonic of the receiver high-frequency oscillator will find its way into the converter when operating in the 28-Mc. band, resulting in a constant signal that has only nuisance value. A low-impedance shielded line feeds the 7.3-Mc. output into the communications receiver. The communications receiver furnishes the necessary selectivity.

The cathode bias of the second r.f. amplifier is varied by the gain control, R_{10}, to avoid blocking by strong signals. The "send-receive" switch, S_1, is used to turn off the converter during transmission periods. The power switch, S_2, is mounted on the gain control and is used to turn off the power to the converter.

The power supply is regulated, using the miniature equivalent of the VR-105, and the stabilized 105 volts is fed to all stages.

The r.f. stages and mixer are built as a separate unit on a strip of alumi-

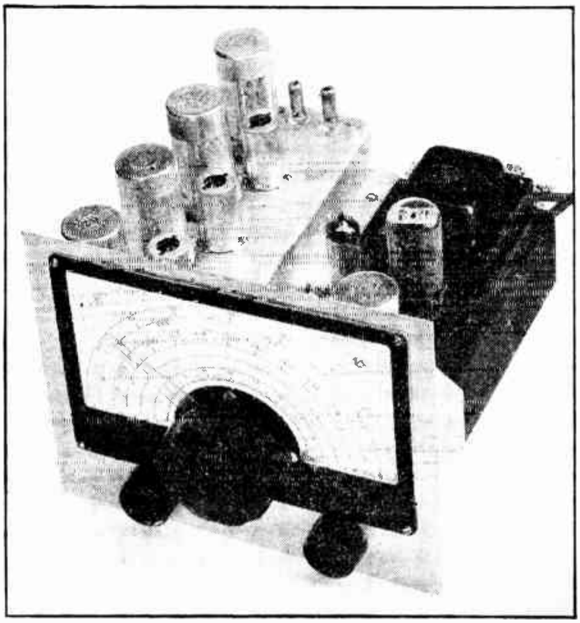


Fig. 1221 — A 28-Mc. converter that uses fixed-tuned r.f. stages and thus eliminates the ganging problem. The knob at left is for the "send-receive" switch, and the right-hand knob is for gain control.

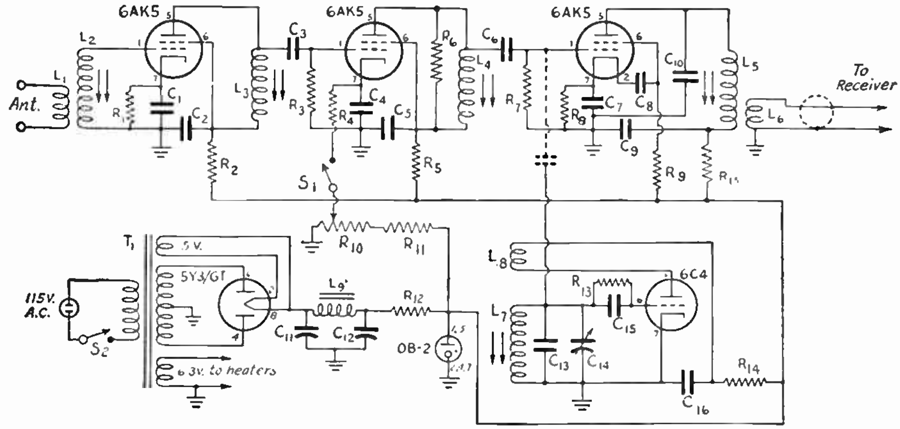


Fig. 1222 — Circuit diagram of the band-pass converter.

C_1, C_2, C_4, C_5, C_7, C_8, C_{16} — 0.001-μfd. postage-stamp mica.
C_3, C_6 — 100-μμfd. postage-stamp mica.
C_9 — 0.01-μfd. mica.
C_{10}, C_{15} — 51-μμfd. ceramic (Erie N150).
C_{11}, C_{12} — 16-μfd. 150-volt electrolytic.
C_{13} — 27-μμfd. ceramic (Erie N150) across C_{14} plus additional capacity mounted in L_7L_8 form. See coil table.
C_{14} — 11-μμfd. midget variable (Hammarlund HF-15 with one stator plate removed).
R_1, R_4 — 180 ohms.
R_2, R_5, R_{11}, R_{15} — 270 ohms.
R_3, R_6, R_8 — 6800 ohms.
R_7 — 1.5 megohms.
R_9 — 1.0 megohm.
R_{10} — 2000-ohm potentiometer, wire-wound.
R_{11} — 10,000 ohms, 2 watts.
R_{12} — 1250 ohms, 10 watts, wire-wound.
R_{13} — 51,000 ohms, 1 watt.
All resistors ½ watt unless otherwise specified.
L_1–L_8 — See coil table.
L_9 — 8-henry 50-ma. filter choke (Stancor C-1279).
S_1 — S.p.s.t. rotary switch.
S_2 — S.p.s.t. switch, mounted on R_{10}.
T_1 — 300-0-300 volts, 50-ma. power transformer, with 5- and 6.3-volt windings (U.T.C. R-6).

num, to furnish a chassis in which the grounds are more certain than they would be on a black-crackled steel chassis, and it also makes a well-shielded amplifier when mounted on the steel chassis. The steel chassis is a standard 7 × 11 × 2-inch affair. A panel is used to support the National ACN dial, and to reduce metal work on the steel chassis the panel is supported away from the chassis by an aluminum bracket on one side and by two of the screws that fasten the dial to the panel. Holes in the chassis allow access to the tuning slugs of the r.f. coils.

The tuning condenser is mounted on a small aluminum bracket fastened to the chassis by two screws and to the condenser by the shaft bushing. This results in a rigid mount that contributes considerably to the mechanical stability of the oscillator.

The construction of the aluminum channel is apparent from Fig. 1221. It is 3 inches wide and 1¼ inches high, and is bolted to the side of the steel chassis and to the top. A small strip of bakelite, supported away from the side by screws and small spacers, is used to support the power-supply end of the filter resistors R_2, R_5 and R_{15}. The ends are fed through small holes in the bakelite and then wrapped around the strip before being soldered together.

◆

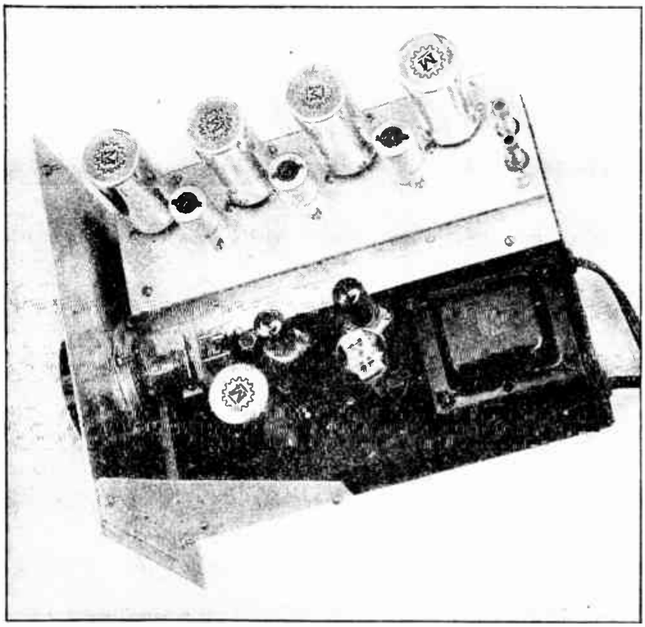


Fig. 1223 — Another view of the converter showing the r.f. subchassis. Note the bracket on the tuning condenser, used to avoid backlash.

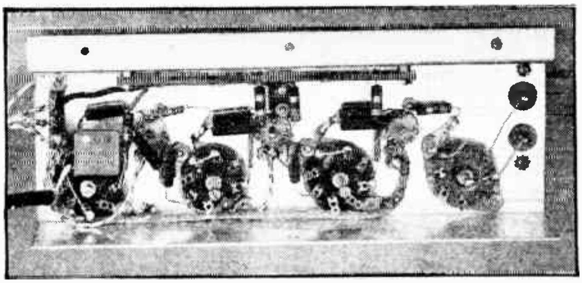


Fig. 1224 – The straightforward arrangement of the r.f. components is shown in this view of the subchassis. The straight side is screwed to the side of the chassis.

In the heater circuits of the miniature tubes, Pin 4 is grounded to a lug under the nut fastening the socket, and Pin 3 is the "hot" heater lead. In the case of the input 6AK5, the hot heater lead was led back in shield braid, and the braid was grounded at the lug grounding Pin 4, and to lugs at two other points along the way. These latter lugs are under the nuts fastening the sockets for L_3 and the output coil, L_5L_6.

The cathode and screen/plate by-pass condensers are grounded to lugs under nuts holding the sockets of their respective plate coils. Since it doesn't matter where the cathode resistors are grounded, they are returned to lugs under the coil sockets ahead of them. Pins 1 and 2 of the coil sockets are grounded to the lugs just mentioned, the No. 3 pins of the coil sockets for L_3, L_4 and L_5 go to the plates of their respective tubes, and the No. 4 pins of the same sockets are connected to the screen pins on the tube sockets. The grid condensers, C_3 and C_6, are tied from Pin 7 on the coil sockets to the grid pins on the tube sockets.

The oscillator and power-supply wiring on the steel chassis is conventional, with the exception of the oscillator coupling condenser. A small National TPB bushing is mounted on the chassis where it will be parallel to the lead on the grid side of R_7. This bushing is connected to the stator of C_{14} and the "hot" side of L_7 by a heavy wire, and coupling is obtained by the capacity between this bushing and the grid lead of the mixer stage. The output cable from L_6 is a length of RG-59/U 70-ohm cable.

If one of the free points on the OB-2 voltage-regulator tube socket is used as a tie-point for C_{12} and L_9, as was done in this case, be sure to clip off the pin on the tube. If this isn't done, a discharge will be obtained inside the tube, since the free pin projects inside the tube envelope and acts as an anode.

The coils for the converter are wound on Millen 74001 tuned plug-in coil forms. The coils are started on the form about ⅛ inch above the lower limit of travel of the iron slug. In the case of L_3 and L_4, one end of the winding is connected to Pin 4 and the other to Pin 7. A jumper is then run from Pin 7 to Pin 3. This jumper has the effect of tapping down the plate on the coil, since the jumper has some reactance at these frequencies. In the case of the oscillator coil, the padding condenser, C_{13}, is mounted inside the coil, although it could be mounted on the coil socket. The tickler, L_8, is wound on the form away from the slug end. The mixer output capacitor, C_{10}, is mounted on the socket. All coils are securely fastened with coil dope, and this is particularly important in the case of the oscillator coil assembly, to insure long-time stability.

After the wiring has been completed and checked, the oscillator should be checked first. Put a voltmeter across R_{14} and see if the voltage increases slightly when the grid of the oscillator tube is touched. If it does, it shows that the circuit is oscillating, and the coil can be tuned to frequency with the iron slug.

Couple the output of the converter to a communications receiver on 7.3 Mc. and adjust the slug of L_5 for maximum noise in the receiver, with power to the converter and the converter gain control at minimum. Some kind of signal will be needed with which to establish the oscillator frequency accurately, and this

COIL DATA FOR THE BAND-PASS CONVERTER

Coil	*14 Mc.*	*28 Mc.*	*50 Mc.*
L_1	13 t. No. 26 d.c.c.	8 t. No. 26 d.c.c.	5 t. No. 26 d.c.c.
L_2	35 t. No. 24 d.c.c.	23 t. No. 24 d.c.c.	8 t. No. 24 d.c.c. spaced wire diam.
L_3, L_4	25 t. No. 24 d.c.c.	8½ t. No. 24 d.c.c.	5 t. No. 24 d.c.c. spaced twice wire diam.
L_5	37 t. No. 26 enam.	Same	Same
L_6	9 t. No. 26 enam.	Same	Same
L_7	4 t. No. 24 d.c.c. spaced to occupy ¼ inch	7 t. No. 20 enam.	2 t. No. 24 d.c.c. spaced wire diam.
L_8	3 t. No. 26 d.c.c.	3 t. No. 26 d.c.c.	2 t. No. 24 d.c.c.
C_{13}	150 μμfd.	27 μμfd.	22 μμfd.

L_1 wound over ground end of L_2, tape insulation. L_8 spaced from L_7 by washer thickness. All coils close-wound unless otherwise specified. All coils wound on Millen 74001 permeability-tuned forms.

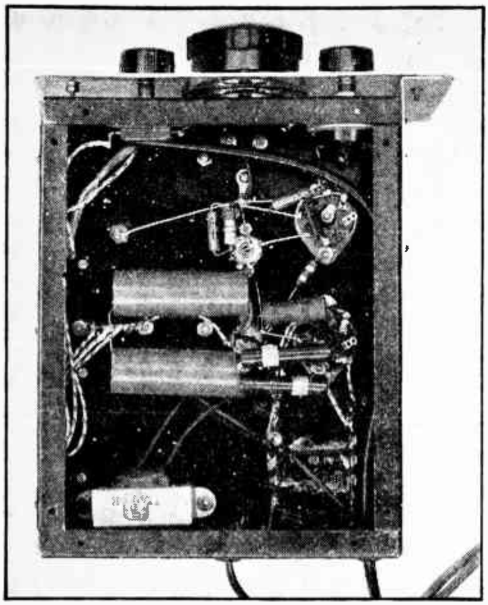


Fig. 1225 — A view underneath the chassis shows the polystyrene bushing used to couple from the oscillator to the mixer. The panel is mounted away from the chassis to simplify mounting of the dial. The tuning screws of the r.f. coil can be seen projecting through holes in the chassis.

signal can be a harmonic from the station transmitter or a test generator. For 28-Mc. alignment, set the signal source at about 28.5 Mc. and the tuning dial at 35 and adjust the slug on the oscillator coil until the signal is heard. Short the input of the receiver with a carbon resistor equal in value to the impedance of the antenna line. Having established the tuning range — and checking it at other points if available — peak L_2, L_3 and L_4 on noise. Tuning across the band, the output noise should peak near the center of the range and fall off slightly at either end. By increasing the inductance of L_4 — running the slug in — and decreasing the inductance of L_3, it will be possible to get practically uniform noise output over the entire range. It will be found that L_2 tunes very broadly when loaded by the resistor or the antenna, and its resonance should be checked with this load disconnected, to make certain that the coil can be made to tune through resonance. A sharp increase in the noise will serve as an indication, and it may be found necessary to retard the gain control for this test, to prevent oscillation in the r.f. stages.

If any queer burbles or sudden peaks of noise are encountered, it indicates regeneration in the r.f. stages. If this is encountered, the r.f. stages can be worked on while removed from the chassis, since there will be enough stray oscillator output to the mixer to receive signals, and the various plate- and heater-supply leads can be investigated with a 0.001-μfd. mica condenser until the source of feed-back is found. Poor grounds can also give trouble.

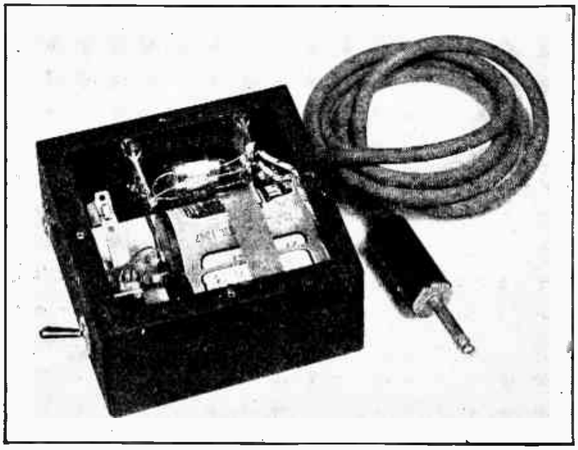
Under normal conditions, the gain of the communications receiver following the converter will have to be reduced considerably, since the gain of the converter runs around 40 db. It will be found to require very little antenna for normal pick-up, but in order to give it every break it should be used with the best antenna available. Some experiment with the input coupling may be necessary if a tuned antenna is used, but this might be only a tuned circuit with a link line running to the converter input.

An Audio Noise Limiter

If one is bothered by ignition and other pulse-type noise on the higher frequencies, the addition of a noise limiter to the output of the receiver will result in improved reception and will allow the reception of some weak signals that might otherwise be lost. The limiter shown in Fig. 1226 is plugged in to the receiver 'phone jack and the headphones are plugged into the limiter, so that no work on the receiver is required. The limiter will also keep the strength of c.w. signals at a constant comfortable level and will do much to relieve

Fig. 1226 — A crystal-diode noise limiter for use between receiver and headphones. Built in a 4 by 4 by 2-inch box, it contains the limiter crystals, bias cells, headphone jack, and on-off switch, and is provided with a cord and plug to connect to the receiver headphone output.

Although primarily intended for c.w. reception, the limiter also is highly effective on 'phone signals when the audio volume level is properly set and the r.f. gain is automatically controlled.



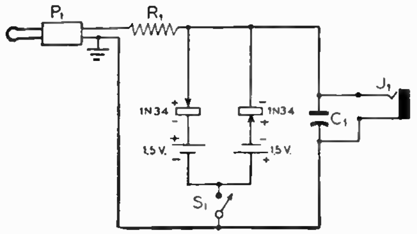


Fig. 1227 — Practical clipper circuit for headphone reception.

C_1 — 0.005-μfd.
J_1 — Single-circuit jack.
P_1 — Headphone plug.
R_1 — 15,000 ohms, ½ watt.
S_1 — S.p.s.t. toggle.

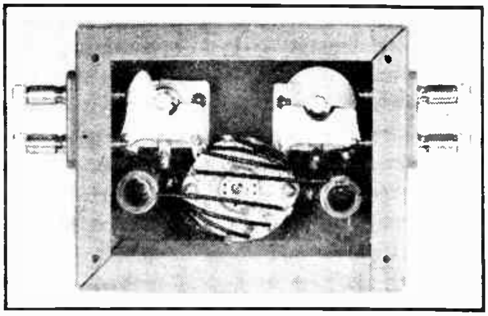


Fig. 1228 — Rear view of the antenna-coupling unit. The two coils can be seen directly below the two condensers.

the operating fatigue resulting from long hours of listening to crackles, key clicks, blocking signals and the like.

As can be seen from the wiring diagram in Fig. 1227, two 1N34 crystal diodes, individually biased by 1½-volt dry cells, are used to short-circuit any signal coming through the 'phone circuit that has an amplitude greater than about 3 volts, peak-to-peak. Hence if the audio gain of the receiver is adjusted to give a signal of this amplitude — comfortable headphone volume — noise peaks of greater amplitude will be short-circuited and not heard in the headphones. A 6AL5 twin diode can be substituted for the two 1N34 crystals, but a heater supply will be required and it is generally more convenient to build the limiter as shown. No current is drawn from the two cells used for bias, and they will last their shelf life.

The limiter can be built in a 4 × 4 × 2-inch cabinet, as shown in Fig. 1226. By removing the two sides of the cabinet, all of the components can be mounted in the frame. The two dry cells can be taped together and then held in place by heavy leads soldered to them, or special clips can be made of spring brass. The two 1N34 crystal diodes are best mounted on tie-points, and the pigtails of the diodes should be held in a pair of long-nose pliers while soldering to them, because too much heat from the soldering iron may decrease the effectiveness of the crystal. The pliers conduct the heat away that might otherwise reach the crystal.

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. A compact unit for this purpose is shown in Fig. 1228. The wiring diagram, Fig. 1229, shows that the unit is a simple pi-section coupler. By proper selection of the condenser and inductance values, a match can be obtained over a wide range of values. It can be placed close to the receiver and left connected all of the time, since it will have little or no effect on the lower frequencies. A short length of 300-ohm Twin-Lead is convenient for connecting the coupler to the receiver.

The antenna coupler is built in a 3 × 4 × 5-inch metal cabinet. All of the components except the two pairs of terminals are mounted on one panel. The condensers are mounted off the panel by the spacers furnished with the condensers, and a clearance hole for the shaft prevents any short-circuit to the panel. The coils, wound on National PRD-2 polystyrene forms, are fastened to the panel with brass screws, and the coils should be wound on the coils as far as possible away from the mounting end. If this still leaves the coil ends within ½ inch of the panel, the forms should be spaced away from the panel by National XP-6 buttons. The switch should be wired so that the

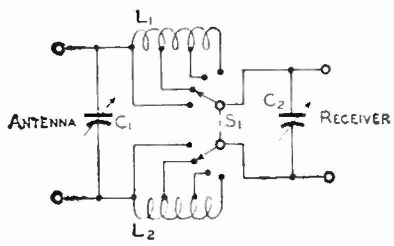


Fig. 1229 — Circuit diagram of the coupling unit.

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

L_1, L_2 — 30 turns No. 18 d.c.c. close-wound on ½-inch diameter polystyrene form, tapped at 2½, 6½ and 14½ turns.

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

switching sequence puts in, in each coil, 0 turns, 2½ turns, 6½ turns, 14½ turns and 30 turns. All of the wiring, with the exception of the input and output terminals, can be done with the panel removed from the box.

The unit is adjusted for maximum signal by switching to different coil positions and adjusting C_1 and C_2. It will not be necessary to retrim the condensers except when going from one end of a band to the other, and when the unit is not in use, as on 7 and 3.5 Mc., the coils should be switched out of the circuit and the condensers set at minimum.

Chapter Thirteen

Transmitter Construction

In the descriptions of apparatus to follow, not only the electrical specifications but also the manufacturer's name and type number have been given for most components. This is for the convenience of the builder who may wish to make an exact copy of some piece of equipment. However, it should be understood that a component of different manufacture, provided it is of equivalent quality and has the same electrical specifications, may be substituted in most cases.

A Beginner's Transmitter

One of the simplest satisfactory transmitters for amateur use is shown in the photographs of Figs. 1301 and 1303. The circuit diagram appears in Fig. 1302. The arrangement consists of a Pierce crystal oscillator capacitance-coupled to an output stage which may be used either as a straight amplifier at the crystal frequency or as a frequency doubler to deliver output at twice the crystal frequency. This combination has the advantage over a simple oscillator transmitter in that the oscillator is isolated from the effects of tuning and loading. Type 6L6, 6V6, or 6F6 tubes, or their glass equivalents, may be used in either the oscillator or amplifier with only a slight difference in performance at the supplied plate voltage.

By the use of the proper coil at L_1, output may be obtained at 3.5 or 7 Mc. with a 3.5-Mc. crystal or at 7 or 14 Mc. with a 7-Mc. crystal. The amplifier input is not tuned so that neutralization of the output stage is unnecessary. C_2 provides regeneration; its value should not depart appreciably from that specified. The output tank circuit is in the form of a pi-section filter which makes it possible to use the transmitter with a wide variety of antenna systems.

Parallel plate feed is used in the output stage to remove plate voltage from the tuning condensers and the coil. Plate voltage for the oscillator is reduced by the series resistor, R_8, while screen voltage is obtained from the voltage divider made up of R_2 and R_3. In the amplifier section, the screen voltage is obtained from the second voltage divider consisting of R_6 and R_7. Grid bias for the oscillator is obtained from the grid leak, R_1, alone, while a combination of cathode resistor (R_5) and grid leak (R_4) is used for the amplifier. A 60-ma. dial lamp serves as a resonance indicator in tuning up the transmitter.

Construction — The chassis or frame is made entirely from lattice strip, 1⅝ inches wide and ¼ inch thick. The sketch of Fig. 1304 shows how the strips are fastened together with 1-inch wire brads. The 1¼-inch spacing between the top strips is appropriate for

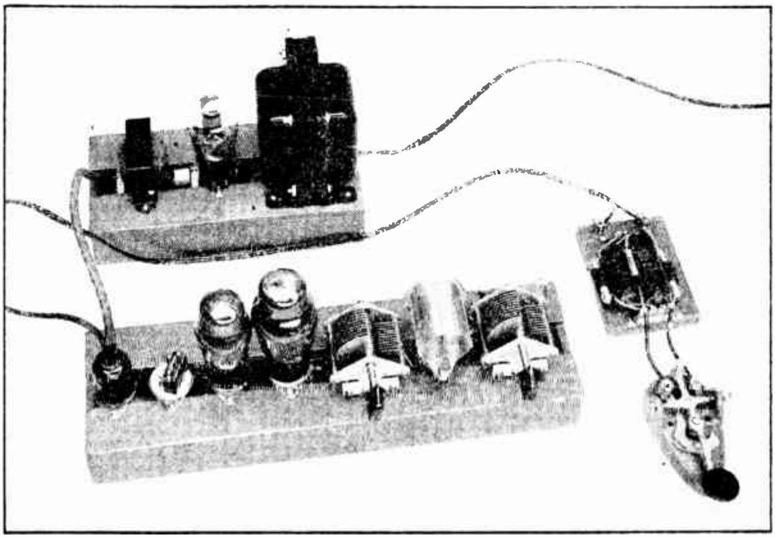


Fig. 1301 — The complete beginner's transmitter. In the r.f. unit in the foreground, left to right, are the 5-prong socket for the power plug, octal sockets for the crystal, oscillator tube and amplifier tube, and the output tank condensers, C_9 and C_{10}, with the coil L_1 in between. On the power-supply chassis at the rear are the filter choke L_2, the Type 80 rectifier tube and the power transformer. The filter condensers, C_{11} and C_{12}, and the bleeder resistor, R_9, are underneath. The key-click filter is to the right.

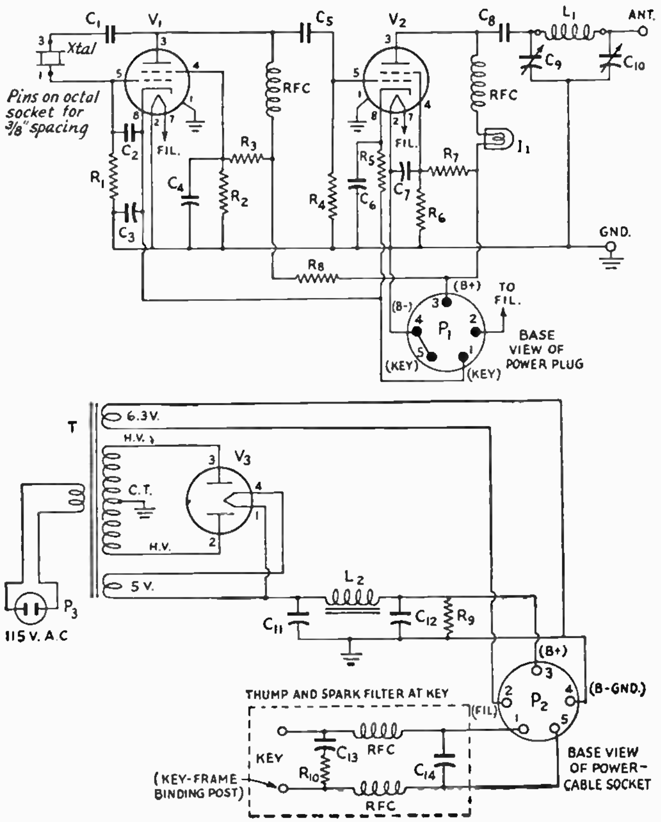


Fig. 1302 — Circuit diagram of the beginner's transmitter and power supply.

C_1, C_8 — 0.001-μfd. mica.
C_2, C_5 — 100-μμfd. mica.
C_3, C_4, C_6, C_7 — 0.01-μfd. paper.
C_9, C_{10} — 250-μμfd. variable (National TMS 250).
C_{11}, C_{12} — 16-μfd. 475-volt electrolytic.
C_{13} — 1-μfd. 400-volt paper.
C_{14} — 0.5 μfd. 400-volt paper.
R_1, R_3 — 47,000 ohms, 1 watt.
R_2, R_6 — 0.1 megohm, 1 watt.
R_4 — 22,000 ohms, ½ watt.
R_5, R_{10} — 330 ohms, 1 watt.
R_7, R_8 — 15,000 ohms, 2 watts.
R_9 — 20,000 ohms, wire-wound, 10 watts.
L_1 — 3.5 Mc.: 32 turns No. 20 d.s.c., 1½-inch diam., close-wound.
— 7 Mc.: 20 turns No. 20 enam., 1½-inch diam., 1½ inches long.
— 14 Mc.: 10 turns No. 18 enam., 1½-inch diam., 1 inch long.
(B & W JEL80, "40" or "20" coils may be substituted.)
L_2 — Filter choke, 10 hy., 130 ma. (Stancor C-2303).
I_1 — 60-ma. dial lamp.
P_1 — 5-prong chassis-mounting male plug.
P_2 — 5-prong female cable plug.
P_3 — A.c. line-cord plug.
RFC — 2.5-mh. r.f. choke.
T — Power transformer; 350 volts each side of center; 5 v., 3 amp.; 6.3 v., 4.5 amp. (Stancor P-4080).
V_1, V_2 — 6V6, 6L6, 6F6 or glass equivalents.
V_3 — Type 80 rectifier.

Millen sockets, but it can be changed to suit sockets of other dimensions, of course.

The completed chassis was given a couple of coats of grey Duco. The sockets are fastened in place by means of small wood screws and are orientated so that most-convenient connections may be made. The power-plug socket has its metal-ring key to the left, the oscillator tube socket key is to the right, the amplifier tube socket toward the front and the coil socket toward the left.

All wiring is done underneath. The ground wire is a piece of No. 14 bare wire which runs the length of the chassis from the No. 4 prong on the power-supply socket to the rotor of C_{10}. To this wire all ground connections shown in the diagram are made. Connections to by-pass condensers and r.f. chokes should be as short as possible, the by-pass condensers being connected to the nearest point on the ground wire. A pair of fiber lug strips provide anchorage for resistors and r.f. chokes. "Hot" r.f. leads (those from the plates and control grids of the tubes and the connections between the tuning condensers and the coil) should be short and direct instead of going around right-angle bends. The output terminals are a pair of Fahnestock clips fastened to the two sides of C_{10}.

Homemade coils may be constructed by

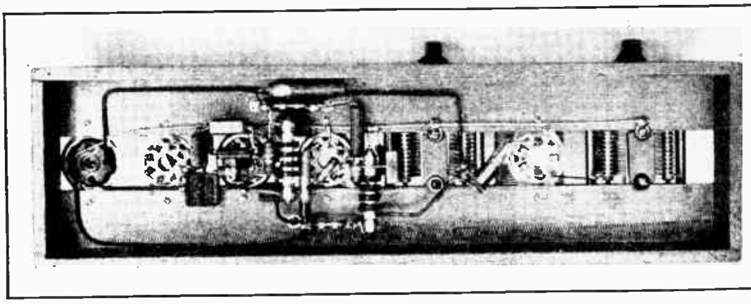


Fig. 1303 — Bottom view of the beginner's transmitter. The by-pass condensers, r.f. chokes and resistors are grouped around the tube sockets. The ground wire mentioned in the text runs along the top edge of the lower chassis strip. The indicator lamp, I_1, is wired in the B+ line just below the amplifier plate r.f. choke. It is placed underneath the chassis where it can be viewed from above through the opening between the chassis strips. The r.f. choke to the right is in the amplifier and the one to the left is in the oscillator circuit.

winding them, according to the dimensions given under Fig. 1302, on Hammarlund 1½-inch diameter 5-prong coil forms. Those shown in the photograph are the B & W JEL series. The link winding is not used.

Inexpensive components are used in the power supply. The transformer is a broadcast-receiver replacement type as are the filter components. The chassis is similar to that used for the transmitter, the only difference being in the length — 9½ inches instead of 15½ inches. The filter condensers, and the bleeder resistor, R_9, are placed underneath.

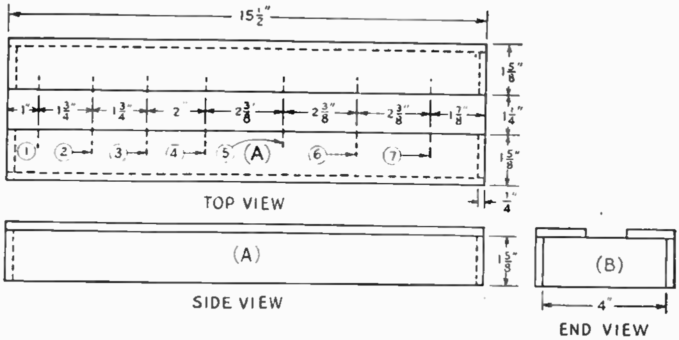


Fig. 1304 — Sketch showing the important dimensions of the beginner's transmitter chassis. The center lines are numbered as follows: 1 — power plug, 2 — crystal socket, 3 — oscillator-tube socket, 4 — amplifier-tube socket, 5 — tuning condenser, C_9, 6 — coil socket, 7 — coupling condenser, C_{10}.

The key-click filter is a separate unit assembled on a small piece of ¼-inch wood. The connecting leads and the leads to the key should be short if the filter is to be effective. The side of the filter connected to power-plug Pin 5 should be connected to the frame of the key.

Adjustment — The transmitter should first be tuned up without the antenna connected. It should be remembered that only the second harmonic of crystals between 3500 and 3650 kc. and between 7000 and 7200 kc. are useful in the higher-frequency amateur bands. With a suitable crystal and coil plugged in, the power supply may be plugged in and the key closed after allowing time for the heaters of the tubes to come up to temperature. The indicator lamp should glow brightly when the key is closed. Setting C_{10} at about half capacitance, C_9 should be adjusted as I_1 is watched for a dip in illumination. If this dip cannot be found anywhere within the range of C_9, another setting of C_{10} should be tried. As soon as the dip has been found, the antenna may be connected, and the tuning process repeated as before. With the antenna connected the dip at resonance will not be so pronounced. In fact, when the amplifier is loaded properly, the dip should be just noticeable — just enough to indicate that the output circuit is tuned to resonance. The proper loading point may be found by adjusting C_{10} at several fixed settings and rotating C_9 through its range for each setting of C_{10}. As the proper point is approached, the capacitance of C_{10} should be adjusted in smaller steps. In most cases the loading will increase as the capacitance setting of C_{10} is decreased. Near maximum loading, the adjustment is fairly critical. With antennas of certain dimensions, it may be necessary to short-circuit a few turns on L_1 to obtain maximum loading in the 3.5-Mc. band with the B & W coil.

While the best antenna within the limits of cost and space should be used, the output circuit provides means of feeding power into a wire of random length; it is not necessary that its length be a multiple of a half wavelength. With the power supply described, an output of about 10 watts should be possible at the crystal fundamental; and 5 or 6 watts when the output stage is used as a frequency doubler. If a milliammeter is connected in series with the key, it should show a reading of about 20 ma. with the amplifier tuned to resonance and unloaded at the crystal fundamental and about 40 ma. when doubling. Loaded, the plate current should run between 70 and 80 ma. With a power-supply voltage of 350, the oscillator plate voltage should be 170, the oscillator screen voltage 90 and the amplifier screen voltage 220 with the amplifier loaded and tuned to resonance.

A Self-Contained 60-watt Transmitter for 3 Bands

The diagram of Fig. 1307 shows the circuit of a simple two-stage transmitter. The rig, shown in Fig. 1305, is enclosed in a cabinet, complete with power supply and antenna tuner.

A 6V6GT Tri-tet oscillator drives an 807 output stage directly with simple capacitive coupling. Any one of ten crystals may be selected from the front of the panel by the crystal switch, S_1. A pair of terminals also is provided at the rear for VFO connection. Bands are changed by means of a system of plug-in coils.

The oscillator circuit operates with either 3.5- or 7-Mc. crystals. In either case, oscillator output may be obtained at the crystal fundamental frequency or its second harmonic. While the output stage may be used as a frequency doubler with fair efficiency, this sort of operation is not recommended unless the unit is to be used as an exciter for a following amplifier.

Parallel plate feed is used in both stages to permit mounting the tuning condensers, C_2 and C_3, directly on the metal chassis without insulation. The v.h.f. choke RFC_2 and the screen resistor, R_7, are necessary to suppress h.f. parasitic oscillations.

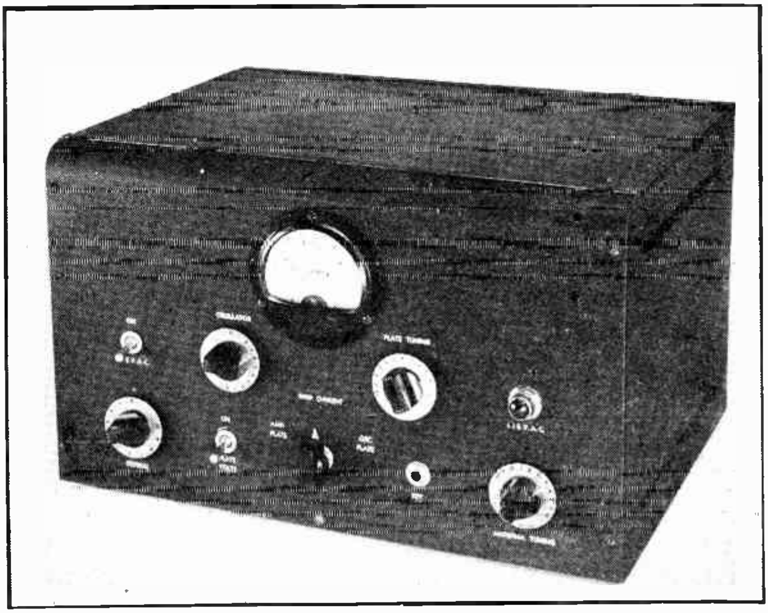


Fig. 1305 — A two-stage low-power transmitter for three bands. To either side of the milliammeter are the oscillator and amplifier plate-tuning controls. Along the bottom are the crystal switch, the plate-voltage switch, the meter switch, the key jack and the antenna tuning control.

The s.p.d.t. toggle switch, S_2, makes it possible either to key both stages simultaneously for break-in work on the lower frequencies, or the output stage alone at 14-Mc. frequencies where oscillator keying chirp may become noticeable. The unit includes a link-coupled antenna tuner, L_4C_4.

The self-contained power supply is built around an inexpensive multiwinding transformer, T_1. The separate filament transformer, T_2, makes it possible to cut off the plate voltage without turning off the heaters of the tubes. A condenser-input filter is used to boost the output voltage to 600 under load. Voltage for the plate of the oscillator and the screen of the 807 is kept from soaring when the key is open by a pair of voltage-regulator tubes. This operating voltage of 250 is dropped to 150 volts for the screen of the 6V6GT by the series resistor, R_3.

The milliammeter may be switched to read oscillator plate current and 807 grid or plate current by the double-gang switch, S_3, which connects the meter across the shunting resistors, R_4, R_6 and R_8. R_4 and R_8 are adjusted to multiply the 10-ma. basic meter-scale reading by 10 and 20, making the full-scale reading 100 and 200 ma. respectively when checking plate currents, while the resistance of R_6 is sufficiently high to have negligible effect upon the meter reading when measuring the grid current of the amplifier.

Construction — Reference should be made to the photographs of Figs. 1305 through 1310 for constructional details. The transmitter is built on a 10 × 14 × 3-inch chassis which fits a standard 9 × 15 × 10¾-inch cabinet. The r.f. section occupies the front half of the chassis, while the power-supply components are lined up at the rear.

All tube and coil sockets are submounted. The cathode coil, L_1, requires a 4-prong socket; octals are needed for the 6V6GT, the oscillator plate coil, L_2, the rectifier and the two VR tubes; L_3 and L_4 require 5-prong sockets.

The oscillator and amplifier groups are separated by a small baffle shield cut from sheet aluminum. It is 4 inches high and 5 inches long and has a cut-out in front for the meter. It is spaced 8 inches in from the right-hand end of the chassis. The line of ten Millen crystal

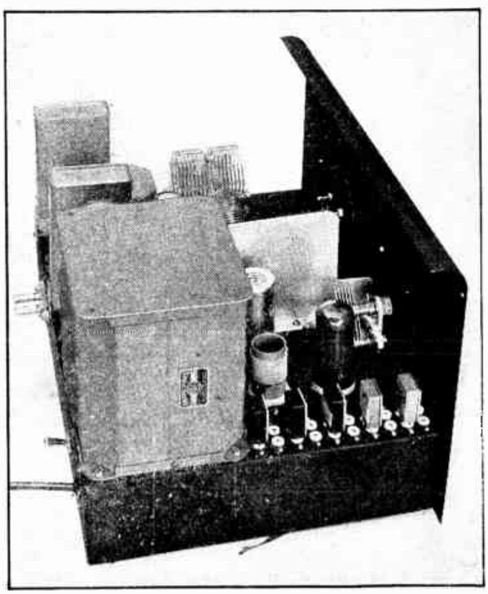


Fig. 1306 — The oscillator section of the low-power transmitter, showing the line of crystal sockets, the cathode coil, the shielded plate coil and the 6V6GT.

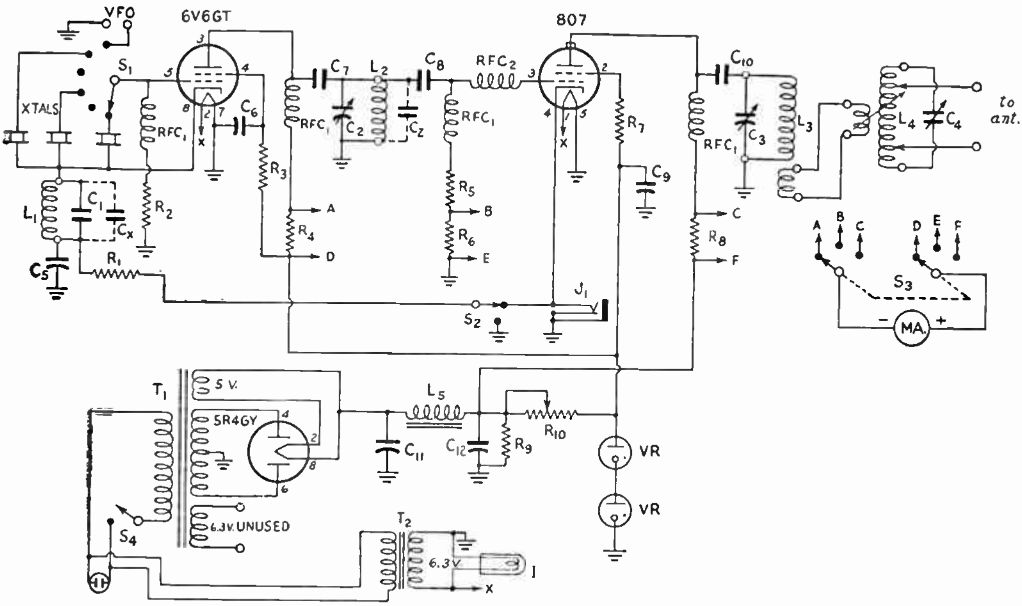


Fig. 1307 — Circuit diagram of the 3-band low-power transmitter.

C_1, C_8 — 100-μμfd. mica.
C_x — 100-μμfd. mica (see text).
C_2 — 50-μμfd. variable (National ST-50).
C_z — 22-μμfd. mica (see text).
C_3, C_4 — 150-μμfd. variable (National ST-150).
C_5, C_6, C_9 — 0.01-μfd. paper.
C_7, C_{10} — 0.001-μfd. mica.
C_{11}, C_{12} — 4-μfd. 1000-volt paper.
R_1 — 220 ohms, 1 watt.
R_2 — 47,000 ohms, ½ watt.
R_3 — 40,000 ohms, 5 watts.
R_4 — 100-ma. meter shunt (see text).
R_5 — 15,000 ohms, 1 watt.
R_6 — 47 ohms, ½ watt.
R_7 — 47 ohms, 1 watt.
R_8 — 200-ma. meter shunt (see text).
R_9 — 50,000 ohms, 25 watts.
R_{10} — 10,000 ohms, 25 watts.
L_1 — Oscillator cathode
 1A (3.5-Mc. crystals) — 14 turns No. 22 d.c.c., 1-inch diam., ⅞ inch long, 100-μμfd. mica, C_x, connected in parallel.
 1B (7-Mc. crystals) — 10 turns No. 22 d.c.c., 1-inch diam., ⅞ inch long.
L_2 — Oscillator plate
 2A (3.5 Mc.) — 80 turns No. 26 d.s.c., ½-inch diam., close-wound, C_z connected in parallel.
 2B (7 Mc.) — 40 turns No. 24 d.c.c., ½-inch diam., close-wound.
 2C — (14 Mc.) 25 turns No. 18 d.c.c., ½-inch diam., 1⅜ inches long.
L_3 — Amplifier plate
 3A (3.5 Mc.) — 24 turns 1½-inch diam., 1⅜ inches long (B & W JEL80 with 16 turns removed), 3-turn link.
 3B (7 Mc.) — 18 turns 1½-inch diam., 1¾ inches long (B & W JEL40), 2-turn link.
 3C (14 Mc.) — 12 turns 1½-inch diam., 2 inches long (B & W JEL20), 2-turn link.
L_4 — Antenna coil
 4A (3.5 Mc.) — 30 turns 1¾-inch diam., 2 inches long, 3-turn variable link at center (B & W JVL80 with 5 turns removed from each end).
 4B (7 Mc.) — 24 turns 1¾-inch diam., 2⅜ inches long, 3-turn link at center (B & W JVL40).
 4C (14 Mc.) — 14 turns 1¾-inch diam., 2¼ inches long, 3-turn link at center (B & W JVL20).
L_5 — 6-henry 175-ma. filter choke.
I — 6.3-volt signal lamp.
J_1 — Closed-circuit jack.
MA — 0–10 ma. meter.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 14 turns No. 20, 5⁄16-inch diam., ¾ inch long.
S_1 — 11-point tap switch, ceramic insulation.
S_2 — S.p.d.t. toggle.
S_3 — Double-gang 3-position rotary switch.
S_4 — S.p.s.t. toggle.
T_1 — 600 volts each side of center, 200 ma.; 5 volts, 3 amp. (UTC S-41).
T_2 — 6.3 volts, 3 amp. (UTC S-55).
VR — Voltage-regulator tubes — VR-150 and VR-105 types in series to give 255 volts.

sockets is placed as close to the left-hand edge of the chassis as possible. Each of these requires two clearance holes and a mounting-screw hole between.

Alongside the crystal row are the 6V6GT oscillator tube and its cathode coil, L_1, followed by the plate coil, L_2, and the oscillator tuning condenser, C_2. The latter is mounted directly on the chassis 4⅝ inches from the left-hand edge. The oscillator grid and plate chokes are mounted underneath.

On the other side of the baffle shield are the 807 with its plate-circuit choke and blocking condenser, C_{10}, the output tank condenser and coil, C_3 and L_3, and the antenna-coupler coil, L_4. The antenna tuning condenser, C_4, is mounted under the chassis. The socket for the 807 is spaced as far below the chassis level as possible, without protruding from the bottom, by means of brackets cut from strip metal. The purpose of this is to provide a shield between the input and output sections of the tube. A 1⅞-inch hole is required to clear the tube envelope. C_3 is mounted directly on the chassis with its shaft 4⅝ inches from the right-hand end of the chassis to balance the shaft of the oscillator plate-tank condenser.

The antenna tuning condenser, C_4, must be

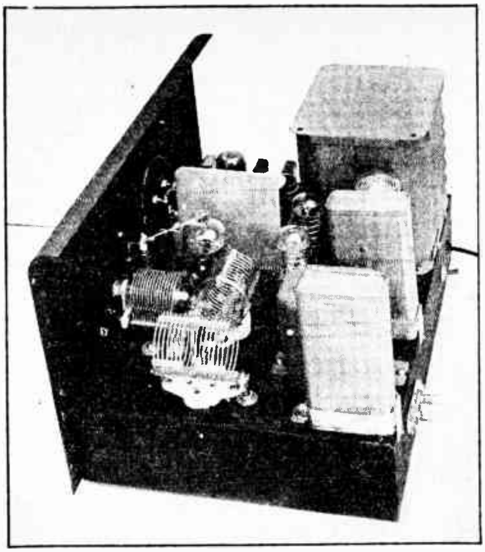


Fig. 1308 — Looking into the amplifier end of the 807-transmitter chassis. The 807 socket is spaced below the chassis to provide shielding between the input and output sections. The coil in the foreground is in the antenna tuner, while the one behind it is the amplifier plate tank coil.

insulated from the chassis. This is done by means of an aluminum angle bracket and a pair of polystyrene feed-through buttons. The condenser is placed so that its shaft comes 1 5/8 inches from the end of the chassis to balance the shaft of the crystal switch at the opposite end. The antenna coil is mounted at right angles to L_3.

The meter switch, S_3, is mounted at the center between the front edge of the chassis and the bottom part of the 807. The key jack and power switch, S_4, are spaced equally to either side of the center of the front edge of the chassis.

The power-supply components are placed as close as possible to the rear edge of the chassis, with the transformer T_1 at the left followed by the rectifier and voltage-regulator tubes, the input condenser, C_{11}, the filter choke, L_5, and the output condenser. A large cut-out is required for the transformer terminals and if filter condensers of the type shown are used, holes for the terminals must be provided in addition to the mounting-screw holes. The leads to the filter choke are fed down through a grommet-lined hole next to the choke. The key switch, S_2, and the antenna terminals are mounted in the rear edge of the chassis where the power cord also enters.

Underneath the chassis, the power wiring was done first, keeping it bunched and close to the chassis wherever possible. The separate filament transformer, T_2, is fastened to the left-hand end of the chassis. By-pass condensers and r.f. chokes should be placed close to the tube terminals to which they connect. The by-pass condensers should be grounded to the chassis at the nearest available point. The coupling and blocking condensers, C_7, C_8 and C_{10}, should be well spaced from the chassis. The same applies to all r.f. wiring, which should also be kept short and direct between points of connection. The length of leads to resistors is not important. In some cases it may be convenient to use fiber lug strips as anchorages or supports for small resistors and r.f. chokes.

The meter shunts, R_4, R_6 and R_8, are mounted directly on the meter switch. R_4 and R_8 are made from No. 30 magnet wire. Approximately 7 feet will be required for R_8 and 14 feet for R_4. Before the meter is mounted in the panel, it should be connected in series with a 3-volt battery and a variable resistance of about 500 ohms. A resistor with a slider will serve the purpose if no other is available. The resistance should be adjusted until the meter reads full scale. When the shunting wire, cut to a length of two or three feet more than that required is connected across the meter terminals, the reading will drop. The length of the wire should be adjusted, bit by bit, until the reading drops to 1 ma. for R_4 and to 1/2 ma. for R_8. The wire then may be wound on a small form for compactness. A 1/2-watt resistor of 100 ohms or more makes a good form and its resistance does not affect the calibration of the shunt to any practical degree.

The link line between the output tank circuit and the antenna tuner, and the connections between the latter and the antenna terminals at the rear, should be made with rigid wire spaced well away from the chassis and surrounding components.

Coils — The output and antenna tank coils, L_3 and L_4, are of the B & W JEL and JVL series respectively.

Some of these require pruning, as indicated in the coil table, to provide the correct L/C ratio. The antenna-tuner coil, L_4, requires an extra pair of contacts for the tap leads. Since a center-tap is not required, it may be cut free from the base pin so that this pin may be used for one of the tap contacts. The other tap contact is provided by drilling out the tubular rivet at one of the ends of the coil-supporting base strip and substituting a banana plug as shown in Fig. 1309. A jack for this plug then is mounted in the chassis close to the coil socket

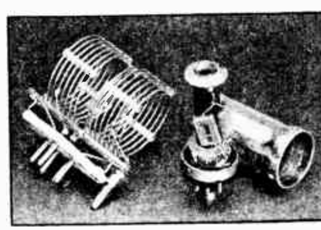


Fig. 1309 — The antenna coil for the 2-stage transmitter requires the addition of an extra contact which is provided by the banana plug. To the right is the 3.5-Mc. oscillator plate coil with the mica padding condenser connected across the winding.

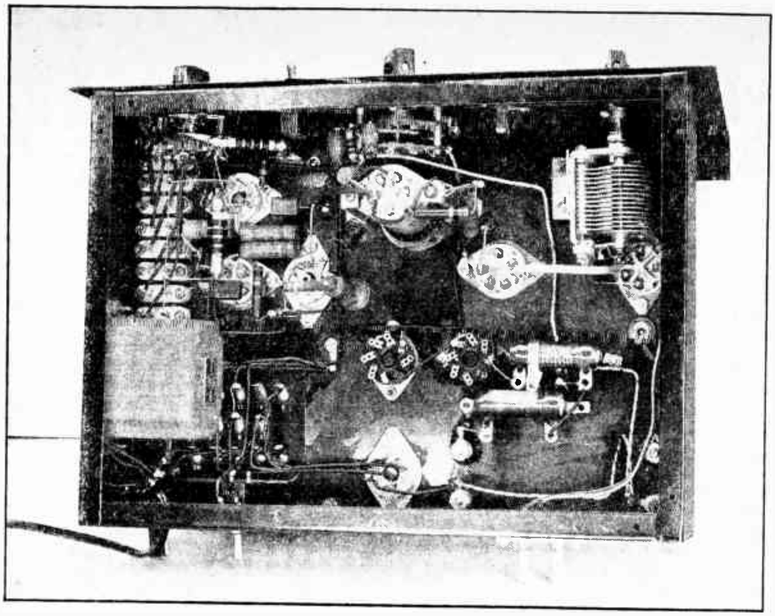


Fig. 1310 — Bottom view of the low-power transmitter, showing the mounting of the 807 socket at the upper center and the location of by-pass condensers, resistors and r.f. chokes. The separate filament transformer is fastened to the left-hand edge of the chassis. The antenna tuning condenser is in the upper right-hand corner, supported on an aluminum angle bracket which is insulated from the chassis by polystyrene buttons.

by drilling out a pair of polystyrene button-type feed-through insulators to fit the jack and setting them in the chassis.

The two cathode coils for L_1 are wound on Millen 4-prong 1-inch forms. The one to be used with 3.5-Mc. crystals requires a 100-μμfd. mica condenser, C_X, connected across it in addition to C_1. This condenser is mounted inside the form so that it is connected in the circuit along with the coil when the latter is plugged in.

The oscillator plate coils are wound on Millen octal-base shielded plug-in forms. If the forms are of the type with iron-core slugs, these should be removed. The 3.5-Mc. coil requires an extra padding condenser, C_Z, of 22 μμfd. This may be a mica condenser soldered across the winding as shown in the photograph of Fig. 1309.

Adjustment — Since the tuning of the cathode tank circuit is fixed, only three circuits, including the antenna circuit, need adjustment. The coil table shows which coils should be plugged in to obtain output depending upon the crystal frequency and the output frequency desired. For initial testing it is well to use a combination giving output in the 3.5- or 7-Mc. band. Before turning on the power supply, a key connected to a plug should be inserted in the key jack and the key switch, S_2, should be thrown to the amplifier-keying side. This will permit the oscillator to operate alone.

COIL TABLE — 60-WATT RIG

Xtal f.	*Output f.*	L_1	L_2	L_3	L_4
3.5 Mc.	3.5 Mc.	1A	2A	3A	4A
3.5 Mc.	7 Mc.	1A	2B	3B	4B
7 Mc.	7 Mc.	1B	2B	3B	4B
7 Mc.	14 Mc.	1B	2C	3C	4C

When the power plug is inserted, the heaters of the tubes should warm up. The VR tubes should glow as soon as the power switch, S_4, is closed. If they do not, the resistance of R_{10} should be reduced until they do.

With the high voltage applied and the meter switched to the first position for oscillator plate current, the meter should read between 35 and 50 ma. As C_2 is adjusted, a point will be found where the plate current dips to a minimum (between 10 ma. and 30 ma. depending upon the frequency), rising on either side. If L_2 has been made close to specifications, this resonance point should be found with about 60 per cent of maximum capacitance in use at C_2 for 3500 kc., 70 per cent for 7000 kc. and 30 per cent for 14,000 kc. If the plate circuit is tuned to a harmonic of the crystal frequency, the increase in current either side of the minimum should be smooth. However, if the plate circuit is tuned to the crystal frequency, the plate current may jump suddenly to a high value when it is tuned to the high-capacitance side of the minimum plate-current point. This indicates that the circuit has stopped oscillating. C_2 should be set sufficiently to the low-capacitance side of the minimum to insure reliable starting of the oscillator when the power is switched on or when the amplifier is keyed.

When VFO input is used, the cathode tank circuit should be shorted out. Otherwise the adjustment is the same except that the oscillator plate circuit may be tuned for maximum amplifier grid current at the fundamental as well as at the harmonic.

The amplifier should be tuned up first with the antenna coil out of its socket. With the meter switched to the second position where it reads amplifier grid current, a reading of 3 to 9

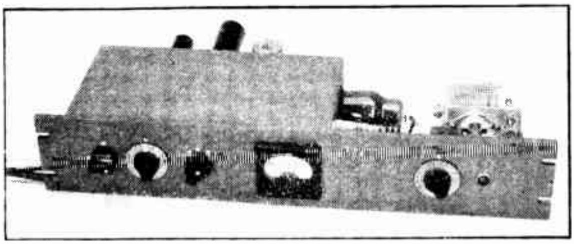


Fig. 1311 — Front view of the plug-in coil transmitter-exciter. The crystal-switch knob is at the left, 2½ inches in from the end of the panel, the dial for the buffer-doubler tuning condenser next, 4¼ inches in, followed by the meter-switch knob, 7 inches in from the edge. The meter is at the center and the output tank-condenser control at the right, 4¼ inches from the end of the panel.

ma. should be obtained when the key is closed. If no grid-current reading is obtained, it is probable that the oscillator stopped when the key was closed. In this case, the tuning of the oscillator should be readjusted. In this instance, at least, it has been found that best keying is obtained when the oscillator plate circuit is detuned to the low-capacity side of resonance to a point where the oscillator plate current remains constant with the key open and closed. This refers only to amplifier keying when the oscillator plate circuit is tuned to the crystal fundamental, of course. Readings of 5 to 10 ma. or more should be obtained in all cases. The key should not be held closed for periods longer than necessary to obtain the reading, until the amplifier plate circuit is tuned to resonance.

With the meter switch thrown to the last position, where it reads amplifier plate current, a reading of 100 ma. or more should be obtained. As C_3 is turned through its range the plate current should dip to a minimum of between 10 and 15 ma. With the L_3 coils altered as indicated in the coil table, resonance should occur at approximately 90 per cent for 3500 kc., 30 per cent for 7 Mc. and 15 per cent for 14 Mc.

The antenna should now be connected to the antenna terminals and the antenna coil plugged in. The adjustable link of the antenna coupler should be swung about half-way out and the taps should be placed on the outside turns of L_4. With the key closed, C_4 should be swung through its range. At some point the amplifier plate current should increase to a maximum, decreasing on either side. Leaving C_4 at the point where maximum plate current is obtained, C_3 should be readjusted for a minimum point which, of course, will be higher than the unloaded minimum obtained before. The adjustments of C_3 and C_4 should be juggled around until a point is reached where any change in C_3 will cause an increase in plate current, while any adjustment of C_4 will cause a decrease in plate current. If the plate current at this point is less than the maximum rated plate current for the tube, the link coupling should be closed up. If it is greater than 100 ma., the coupling should be reduced. If it is found that the link adjustment is insufficient to bring the plate current to the desired value, the taps should be moved in a turn at a time, keeping them always equidistant from the ends of the coil. It should be remembered that the tap adjustments as well as any change in the position of the link may affect the tuning of the amplifier plate circuit, so it should be retuned to obtain minimum plate current as a final adjustment. This minimum should, of course, be the rated plate current of 100 ma. when the amplifier is fully loaded. The dip in plate current at resonance naturally will be very slight when the amplifier is operating under full load.

A 75-Watt Plug-In Coil Transmitter-Exciter

The compact 75-watt transmitter unit shown in the photographs of Figs. 1311 and 1312 consists of three stages. The circuit diagram appears in Fig. 1313. A 6V6 Pierce crys-

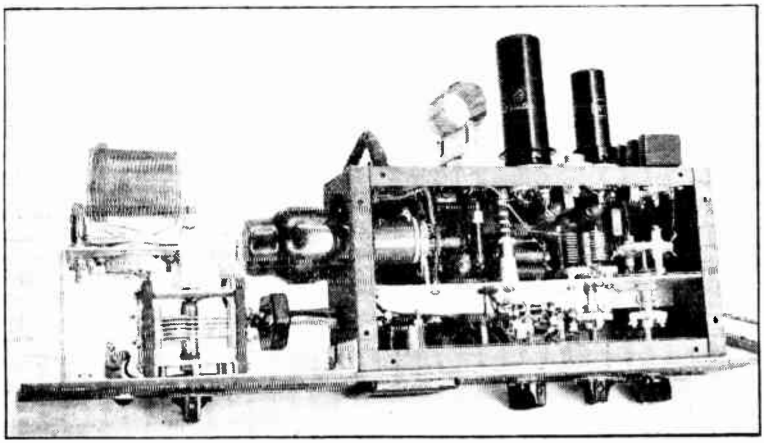


Fig. 1312 — Bottom view of the compact 75-watt transmitter-exciter. The output tank-circuit components are to the left. The chassis to the right is divided by the aluminum subpanel to which the 807 socket, C_1, and the crystal switch are attached.

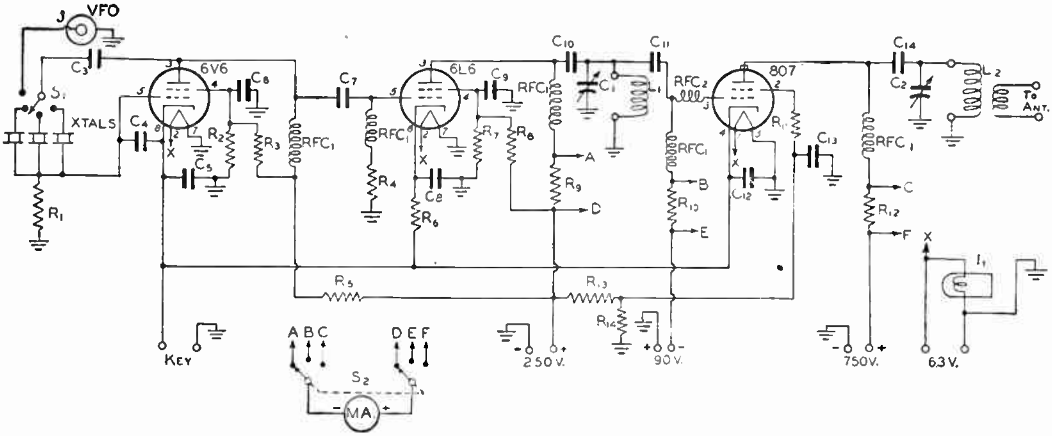


Fig. 1313 — Circuit diagram of the 75-watt plug-in-coil transmitter-exciter.

C_1 — 100-μμfd. variable (Millen 22100).
C_2 — 250-μμfd. variable (Cardwell XR-250-PS).
C_3, C_{10}, C_{14} — 0.001-μfd. mica.
C_4, C_7, C_{11} — 100-μμfd. mica.
C_5, C_6, C_8, C_9, C_{12}, C_{13} — 0.0047-μfd. mica.
R_1, R_3 — 47,000 ohms, 1 watt.
R_2, R_4, R_7, R_{14} — 0.1 megohm, 1 watt.
R_5, R_8 — 15,000 ohms, 2 watts.
R_6 — 330 ohms, 1 watt.
R_9 — "10 times" multiplier, copper-wire meter shunt (see text).
R_{10} — 47 ohms, 1 watt.
R_{11} — 47 ohms, carbon, noninductive, 1 watt.
R_{12} — "20 times" meter shunt (see text).
R_{13} — 2500 ohms, 10 watts.
L_1 — 3.5 Mc. — 15 μhy. (National AR16-40E).
7 Mc. — 5 μhy. (National AR16-20E).
14 Mc. — 2 μhy. (National AR16-10E).
L_2 — 3.5 Mc. — 12 μhy. (B & W JEL40).
7 Mc. — 4 μhy. (B & W JEL20).
14 Mc. — 3 μhy. (B & W JEL15).
28 Mc. — 0.5 μhy. (B & W JEL10 pruned to 4 turns to tune to band).
J — Concentric-cable connector.
MA — Milliammeter, 0–10 scale.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 11 turns No. 20, 5/16-inch diam., 3/4 inch long.

tal oscillator with a crystal-switching system drives a 6L6 buffer-doubler which, in turn, drives an 807 in the output stage, which may be used either as a straight amplifier or as a second frequency doubler. The milliammeter may be switched to read buffer-doubler plate current, amplifier grid current or amplifier plate current. Plate voltage for the oscillator and buffer-doubler stages is obtained from a 250-volt power supply which also provides screen voltage for all tubes through individual voltage dividers. The output stage requires a separate 600- to 750-volt plate supply. R_9, R_{10} and R_{12} are shunts across which the meter is switched. The resistance of R_{10} is high enough to have negligible effect upon the meter reading. R_9 and R_{12}, however, are of lower resistance to give a scale multiplication of 10 in the case of R_9 and 20 in the case of R_{12}. Since the 807 grid current is small, batteries form the most convenient source of biasing.

Construction — The transmitter is built as a standard rack unit with a 3½-inch panel. At the left-hand end is a 5 × 10 × 3-inch chassis which houses everything except the output tank circuit. At the right-hand end of the back edge of the chassis, as shown in Fig. 1312, is a vertical row of three Millen crystal sockets. There is space for two additional sockets if they are desired. The crystal sockets are followed, from right to left, by the 6V6 oscillator tube, the 6L6 buffer-doubler and its tank coil, L_1. The coil socket is mounted flush on the metal by cutting clearance holes for the terminals in the chassis. Between the two tube sockets is an Amphenol cable connector for VFO input and the power cable enters the chassis at the left-hand end. A separate well-insulated wire is brought out for the plate voltage for the 807.

The upper half of the 807 protrudes from the left-hand end of the chassis in Fig. 1312. Its socket is mounted on metal angle pieces fastened to the back of the chassis and the aluminum subpanel which partitions the chassis. By-pass condensers, resistors and r.f.

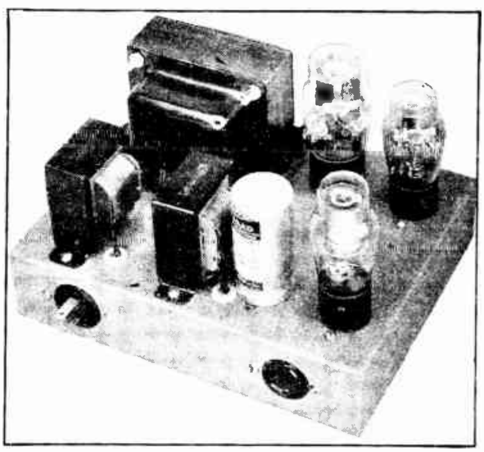


Fig. 1314 — A combination power-supply unit delivering 250 or 300 volts for exciter plate supply and 75 volts of fixed bias. If desired, the components may be combined for a high-voltage plate supply on a single chassis. The circuit diagram of the combination unit is shown in Fig. 1315.

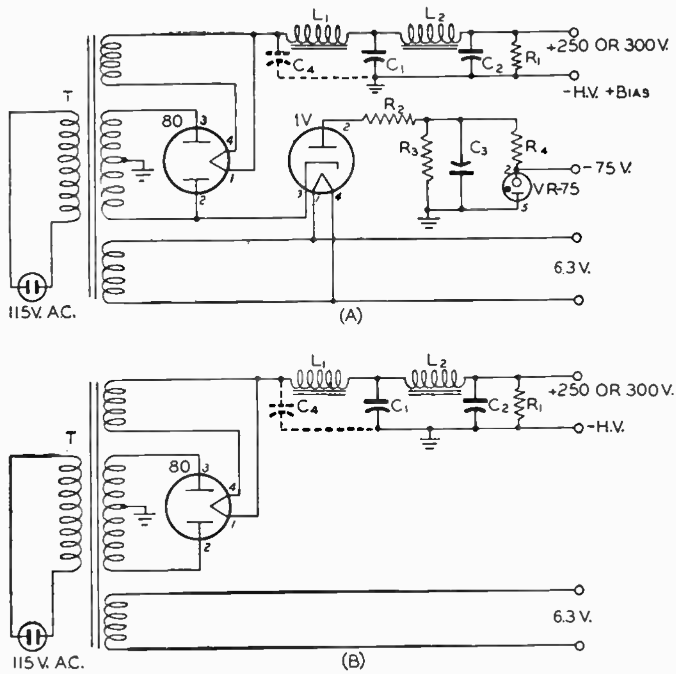


Fig. 1315 — Circuit diagram of the combination plate, screen and grid-bias power supply in Fig. 1314.

C_1, C_2 — Sections of 8-μfd. 450-volt dual electrolytic.
C_3 — 8-μfd. 450-volt paper.
C_4 — Same as C_3 (used only for 300-volt output).
R_1 — 20,000 ohms, 10 watts.
R_2, R_3 — 22,000 ohms, 2 watts.
R_4 — 15,000 ohms, 2 watts.
L_1, L_2 — 6-hy. 80-ma. 138-ohm filter choke (Thordarson T-57C51).
T — 300 volts r.m.s. each side of center-tap, 90 ma.; 5 volts, 3 amp.; 6.3 volts, 3.5 amp. (Thordarson T-13R13).

If desired, the bias branch may be omitted, as shown in the alternative diagram at B. All values remain as above.

chokes are mounted close to the sockets of the tubes with which they are associated and this wiring is done before inserting the subpanel. The subpanel carries the crystal switch, the buffer-doubler tank condenser, C_1, the plate choke for the 6L6 and one of the two angle pieces supporting the 807 socket. The milliammeter and the meter switch are mounted on the front panel with clearance holes cut in the front edge of the chassis. Flexible shaft couplings connect the crystal switch and the buffer-doubler tuning condenser with their control knobs. The back of the left-hand end of the panel (Fig. 1312) is covered with a sheet of aluminum and the output tank condenser is mounted directly on this. An aluminum bracket, fastened to the panel at one end and to the rear of the condenser at the other, supports the socket for the tank coil, L_2, and the link output terminals. The plate r.f. choke and blocking condenser, C_{14}, are just below the 807 in Fig. 1312.

The meter-shunting resistors, R_9 and R_{12}, are wound with No. 30 copper wire, around a small-diameter form. The proper length of wire may be determined by adjusting a variable resistance in series with 1.5 or 3 volts of battery until it reads full scale and then shunting various lengths of the No. 30 wire across the meter terminals until the meter-reading drops to one-tenth in the case of the 10-times shunt and to one-twentieth of full-scale reading for the 20-times shunt, remembering that the shorter the shunting wire, the lower the meter will read when shunted.

The adjustment of the transmitter is simply a matter of plugging in the proper coils and crystal for the desired output frequency and tuning the two tank circuits to resonance. In some instances, it may be possible to find two points of resonance, one at the fundamental and one at the second harmonic, but these can be identified by noting whether the condenser is near maximum or minimum capacitance.

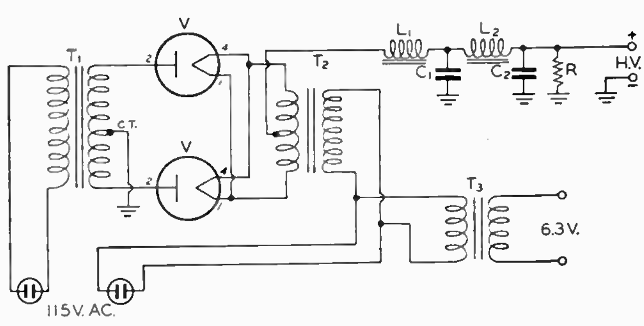


Fig. 1316 — Circuit of the power supply in Fig. 1317.

C_1 — 2-μfd. 1000-volt paper (Sprague OT21).
C_2 — 1-μfd. 1000-volt paper (Sprague OT41).
R — 20,000 ohms, 50 watts.
L_1 — Input choke, 6–19 hy., 300 ma., 125 ohms (Kenyon T-510).
L_2 — Smoothing choke, 11 hy., 300 ma., 125 ohms (Kenyon T-166).
T_1 — 925 or 740 volts r.m.s. each side of center-tap, 300-ma. d.c. (Kenyon T-656).
T_2 — 2.5 volts, 10 amp., 2000-volt insulation (Kenyon T-352).
T_3 — 6.3-volt 3-ampere filament transformer.
V — Type 866 jr. rectifier.

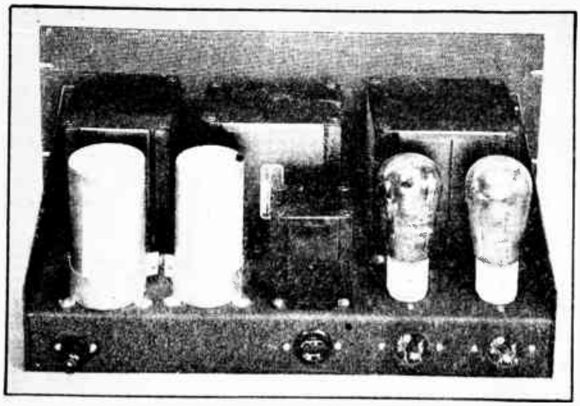


Fig. 1317 — This power-supply unit delivers either 620 or 780 volts at full-load current of 260 ma. with 0.4-per-cent ripple and regulation of 22 per cent. Voltage is changed by a tap on the plate-transformer primary winding. The filter chokes are at the left and the plate power transformer at the right on the panel side of the chassis. The can-type 1000-volt filter condensers are at the left in front and the rectifier tubes at the right, with the rectifier filament transformer in between. All exposed component terminals are underneath the chassis. The panel is 8¾ × 19 × 3 inches. The 2.5-volt 10-ampere rectifier filament transformer should have 10,000-volt insulation. A 6.3-volt filament transformer is included for heating the filaments of r.f. tubes. This transformer is mounted underneath the chassis; its output terminals are brought out to a standard a.c. receptacle in the rear. The circuit diagram is shown in Fig. 1316.

With a 250-volt supply, the combined plate and screen current of the 6L6 should be about 20 ma. when working as a straight amplifier and 30 ma. when operating as a doubler. The maximum plate current of the unloaded 807 will vary between 50 and 60 ma. when the tube is doubling frequency and between 10 and 15 ma. when working as a straight amplifier. When the stage is operated at 750 volts, and loaded to a plate current of 100 ma., the grid current should run at least 3 ma. as a straight amplifier or 6 ma. as a doubler.

The supply shown in Figs. 1314 and 1315 will provide 250 volts for the first two stages and bias for the grid of the 807. The 750-volt supply shown in Figs. 1316 and 1317 may be used for the output stage.

A Combination Low-Voltage Plate or Screen Supply and Fixed-Bias Pack

Figs. 1314 and 1315 illustrate a combination pack which will deliver 250 or 300 volts, 75 ma., for supplying plate voltage for receiving-tube exciter stages as well as screen and fixed-bias voltage for a beam-tube driver stage.

The circuit diagram is shown in Fig. 1315-A. In addition to the usual full-wave rectifier circuit employing a Type 80 tube, a 1V half-wave rectifier also is connected across one half of the transformer secondary in reverse direction to provide a negative biasing voltage which is held constant at 75 volts by the VR-75-30 regulator tube. With the dropping resistor shown, the regulator tube will pass a grid current of 25 ma. without overload. The 1V rectifier is indirectly heated, so that it may be operated from the same 6.3-volt winding provided to supply the r.f. tubes in the transmitter.

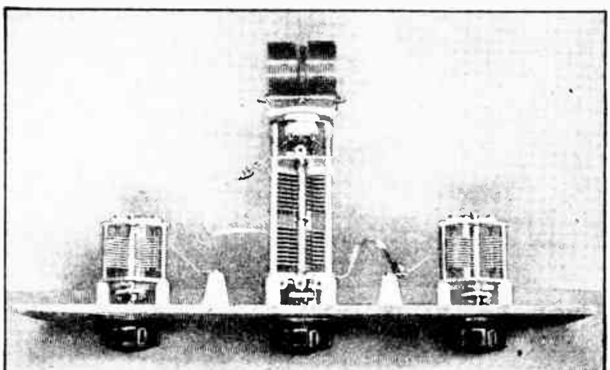
The output voltage at a normal load current of about 75 ma. can be increased from 250 to about 300 by the addition of an input filter condenser, C_4, the connections for which are shown by dotted lines.

If the bias section is not needed, plate or screen voltage may be obtained with the simplified circuit shown in Fig. 1315-B, eliminating the bias section.

A Low-Power Antenna Tuner for Rack Mounting

In the rack-mounted low-power antenna tuner shown in Fig. 1318, separate series and parallel condensers are used. This arrangement, while requiring three variable condensers, has the advantage that no switching is necessary when changing over from series to parallel tuning. It also makes possible the use of the tuner to cover a considerably wider range of antenna and transmission-line conditions, because the series condensers can be adjusted in conjunction with the parallel condenser to shorten the electrical length of the feeders whenever this is required to make

Fig. 1318 — A rack-mounting antenna tuner for low-power transmitters. C_1 is in the center, with C_2 and C_3 on either side. All of the components are mounted directly on the 5¼-inch panel. The variable condensers are mounted on the assembly rods on National Type GS-1 insulating pillars which are fastened to the condenser end plates with machine screws from which the heads have been removed. Small Isolantite shaft couplings are used to insulate the controls. Clips with flexible leads are provided for the split-stator condenser, C_1, so that its sections may be connected either in parallel or in series to form either a high- or low-capacitance tank circuit as required.



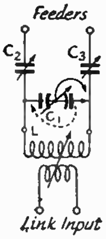


Fig. 1319 — Circuit of the rack-mounting antenna tuner for use with transmitters having final amplifiers which are operated at less than 1000 volts on the plate.

All coils are 1¾ inches in diameter and 2¼ inches long, with the variable link located at the center. For series tuning, use the coil specified for the next-higher frequency band, which will be approximately correct.

C_1 — 100 μμfd. per section, 0.045-inch spacing (National TMK-100-D) for high voltages; receiving type for low voltages (Hammarlund MCD-100).
C_2, C_3 — 250 μμfd., 0.026-inch spacing (National TMS-250) for high voltages; receiving type for low voltages (Hammarlund MC-250).
L — B & W JVL-series coils. Approximate dimensions for parallel tuning for each band are as follows:
3.5-Mc. band — 40 turns No. 20.
7-Mc. band — 21 turns No. 16.
14-Mc. band — 14 turns No. 16.
28-Mc. band — 8 turns No. 16.

parallel tuning effective. In addition, the series condensers also are useful in that they provide a measure of control over the amplifier loading when parallel tuning is used.

Clips with flexible leads attached are provided for the parallel condenser, C_1, so that the sections may be connected either in parallel or in series to form either a high- or low-capacity tank circuit, as required. When the high-*C* parallel tank is desired, the two stators are clipped together, as shown by the dotted lines in the circuit diagram of Fig. 1319, and the rotor is connected to the opposite feeder. When the two sections are connected in series, for low-*C* operation, the break-down voltage is increased.

Below the circuit diagram, Fig. 1319, two sets of variable condensers are suggested. The smaller receiving-type condensers with 0.03-inch air gap should be satisfactory for low-power transmitters operating at plate potentials of 400 to 450 volts, while larger condensers with 0.045-inch spacing will be required for transmitters using plate voltages up to about 750 or 1000.

A Three-Stage 100-Watt Transmitter for Five Bands

The three-stage transmitter shown in Figs. 1320, 1322 and 1323 is designed to use a single 1000-volt 100-ma. tube such as the 1623, 809, HY40, or higher-voltage tubes at reduced ratings, in the output stage.

Referring to the circuit diagram of Fig. 1321, a 6L6, operating at a plate voltage of 400 but at reduced input, is used in the Tri-tet oscillator circuit. A potentiometer in the screen circuit provides a means of varying the screen voltage and, ultimately, the excitation to the final amplifier. The 2E25 buffer-doubler circuit is capacitively-coupled to the oscillator. This second stage makes it possible to obtain excitation for the final amplifier in a third band from a single crystal, operation in the second band being available by doubling frequency in the oscillator itself. Parallel plate feed is used in the second stage to permit series grid feed to the final amplifier, thereby avoiding the probability of low-frequency parasitic oscillations.

The neutralized final amplifier is directly coupled to the driver stage. C_8 and L_5 form a trap for v.h.f. parasitic oscillations.

The meter switch, *S*, shifts the milliammeter to read oscillator cathode current, driver screen current, driver cathode current, final-amplifier grid current and final-amplifier cathode current. The individual filament transformers permit independent metering of the cathode currents of the last two stages.

Power supply — This transmitter is designed to operate from the combination 1000-volt and 400-volt plate supply shown in Figs 1324 and 1325. Both fixed bias of 75 volts for the 2E25 and cut-off bias for the final amplifier may be obtained from the unit shown in Figs. 1326 and 1327. For the 1623 tube, resistors R_2

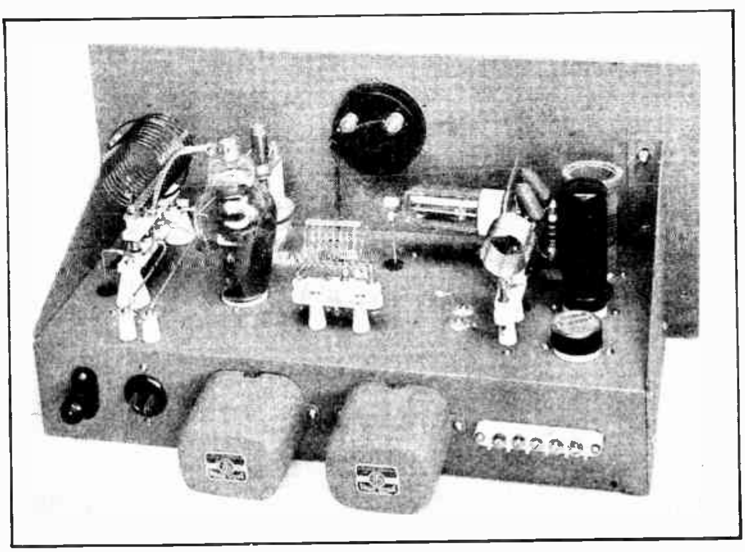


Fig. 1320 — On top of the chassis of the 100-watt transmitter, the cathode coil, L_1, the 6L6 and the crystal are in line at the right-hand end. The 2E25 is mounted horizontally on a small panel which also provides mounting space for the filament and screen by-pass condensers, the coupling condenser, C_7, the grid leak, R_5, and the grid choke. L_2 is just to the left of the 6L6 and to the right of C_2 underneath. L_3 is in the center at right angles to L_2 and L_4 and just to the rear of C_3 underneath. The 1623 socket is submounted to lower the plate terminal. The neutralizing condenser, C_9, is directly in front of the tube. RFC_2 is just to the left of L_4. The two filament transformers are mounted on the rear edge.

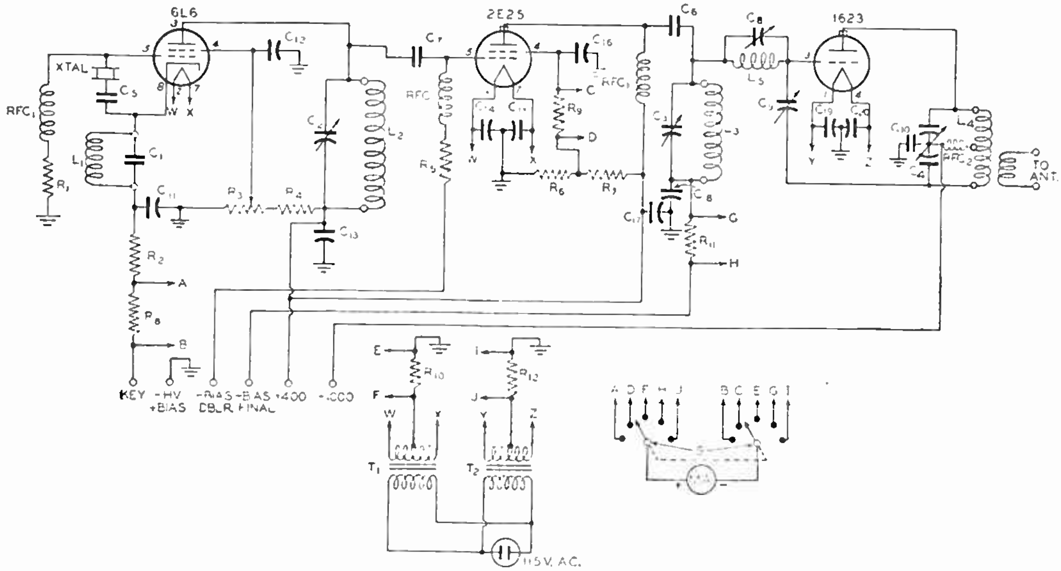


Fig. 1321 — Wiring diagram of the three-stage five-band 100-watt transmitter for 1000-volt operation.

C_1 — 100-µµfd. mica.
C_2, C_3 — 150-µµfd. variable (National ST-150).
C_4 — 100 µµfd. per section, 0.05-inch spacing (Hammarlund HFBD-100-C).
C_5, C_6 — 0.001-µfd. mica.
C_7 — 100-µµfd. mica.
C_8 — 6–60-µµfd. mica trimmer (two National M-30s in parallel).
C_9 — Neutralizing condensers (National NC-800).
C_{10} — 0.001-µfd. 5000-volt mica.
C_{11}, C_{12}, C_{13}, C_{14}, C_{15}, C_{16}, C_{17}, C_{18}, C_{19}, C_{20} — 0.01-µfd. paper.
R_1 — 0.1 megohm, ½ watt.
R_2 — 330 ohms, 1 watt.
R_3 — 20,000-ohm 10-watt potentiometer (Mallory E20MP).
R_4 — 25,000 ohms, 10 watts.
R_5 — 47,000 ohms, 1 watt.
R_6 — 20,000 ohms, 10 watts.
R_7 — 10,000 ohms, 10 watts.
R_8, R_9, R_{10}, R_{11}, R_{12} — 22 ohms, 1 watt.
L_1 — 1.75-Mc. crystals — 32 turns No. 24 d.s.c., close-wound.
3.5-Mc. crystals — 9 turns No. 22, 1 inch long; 100-µµfd. mica in coil form, connected across winding.
7-Mc. crystals — 6 turns No. 22, ⅝ inch long.
All on Hammarlund 1½-inch diam. forms.
L_2, L_3 — 1.75-Mc. — 56 turns, 1¼-inch diam., 1¾ inches long, 54 µh. (National AR80, no link).
3.5 Mc. — 28 turns, 1¼-inch diam., 1½ inches long, 15 µh. (National AR40, no link).
7 Mc. — 14 turns, 1¼-inch diam., 1¼ inches long. 4.2 µh. (National AR20), no link.
14 Mc. — 8 turns, 1¼-inch diam., 1½ inches long, 1.25 µh. (National AR10, no link).
28 Mc. — 4 turns, 1-inch diam., ¾-inch long, 0.5-µh. (National AR5, turns close, no link).
L_4 — 1.75 Mc. — 40 turns, No. 18, 2½-inch diam., 2½ inches long, 78 µh. (B & W BCL160). An 80-µµfd. fixed air padder (Cardwell JD-80-OS) is placed in right-rear corner of chassis and attached to coil with flexible leads and clips.
3.5 Mc. — 32 turns No. 16, 2½-inch diam. 2¾ inches long, 39 µh. (B & W BCL80).
7 Mc. — 20 turns No. 14, 2-inch diam., 2½ inches long, 12 µh. (B & W BCL40).
14 Mc. — 8 turns No. 14, 2-inch diam., 2 inches long, 2.5-µh. (B & W BCL20). One turn removed from each end.
28 Mc. — 4 turns No. 12, 2-inch diam., 1¾ inches long, 0.7 µh. (B & W BCL10). One turn removed from each end.
L_5 — 5 turns No. 14, ½-inch diam., ½ inch long.
MA — 0–200 d.c. meter.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 1-mh. 300-ma. r.f. choke (National R-300U).
S — Double-gang 5-circuit switch (Mallory 3226J).
T_1, T_2 — Filament transformer, 6.3 volt, 3 amp. (UTC S55).

and R_3 should be 6000 ohms and 7000 ohms, respectively.

Tuning — Coils for the desired output frequency, consistent with the crystal frequency, should be plugged in the various stages, bearing in mind that frequency may be doubled in the plate circuit of the oscillator and again in the second stage, if desired. It should also be remembered that the selection of the cathode coil, L_1, depends upon the crystal frequency and not necessarily the output frequency of the oscillator, the same cathode coil being used for both fundamental and second-harmonic output from the crystal stage. Since much better efficiencies can be obtained with the 2E25 operating as a straight amplifier, it is advisable to avoid doubling in this stage.

The first two stages should be tested first, with all voltages applied except the plate voltage for the final amplifier. Tuning the oscillator to resonance, with the key closed, should cause a slight dip in cathode current accompanied by an abrupt rise in the screen and cathode current of the second stage. Tuning the 2E25 plate circuit to resonance should produce a good dip in cathode current, with a simultaneous reading of maximum grid current to the final amplifier.

The amplifier should then be neutralized and tested for parasitic oscillation. The latter is done by shifting the final-amplifier plate-voltage lead to the 400-volt tap and turning

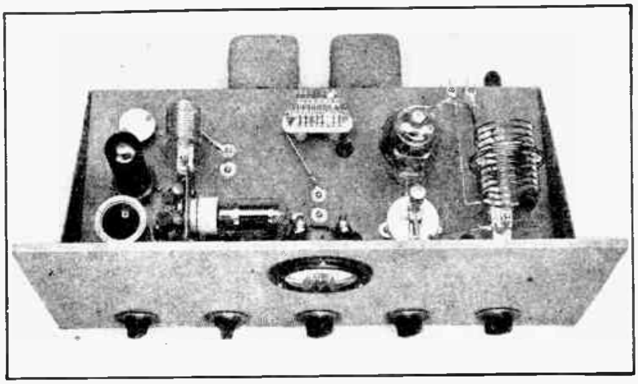


Fig. 1322 — All controls for the 100-watt five-band transmitter are below the chassis level. From left to right, they are the oscillator screen-voltage potentiometer, the oscillator plate-tank condenser, the buffer-doubler plate-tank condenser, the meter switch and the final-amplifier plate-tank condenser. The panel is of standard rack width and is 8¾ inches high.

off the bias supply. No plate voltage should be applied to the exciter stages. C_4 is then varied through its entire range for several settings of C_3. If at any point a change in the final-amplifier cathode current is observed, C_8 should be adjusted to eliminate it. During this process, plate voltage should not be applied long enough to cause appreciable heating of the tube.

Normal operating voltages may now be replaced and the final amplifier tuned up in the usual mannner. A plate current of 100 ma. will indicate normal loading of the final amplifier. (Plate current will be the difference between grid and cathode currents under operating conditions.) With all stages tuned and the amplifier loaded normally, the oscillator cathode current should run between 16 and 30 ma., 2E25 screen current between 6 and 11 ma., 2E25 cathode current between 45 and 70 ma., 2E25 grid voltage between 125 and 260 volts, oscillator screen voltage between 100 and 250 volts, and 2E25 screen voltage between 210 and 250 volts, exact values depending upon whether the stage is operating at the fundamental or doubling frequency. Excitation should be adjusted to keep the amplifier grid current between 20 and 25 ma., when the grid voltage should measure 130 to 150 volts. Power output of 65 to 75 watts should be obtainable on all bands. The oscillator circuit may be arranged for optional VFO input by short-circuiting the cathode circuit.

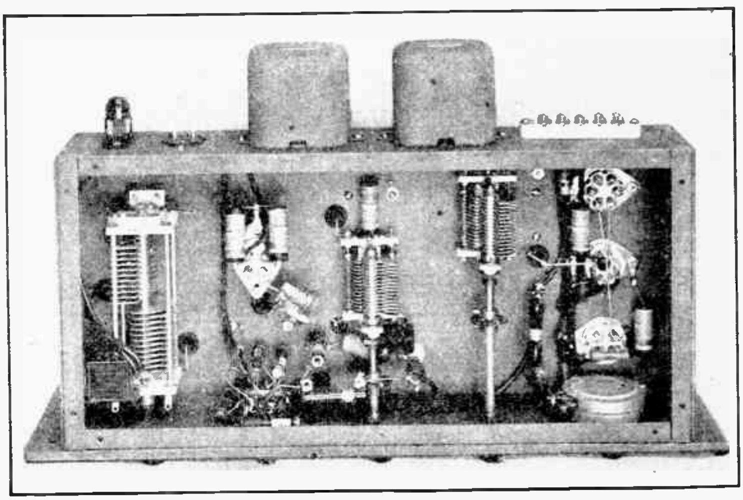
If the output stage is to be plate-modulated, the plate voltage should be reduced to 750. Operating data for suitable tubes of other types will be found in the tables in Chapter Twenty.

A suitable antenna tuner is the one shown in Fig. 1318. The larger variable condensers should be used.

A Simple Combination Bias Supply

Fig. 1326 shows the circuit diagram of the simple transformerless bias unit, pictured in Fig. 1327, which may be used to supply cut-off bias voltages up to 100 volts or so. Through grid-leak action it will also provide the additional operating bias voltage required, if the resistor values are correctly proportioned. The circuit also includes a second branch, consisting of R and a VR-75-30 voltage-regulator tube, supplying regulated voltage. This branch may not be required in all cases, but will be found convenient in many applications for providing fixed cut-off or protective bias for a

Fig. 1323 — Underneath the 8 × 17 × 3-inch chassis of the 100-watt transmitter. C_2 to the right and C_3 in the center are insulated from the chassis by polystyrene button insulators. C_4 to the left also is insulated and is spaced from the chassis to bring all shafts at the same level. Leads to the coils immediately above the tank condensers pass through large grommeted clearance holes. Meter-shunt resistances are soldered directly to the switch terminals. R_3 at the right is insulated from the chassis by extruded bakelite washers. The v.h.f. parasitic trap is suspended in the amplifier grid lead to the left of C_3. Insulating couplings are required for C_2 and C_3.



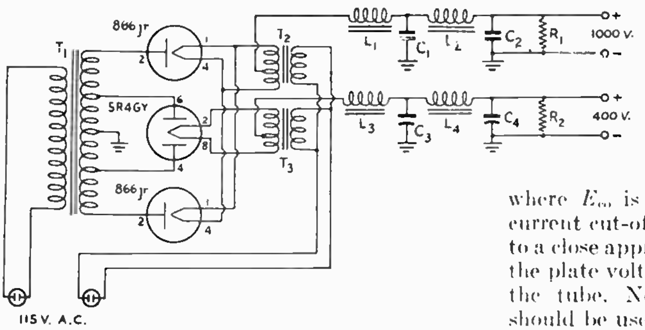


Fig. 1324 — Circuit diagram of the combination 1000- and 400-volt supply for the 100-watt transmitter.

C_1, C_2 — 2-μfd. 1000-volt paper (Mallory TX805).
C_3 — 4-μfd. 600-volt electrolytic (C-D 604).
C_4 — 8-μfd. 600-volt electrolytic (C-D 608).
R_1 — 20,000 ohms, 75 watts.
R_2 — 20,000 ohms, 25 watts.
L_1, L_3 — 5/20-hy. swinging choke, 150 ma. (Thordarson T-19C39).
L_2, L_4 — 12-hy. smoothing choke, 150 ma. (Thordarson T-19C46).
T_1 — High-voltage transformer, 1075 and 500 volts r.m.s. each side, 125- and 150-ma. simultaneous current rating (Thordarson T-19P57).
T_2 — 2.5 volts, 5 amp. (Thordarson T-19F88).
T_3 — 5 volts, 4 amp. (Thordarson T-63F99).

low-power stage independent of the main output voltage.

Adjustment — The voltage-divider resistances, R_2 and R_3, are combined in a single resistor with two sliding taps. One of these taps alters the total resistance by short-circuiting a portion of the resistance at the negative end, while the other adjusts the cut-off voltage. The method of determining the values of resistance in each section is as follows:

The bias section, R_2, is adjusted to equal the recommended grid-leak resistance for the tube or tubes in use. The value of resistance between the biasing tap and the short-circuiting tap is determined by the following formula:

$$R_3 = \frac{160 - E_{co}}{E_{co}} \times R_2,$$

where E_{co} is the voltage required for plate-current cut-off. This value may be determined to a close approximation for triodes by dividing the plate voltage by the amplification factor of the tube. No supplementary grid-leak bias should be used in the stage being supplied by the pack.

The resistance in each section should be first set at the values determined by the formula. The biased amplifier should then be turned on, without excitation. If the plate current is not almost completely cut off, or at least reduced to a safe value, the biasing tap should be moved upward (in the negative direction). With the amplifier in operation and drawing rated grid current, the biasing voltage should be meas-

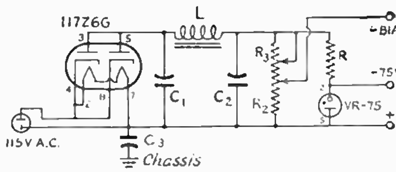


Fig. 1326 — Circuit diagram of the transformerless bias supply with voltage-regulated output, shown in Fig. 1327.

C_1, C_2 — 16-μfd. 450-volt electrolytic.
C_3 — 0.01-μfd. paper.
R — 7500 ohms, 10 watts.
$R_2 + R_3$ — 15,000-ohm 50-watt wire-wound resistor with two sliders.
See text for details of adjustment and operation.
L — 60-ma. replacement filter choke.

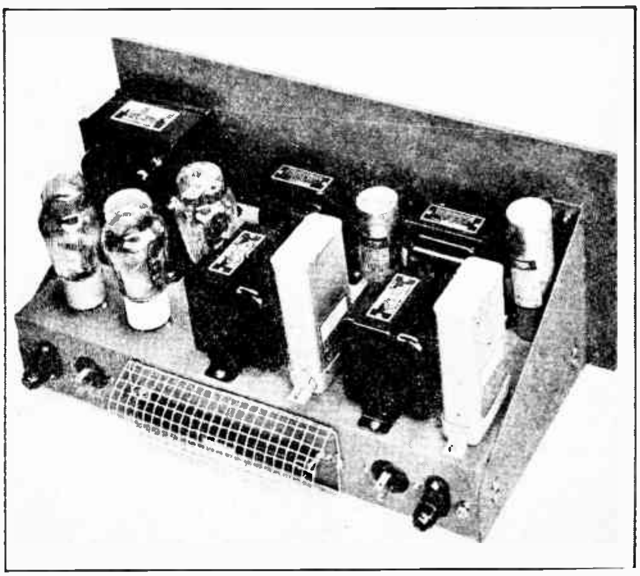


Fig. 1325 — This power supply makes use of combination transformers and a dual filter system, delivering 1000 volts at 125 ma. and 400 volts at 150 ma., or 400 volts and 750 volts simultaneously, depending upon the transformer selected. The circuit diagram is given in Fig. 1324. The 1000-volt bleeder resistor is mounted on the rear edge of the chassis, with a protective guard made of a piece of galvanized fencing material to provide ventilation. Millen safety terminals are used for the two high-voltage terminals. Ceramic sockets should be used for the 866 jrs. The chassis measures 8 × 17 × 3 inches and the standard rack panel is 8¾ inches high.

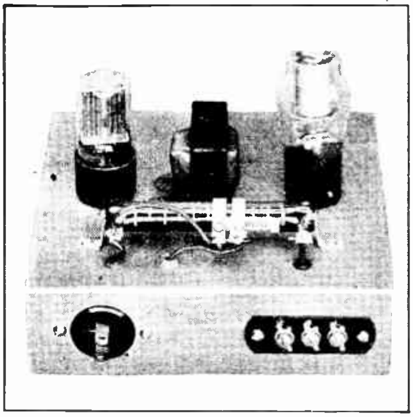


Fig. 1327 — A transformerless combination bias-supply suitable for supplying bias for r.f. stages requiring 125 volts or less for cut-off. A second branch, controlled by a VR-75-30 regulator tube, provides 75 volts fixed bias for a second stage whose grid current does not exceed 20 ma. The unit above is constructed on a 7×7-inch chassis, although the components may easily be fitted into any spare space on another power-supply chassis. The regulated VR-tube branch may be omitted if not required. The circuit diagram is shown in Fig. 1326.

ured, using a high-resistance voltmeter. If the grid voltage is higher than that recommended in the tube operating tables, *both* the biasing tap and the short-circuiting tap on the upper section should be moved, bit by bit, toward the positive end until the correct operating bias is obtained. The bias voltage should then be measured again. A final adjustment may be necessary to again arrive at cut-off voltage without excitation.

Fig. 1327 shows the components assembled separately on a small chassis. They may, however, be combined with plate-supply components on a single chassis, since little additional space will be required.

It will be noticed in the circuit diagram that a polarized plug is used in the line and that the only connection between the circuit and the chassis is through the condenser, C_3. This is to prevent short-circuiting the power line, should an ordinary plug be used and be inserted incorrectly in the socket. The polarized plug should be connected so that the grounded side of the power line is connected to the positive side of the bias supply.

A Four-Band 125-Watt Transmitter

Figs. 1328 and 1330 show two views of a simple 125-watt 4-band transmitter. As the circuit diagram of Fig. 1329 shows, it consists of an RK-4D32 beam tetrode with a two-stage driver using 7C5 receiving-type tubes. 3.5-Mc. crystals may be used for 3.5-, 7- and 14-Mc. work, or 7-Mc. crystals for 7-, 14- and 28-Mc. operation. When the output stage is operated at the crystal fundamental frequency, the doubler tube and coil are removed from their sockets and a jumper connecting the grid and plate terminals is inserted in the tube socket. To obtain the required Q in the output tank circuit, the coil is tapped, rather than use the large tank capacitance which would otherwise be necessary.

Series plate feed is used in all stages. Screen voltage for the 7C5s is taken from individual voltage dividers, while a series resistor is used in the output stage so that it can be plate-screen modulated if desired. The oscillator is keyed when the doubler stage is not in the circuit, otherwise the doubler is keyed. If oscillator keying on all bands is desired, the

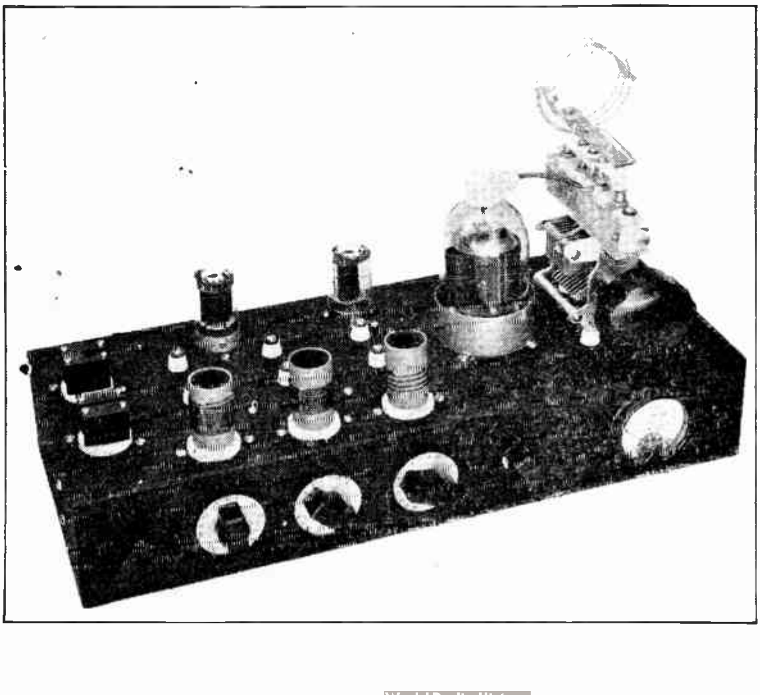


Fig. 1328 — A 125-watt transmitter for 3.5 to 30 Mc. The crystal switch is to the left and the meter switch is to the left of the meter. The crystals are plugged into the two tube sockets to the left in pairs.

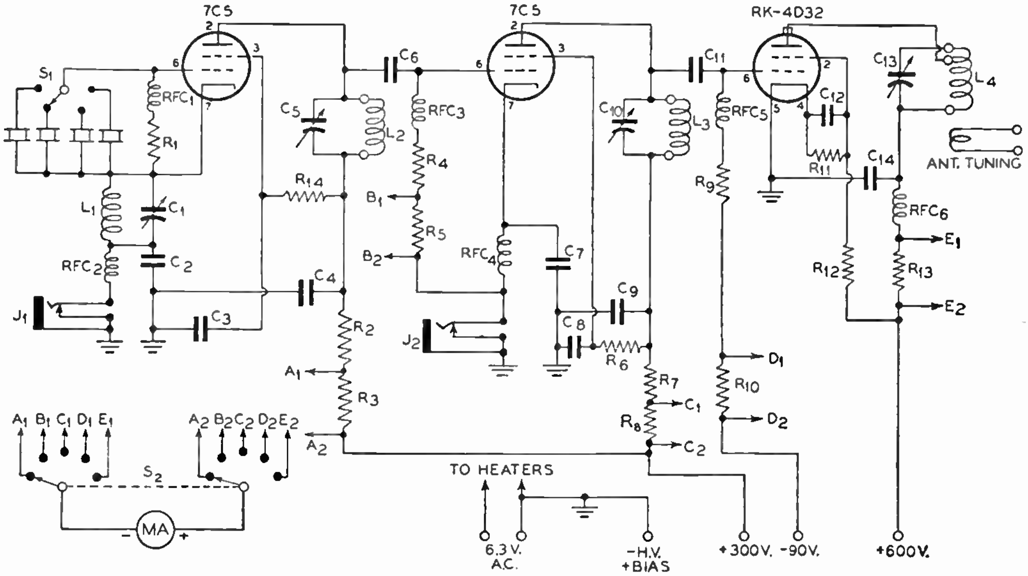


Fig. 1329 – Circuit diagram of the 125-watt 4-band transmitter.

C_1 — 140-μμfd. receiving-type variable (Hammarlund MC-140S).
C_2, C_3, C_4 — 0.002-μfd. mica.
C_5, C_{10} — 100-μμfd. receiving-type variable (Hammarlund MC-100S).
C_6, C_{11} — 100-μμfd. mica.
C_7, C_9 — 0.001-μfd. mica.
C_8 — 500-μμfd. mica.
C_{12}, C_{14} — 0.001-μfd. 1200-volt mica.
C_{13} — 100-μμfd. 1500-volt variable (National TMK-100).
R_1 — 0.1 megohm, 1-watt composition.
R_2 — 680 ohms, 1-watt composition.
R_3, R_5, R_8, R_{10} — 68 ohms, ½-watt composition.
R_4 — 47,000 ohms, 1-watt composition. (See text.)
R_6 — 10,000 ohms, 1-watt composition.
R_7 — 470 ohms, 1-watt composition.
R_9 — 2200 ohms, 1-watt composition.
R_{11} — 25,000 ohms, 10-watt wire-wound.
R_{12} — 12,500 ohms, 20-watt wire-wound (two 10-watt 25,000-ohm resistors in parallel).
R_{13} — See text.
R_{14} — 22,000 ohms, 1-watt composition.

L_1 — For 3.5-Mc. crystals — 16 turns No. 22 d.s.c. wire, 1-inch diameter, ⅞ inch long.
For 7-Mc. crystals — 12 turns No. 22 d.s.c. wire, 1-inch diameter, 1 inch long.

L_2 — 3.5 Mc. — 40 turns No. 22 d.s.c., 1-inch diameter, close-wound.
7 Mc. — 18 turns No. 20 enam., 1-inch diameter, 1¼ inches long.
14 Mc. — 10 turns No. 20 enam., 1-inch diameter, ¾ inch long.
L_3 — 7 Mc., 14 Mc. — Same as L_2.
28 Mc. — 4 turns No. 20 enam., 1-inch diameter, ½ inch long.
L_4 — 3.5 Mc. — 24 turns 2 inches diameter, 1¼ inches long (24 turns removed from B & W BEL-80), tapped at 6 turns from plate end.
7 Mc. — 12 turns 2 inches diameter, 1¼ inches long (10 turns removed from B & W BEL-40), tapped at 3 turns from plate end.
14 Mc. — 8 turns 2 inches diameter, 2 inches long (B & W BEL-20), tapped at 2 turns from plate end.
28 Mc. — 3 turns 2 inches diameter, 1 inch long (B & W BEL-10), no tap.
J_1, J_2 — Closed-circuit jack.
MA — 0–200 d.c. milliammeter.
RFC_1, RFC_2, RFC_3, RFC_4, RFC_5 — 2.5-mh. r.f. choke (Hammarlund CHX).
RFC_6 — 1-mh. r.f. choke (National R-300).
S_1 — Single-circuit 4-position ceramic rotary switch.
S_2 — Two-circuit 5-position ceramic rotary switch.

47,000-ohm grid-leak resistor, R_4, should be replaced with a 33,000-ohm unit and a 45-volt battery connected in series between the lower end of R_5 and the keying jack, J_2. The milliammeter, MA, has a scale of 0–200 ma. and can be switched to read oscillator current, doubler current or amplifier grid or plate current. The shunt R_{13} is wound with No. 30 copper wire, using a ½-watt 68-ohm resistor as a form. The length of the wire used in the shunt is adjusted to give a meter-scale multiplication of two so that the full-scale reading becomes 400 ma. when the meter is switched in this position.

The unit is built on a 7 × 7 × 3-inch chassis. Two octal sockets are mounted at the left-hand end of the chassis to serve as mountings for four crystals. The crystal switch, S_1, is mounted underneath centrally between the two sockets. The sockets for L_1, L_2 and L_3 are lined up along the front edge, while their respective tuning condensers are placed to the rear underneath. They are insulated from the chassis by means of small feed-through insulators and the shafts are fitted with insulating couplings. The two tube sockets are just to the rear of the condensers. A refinement which is not strictly necessary at the frequencies at

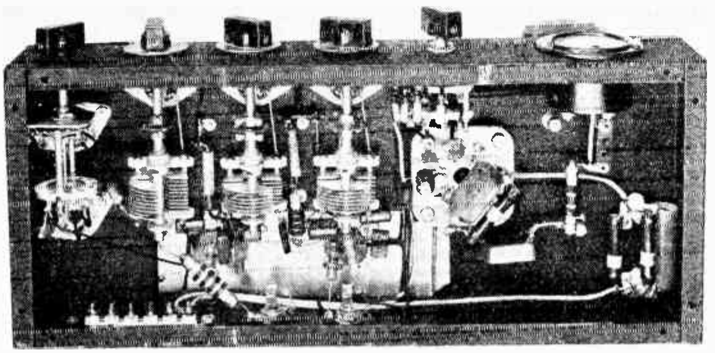


Fig. 1330 — Bottom view of the 4D32 transmitter. A terminal strip set in the back edge of the chassis is provided for power-supply connections. The two jacks, also set in the rear edge, are for the key.

which this transmitter operates is the copper-sheet ground plate which surrounds the 7C5 sockets and to which all r.f. ground connections are made.

The sockets for the 4D32 and the output tank circuit are to the right with the coil socket mounted on top of the condenser. In mounting the coil, the top plate of the tank condenser is replaced by a strip of aluminum bent up at one end to form a "J." One of the outside coil jacks passes through a hole in the end of the "J," while a small ½ × ¾-inch cone insulator is used to support the "hot" end of the coil jack bar. The tube socket is spaced ¼ inch below the chassis to provide clearance for the screws which fasten the tube shield to the chassis. This shield is a 2½-inch coil shield cut down so that it comes up 1¼ inches from the chassis. C_{13} is insulated from the chassis by mounting it on small feed-through insulators. Care should be used in selecting a well-insulated dial and coupling for this condenser since the shaft carries the full high voltage.

The driver coils are wound on Millen 1-inch diameter forms. The standard BEL-series B & W coils used in the output stage must be altered slightly to provide for the tap. A fifth plug is added at the empty center hole. The link connection, normally near the plate end of the coil, is shifted to this center plug, while the tap is connected to the plug normally connected to the link.

The power supply shown in Fig. 1315 will provide plate voltage for the 7C5s and biasing voltage for the 4D32 if a VR-90 is substituted for the VR-75. The high-voltage supply shown in Fig. 1317 is suitable for the final stage.

The plate current to the crystal oscillator should run around 20 ma. and the doubler plate current about 40 ma. Grid current to the doubler should be about 2 ma. and to the final at least 6 ma. under load.

A 100-Watt Output Bandswitching Transmitter or Exciter

The transmitter pictured in Figs. 1331, 1333 and 1344 incorporates bandswitching over all bands from 3.5 to 28 Mc. It consists of a 6V6 Tri-tet oscillator which gives either fundamental or second-harmonic output from a 3.5-Mc. crystal, a 6N7 dual-triode frequency multiplier with its first triode section operating as a doubler from 7 to 14 Mc. and the second section doubling from 14 to 28 Mc., and a final stage with two 807s in parallel. The Tri-tet cathode coil may be cut in or out of the circuit as desired, so that the 6V6 may be used as a straight tetrode crystal oscillator on either 3.5 or 7 Mc. Provision is made for crystal switching, six crystal sockets being included, and a seventh switch position is used for external VFO input. The power output on all bands is in excess of 100 watts when the 807s are operated at ICAS c.w. telegraph ratings.

The circuit diagram of the transmitter is given in Fig. 1332. The switching circuit is so arranged that the grids of unused 6N7 triode sections are disconnected from the preceding stage and grounded; thus excitation is not applied to idle doubler tubes. Only one coil is used in the 6V6 stage to cover both 3.5 and 7 Mc.; for 3.5 Mc. an air padding condenser, C_2, is switched in parallel with the 7-Mc. tank circuit to extend the tuning range to 3.5 Mc.

Capacity coupling between stages is used throughout. The plates of the first three stages are parallel-fed so that the plate tuning condensers can be mounted directly on the metal chassis. Coupling to the 807 grids is through a tap on each plate coil; this "tapping down" not only provides the proper load for the various driver stages but also helps overcome the effect on the driver tuning ranges of the rather large shunt capacitance resulting from operating the two beam tetrodes in parallel. Series feed is used in the plate circuit of the 807s, the tank condenser being of the type that is insulated from the chassis. Operating bias for the 807s is obtained from a grid-leak resistor, and the screen voltage is obtained through a dropping resistor from the plate supply.

Plate currents of all tubes are read by a 0–100 d.c. milliammeter which can be switched to any plate circuit by means of S_4. One switch position is provided for checking the final-stage grid current. The d.c. cathode returns of both the 6V6 and the 807s are brought out to terminals so that a choice of keying is offered. If the 6V6 cathode lead is grounded, the amplifier alone may be keyed in the cathode circuit; if the two cathode returns are connected together, the oscillator and amplifier may be keyed simultaneously for break-in operation. (The oscillator alone cannot be keyed with the 807 cathodes grounded, because without fixed bias on the latter tubes the plate input would be excessive under key-up conditions.)

To prevent parasitic v.h.f. oscillations, small chokes (RFC_5 and RFC_6) are connected in the grid leads to the 807s, and a 68-ohm resistor is connected in each screen lead. These suppressors are mounted as closely as possible to the tube sockets. A parasitic trap, L_5C_7, is connected in the common plate lead to the 807s. Because of the high power sensitivity of the paralleled 807s and the fact that the grid-plate capacitance is doubled by the parallel connection, the tubes may oscillate in t.p.t.g. fashion at the operating frequency if the amplifier is run with no load on the plate tank. However, this tendency toward oscillation disappears with a small load (less than one-fourth rated plate current) and the amplifier is perfectly stable under normal loading conditions.

As shown in Fig. 1333, the amplifier plate coils are mounted on an aluminum bracket supported by the main chassis. The bracket dimensions are 6½ inches long by 4 inches wide on top, with mounting legs 2½ inches high. Half-inch lips bent outward from the bottoms of the legs provide means for mounting to the chassis. The amplifier bandswitch, S_3, is mounted underneath the coil bracket, with the two switch wafers spaced out so they are approximately two inches apart. This brings the plate switch section directly under the 28-Mc. tank coil so that the shortest leads can be obtained at the highest frequency. The output link connection runs from the other switch section (at the front) through a length of 300-ohm feeder to terminals on the rear wall of the chassis. Because of the low ratio of plate voltage to plate current, a rather low L/C ratio must be used in the plate tank circuit to secure a reasonable Q. The standard coils used are therefore modified to the dimensions given in Fig. 1332. Other types of manufactured coils (100-watt rating) may be used if desired, provided turns are taken off to bring the 3.5-Mc. band near maximum capacitance on the 150-$\mu\mu$fd. tank condenser, the 7-Mc. band at 65 to 70 per cent of maximum, and the 14-Mc. band to approximately 30 per cent of maximum. The 28-Mc. band may tune at nearly minimum capacitance, since the minimum circuit capacitance is fairly large.

In the bottom view, Fig. 1334, the meter switch with its shunting resistors is at the left. The driver bandswitch, S_2, is in the center; the section nearest the panel is for C_2, the rotor of the next section goes to the grid of the 14-Mc. doubler, the rotor of the third section to the 28-Mc. doubler, and the rotor of the last section to the grids of the 807s. In this view the right-hand section of the 6N7 is the 14-Mc. doubler. Grid and plate blocking condensers are supported between the tube-socket terminals and small ceramic pillars which serve as tie-points for r.f. wiring. The coil taps to the 807 switch drop through holes in the chassis directly below the proper prongs on the coil sockets. The crystal switch, crystal-holder assembly, oscillator cathode tuned circuit, and shorting switch, S_5, are in the upper left-hand corner. The crystal sockets (for the new small crystals) are mounted in a row on a 1½ × 3-inch piece of aluminum secured to the chassis by mounting pillars of square aluminum rod. The spare crystal socket on top of the chassis is for old-type crystal holders with ¾-inch pin spacing. In general, chokes and by-pass condensers are grouped as closely as

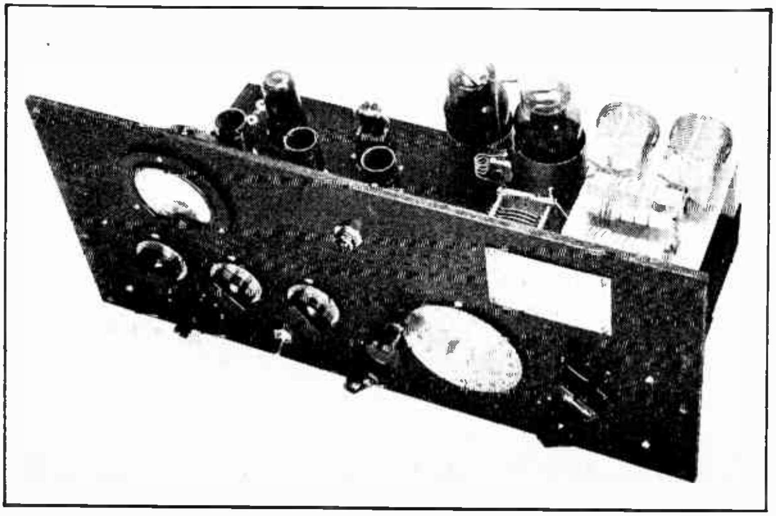


Fig. 1331 — A 100-watt output transmitter or exciter with bandswitching over four bands. The output stage uses parallel 807s. Crystal switching, with provision for VFO input, and meter switching are incorporated. Tuning controls, from left to right, are crystal oscillator-doubler, 14-Mc. doubler, 28-Mc. doubler, and (large dial) final amplifier. The crystal switch is at the lower-left corner, driver bandswitch in the center, and meter switch at the lower right. The amplifier bandswitch is above the meter switch and to the right of the amplifier tuning dial.

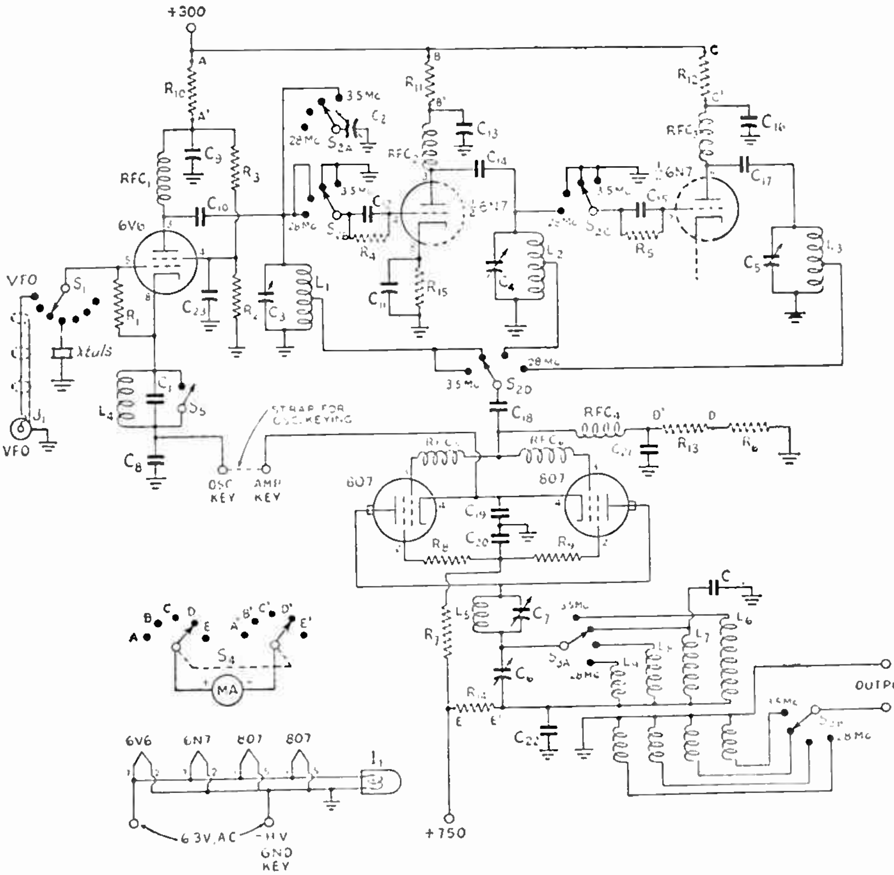


Fig. 1332 — Circuit diagram of the 100-watt bandswitching transmitter.

C — See text.
C_1 — 220-μμfd. mica (mounted inside L_4).
C_2 — 140-μμfd. air padder.
C_3, C_4, C_5 — 100-μμfd. variable (National ST-100).
C_6 — 150-μμfd. variable, 0.05-inch plate spacing (Hammarlund HFB-150-C).
C_7 — 3–30-μμfd. ceramic padder.
C_8, C_{19}, C_{21} — 0.0047-μfd. mica.
C_9, C_{11}, C_{13}, C_{16}, C_{23} — 0.01-μfd. paper, 600 volts.
C_{10}, C_{14}, C_{17} — 0.0022-μfd. mica, 500 volts.
C_{12}, C_{15}, C_{18} — 100-μμfd. mica.
C_{20} — 470-μμfd. mica, 2500 volts.
C_{22} — 0.0022-μfd. mica, 2500 volts.
R_1 — 0.1 megohm, ½ watt.
R_2, R_3 — 47,000 ohms, 1 watt.
R_4 — 47,000 ohms, ½ watt.
R_5 — 22,000 ohms, ½ watt.
R_6 — 12,000 ohms, 1 watt.
R_7 — 25,000 ohms, 10 watts.
R_8, R_9 — 68 ohms, ½ watt.
R_{10}, R_{11}, R_{12}, R_{13}, R_{14} — 25 ohms, ½ watt (*R_{14}* shunted as described below).
R_{15} — 470 ohms, 1 watt.

NOTE: *R_{14}* is shunted by a length of No. 30 copper wire (about 8 or 10 inches) wound around the resistor. The wire length should be adjusted to make the milliammeter read one-fifth its normal value, increasing the full-scale range to 500 milliamperes.

L_1 — 21 turns No. 18 on 1-inch diam. form, length 1 inch; tapped 15 turns from ground.
L_2 — 10 turns No. 18 on 1-inch diam. form, length 1 inch, tapped 7 turns from ground.
L_3 — 5 turns No. 18 on 1-inch diam. form, length 1 inch; tapped 2 turns from ground.
L_4 — 13 turns No. 18 on 1-inch diameter form, length 1 inch.
L_5 — 4 turns No. 18, diam. ⅜ inch, length ⅝ inch, mounted on C_7.
L_6 — 22 turns No. 20, diam. 1½ inches, length 1⅜ inches. Link 3 turns.
L_7 — 13 turns No. 16, diam. 1½ inches, length 1⅜ inches. Link 3 turns.
L_8 — 7 turns No. 16, diam. 1½ inches, length 1⅜ inches. Link 3 turns.
L_9 — 4 turns No. 16, diam. 1½ inches, length 1½ inches. Link 3 turns.

NOTE: *L_1, L_2, L_3* wound on Millen 45004 forms, *L_4* on Millen 45000 form; *L_6, L_7, L_8, L_9* are Coto C1680E, C1640E, C1620E and C1610E, respectively, with turns removed to conform to specifications above.

I_1 — 6.3-volt pilot lamp.
J — Coaxial-cable socket (Amphenol).
MA — 0–100 d.c. milliammeter.
RFC_1, RFC_2 — 2.5-mh. r.f. choke (National R-100).
RFC_3 — 2.5-mh. r.f. choke (National R-100U).
RFC_4 — 2.5-mh. r.f. choke (Millen 34102).
RFC_5, RFC_6 — 18 turns No. 20 d.c.c., ¼-inch diam., close-wound on 1-watt resistor (any high value of resistance may be used).
S_1 — Ceramic wafer switch, 7 positions.
S_2 — Four-gang 6-position ceramic wafer switch (4 positions used).
S_3 — Two-gang 4-position ceramic wafer switch (Yaxley 162C).
S_4 — Two-gang 6-position ceramic wafer switch (5 positions used).
S_5 — S.p.s.t. toggle switch.

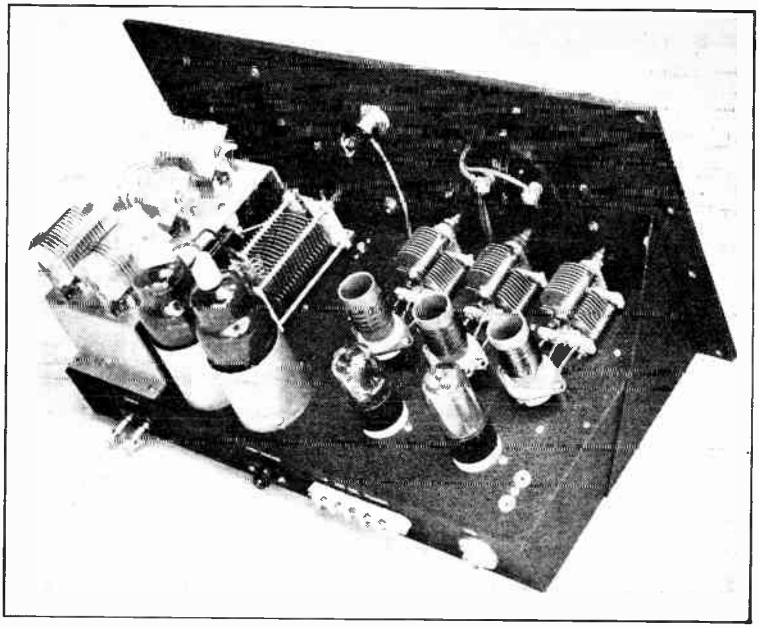


Fig. 1333 — Top view of the 100-watt band-switching transmitter. The oscillator and doubler coils are of the plug-in type for convenience in mounting and adjustment, but do not need to be changed to cover the frequency range from 3.5 to 30 Mc. The cable terminal on the chassis wall at the right is for VFO input; r.f. output terminals are at the extreme left.

possible about the tube sockets with which they are associated, to keep r.f. leads short. In the 807 circuit, the screen by-pass condenser, C_{20}, is mounted vertically from a small metal angle between the two tube sockets, and all grounds for the cathode, screen and grid circuits are brought to a common point between the two sockets.

The condenser, C, across only the 7-Mc. 807 tank coil is actually a 1 × 1-inch piece of copper with a short tab at one end. The tab is soldered to the plate lead from the coil just under the coil bracket and then bent so that the 1 × 1 portion is parallel to the bracket and separated from it by about ⅛ inch. The coil by itself resonated with the stray capacitance at 28 Mc. and absorbed considerable energy when the transmitter was operating on that band; the small capacitance detunes it and prevents such absorption. It may not be needed with other types of coils or with slightly different construction.

Preliminary tuning should be done with the plate voltage for the 807 disconnected. Set S_2 and S_3 for 28-Mc. output, set S_4 to read oscillator plate current, and close the key, if oscillator keying is being used. With a 3.5-Mc. crystal, make sure S_5 is open; with a 7-Mc. crystal S_5 should be closed. Rotate C_3 for a small kick in the plate current that indicates resonance at the crystal harmonic, in the case of the Tri-tet, and for the marked dip in plate current that indicates oscillation with the tetrode oscillator. The current should be in the vicinity of 16 to 18 ma. Switch the meter to the 14-Mc. doubler and adjust C_4 to obtain minimum plate current. This should be about 15 ma. Check the 28-Mc. doubler plate current similarly; it should be between 25 and 30 ma. at resonance. The final-amplifier grid current should be 7 to 8 ma.

Next, connect a 70-ohm dummy antenna or

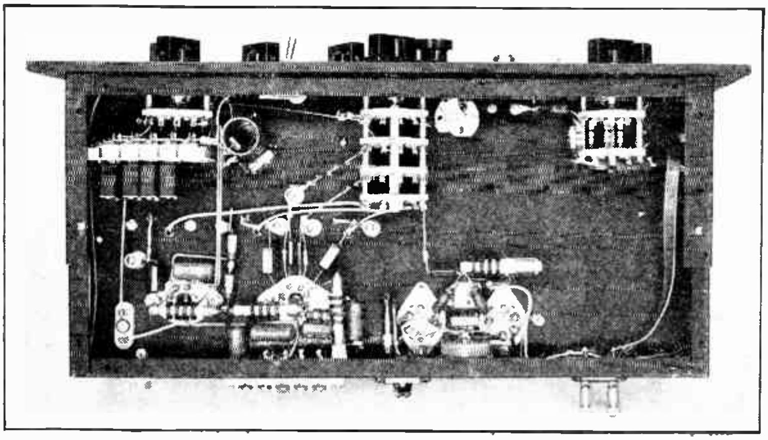


Fig. 1334 — Bottom view of the 100-watt bandswitching transmitter. The chassis dimensions are 8 × 17 × 2 inches and the panel (of crackle-finished Masonite) is 8¾ × 19 inches. Parts layout is described in the text. The 750-volt lead is brought through a Millen safety terminal, and all other power and keying connections go to a ceramic terminal strip at the rear. The connection between the crystal switch and the VFO input socket is through a short length of RG/58U cable lying in the corner of the chassis.

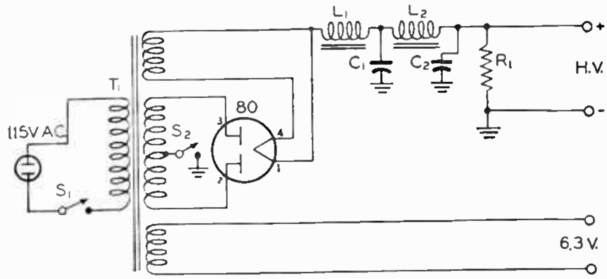


Fig. 1335 — Circuit diagram of a 250- to 300-volt 100-ma. power supply.
C_1, C_2 — 8-μfd. 600-volt electrolytic.
R_1 — 20,000 ohms, 25 watts.
L_1, L_2 — 30-hy. 110-ma. filter choke (Stancor C-1001).
S_1, S_2 — S.p.s.t. toggle switch.
T_1 — 440 volts each side of center, 130 ma.; 5 v., 3 a.; 6.3 v., 3.5 a.

100-watt lamp to the output terminals, set C_6 near minimum capacity, and apply plate voltage to the 807s. Adjust C_6 for minimum plate current, which should be about 200 ma. with this load. Readjust the driver circuits for maximum grid current to the 807s.

Tuning procedure for other bands is much the same, except that the amplifier cannot be loaded to full input on the lower frequencies by either the dummy antenna or lamp, with the links furnished with the coils specified. In such cases an antenna should be used to load the transmitter after it has been determined that the various stages are working properly. On 3.5 Mc., C_2 should be adjusted so that a crystal on 3500 kc. can be made to oscillate with C_3 set near maximum capacity. Generally, C_2 will be set at approximately full capacity.

The transmitter requires a power supply delivering 60 to 70 ma. at 300 volts for the oscillator and doublers, and one delivering 200 ma. at 750 volts for the 807s. The supplies of Figs. 1317 and 1335 are suitable.

A Two-Stage High-Power Transmitter

The photographs of Figs. 1336, 1338 and 1339 show a two-stage transmitter capable of handling a power input of 900 watts on c.w. or 675 watts on 'phone. The circuit diagram is shown in Fig. 1337. It is a simple arrangement in which a 6L6 Tri-tet crystal oscillator drives an Eimac 4-250A in the output stage, either at the crystal fundamental or at the second harmonic so that the transmitter will cover two bands with a single crystal of proper frequency without doubling in the output stage. Through the use of plug-in coils and a selection of crystals, the transmitter may be used in all bands between 3.5 Mc. and 28 Mc. inclusive.

Any one of four crystals may be selected by means of S_1, although more crystal positions may be added. R_4, R_5 and R_6 are metering resistors across which the milliammeter is switched to read combined oscillator screen and plate currents, amplifier grid current or amplifier cathode current. R_5 has sufficient resistance to have no practical effect upon the meter reading, but the other shunts which are made from copper wire are adjusted to give a meter-scale multiplication of 10, making the full-scale reading 500 ma. The diagram shows both stages keyed simultaneously. If amplifier keying only is desired, R_1 should be connected to ground instead of to the key terminal.

Construction — The transmitter is built on a 10 × 17 × 3-inch chassis with a 10½-

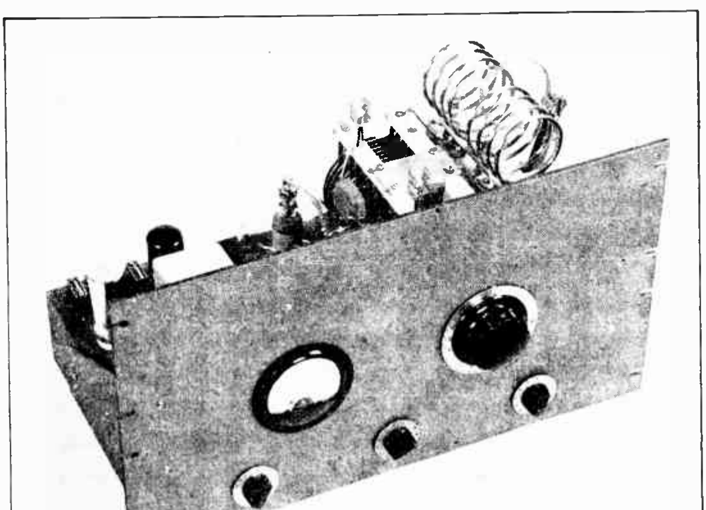


Fig. 1336 — Front view of the 4-250A transmitter. Along the bottom of the panel, from left to right, are the controls for the oscillator tuning condenser, the crystal switch and the metering switch. The large dial is for the output tank condenser.

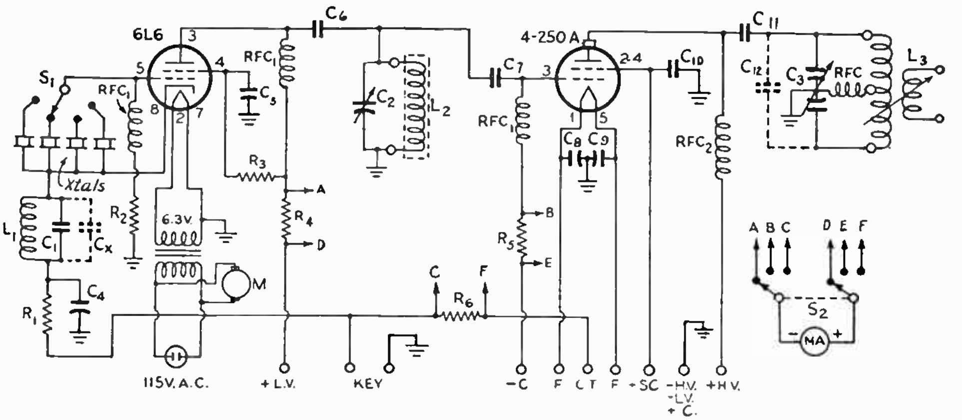


Fig. 1337 — Circuit diagram of the two-stage high-power transmitter.

C_1 — 100-μμfd. mica.
C_2 — 100-μμfd. variable (National ST-100).
C_3 — 50-μμfd.-per-section 0.171-inch-plate-spacing variable (Millen 14050).
C_4, C_5, C_8, C_9 — 0.01-μfd. paper.
C_6 — 0.0015-μfd. mica.
C_7 — 100-μμfd. mica, 5000 volts.
C_{10} — 0.001-μfd. mica, 5000 volts.
C_{11} — 0.001-μfd. mica, 10,000 volts.
C_{12} — Vacuum-type padding capacitor, 25 μμfd., 16,000 volts (GE GL-122).
C_x — 100-μμfd. mica (for 3.5-Mc. crystals only).
R_1 — 220 ohms, 1 watt.
R_2 — 47,000 ohms, ½ watt.
R_3 — 5,000 ohms, 10 watts.
R_4, R_6 — 58 inches No. 22 copper wire wound on small-diam. form.
R_5 — 47 ohms, ½ watt.
L_1 — 3.5-Mc. crystals: 22 turns No. 22 d.s.c., ½-inch diam., close-wound. C_x connected across winding.
7-Mc. crystals: 12 turns No. 22 d.s.c., ½-inch diam., close-wound.
14-Mc. crystals: 6 turns No. 20 d.s.c., ½-inch diam., ⅝ inch long.
L_2 — 3.5 Mc.: 40 turns No. 22 d.s.c., 1-inch diam., close-wound.
7 Mc.: 20 turns No. 22 d.s.c., 1-inch diam., close-wound.
14 Mc.: 9 turns No. 22 d.s.c., 1-inch diam., ¾ inch long.
28 Mc.: 5 turns No. 20 enam., ⅝-inch diam., ⅜ inch long (on Millen Type 45500 threaded ceramic form).
L_3 — B & W TVH-series coils.
M — Fan motor (Barber-Colman Type d Yab 569-1 with Type Yab 355-2 2½-inch fan, Rockford, Ill.).
MA — 0–50 milliammeter.
RFC, RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — Hammarlund CH-500 r.f. choke.
S_1 — 4-position ceramic tap switch.
S_2 — Double-gang 3-position switch.

inch standard rack panel. The mechanical arrangement shown in the photographs should be followed as closely as possible, since upon the placement of parts may depend the stability of the amplifier. The oscillator-circuit components are grouped at the left-hand end of the chassis. The Millen crystal sockets are lined up with their centers 1½ inches in from the rear edge of the chassis in the left-hand corner. The sockets for the 6L6 and the plug-in cathode coil, L_1, are in line with their centers, 3½ inches from the back edge of the chassis, while the oscillator plate coil is in line with the 6L6, 6 inches from the rear edge of the chassis

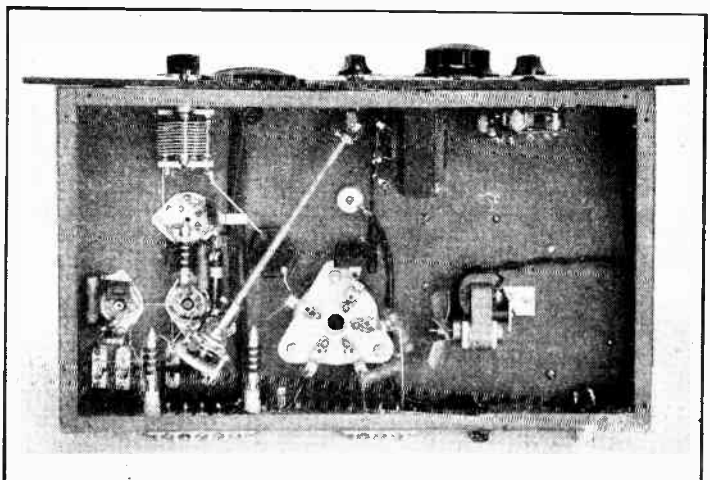


Fig. 1338 — Bottom view showing the arrangement of parts under the chassis. Mounted off the rear edge of the chassis are the oscillator (left) and amplifier (right) grid chokes. The oscillator plate choke is above. The condenser under the crystal-switch control shaft is the coupling condenser, C_7. The oscillator tuning condenser, C_2, the 6.3-volt filament transformer and the metering switch are along the front edge of the chassis. The ventilating fan is to the right of the tube socket.

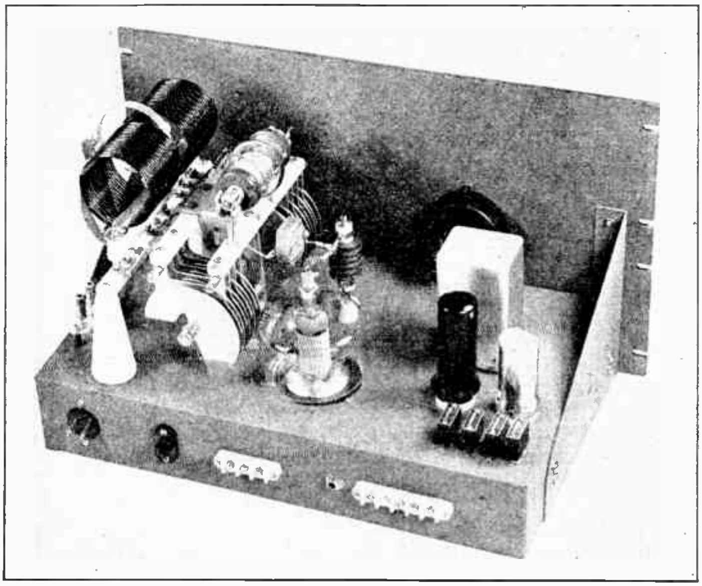


Fig. 1339 — Rear view of the two-stage high-power transmitter, showing the vacuum-type padding condenser in place on top of the tank condenser.

and 3½ inches from the left-hand end. The crystal switch is placed near the 6L6 socket and set at an angle with respect to the edges of the chassis. It is controlled by a knob at the center by means of a long ¼-inch shaft, which runs diagonally across the chassis, and a Millen 39005 all-metal flexible shaft coupling of the "universal-joint" type.

The socket for the 4-250A is centered 7¾ inches from the left-hand end of the chassis and 3 inches from the rear edge. It is spaced 1⅛ inches below the chassis on metal pillars so that the base of the tube is shielded from the plate. A spring contact is fastened to the socket so that the metal ring around the base of the tube will be grounded when the tube is inserted in the socket. The 4-250A requires a small amount of forced-air cooling. This is supplied by a small fan directed at the base of the tube. A bottom plate should be used on the chassis so that the air will be forced up around the envelope of the tube. The amplifier plate-tank condenser is placed with its shaft 5¼ inches in from the right-hand edge of the chassis, while the coil base assembly is elevated on 3-inch cone insulators centered 2½ inches from the edge. The clips for the padding condenser, C_{12}, required for the 3.5- and 7-Mc. bands, are mounted on top of the condenser on 1-inch tubular spacers. A pair of long 6-32 mounting screws, passing through the spacers, serve to make the connection between the stators of C_3 and the terminals of C_{12}. The Hammarlund CH-500 r.f. choke, RFC_2, is mounted alongside the tank condenser, near the center, with the plate blocking condenser, C_{11}, fastened to the top.

Plate voltage is fed from a Millen safety terminal in the rear edge of the chassis to the bottom end of the r.f. choke through a Millen 32101 steatite bushing. The hole for the safety terminal should have a clearance of about ¹⁄₁₆ inch around the part which goes through the chassis, to decrease the danger of a voltage break-down at this point. The link output terminals are in the right-rear corner, insulated

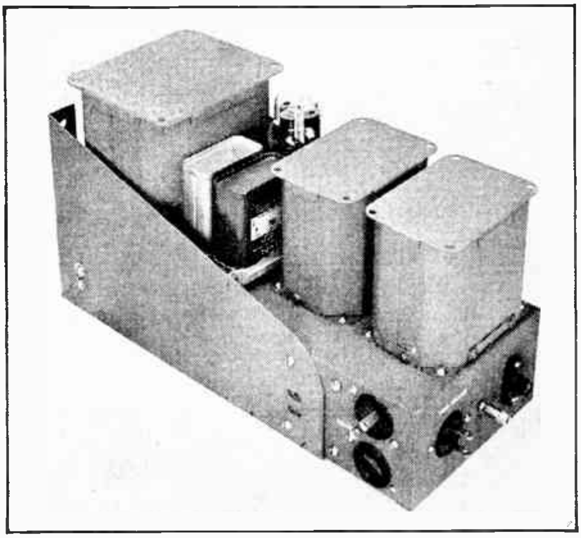
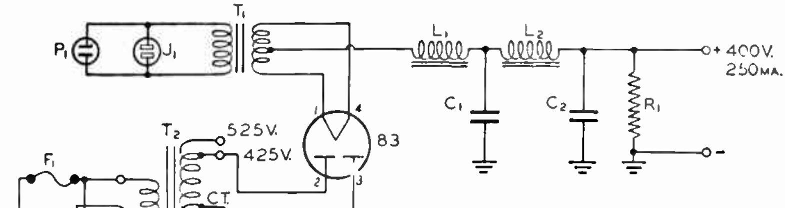


Fig. 1340 — A 400-volt 250-ma. power supply. A 6 × 14 × 3-inch chassis is used, with all wiring, the filament transformer for the 83 rectifier, and the bleeder resistor mounted beneath the chassis. The fuse, pilot lamp, and the on-off switch (not visible in this view) are mounted on the front chassis wall. A.c. input to the high-voltage transformer and the filament transformer are at the rear of the chassis, as are the safety terminal for the B+ output and the binding post for ground connection.



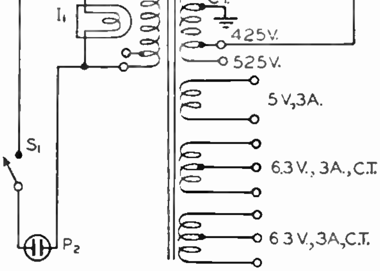


Fig. 1341 — **Schematic diagram of the 400-volt 250-ma. power supply.**

C_1 — 2-μfd. 600-volt oil-filled capacitor (Aerovox Type 609).

C_2 — 4-μfd. 600-volt oil-filled capacitor (Aerovox Type 609).

R_1 — 25,000 ohms, 20 watts.

L_1 — Swinging choke, 5–25 hy., 225 ma., 120-ohm d.c. resistance (UTC S32).

L_2 — Smoothing choke, 20 hy., 225 ma., 120-ohm d.c. resistance (UTC S31).

F_1 — 2-amp. fuse (Littelfuse Type 3AG), and fuse-holder assembly (Littelfuse 1212B).

I_1 — 110-volt a.c. pilot-lamp-and-socket assembly.

J_1 — Panel-mounting a.c. receptacle (Amphenol 61 F1).

P_1, P_2 — Panel-mounting a.c. plug (Amphenol 61 M1).

S_1 — S.p.s.t. toggle switch.

T_1 — Filament transformer, 5 volts, 4 amp. (Thordarson T-63F09).

T_2 — Power transformer, 525 or 425 volts a.c. each side of center-tap, 250-ma. rating. Filament windings: 5 v. 3 amp.; 6.3 v. 3 amp. c.t.; 6.3 v. 3 amp. c.t. (UTC S40).

from the chassis on a National FWG polystyrene terminal strip.

Underneath, at the amplifier end of the chassis, are the metering switch, S_2, and the 6.3-volt filament transformer. An external filament transformer is required for the 4-250A. It should have a rating of 5 volts, 15 amperes.

On the panel, the milliammeter is placed to balance the amplifier tuning dial, the meter-switch knob to balance that of the oscillator tuning condenser, while the crystal switch is at the center, near the bottom edge. Along the rear edge of the chassis, from left to right, as viewed from the rear, are a terminal strip for making connections to the oscillator supply, the amplifier screen-voltage dropping resistor, and to the biasing-voltage source, if one is used; the key jack, filament terminals for the 4-250A including a center-tap connection, a safety terminal for the high-voltage connection, and a male plug for the 115-volt line to the 6.3-volt filament transformer and the fan motor.

The cathode coils, L_1, are wound on Millen octal-base shielded forms without tuning slugs. A change in cathode coils is required only with a change in the band in which the *crystal* lies. The coil for use with 3.5-Mc. crystals requires an additional 100-μμfd. mica condenser, C_x, connected across the winding as shown by the dotted lines in Fig. 1337. This condenser is placed inside the plug-in shield along with the 3.5-Mc. coil. The 100-μμfd. capacitor, C_1, which is connected permanently in the circuit, is sufficient for use with 7- and 14-Mc. crystals. Since larger coils are desirable for the plate circuit of the oscillator, the coils for L_2 are wound on 1-inch diameter forms enclosed in National Type PB-10 plug-in shield cans. The shield should be grounded to the chassis through one of the available pins in the base.

External connections to the unit are indicated in Fig. 1337. If both stages are to be keyed as shown, no fixed bias is necessary and all that is required is a grid leak of 5000-ohm 5-watt size, connected across the biasing terminals. This biasing system will serve also in case only the amplifier is to be keyed. Keying of the oscillator alone is not recommended because of the effects of soaring screen voltage, which makes it impossible to cut off plate and screen currents in this unit without exceeding the normal operating bias. For this reason, it is

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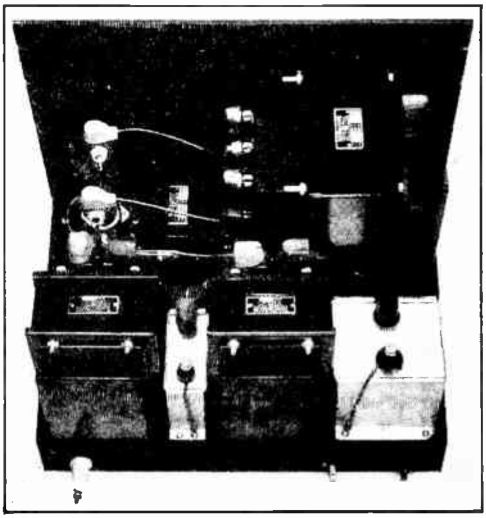


Fig. 1342 — **This power-supply unit delivers 2025 or 2480 volts at full-load current of 450 ma., with ripple of 0.5 per cent and regulation of 19 per cent. Voltages are selected by taps on the secondary. All exposed high-voltage terminals are covered with Sprague rubber safety caps and the tube plate terminals with moulded caps. The rectifier tubes are placed away from the plate transformer to avoid induction troubles. The panel is 14 × 19 inches and the chassis 13 × 17 × 2 inches. The exposed high-voltage terminal should be covered with a rubber-tubing sleeve. The circuit is shown in Fig. 1343.**

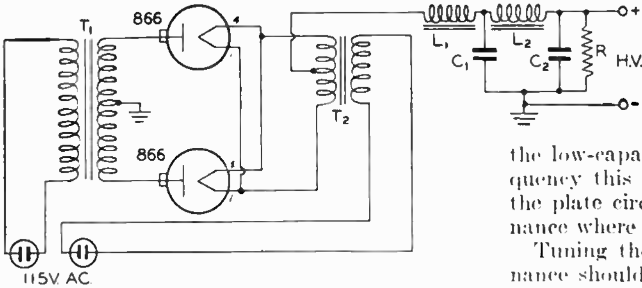


Fig. 1343 — Circuit diagram of the 2500-volt 150-ma. power supply.
C_1 — 1-μfd. 2500-volt oil-filled (GE Pyranol).
C_2 — 4-μfd. 2500-volt oil-filled (GE Pyranol).
R — 50,000 ohms, 200 watts.
L_1 — Input choke, 5–20 hy., 500 ma., 75 ohms (Thordarson T-19C38).
L_2 — Smoothing choke, 12 hy., 500 ma., 75 ohms (Thordarson T-19C45).
T_1 — 3000 or 2150 volts r.m.s. each side of center, 500-ma. d.c. (Thordarson T-19P68).
T_2 — 2.5 volts, 10 amp., 10,000-volt insulation (Thordarson T-64F33).
The voltage regulation may be improved by the use of a lower value of bleeder resistance, *R*, although at some sacrifice in maximum permissible load current. This circuit is also used for the 1500-volt supply shown in Fig. 1360.

highly advisable to use an overload relay in the plate-supply circuit of the amplifier, to protect the tube in case the oscillator fails to function. The circuit diagram of a suitable low-voltage supply for the oscillator delivering 350 to 400 volts is shown in Figs. 1340 and 1341. The high-voltage supply of Figs. 1342 and 1343 may be used for the output stage. The screen voltage-dropping resistor is not included in the unit because of the heat generated. It should be located externally, possibly in the power-supply unit, and should consist of two 50,000-ohm 160-watt resistors in parallel.

Adjustment — After the proper coils for the desired band have been plugged in and the crystal switch turned to select the proper crystal, the key may be closed with the low-voltage supply turned on, but with the high-voltage supply turned off. The combined oscillator plate and screen current at resonance should be between 35 and 75 ma., depending upon the crystal frequency and whether or not the oscillator is doubling frequency. If the oscillator is operating at the crystal fundamental frequency, oscillation will cease abruptly when the plate tank circuit is tuned to the high-capacitance side of resonance. For reliable operation this circuit should be tuned slightly to the low-capacitance side. When doubling frequency this characteristic disappears so that the plate circuit may be tuned to exact resonance where maximum output should occur.

Tuning the oscillator plate circuit to resonance should result in a grid-current reading when the meter is switched to the second meter-switch position. The reading will vary between 30 and 35 ma. to 50 ma. or more, depending upon the frequency and whether the oscillator is doubling frequency or working "straight through." The potential of the high-voltage supply should be reduced during preliminary adjustments. If no other means of reducing the voltage is available a 200-watt 115-volt lamp may be connected in series with the primary winding of the high-voltage transformer. The plate circuit of the amplifier should be tuned to resonance first with the antenna link swung out to the minimum-coupling position. The output tank circuit of the amplifier may be coupled through the link coil, either directly to a properly-terminated low-impedance transmission line, or through an antenna tuner such as the one shown in Figs. 1344 and 1345 to any type of antenna system. With the antenna system connected and the link swung in for maximum coupling, the plate current should increase when the antenna system is tuned through resonance. Every adjustment of the coupling or tuning of the antenna system should always be followed by a readjustment of the tuning of the amplifier tank circuit for resonance. As the loading is increased the plate current at resonance will increase. The loading may be carried up to the point where the plate current (cathode current, minus grid and screen currents) is 300 ma. at 3000 volts.

A Wide-Range Antenna Coupler

The photograph of Fig. 1344 shows the constructional details of a wide-range antenna coupler suitable for use with high-power transmitters. Various combinations of parallel and

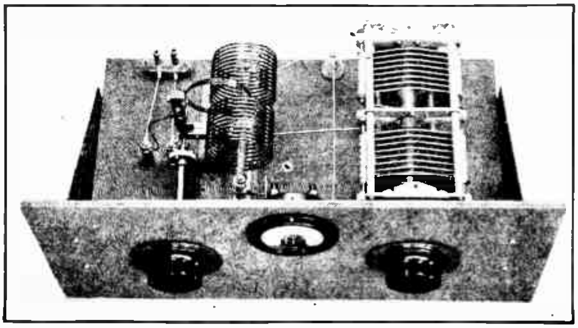


Fig. 1344 — Wide-range antenna coupler. The unit is assembled on a metal chassis measuring 10 × 17 × 2 inches, with a panel 8¾ × 19 inches in size. The variable condenser is a split-stator unit having a capacity of 200 μμfd. per section and 0.07-inch plate spacing (Johnson 200ED30). The plug-in coils are the B. & W. TVL series. The r.f. ammeter has a 1-ampere scale. If desired, the coils may be wound with fixed links on standard transmitting ceramic forms. The links will have to be provided with flexible leads which can be plugged into a pair of jack-top insulators mounted near the coil jack strip, unless a special mounting is made providing for the seven plug-in connections required.

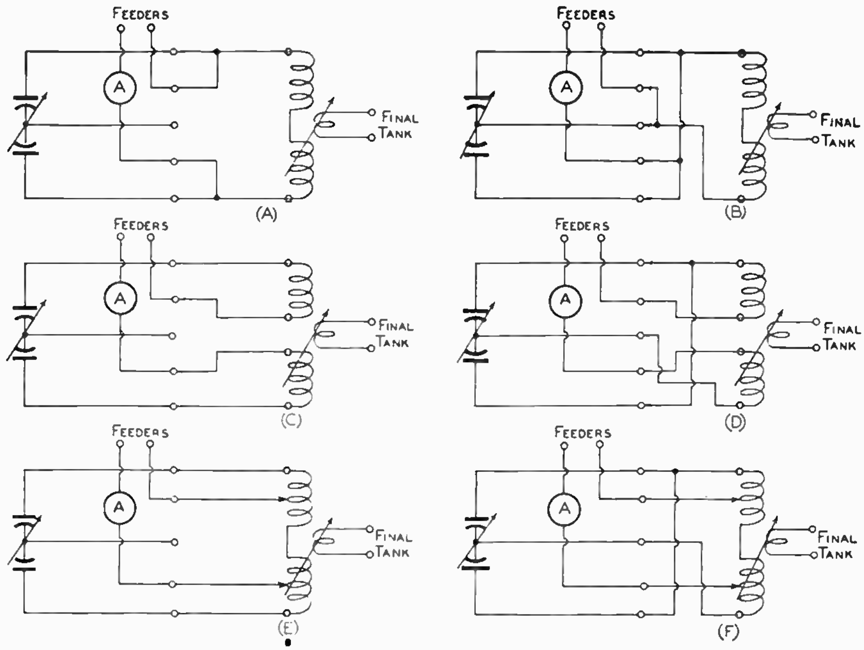


Fig. 1345 — Circuit diagram of the wide-range antenna coupler for use with the bandswitching amplifier. A — parallel tuning, low *C*. B — parallel tuning, high *C*. C — series tuning, low *C*. D — series tuning, high *C*. E — parallel tank, low-impedance output, low *C*. F — parallel tank, low-impedance output, high *C*. For single-wire matched-impedance feeders, the arrangements of E or F would be used with a single tap instead of the double tap shown. For simple voltage-fed antennas, the arrangement of A would be used with the antenna connected to the terminal with the ammeter. After the inductance required for each of the various bands has been determined experimentally, the connections to the coils can be made permanent. Then it will be necessary merely to plug in the right coil for each band, tune the condenser for resonance, and adjust the link for loading.

series tuning, with high- and low-*C* tanks and high- and low-impedance outputs, are available. Diagrams of the various circuit combinations possible with this arrangement are given in Fig. 1345.

A separate coil is used for each band, and the desired connections for series or parallel tuning with high or low *C*, or for low-impedance output with high or low *C*, are automatically made when the coil is plugged in. Coil connections to the pins for various circuit arrangements are shown in Fig. 1345.

The tuning condenser specified, together with a set of standard plug-in transmitting coils, should cover practically all coupling conditions likely to be encountered.

Because the switching connections require the use of a central pin, a slight alteration in the B & W-coil mounting unit is required. The central link mounting unit should be removed from the jack bar and an extra jack placed in the central hole thus made available. The link assembly should then be mounted on a 2-inch cone insulator to one side of the jack bar.

Correspondingly, the central nut on each coil plug base must be removed and a Johnson tapped plug, similar to those furnished with the coils, substituted. An extension shaft may then be fitted on the link shaft and a control brought out to a knob on the panel.

The split-stator tank condenser is mounted by means of angle brackets on four 1-inch cone-type ceramic insulators, and an insulated flexible coupling is provided for the shaft.

If desired, the coils may be wound with fixed links on ceramic transmitting coil forms. The links should be provided with flexible leads which can be plugged into a pair of jack-top insulators mounted near the coil jack strip, unless a special mounting is made providing for seven connections.

The unit as described should be satisfactory for transmitters operating at a plate voltage of up to 1500 with modulation and somewhat more on c.w. For appreciably higher voltages, a tank condenser with larger plate spacing should be used.

ℂ A Medium-Power Bandswitching Transmitter

The transmitter illustrated in Figs. 1346 through 1351 combines complete bandswitching from 80 meters through 10 meters with moderately high power. A 4-125A beam tetrode is used in the output stage, driven by frequency-multiplying stages which, because of the low driving-power requirements of the final, can loaf along at considerably below ratings. The final can be operated at 375 watts input for c.w. operation, or 300 watts in 'phone service.

As shown in Fig. 1348, a Pierce crystal oscillator is used, operated at low plate voltage

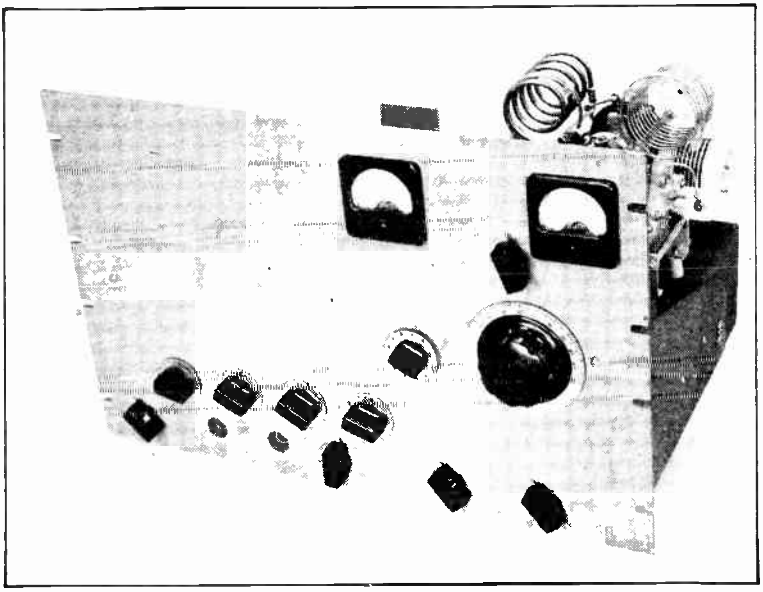


Fig. 1346 — **Front view of the bandswitching transmitter. The controls along the bottom of the panel are, left to right, the crystal selector switch, oscillator key jacks, low-power-stage bandswitch, plate-meter switch and grid-meter switch. Above, in the same order, are the tuning controls for the 6V6, the two sections of the 6N7, and the 807 plate. The 270-degree knob to the left of the main tuning dial is the excitation control. The knob between the meters controls bandswitching in the output stage. The plate meter is on the left.**

to permit maximum frequency stability. Ten-meter output can be obtained with 80-, 40-, or even 160-meter crystals, and output in the 11-meter band can also be obtained with suitable crystals.

The output of the crystal oscillator is coupled to the grid of a 6V6 which acts as either a straight amplifier or as a doubler, depending upon the fundamental frequency of the crystal and the position of the bandswitch. The plate tank coil for this stage is tapped, with the entire coil being used when the bandswitch is set for 80-meter output, and only a portion of it when output at higher frequencies is desired.

Plate voltage for the 6V6 is dropped to about 360 by R_{10}. The screen voltage of the tube is made adjustable by means of a 75,000-ohm wire-wound potentiometer, the excitation control, which, with the usual dropping resistor, forms a voltage divider across the plate supply. By changing the screen voltage, the output of the tube is adjusted to whatever level is required for adequate drive to the 4-125A.

When the bandswitch is set in either the 80- or 40-meter positions, the output of the 6V6 is fed to the grid of the 807. For 20- and 10-meter operation, the output of the 6V6 is switched to the grid of the first section of the 6N7 frequency multiplier. The 6N7 stages are arranged so that the grid not in use is grounded. For 20-meter operation only the first section of the 6N7 is used while for 10-meter operation both sections are used, operating as doublers from the 40-meter output of the 6V6.

The 807 operates straight through on all frequencies. In this stage the 80- and 40-meter ranges are covered by one coil, wound on a ceramic form and housed in a shield can above deck. The 20- and 10-meter ranges are covered by an air-wound coil, the plug-in type being used solely to permit removal of the 807 tube from its socket. Bandswitching in the 807 stage is accomplished by a ceramic switch similar in construction and contact arrangement to the multiple-section switch used in the earlier stages, and ganged to it through a right-angle drive mechanism. The screen circuit of the 807 includes a parasitic-suppressing resistor, R_{18}, inserted ahead of the usual screen by-pass condenser. Bias for the 807 is obtained from two series-connected 45-volt Mini-Max batteries.

With about 425 volts on the plate and 325 on the screen, the 807 delivers more than enough drive for the 4-125A final on all bands.

The circuit of the 4-125A final amplifier is designed to permit plate-and-screen modulation of the tube if 'phone operation is desired. Hence, a screen dropping resistor is used to furnish screen voltage from the plate supply. It is necessary to drop the screen voltage to 350 or 400 from whatever potential is used on the plate. Space limitations do not permit mounting a single 100-watt resistor inside the chassis, so two 50-watt units are mounted side by side and connected in series to obtain the required 100-watt rating.

Operating bias for the final is obtained by means of a grid resistor, no fixed bias being required. To keep the screen voltage from soaring to the full plate-supply value under key-up conditions or in the event of excitation failure, a 6Y6 tube is used as a protective device. The 6Y6 is triode-connected, with its plate connected to the screen end of the screen dropping resistor, and its grid connected to the grid side of the grid leak for the 4-125A. When excitation is present, about 200 volts of bias is applied to the grid of the 6Y6 from the IR drop across the grid resistor — more than enough to keep the tube nonconductive. However, when the key is up, excitation is removed, and the 6Y6 grid is without bias. Thus it draws plate current through the screen dropping resistance. The current drawn, in the neighborhood of 20 or 30 ma., is sufficient to reduce the screen voltage on the 4-125A to a very low value. As a result, the final plate current falls to 8 or 10 ma. — much better than relatively enormous amounts of fixed bias could do under similar conditions.

Three coils are used in the plate circuit of the final. The first, wound on a ceramic form, is used for 80- and 40-meter operation. A commercial air-wound coil with the plug strip removed is used in the 20-meter tank. The 10-meter coil is made of ¼-inch copper tubing. Bandswitching in the final amplifier is accomplished by a pair of ganged single-pole four-position switches of the heavy-duty type. Particular care should be taken to insure good insulation in mounting both switches because the r.f. potentials encountered are very high, especially when the final is unloaded during tune-up.

The use of fixed links for output coupling, a mechanical necessity, requires that an antenna tuning unit having a variable link be employed for proper adjustment of loading. The unit described in Figs. 1344 and 1345 will be suitable. The meters are switched across 22-ohm ½-watt resistors by double-pole ceramic switches. Both meters are 0–50-ma. range, additional shunts being used to extend the ranges to 100 ma. for the 807 plate circuit and to 500 ma. for the 4-125A cathode. The shunts and the 22-ohm resistors are mounted on the switch contacts. The shunt for the 807 stage is wound with resistance wire, but if this type of wire is not available a suitable length of No. 30 insulated wire may be used. A short length of the latter is all that is required for the shunt for the 4-125A. The metering circuits are arranged as follows:

PLATE METER

Position	*Circuit Read*	*Scale*
AB	6F6 plate and screen	50 ma.
CD	6Y6 plate and screen	50 ma.
EF	6N7 plate (20 meters)	50 ma.
GH	6N7 plate (10 meters)	50 ma.
IJ	807 plate	100 ma.
KL	4-125A cathode	500 ma.

GRID METER

Position	*Circuit Read*	*Scale*
MN	807 control grid	50 ma.
OP	807 screen	50 ma.
QR	4-125A control grid	50 ma.
ST	4-125A screen	50 ma.

The physical layout of the rig is shown in the photographs. The entire transmitter is built on a standard 17 × 13 × 4-inch steel chassis, with a 19 × 12¼-inch Masonite panel to permit rack mounting. While maximum use of the

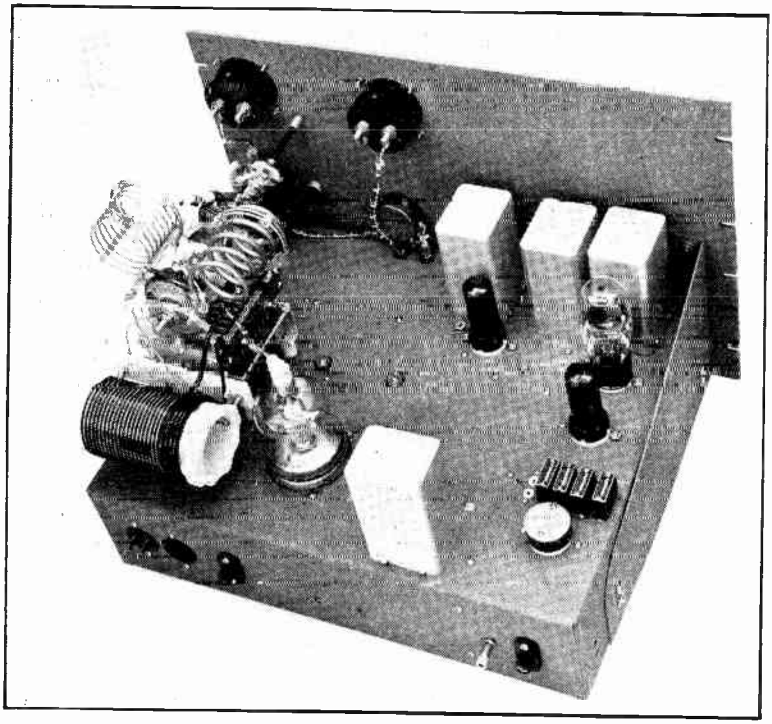
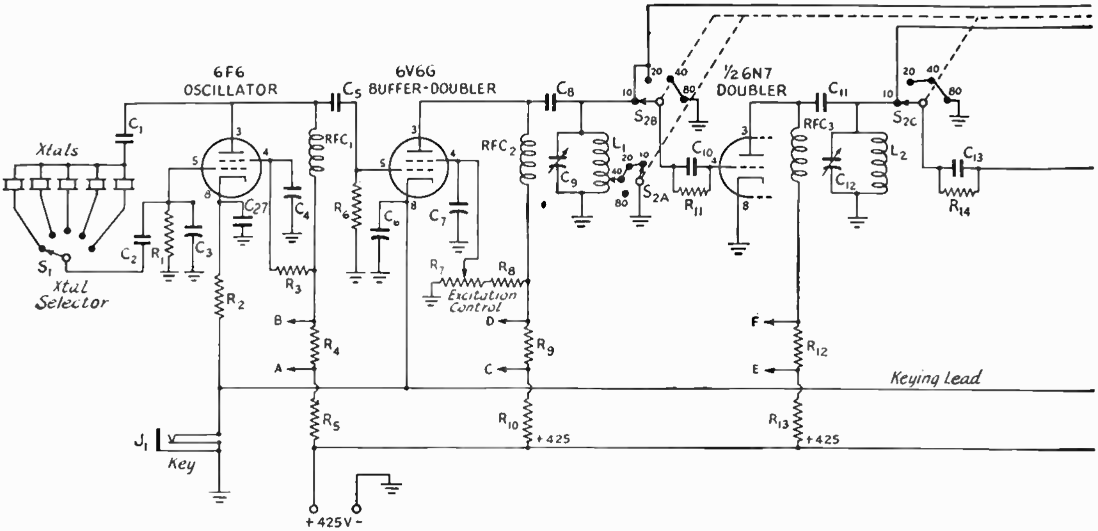


Fig. 1347 — Rear view of the band-switching transmitter showing placement of parts mounted above the chassis. Adequate space for the later addition of a VFO unit is available in the center of the chassis.



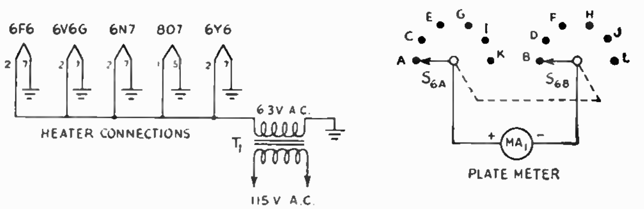


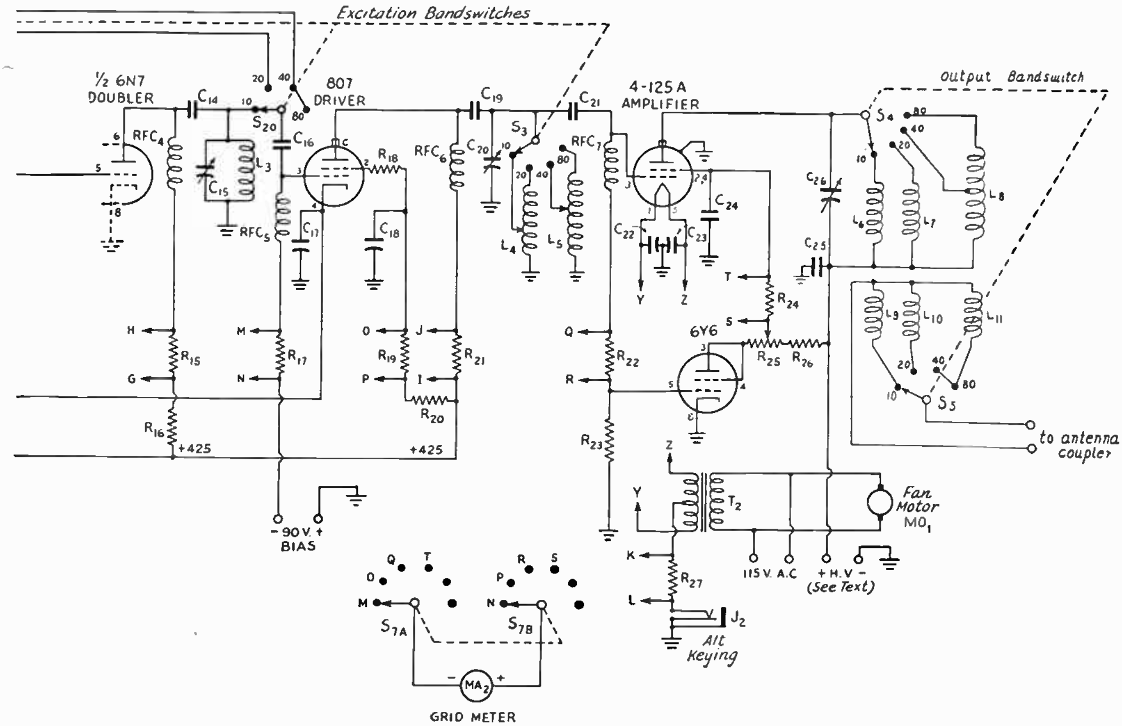
Fig. 1348 — Circuit diagram of the bandswitching transmitter.

C_1, C_2, C_{16} — 0.001-μfd. mica.
C_3, C_4, C_6, C_7, C_{17}, C_{18}, C_{22}, C_{23}, C_{27} — 0.01-μfd. 600-volt paper.
C_5, C_8, C_{21} — 150-μμfd. mica.
C_9 — 140-μμfd. receiving variable (Hammarlund MC-140-S).
C_{10}, C_{13} — 100-μμfd. mica.
C_{11}, C_{14} — 0.0022-μfd. mica.
C_{12}, C_{15} — 50-μμfd. receiving variable (Hammarlund MC-50-S).
C_{19} — 0.001-μfd. mica, 1200 volts working.
C_{20} — 100-μμfd. receiving variable (National ST-100).
C_{24} — 220-μμfd. mica, 5000 volts d.c. working.
C_{25} — 0.001-μfd. mica, 5000 volts working.
C_{26} — 120-μμfd. variable, 0.10-in. air gap (Cardwell XE-120-XS).
R_1 — 47,000 ohms, ½ watt.
R_2 — 1000 ohms, ½ watt.
R_3 — 68,000 ohms, ½ watt.
R_4, R_9, R_{12}, R_{15}, R_{17}, R_{19}, R_{22}, R_{24} — 22 ohms, ½ watt.
R_5 — 50,000 ohms, 3 watts (three 0.15-megohm 1-watt units in parallel).
R_6 — 0.1 megohm, ½ watt.
R_7 — 75,000-ohm wire-wound potentiometer.
R_8 — 50,000 ohms, 10 watts.
R_{10} — 15,000 ohms, 10 watts.
R_{11}, R_{14} — 22,000 ohms, ½ watt.
R_{13} — 7500 ohms, 10 watts.
R_{16} — 5000 ohms, 10 watts.
R_{18} — 68 ohms, ½ watt.
R_{20} — 30,000 ohms, 10 watts.
R_{21} — Meter shunt; see text.
R_{23} — 20,000 ohms, 5 watts.
R_{25} — 50,000 ohms, 50 watts, with slider.
R_{26} — 50,000 ohms, 50 watts.
R_{27} — Meter shunt: see text.
L_1 — 31 turns No. 22 d.s.c., 12½ turns ½ inch long between ground end and tap: 18½ turns closewound between tap and plate end. Wound on 1-inch diam. form (Millen 45000).
L_2 — 11 turns No. 22 d.s.c., 1 inch long on 1-inch diam. form (Millen 45000).
L_3 — 4 turns No. 20 d.s.c., 1 inch long on 1-inch diam. form (Millen 45000).

below-chassis space is required, there is enough space left above deck and on the front panel to permit the subsequent addition of a VFO unit if desired. There is adequate space on the panel for a National Type ACN dial, and clearance is provided between two of the coil shields for a shaft to tune the VFO.

The tube line-up, shown in Fig. 1347, has the 6F6 Pierce oscillator located about halfway back along the right-hand chassis edge, the 6V6 buffer-doubler immediately behind the oscillator, the 6N7 frequency multiplier to the left of the 6V6, the 4-125A final in the left foreground, and the 6Y6 screen-protecting tube in the corner, near the front panel. The 807 driver stage, mounted below the chassis, is visible in Fig. 1351. This view also shows the arrangement of the bandswitching system used for the low-power stages. A four-section ceramic switch is ganged to a similar single-section switch through a Millen right-angle drive mechanism.

The two ceramic switches at the lower left in the bottom view are for switching the meters. The small fan near the submounted socket for the 4-125A serves the dual purpose of cooling the final-amplifier tube base seals and the screen dropping resistors. The 807 driver is mounted parallel to the chassis surface in a Millen shield-and-socket assembly to prevent



L_4 — 35 turns No. 20 d.s.c., 16 turns ⅞ inch long between ground end and tap, 19 turns close-wound between tap and plate end. Wound on 1¼-inch diam. ceramic form (National XR-16).

L_5 — 7 turns No. 18 bare, 1³⁄₁₆-inch diam., 1⅜ inches long, tapped 3 turns from ground end. (National AR-16-10E with link and 1 turn of coil removed. Link connection on plug-in base used to bring out tap.)

L_6 — 5 turns ¼-inch copper tubing, 1½-inch i.d., 3⅛ inches long.

L_7 — 8 turns No. 14 bare tinned, 2-inch diam., 2 inches long (B & W 20BEL with 2 turns removed.)

L_8 — 26 turns No. 14 enam. tapped 15 turns from plate end, 3½ inches long, 2½-inch diam. ceramic form (National XR-10A).

L_9 — 2 turns No. 14 bare tinned, 2½-inch diam., wound over end of L_6 and spaced ¼ inch from it.

L_{10} — 2 turns No. 14 bare tinned, 2⅜-inch diam., wound over end of L_7 and spaced ¼ inch from it (Part of B & W 20BEL assembly.)

L_{11} — 4 turns No. 14 bare tinned, wound over ground end of L_8 and insulated from it by spaghetti tubing.

J_1, J_2 — Closed-circuit jack.

MA_1, MA_2 — 0–50 ma., 2-inch-square case.

MO_1 — Fan-and-motor assembly (Barber-Colman Type Yab 569–1, with Type Yab 355-2 2½-inch fan).

RFC_1 to RFC_7, inc. — 2.5-mh. r.f. choke (Millen 34102).

S_1 — 5-position single-pole ceramic switch (Centralab 2501).

S_2 — 4-section single-pole 4-position ceramic switch (Mallory 164-C).

S_3 — Single-section single-pole 4-position ceramic switch (Mallory 161-C).

S_4, S_5 — Single-pole 4-position ceramic switch, heavy-duty contacts (Ohmite T-504).

S_6, S_7 — Two-section double-pole 6-position ceramic (Centralab 2511).

T_1 — Filament transformer, 6.3 volts, 4 amp. (Stancor P-4019).

T_2 — Filament transformer, 5 volts, 10 amp. (Stancor P-6135).

feed-back from plate to grid, and a second shield plate runs from the 807 socket to the rear wall of the chassis to prevent stray coupling from the 807 plate circuits to the oscillator and doubler circuits. The crystal selector switch is mounted on a bracket bolted to the right-hand chassis edge, close to the oscillator tube and crystal sockets. The terminal board mounted near the meter switches holds all the plate and screen dropping resistors. The filament transformers and bias batteries are mounted near the left-hand edge of the chassis. Tuning condensers for the 6V6 and the 6N7 stages are mounted along the front edge of the chassis, while the tuning condenser for the 807 plate circuit is mounted near the rear, between the 807 plate cap and the grid connection of the 4-125A. The shaft for this condenser is brought out to the front panel at an angle by means of two National couplings of the "universal-joint" type.

Plate voltage for the low-power stages is brought in through a safety connector mounted on the rear chassis wall near the shield partition. The high voltage is brought in through a similar connector near the 4-125A and its screen dropping resistors. Power for the filament transformers and the fan is supplied through a male connector mounted on the left side of the rear chassis wall, with a female connector wired in parallel mounted alongside to permit the 115-volt source to be transferred elsewhere if desired.

All of the coils in the low-power stages, except that used for the 20- and 10-meter ranges in the 807 stage, are mounted above deck in National Type RO shield cans. The location of the three coils and the output links used in the final amplifier is shown in Fig. 1347.

The most important mechanical consideration in building the transmitter is the proper location and ganging of the bandswitches for the low-power stages. The usefulness of the rig will be greatly impaired if the switching system becomes balky or develops slippage. Thus any amount of time spent in properly mounting the switches, and the right-angle drive shaft which connects them, is worth while.

Fig. 1349 shows the location of the more important holes to be drilled in the chassis. The holes marked with an asterisk are those involved in mounting the right-angle drive mechanism, and are critical. The others are less critical and are included only to serve as a guide in construction.

Drill the holes for the posts which support the right-angle drive first. These posts are supplied by the manufacturer, and can be removed to facilitate mounting by releasing the Allen set-screws. Extreme care should be taken to insure that the holes drilled for the posts are lined up at exactly right angles to the front edge of the chassis, otherwise the entire switching system will be askew. After the holes are drilled, insert the posts, tighten them so that they are firm, and slide the drive mechanism on them with the "U"-shaped opening pointing in the direction shown in Fig. 1351.

When certain that the posts are placed correctly and that the gear box will slide on them with ease, remove the two short shafts that hold the bevel gears inside the frame of the drive unit. Replace one of these shafts with a 9¾-inch length of ¼-inch brass or aluminum shafting. This piece is to be the main drive shaft which runs through the front panel, through the right-angle drive assembly to the single-section bandswitch S_3, which is mounted at the rear of the chassis, near the 807. The other shaft is replaced by the shaft of the four-section bandswitch, S_2. Saw off all but ⅞ inch of this shaft, measuring from the point where the shaft enters the bushing on the front of the switch. Insert it in the drive mechanism and replace the gear so that it meshes with the gear on the other shaft of the drive.

The rear of the four-section bandswitch should be supported by a bracket made of

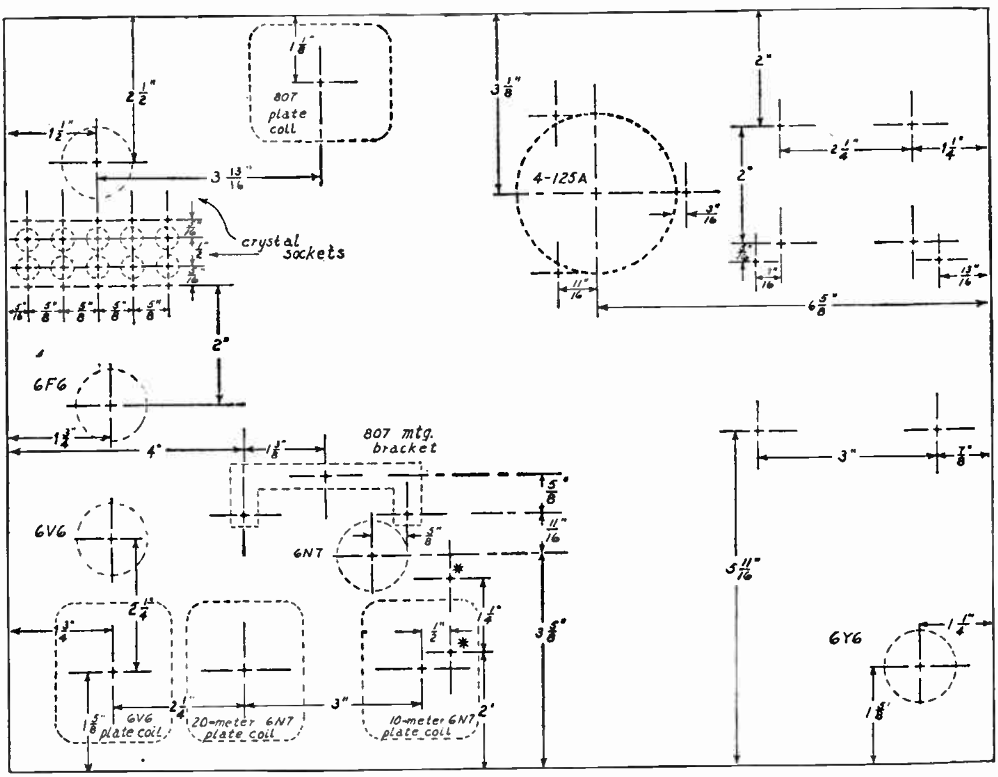


Fig. 1349 — Layout of the top surface of the chassis. The holes marked with an asterisk are for mounting the right-angle drive. Others, which are included for the convenience of the constructor, are less critical and may be rearranged slightly to suit individual needs.

⅛-inch aluminum. The dimensions of this bracket are shown in Fig. 1350. The rear section of the bandswitch is held ⅜ inch away from the bracket by ¼-inch spacers plus a couple of fiber washers.

The single-section bandswitch used in conjunction with the 807 stage is supported from the rear as shown in Fig. 1351. The rear of the ceramic switch wafer should be held about 1⅛ inch from the chassis wall by small metal spacers.

After the low-power bandswitching system has been installed and is operating satisfactorily, the mounting holes can be drilled for the other parts to be located below deck. The location of these parts is not critical, and can be determined from the photographs.

The fan motor is mounted on one of the brackets supplied with the Millen 807-tube shield-and-bracket assembly. The bracket itself is bolted to the chassis with screws which pass through small rubber grommets. This mounting, which reduces the amount of vibration transferred to the chassis, will be a necessity if the addition of a VFO to this transmitter is contemplated. A bottom plate for the chassis, with a few ventilating holes drilled near the rear of the fan motor, should be used to insure maximum effectiveness of the fan. Considerable heat is generated within the transmitter, and care must be taken to insure an adequate flow of air around the tube base to avoid cracking the seals.

The socket for the 4-125A is mounted below the chassis on ½-inch spacers. Small spring contacts, made from shim stock or thin phosphor-bronze and formed to contact the grounding ring on the base of the tube, are fastened under the screws that hold the socket in place. The tube itself is inserted in the socket through a 2¾-inch hole in the chassis. This arrangement provides the necessary shielding between plate and grid circuits to prevent oscillation.

The final tank assembly is constructed as a single unit, removable from the chassis, and built entirely on the framework of the tuning condenser. The 80-and-40-meter coil form is mounted on the rear frame of the condenser, and held away from the frame by ½-inch spacers. The 20- and 10-meter coils are mounted on brackets made of ¼-inch polystyrene, and are positioned so that the links are nearest the front panel. The brackets are bolted to the frame of the tuning condenser. The two heavy-duty switches are also supported by these brackets. The shafts of these switches are ganged by an insulated coupling. The entire tuning-condenser-and-tank-coil assembly is supported by 1-inch ceramic stand-off insulators and "U"-shaped brackets which provide 1-inch clearance between the rotor plates of the condenser and the chassis.

The output connectors are banana jacks mounted on a piece of ¼-inch polystyrene which replaces one of the two Mycalex bars on the tuning condenser. The centers of the jacks

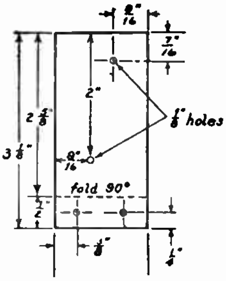


Fig. 1350 — Dimensions of the angle bracket used to support the rear of the four-section bandswitch.

are spaced ¾ inch to fit a standard banana-plug assembly.

The winding specifications for the coils used in the low-power stages are given in Fig. 1348. These coils should be wound and mounted before any of the wiring around the bandswitch is started, otherwise the coil leads will be inaccessible. The coil forms used in the 6V6 and 6N7 stages are mounted about ½ inch above the chassis by small spacers. The ceramic form used in the 80-and-40-meter coil for the 807 is held away from the chassis by small angle brackets. Where it is necessary to run leads from the coils through the chassis, Millen ceramic bushings are used. Wiring will be simplified if the bandswitch assembly is removed temporarily while the connections around the sockets are made. Some of the wiring on the bandswitch itself can be done while the switch is out of the chassis. All wiring in the bandswitch assembly is done with No. 16 bare tinned wire.

A ¾-inch ceramic feed-through bushing is used to carry the high-voltage lead through the chassis to the plate connection of the final tube. The junction of the two screen dropping resistors is mounted on this bushing with a National GS-10 stand-off insulator to prevent shorting. The other ends of the screen resistors are supported by two more of these stand-offs from the rear chassis wall.

The low-power stages can be operated from the supply shown in Figs. 1340 and 1341. The final amplifier is designed for use with 1800 to 2000 volts on its plate, but as much as 2500 volts can be used if only c.w. operation is planned. If it is desired to run the final in 'phone service at more than 2000 volts, a tuning condenser with wider spacing between plates will be required. It should also be noted that the 807 stage is designed for operation at no more than 450 volts, as this is more than enough to secure the output required to drive the final. If operation at higher voltage is planned, a larger tuning condenser will be required in this stage.

After checking the wiring carefully, power may be applied to the low-power stages of the transmitter. Turn the excitation control to maximum and set both bandswitches to the 10-meter position, one meter switch to the

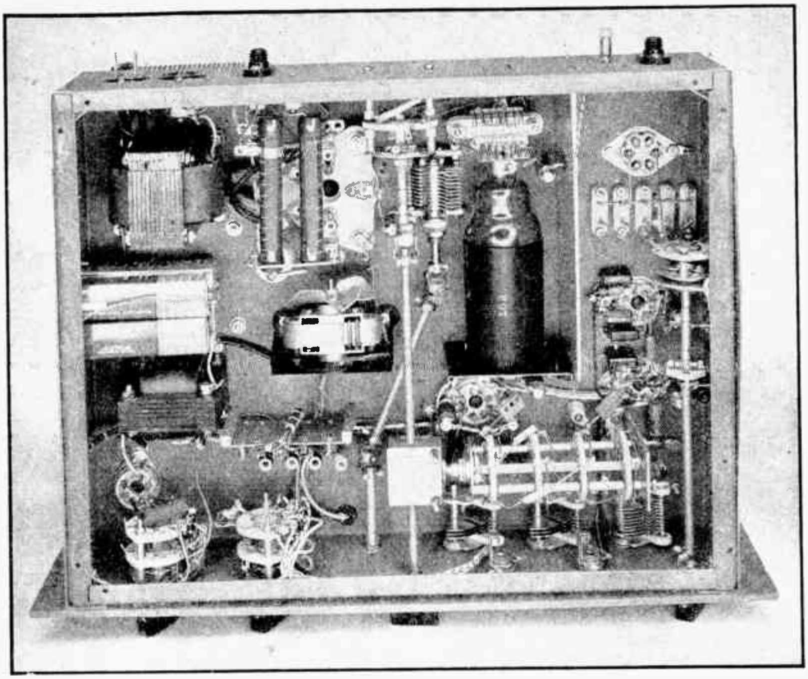


Fig. 1351 — Bottom view of the chassis of the band-switching transmitter showing location of parts and wiring. The crystal selector switch and the sockets for the 6F6 oscillator and the 6V6 buffer-doubler are on the right. The 6N7 socket is visible between the four-section bandswitch and the 807 mounting bracket. The 6Y6 protective-tube socket, the meter switches and the terminal board for the plate- and screen-dropping resistors are in the lower left-hand corner.

position which reads plate-and-screen current in the 6V6 stage, and the other to read grid current in the 4-125A. Tune the 6V6 plate circuit to resonance as indicated by a slight dip in the meter reading. Turn the meter switch to read plate current in the 20-meter section of the 6N7. The dip in plate current as this stage is tuned to resonance should be pronounced. A similar procedure is followed in tuning the 10-meter section of the 6N7. Plate current in this stage will be considerably higher than in the 20-meter section. The dip in plate current as the 807 plate circuit is resonated should also be pronounced, dropping from about 80 or 90 ma. to 30 or 40 at resonance. Grid current to the final stage should be measured at this time. If everything is as it should be, there should be at least 10 ma. of grid current. If more than 10 ma. is indicated, back off the setting of the excitation control until it falls to 10 ma. The control exerted by this potentiometer is not linear, and it may be found that there is little or no change in grid current over a considerable portion of the adjustment; in fact, the grid current may increase somewhat at first as the control is backed off. This is an indication that the drive to the 807 grid is excessive, causing its screen current to rise higher than normal, and reducing the output of that stage.

Once the low-power stages have been adjusted to give the rated amount of grid drive to the 4-125A, plate voltage may be applied to the final. When tuning up, reduced plate voltage is advisable to prevent the tube from being damaged should the final tank coil fail to resonate. A dummy load — a 200-watt lamp bulb, for example — should be connected to the output terminals of the transmitter before plate voltage is applied to the final. This is a "must," since the screen dropping resistance must be adjusted to provide rated screen voltage with the final loaded. Adjustment under any other condition will be useless. Tune the final tank circuit to resonance. The plate current at this time, with the load coupled to the final, should not exceed 150 ma. It should be remembered that the meter reads combined plate and screen current, so the screen current, which can be read on the other meter, must be subtracted from the indicated value to get the true plate current.

If plate current is too high the probable reason is excessive screen voltage. The slider on the screen dropping resistor should be adjusted to apply 350 volts to the screen when the tube is operating at full plate voltage, with rated grid drive, rated screen current, and working at full load. Be sure to remove the plate voltage from the final before adjusting the resistor! If the plate current is excessive after the screen voltage is set at the right value, decrease the loading on the final, remembering that with a change in loading the screen current changes and therefore the screen voltage will have to be readjusted. Too much screen voltage will result in excessive plate current and consequent overheating of the plate. Too much grid drive will cause a sharp drop in output because the screen current increases with grid excitation, in turn reducing the screen voltage. Optimum grid drive can best be determined under actual on-the-air conditions, using feeder current as an indication of maximum output.

Operating Voltages & Currents in the 4-125A Bandswitching Transmitter.

Conditions of measurement: Transmitter tuned for 10-ma. grid drive to 4-125A, 28-Mc. output. Supply voltage, 430. Readings obtained with 20,000-ohm-per-volt meter.

Tube	*Element*	*Volts*	*Ma.*
6F6	Plate Screen	128 70	6
6V6	Plate Screen	360 9	6
6N7	Plate 1 Plate 2	340 300	12 28
807	Plate Screen Grid	430 340 −90	28 3 *
4-125A	Plate Grid Screen	2000 −200 350	150 10 30

Note: Plate current and screen voltage and current, in the 4-125A, are dependent upon external loading.
* Less than 1-ma. grid drive to 807 required to produce 10-ma. drive to grid of 4-125A.

Those who plan to use the rig on c.w. only may find the use of a fixed screen supply more satisfactory than the screen-dropping-resistor method, although means must be provided to remove screen voltage whenever plate voltage is removed to prevent damage to the tube.

The accompanying table gives representative voltages and currents measured under operating conditions. Some variation from these figures can be expected depending upon the actual supply voltages used, but they will serve as a general indication, useful in checking performance.

The transmitter may be operated with a separate VFO unit as its frequency control by removing the crystal-oscillator tube from its socket and feeding the output of the VFO between the plate pin of the oscillator socket and ground. Adequate drive for either 'phone or c.w. operation can be obtained on all bands with a VFO such as that described in Fig. 1352.

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It should be noted that this method is satisfactory only in cases where direct coupling will not short-circuit the plate supply. In other cases, the VFO should be coupled through a 0.001-μfd. blocking condenser.

A Simple VFO Crystal Substitute

Figs. 1352, 1354 and 1355 show different views of a VFO unit with sufficient power output to drive the average crystal-oscillator tube. As the circuit diagram of Fig. 1353 shows, it consists of a 6SK7 ECO followed by a pair of 6F6s as isolating amplifiers. The primary frequency range covered by the oscillator is 3500–4000 kc., but this range may be shifted lower to cover 3395–3800 kc. for multiplying to cover the frequencies in the 10- and 11-meter bands by readjustment of the band-setting condenser, C_2. Plate and screen voltages are provided by a small built-in voltage-regulated power supply. Only the plate of the output tube is operated at the full power-supply voltage, the voltage of the rest of the plates and screens being limited to 150 by the VR tube.

Construction — The oscillator portion is constructed as a separate unit in a standard 3 × 4 × 5-inch steel box. The tuning condenser, C_1, and the coil form for L_1 are fastened to the rear wall of the box. C_1 is coupled to the National Type AM dial by a short extension shaft and a flexible coupling. The band-setting air condenser, C_2, is mounted against the right side of the box near the lower rear corner where it can be adjusted from the outside with a

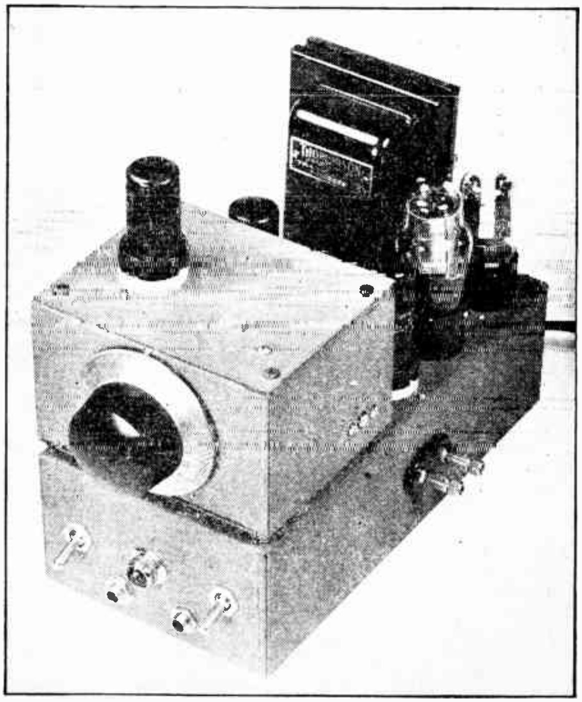

Fig. 1352 — The complete VFO unit. The oscillator is housed in a separate compartment which is shock-mounted on rubber grommets. The oscillator tube is on top of the compartment. To the rear are the two 6F6 amplifier tubes, the VR tube, the rectifier and the power transformer. In front are the stand-by switch, the power switch, pilot lamp and the two keying jacks. The output terminals are to the right.

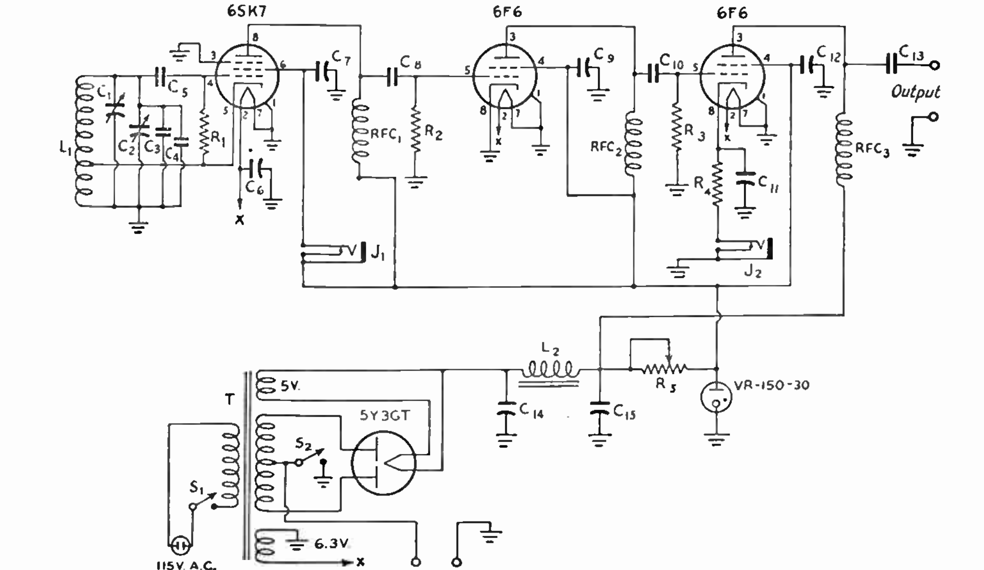

Fig. 1353 — Circuit diagram of the simple VFO.

C_1 — 100-$\mu\mu$fd. variable (Hammarlund MC-100S).
C_2 — 75-$\mu\mu$fd. variable (Hammarlund APC75).
C_3 — 220-$\mu\mu$fd. zero-temp.-coëf. mica.
C_4 — 68-$\mu\mu$fd. zero-temp.-coëf. mica.
C_5, C_8, C_{10}, C_{13} — 100-$\mu\mu$fd. mica.
C_6, C_7, C_9, C_{11}, C_{12} — 0.01-μfd. paper.
C_{14}, C_{15} — 8-μfd. 450-volt electrolytic.
R_1, R_2 — 47,000 ohms, ½ watt.
R_3 — 0.1 megohm, ½ watt.
R_4 — 220 ohms, 1 watt.
R_5 — 5000 ohms, 25 watts.
L_1 — 17 turns No. 20 enam., 1⅛ inches long, 1-inch diam., tapped 5 turns from ground end.
L_2 — 30 hy., 50 ma. (Stancor C-1003).
J_1, J_2 — Closed-circuit jack.
RFC_1, RFC_3 — 2.5-mh. r.f. choke.
RFC_2 — Millen 47002 (½-inch diam. by 2½ inches long) polystyrene form wound full with No. 30 d.s.c. wire.
S_1, S_2 — S.p.s.t. toggle switch.
T — 340 volts each side center, 55 ma.; 5 v., 2 a.; 6.3 v., 1½ amp.

screwdriver to set the beginning of the tuning range. The tube is mounted externally on top of the box where it will be well ventilated and where its heat will have minimum effect upon the tuned circuit. The coupling lead between the plate of the oscillator tube and the grid of the first 6F6 is made with flexible wire passed through National TPB polystyrene bushings, one in the oscillator compartment and one in the base chassis, the rigid wire which comes with the bushing having first been removed by warming with a soldering iron. The power and keying leads are brought out in a similar manner through holes lined with rubber grommets. The oscillator box is shock-mounted by means of long machine screws at each corner of the bottom plate. The screws pass through grommet-lined holes in the top of the chassis.

The base chassis is 5 × 10 × 3 inches. The two 6F6s are mounted on either side of the chassis immediately behind the oscillator compartment. Underneath, the filter choke is fastened against the side of the chassis in the left rear near the two filter condensers, C_{14} and C_{15}. The two plate r.f. chokes, RFC_2 and RFC_3, are mounted near their associated tube sockets. On the front edge are the control switches, S_1 for power, and S_2 which is the stand-by switch, cutting off plate voltage to all stages. Terminals in parallel with S_2 are mounted in the rear edge of the chassis to connect to a send-receive relay if this is found desirable. The output terminals are set in the right-hand side.

Adjustment — The resistance of R_5 should be adjusted experimentally so that the VR

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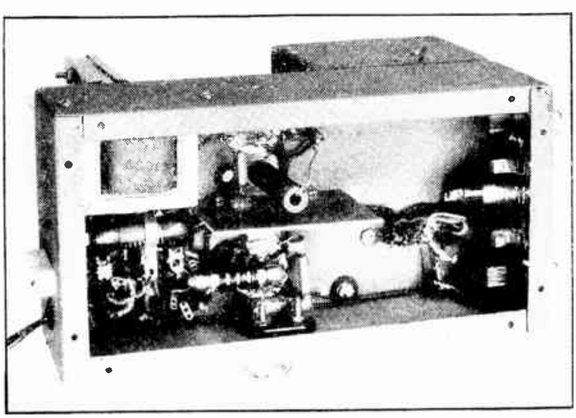

Fig. 1354 — Bottom view of the VFO unit showing the filter choke and the various r.f. chokes and by-pass condensers associated with the amplifiers.

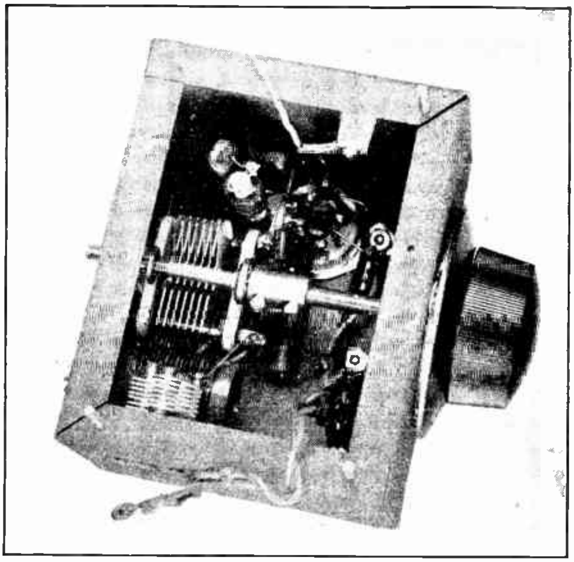

Fig. 1355 — Bottom view of the oscillator compartment. The tuning condenser and the coil are fastened to the rear wall of the box, while the air trimmer is mounted on the lower end in the photograph. The small cone insulator supports the coupling lead to the first amplifier stage.

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tube is ignited with the key either closed or open. If the glow disappears when the key is closed, the resistance of R_5 should be reduced. With the dial set for maximum capacitance of C_1, C_2 should be adjusted with a screwdriver to set the frequency at 3500 kc. (3395 if the VFO is to be used for 10- and 11-meter operation). C_1 should then cover the range to 4000 kc. (or 3800 kc.).

Coupling to the crystal oscillator in most transmitters is simply a matter of running a wire from the "hot" output terminal (the terminal connected to the plate of the output tube through C_{13}) to the grid of the oscillator tube, and the other output terminal to the chassis of the transmitter. In Tri-tet and grid-plate oscillator circuits, the cathode tanks should be short-circuited. In triode or tetrode crystal-oscillator circuits using parallel plate feed, it may be necessary to shift to series feed to prevent low-frequency parasitic oscillation because of the r.f. chokes in both the input and output circuits. In Pierce circuits, the oscillator tube may be fed as a grounded-grid amplifier by connecting the output terminals of the VFO in series between the cathode and the biasing resistor and by-pass. As an alternative, in this type of circuit, the oscillator tube may be eliminated and the VFO fed to the grid of the next tube.

Keying — Best keying characteristics will be obtained by keying the output stage although a second keying jack, J_1, is included for use if break-in operation is necessary. Since the key would be at 150 volts above ground, a keying relay or vacuum-tube keyer should be used here to avoid the danger of shock. In keying the oscillator, any key-click-filter lag should be kept at the minimum required for satisfactory click suppression, to avoid chirps. Usually, r.f. chokes only at the relay terminals will be sufficient. As much lag as is desired can be used when keying the output stage, since keying at this point does not affect the frequency.

The oscillator draws 8 ma. in the plate circuit and 3 ma. in the screen circuit. The plate current of the first amplifier should run about 15 ma. with the oscillator key closed and 32 ma. when excitation is removed. The output-stage currents should be 17 ma. with excitation and 25 ma. without excitation.

A Push-Pull Amplifier for 200 to 500 Watts Input

Figs. 1356, 1357 and 1359 show various views of a compact push-pull amplifier using tubes of the 1500-volt 150-ma. class, although

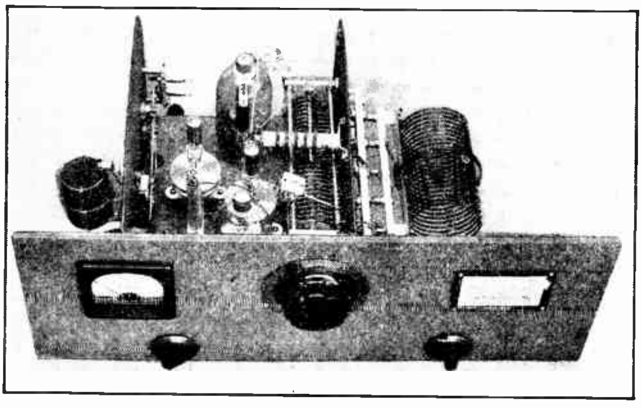


Fig. 1356 — A general view of the compact 450-watt push-pull amplifier, showing the front-panel and top-of-chassis arrangement. Mounted on a standard relay rack, the height is only 7 inches and the depth 9 inches. Grid and plate tank circuits are isolated from each other by the double shielding partitions. On the panel are the 0–100-ma. milliammeter, which is switched to read current in all circuits, the plate-tank tuning dial, and a chart giving coil and tuning data. The small knob at the left below is the grid-circuit tuning control, while the one to the right is for the meter switch. The tube sockets are mounted adjacent to the stator terminals of the plate-tank condenser, C_2, in the center, with the neutralizing condensers between, providing short leads.

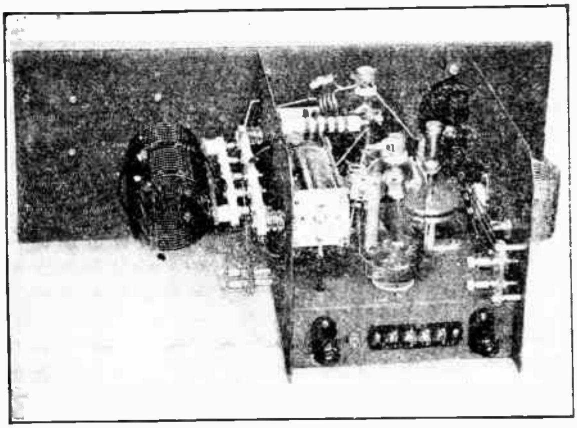


Fig. 1357 — All components of the 450-watt push-pull amplifier are assembled around a small metal chassis 7 × 2 × 9 inches deep. The partitions are standard 6½ × 10-inch interstage shields. The plate tank condenser is mounted on the left-hand partition. The plate tank-coil jack-bar is mounted centrally, opposite the condenser, on spacers which give ½-inch clearance between the strip and the partition. C_{10} is mounted with a small angle bracket on the partition under the center of C_2. The socket for the grid tank coil is mounted just above the chassis line. Millen safety terminals are used for the external high-voltage plate and bias connections.

the design is also suitable for use with tubes of the 1000-volt 100-ma. class. With the lower plate voltages a plate tank condenser with a spacing between plates of 0.05 inch, and smaller tank coils, may be used.

The circuit, shown in Fig. 1358, is quite conventional, with link coupling at both input and output. The tuned circuits, L_3C_6 and L_4C_5, are traps important for the prevention of v.h.f. parasitic oscillations. The 100-ma. meter may be shifted between the grid and cathode circuits for reading either grid current or cathode current. When shifted to read cathode current, the meter is shunted by a resistor, R_2, which multiplies the scale reading by five. This resistor is wound with No. 26 copper wire, the length being determined experimentally to give the desired scale multiplication.

Construction — The mechanical arrangement shown in the photographs results in a compact unit requiring a minimum of panel space. The tank condenser is mounted on the left-hand partition (Fig. 1357) at a height which brings its shaft down 2⅝ inches from the top of the panel. The plate tank-coil jack bar is mounted centrally with the condenser on spacers which give a ½-inch clearance between the strip and the partition. C_{10} is mounted with a small angle on the partition under the center of C_2. Leads from both ends of the rotor shaft are brought to one side of C_{10} for symmetry.

The two tube sockets are mounted in a line through the center of the chassis and at op-

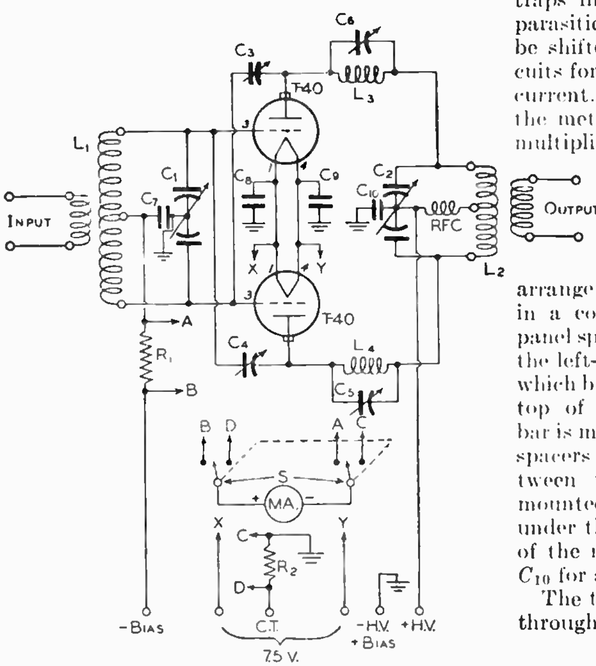


Fig. 1358 — Circuit diagram of the 450-watt push-pull amplifier.

C_1 — 100 μμfd. per section, 0.03-inch spacing (Hammarlund HFAD-100-B).
C_2 — 100 μμfd. per section, 0.07-inch spacing (Hammarlund HFBD-100-E).
C_3, C_4 — Neutralizing condenser (National NC-800).
C_5, C_6 — 3–30-μμfd. mica trimmer (National M-30).
C_7, C_8, C_9 — 0.01-μfd. mica.
C_{10} — 0.001-μfd. mica, 7500-volt rating (Aerovox 1653).
R_1 — 22 ohms, 1 watt.
R_2 — Meter multiplier resistance for 5-time multiplication, wound with No. 26 wire.
L_1 — B & W JCL series, dimensions as follows: *
  3.5 Mc. — 11 turns No. 20, 2⅛ inches long.
  7 Mc. — 26 turns No. 16, 2⅛ inches long.
  14 Mc. — 14 turns No. 16, 1⅞ inches long (remove 2 turns from B & W coil).
  28 Mc. — 6 turns No. 16, 1⅞ inches long (remove 2 turns from B & W coil).
L_2 — B & W TCL series, dimensions as follows: **
  3.5 Mc. — 26 turns No. 12, 3½-inch diam., 4½ inches long.
  7 Mc. — 22 turns No. 12, 2½-inch diam., 4½ inches long.
  14 Mc. — 10 turns No. 12, 2½-inch diam., 4¼ inches long, remove one turn from each end.
  28 Mc. — 4 turns ¼-inch copper tubing, 2½-inch diam., 1½ inches long. Remove one turn from each end.
L_3, L_4 — 4 turns No. 14, ½-inch diam., ¾ inch long.
MA — 100-ma. milliammeter.
RFC — 1-mh. r.f. choke (National R-154U).
S — 2-section 2-position rotary switch.

* All 1½-inch diam., 3-turn links.
** All coils fitted with 2-turn links.

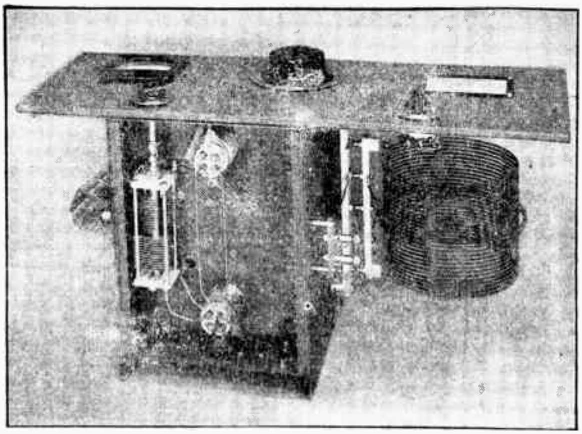


Fig. 1359 — Bottom view of the 450-watt push-pull amplifier. The grid tank condenser is mounted between the two tube sockets which are set below the chassis on brackets. Connections between the condenser terminals and the coil socket above pass through grommet-lined holes in the chassis. The partitions provide shielding between input and output tank coils.

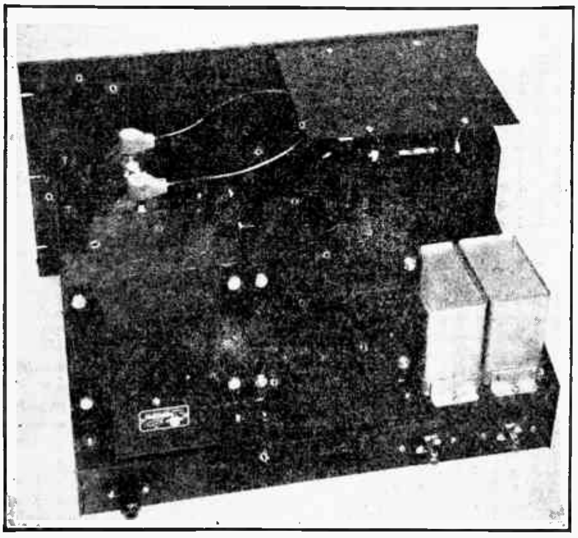
posite ends of the plate tank condenser. They are spaced about one inch below the chassis on long machine screws. The neutralizing condensers are placed between the two tubes, so that the leads from the plate of one tube to the grid of the other are short. The r.f. choke is mounted just above the tank condenser.

The right-hand partition is cut out at the forward edge to clear the meter. This cut-out can be readily made with a socket punch and a hacksaw. The socket for the grid tank coil is mounted 4½ inches behind the panel, just above the chassis line.

The grid tank condenser, C_1, is mounted under the chassis without insulation. Large clearance holes, lined with rubber grommets, are drilled for connecting wires which must be run through the chassis or partitions. The parasitic traps are made self-supporting in the plate leads from the tank condenser to the tube caps. The panel is placed so that the plate tank-condenser shaft comes at the center. The meter switch is mounted to balance the knob controlling C_1.

Power supply and excitation — The T-40 tubes shown in the photographs operate at a maximum plate voltage of 1500 for c.w. work. For this, the unit shown in Fig. 1360 is suitable. The supply shown in Fig. 1326, minus the VR-tube branch, will provide the biasing voltage required for plate-current cut-off. R_2 should have a resistance of 2500 ohms and R_3 of 1500 ohms. A filament transformer delivering 7.5 volts at 5 amperes also will be required. The exciters of Figs. 1311 or 1331 will furnish adequate drive.

Fig. 1360 — This power supply delivers 1500 or 1250 volts at a full-load current of 425 ma., with 0.25-per-cent ripple and regulation of 10 per cent. Voltages are selected by taps on the transformer secondary. The secondary terminal board is covered with a section of steel panel supported by brackets fastened underneath the core clamps and insulating caps are provided for the tube plate terminals. A special safety terminal (Millen) is used for the positive high-voltage connection. The panel is 10½ × 19 inches and the chassis size is 13 × 17 × 2 inches. The circuit for this supply is shown in Fig. 1313. The following values should be used:



C_1, C_2 — 4-μfd. 2000-volt paper (C-D TJU20040).
R — 20,000 ohms, 150 watts.
L_1 — 5/20 hy., 500 ma., 75 ohms (Stancor C1405).
L_2 — 8 hy., 500 ma., 75 ohms (Stancor C1415).
T_1 — 1820 or 1520 volts r.m.s. each side of center-tap. 500-ma. d.c. (Stancor Type P6157).
T_2 — 2.5 volts, 10 amp., 10,000-volt insulation (Stancor Type P3025).

For a 1000-volt supply, the following values are suggested:

C_1, C_2 — 4-μfd. 1500-volt paper (Aerovox 1505).
R — 30,000 ohms, 50 watts.
L_1 — 8/30-hy. filter input choke, 250 ma. (Stancor C-1702).
L_2 — 15-hy. filter smoothing choke, 250 ma. (Stancor C-1703).
T_1 — 1250 or 1000 volts r.m.s. each side of center-tap, 250 ma. (Stancor P-4030).
T_2 — 2.5 volts, 10 amp. (Stancor P-3025).

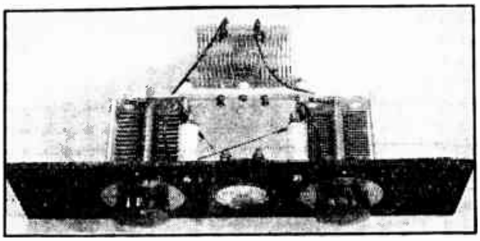


Fig. 1361 — A link-coupled antenna-tuning unit for use with resonant feed systems and medium-power amplifiers. The inductance, with variable link, is mounted on the condenser frames. Clips are provided for changing the number of turns and for switching the condensers from series to parallel. The panel is $5\frac{1}{4} \times 19$ inches.

Tuning — After the amplifier has been neutralized, a test should be made for parasitic oscillation. The bias should be reduced until the amplifier draws a plate current of about 100 ma. without excitation. With C_1 adjusted to various settings, C_2 should be varied through its range and the plate current watched closely for any abrupt change. Any change will indicate oscillation, in which case C_5 and C_6 should be adjusted simultaneously in slight steps until the oscillation disappears. Unless the wiring differs appreciably from the original, complete suppression will be obtained with the two condensers at full capacity. Changing bands should have no effect upon this adjustment.

With normal bias replaced, the amplifier should now be tuned up and the excitation adjusted so that a grid current of 60 ma. is obtained with the amplifier fully loaded. Full loading will be indicated when the cathode-current meter registers 360 ma., which includes the 60-ma. grid current. Under these conditions the biasing voltage should rise to 150 volts, dropping to about 70 volts without excitation when the plate current will fall to almost zero.

If the amplifier is to be plate-modulated, the plate voltage should be reduced to 1250 and the loading decreased to reduce the plate current to 250 ma. The same bias-supply adjustment will be satisfactory for this type of operation but excitation may be reduced to give a grid current of 40 ma., bringing the total cathode current to 290 ma. The antenna tuner shown in Fig. 1361 may be used.

Operating conditions for tubes of other characteristics will be found in Chapter Twenty.

ℂ Antenna Tuner for Medium Power

The antenna tuner shown in Fig. 1361 will usually be satisfactory for amplifiers operating at plate voltages not in excess of 1250.

The two condensers are mounted from the panel by means of insulating pillars taken from National GS-1 insulators, which are fastened to the end plates with small sections of machine screws from which the heads have been cut. The variable link coil is mounted between the two rear end plates. The size of the coil is varied by short-circuiting turns, using clips which are attached to the condensers with flexible leads. As shown by the circuit diagram, Fig. 1362, the condensers are connected in parallel when the second pair of clips connects each rotor to the stator of the opposite condenser. The feeders are connected to the two large stand-off insulators mounted on the panel.

ℂ A Compact 450-Watt Push-Pull Amplifier

The photographs of Figs. 1363, 1365 and 1366 show an amplifier designed along the lines of the type of construction often referred to as "dish type." This style of construction has many advantages, although its use normally is confined to components of moderate physical dimensions and weight.

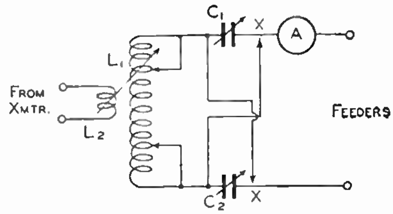


Fig. 1362 — Circuit diagram of the link-coupled antenna tuning unit for use with medium-power transmitters.

C_1, C_2 — 100-μμfd. variable, 0.07-inch spacing (National TNC-100).
L_1 — 22 turns No. 14, diam. $2\frac{3}{4}$ inches, length 4 inches (Coto with variable link).
L_2 — 4 turns, rotating inside L_1.
A — R.f. ammeter, 0–2.5-ampere range for medium-power transmitters.

The tank coils may be mounted so that very little metal of the normal rack structure is in the immediate fields of the tank coils — a condition almost impossible to approach in the usual form of construction with metal panels and side brackets. Plug-in coils are made much more accessible for changing and the direction of "pull" in removing coils is out-

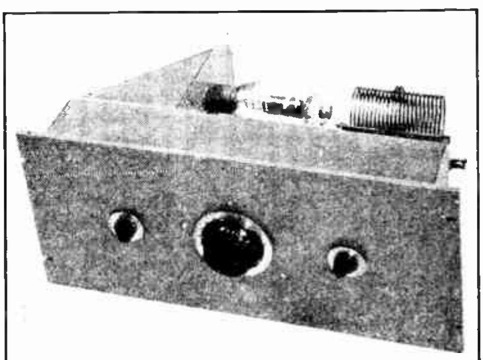


Fig. 1363 — The three controls of the 450-watt "dish-type" amplifier are arranged symmetrically. The meter switch is at the right, the control for the plate tank condenser at the center and the grid-circuit control at the left. The panel which is $8\frac{3}{4} \times 19$ inches is fitted with panel bearings for the condenser-shaft extensions. It is fastened to the chassis by flat-head screws after the bottom edges of the chassis have been drilled and tapped.

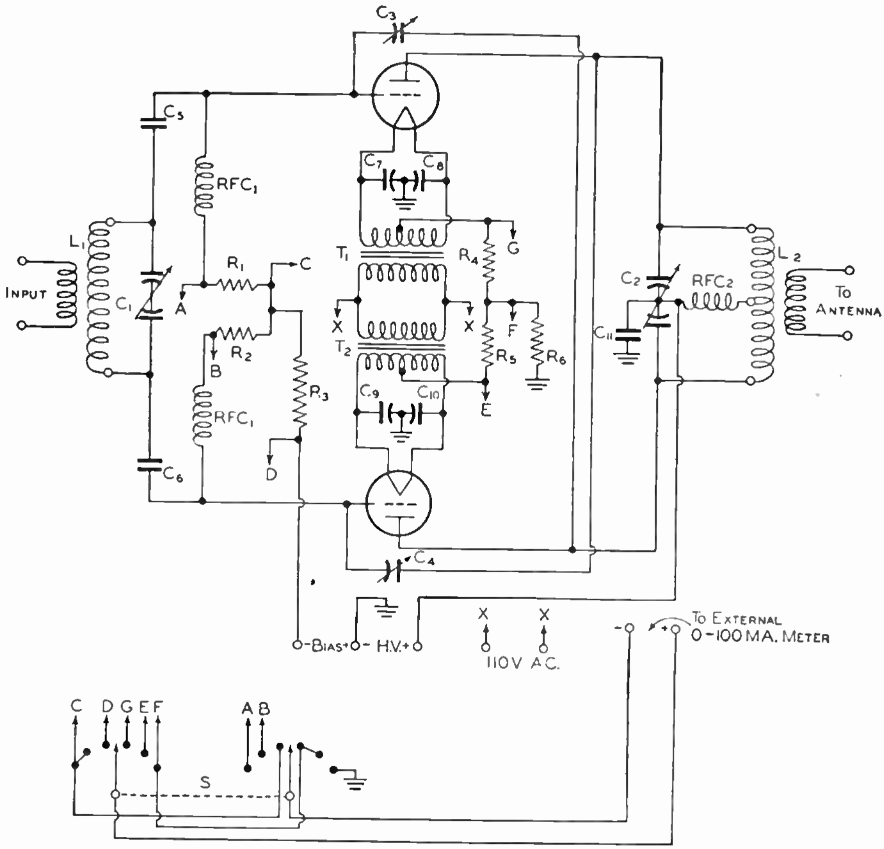


Fig. 1364 — Circuit diagram of the "dish-type" push-pull 150-watt amplifier.

C_1 — 100 $\mu\mu$fd. per section (Hammarlund MCD100M).
C_2 — 100 $\mu\mu$fd. per section (Cardwell MT100GD), 0.07-inch spacing.
C_3, C_4 — Neutralizing condenser, 10 to 15 $\mu\mu$fd. (Hammarlund N10).
C_5, C_6 — 470-$\mu\mu$fd. 600-volt mica.
C_7, C_8, C_9, C_{10} — 0.01-μfd. 600-volt paper.
C_{11} — 0.002-μfd. 5000-volt mica.
R_1, R_2, R_3 — 22 to 47 ohms, 2 watts.
R_4, R_5, R_6 — Cathode-current meter shunts (see text).
L_1 — National AR series coils with center link (variable-link type recommended).

Substitute coils may be wound on 1½-inch diam. form as follows:

3.5 Mc. — 44 turns, 2 inches long.
7 Mc. — 22 turns, 2 inches long.
14 Mc. — 10 turns, 1½ inches long.
28 Mc. — 6 turns, 1½ inches long.

L_2 — B & W TL series with center links.

Substitute coils may be wound as follows on 2½-inch diam. forms:

3.5 Mc. — 36 turns, 4 inches long.
7 Mc. — 18 turns, 4 inches long.
14 Mc. — 10 turns, 3 inches long.
28 Mc. — 6 turns, 3 inches long.

RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 1-mh. r.f. choke (National 154-U).
S — 2-gang 6-position rotary switch (Mallory).
T_1, T_2 — 6.3 volts, 6 amp.

ward away from the rack rather than upward into the next rack unit above. Terminals may be mounted so that the wiring between rack units may be made inconspicuous and so that the chances of personal injury from accidental contact with exposed terminals at the rear are greatly reduced. Lastly, this form of construction usually reduces the required height of the unit which is a particular advantage in table racks where vertical space is at a premium.

The circuit of the amplifier shown in the diagram of Fig. 1364 is standard in every way except in the method of metering. By means of the two-gang six-position switch, it is possible to measure the individual grid and cathode currents of each tube as well as total grid or total cathode currents. To accomplish this, two small filament transformers are used, one for each tube, instead of a single large transformer. The meter is switched across shunting resistances in each circuit to simplify switching. In the cathode circuits, the shunting resistors should be carefully adjusted to provide a scale multiplication of ten, giving a full-scale reading of 1000 ma.

In doing the r.f. wiring, care should be taken to keep it as symmetrical as possible. In forming the long wires between the neutralizing condensers and the tank-condenser stators, the lengths should be made identical. The wire connecting to the rear condenser stator should go directly in a straight line, while the one going to the front stator section may be bent to make up for the difference in distance be-

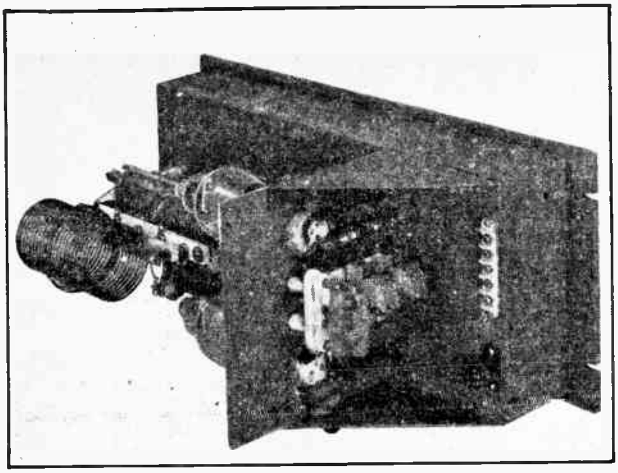


Fig. 1365 — The grid-circuit components of the "dish-type" 450-watt amplifier are mounted on this side of the partition which is braced by standard 5-inch triangular brackets. The tank condenser is mounted by means of a screw in the hole which remains when the shield between the stators is removed. The ceramic terminal strip is for all external connections except for positive high voltage for which a special safety terminal is provided. A large clearance hole should be cut in the chassis for the condenser shaft. The shaft, which should come at the center line of the chassis, should be provided with a flexible insulating coupling.

tween the neutralizing condensers and the two stators. The plate leads to the tubes should be tapped on these long wires at points which will make the wire length between neutralizing condenser and plate and between tank condenser and plate equal on each side.

The positive high-voltage lead, run inside the chassis with high-voltage cable, comes up through a feed-through insulator near the plate choke.

The rotors of the grid tank condenser are not grounded, since experience has shown that an amplifier of this type usually neutralizes more readily without the ground connection and excitation usually divides more evenly between the two tubes.

The leads from the neutralizing condensers to the grid terminals are crossed over before they pass through small feed-through points mounted in the partition. The grid-r.f. chokes are self-supporting between the tube grid terminals and the feed-through points in the chassis which carry the biasing leads inside to the individual grid leaks. Filament wires are run through ⅜-inch holes lined with rubber grommets.

Inside the chassis, the leaks and meter shunting resistances are supported on fiber lug strips. The leads going to the switch should be soldered in place, formed into cables and the other ends connected to the switch on the panel as the last operation before putting the panel in place.

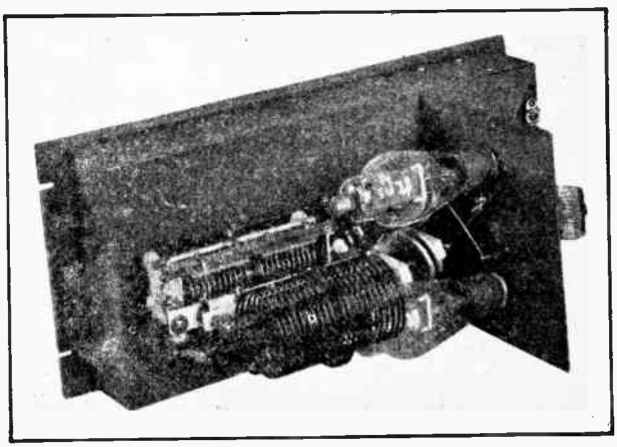
This amplifier is suitable for use with any of the 1000-volt 100-ma. to 1500-volt 150-ma. triodes. Those shown in the photographs are 812s.

For 1500-volt tubes, the power supply shown in Fig. 1360 is suitable for use with this amplifier and bias may be obtained from a unit such as the one shown in Fig. 1327. The bias-supply resistor should be adjusted so that the total grid voltage under operating conditions will not be less than 125 volts without exceeding the maximum grid-current rating of 25 ma. per tube when the amplifier is loaded to rated plate current.

The amplifier requires a driver delivering 25 to 40 watts. Those of Figs. 1311 and 1331 are suitable.

If the layout and wiring have been followed carefully, no difficulties should be encountered

Fig. 1366 — The plate tank-coil jack strip of the 450-watt push-pull amplifier is fastened to the tank-condenser frame with strip-metal brackets. The assembly, mounted on ⅝-inch stand-off insulators is placed at the center of the chassis as far to the left as possible. The condenser shaft is extended at right angles through the bearing in the center of the chassis by means of two Millen 45-degree shaft joints connected together by a short length of bakelite shafting. The sockets for the tubes are submounted on the 6 × 8-inch partition, 3½ inches up from the chassis and 1⅞ inches from each edge and are orientated so that the plates of the tubes will be in a vertical plane.



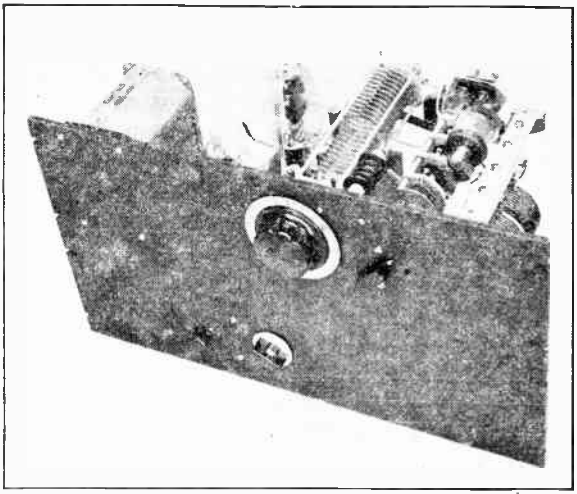


Fig. 1367 — Top view of the band-switching amplifier. The plate-tank switching assembly is to the right.

in neutralizing nor with parasitics. Both grid and plate currents should check the same within ten per cent.

The meter when switched to read grid current forms a good neutralizing indicator. Both neutralizing condensers should be kept at equal settings and adjusted simultaneously until the grid current remains perfectly steady as the plate tank condenser is tuned through resonance. Neutralizing is always done with plate voltage removed.

A suitable antenna tuner will be found in Figs. 1361 and 1362.

A 450-Watt Bandswitching Amplifier

Figs. 1367, 1369 and 1370 show the details of a bandswitching push-pull amplifier for the 3.5-, 7-, 14- and 28-Mc. bands. It is suitable for use with any of the popular 1000- or 1500-volt 100- to 150-ma. triodes. The tubes shown in the photographs are 812s.

As shown in the circuit diagram of Fig. 1368, all of L_1 in the grid tank circuit and all of L_4 in the plate tank circuit are used for 3.5 Mc. Low-frequency padders, C_1 in the grid circuit and C_{10} in the plate, are switched across the coils simultaneously. For 7 Mc., the padding condensers are cut out and L_1 and L_4 are tapped so that only a portion of each coil is in use. At 14 Mc., the coils L_2 and L_3 are used with the padders, while at 28 Mc. the same coils are used without the padders. Links for the two coils in each tank circuit are connected in series.

The components are assembled on a standard 19-inch panel, 10½ inches high. The two tubes, the neutralizing condensers and L_2 are mounted on top of a 5 × 10 × 3-inch chassis fastened to the panel with its center 7 inches from the left-hand edge and its bottom edge ¾ inch above the lower edge of the panel. The

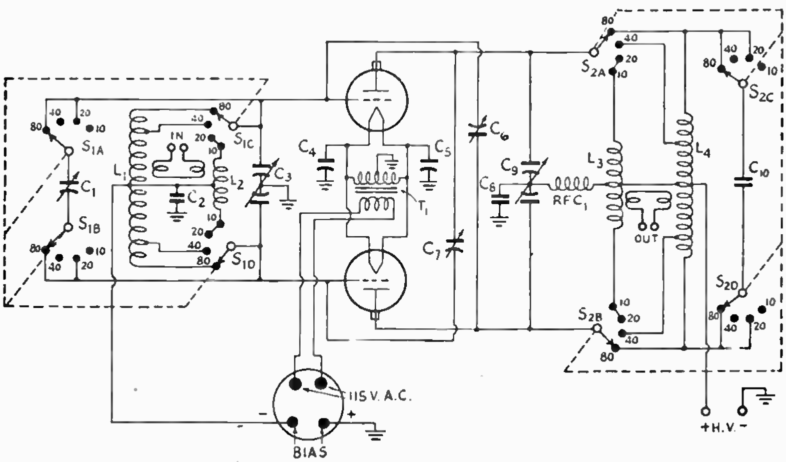


Fig. 1368 — Circuit diagram of the bandswitching push-pull amplifier.

C_1 — 30-μμfd. variable, 0.07-inch spacing (Cardwell ZT-30-AS).
C_2 — 0.001-μfd. mica.
C_3 — 35-μμfd.-per-section variable (Millen 21935).
C_4, C_5 — 0.01-μfd. paper.
C_6, C_7 — Neutralizing condenser (National NC-800).
C_8 — 0.001-μfd. 5000-volt mica.
C_9 — 65-μμfd.-per-section variable (Hammarlund HFBD-65-F).
C_{10} — 50-μμfd. vacuum capacitor (Type GE GL-1L38).
L_1 — B & W 80BCL, tapped at 12th turn from each end.
L_2 — 10 turns No. 14 enameled, 1¼-inch diam., 1 inch long.
L_3 — B & W 10TCL.
L_4 — B & W 80TCL reduced to 24 turns, tapped at 3rd turn from each end.
RFC_1 — 1-mh. r.f. choke (National R-154U)
S_1, S_2 — 4-gang 4-position ceramic rotary switch (Mallory 164-C).
T_1 — 6.3 volts, 8 amp. (UTC S61).

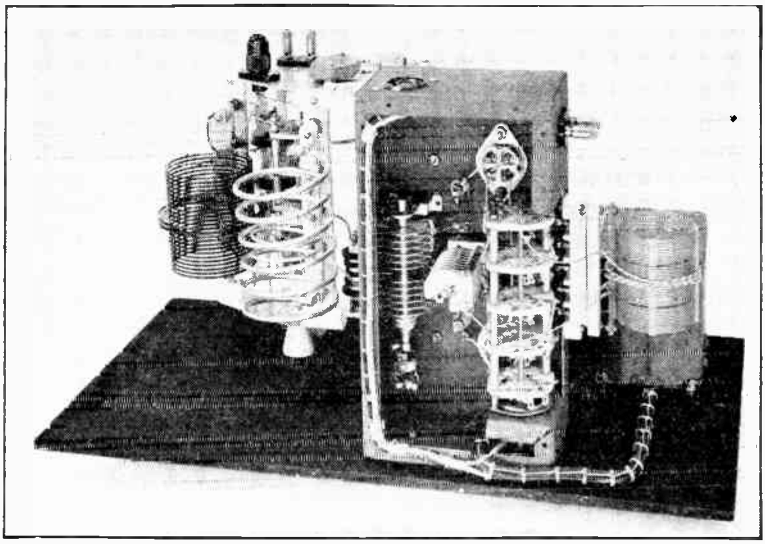


Fig. 1369 — Bottom view of the band-switching amplifier showing the grid-circuit assembly to the right and the plate-circuit group to the left.

tubes are spaced 5¼ inches, center to center, and their sockets are submounted and centered 1¾ inches from the right-hand edge of the chassis as viewed from the rear. L_2 is wound on a polystyrene form mounted on a National AR coil-plug strip. Its socket is centered between the tubes and ⅝ inch from the edge of the chassis. A 5¾ × 2-inch cut-out is made in the outside edge of the chassis to clear the grid bandswitch, S_1. A 1¾-inch piece of the cut-out is left and bent inward at right angles to provide a mounting for the switch. The coil for L_1 is removed from its plug strip and transferred to a Millen plug strip which has the required additional contacts for the 7-Mc. taps. The cut-out is notched at the top to provide clearance for the terminals of the coil socket.

Underneath the chassis, C_9 is mounted vertically on spacers at the center, while the grid tuning condenser, C_3, is mounted as close to the inside edge as possible. Leads between the lower terminals of the neutralizing condensers and the grid terminals of the tube sockets are fed down through the top of the chassis via small feed-through insulators. Link input terminals are mounted on the outside edge of the chassis, near the rear, while 115-volt a.c. and biasing connections are made through a cable socket set in the rear edge. The filament transformer is mounted on the upper right-hand corner of the panel, as

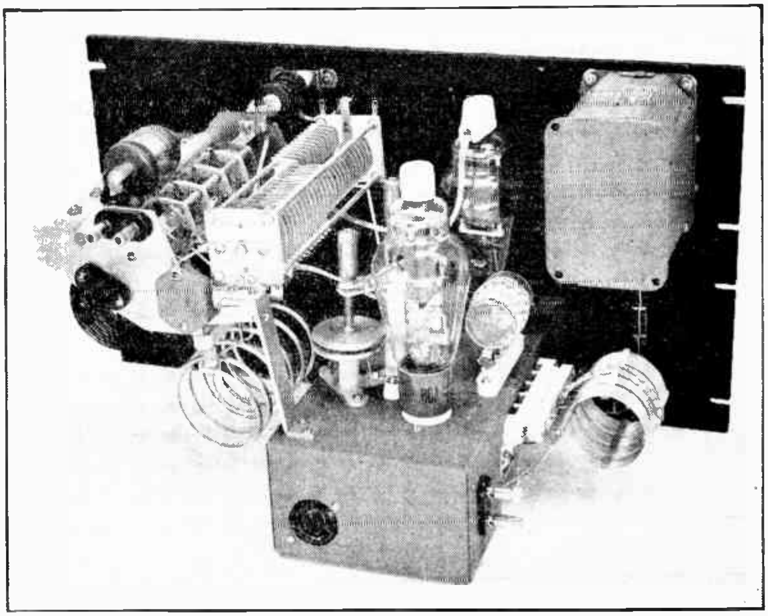


Fig. 1370 — Rear view of the band-switching amplifier.

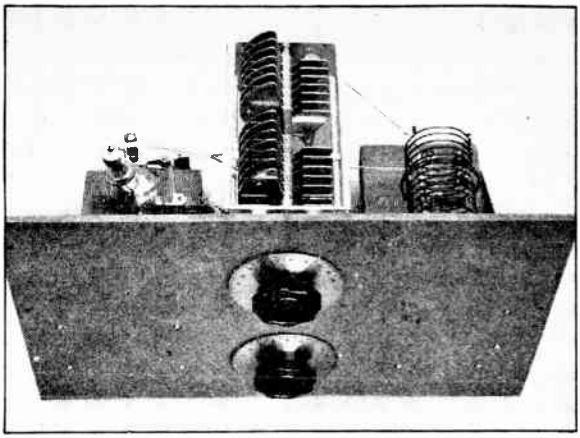


Fig. 1371 — A single-tube high-power amplifier for high-voltage inputs up to 500 watts. The standard rack panel is 12¼ inches high.

viewed from the rear of the unit.

The plate tuning condenser, C_9, is mounted on aluminum-strip brackets fastened to the chassis to bring its shaft 8¾ inches from the right-hand edge of the panel (as viewed from the front) and 2½ inches from the top. Aluminum sheet is cut to form end plates for a subassembly which includes the switch, S_2, the two coil sockets, and a mounting for the padder, C_{10}. As viewed from the rear, L_4 is to the left and L_3 to the right. Pillar-type ceramic insulators form spacers for the mounting angles which support the cartridge-fuse clips in which the vacuum-type padding condenser, C_{10}, is mounted. The assembly is spaced from the panel on 1¼-inch cone stand-offs, placed so that the shaft comes 5½ inches from the right-hand edge of the panel and 2¾ inches below the top edge. The Millen safety terminal for the high-voltage connection, the link output terminals and the insulating condenser, C_8, are fastened to the rear end plate of the assembly. The plate r.f. choke is fastened to the panel between the plate tank condenser and the switch assembly.

For 1500-volt operation the plate supply shown in Fig. 1360 is suitable. The same circuit with the 1000-volt values is appropriate for lower-voltage tubes. If 1500-volt tubes are to be used, the exciter should be capable of delivering 25 to 30 watts. The units shown in Figs. 1311 and 1331 are suggested. The same exciters, with the 807s operated at lower voltage if desired, may be used also to drive the smaller tubes. The antenna coupler of Fig. 1361 is suitable for use with either class of tubes.

A Single-Tube 500-Watt Amplifier

A single-tube amplifier which may be operated at inputs up to 500 watts at voltages as high as 3000 is shown in Figs. 1371, 1372 and 1374. The circuit, shown in Fig. 1373, is strictly

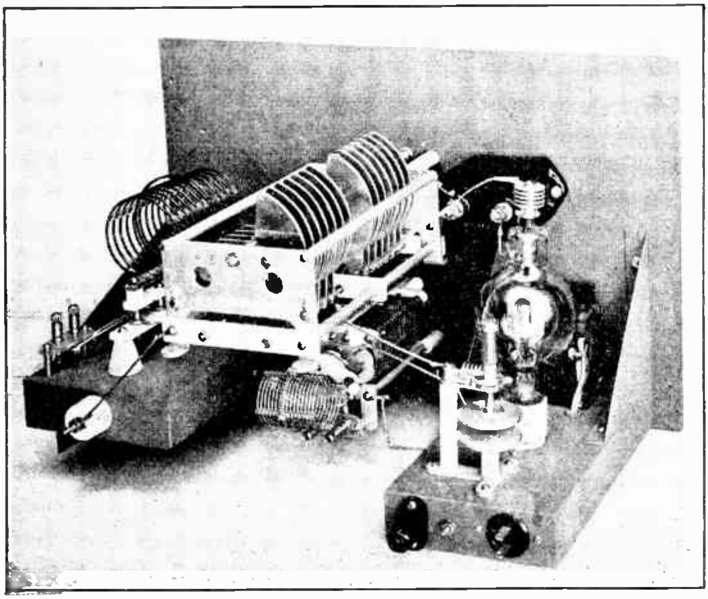


Fig. 1372 — Rear view of the high-power single-tube amplifier. The two tank condensers are mounted, one above the other, in the center of the panel by means of Isolantite pillars from stand-off insulators. Four National Type GS-2 insulators are used to support the plate tuning condenser, while three Type GS-1 insulators are used for the grid tuning condenser. Insulated flexible couplings and panel bearings are used on each shaft to insulate the controls. One of high break-down voltage rating should be used for the plate condenser, and *the panel bearings must be grounded!* The socket for the grid tank coil is mounted, using insulated spacers and a small metal plate as a base, on the rear end plate of C_1. Metal strips, also fastened to the end plate, support the input-link terminal strip. The insulating by-pass condenser, C_4, is mounted just to the right of C_2.

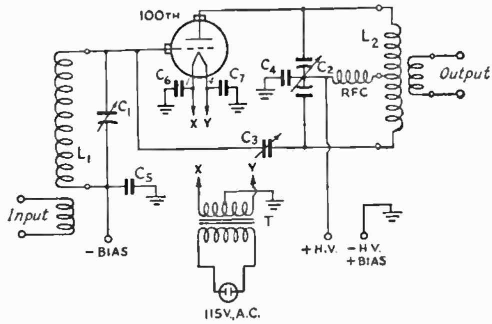


Fig. 1373 — Circuit of the 500-watt input amplifier.

C_1 — 250-μμfd. variable, 0.047-inch spacing (National TMK-250).

C_2 — 100 μμfd. per section, 0.171-inch spacing (National TMA-100-DA).

C_3 — Neutralizing condenser (National NC-800).

C_4 — High-voltage condenser, 0.001-μfd. mica, 12,500-volt rating (Cornell-Dubilier 21A-86).

C_5, C_6, C_7 — 0.01-μfd. mica.

L_1 — 3.5 Mc. — 26 turns No. 16, 1½-inch diam., 2⅛ inches long, 3-turn link (B & W JCL40).
7 Mc. — 16 turns No. 16, 1½-inch diam., 1⅞ inches long, 3-turn link (B & W JCL20).
14 Mc. — 8 turns No. 16, 1½-inch diam., 1⅞ inches long, 3 turn link (B & W JCL10).
28 Mc. — 6 turns No. 16, 1½-inch diam., 1½ inches long, 2-turn link (B & W JCL10, 1 turn removed from each end).

L_2 — 3.5 Mc. — 26 turns No. 12, 3½-inch diam., 4½-inches long, 2-turn link (B & W TCL80).
7 Mc. — 22 turns No. 12, 2½-inch diam., 4½ inches long, 2-turn link (B & W TCL40).
14 Mc. — 12 turns No. 12, 2½-inch diam., 4¼ inches long, 2-turn link (B & W TCL20).
28 Mc. — 6 turns ⅛-inch copper tubing, 2½-inch diam., 4½ inches long, 2-turn link (B & W TCL10).

RFC — 1-mh. r.f. choke, 300 ma. (National R-3001 mounted on GS-1 insulator).

T — Filament transformer, 5 volts, 8 amp. (Thordarson T-19F84).

conventional, with link coupling for both input and output circuits. While a Type 100TH tube is shown in the photographs, almost any other tube of similar physical size and shape which is designed to operate at plate voltages of 3000 or less may be used in the same circuit arrangement.

Power supply and tuning — The plate power supply shown in Fig. 1343 may be used with this unit. Bias may be obtained from the unit shown in Fig. 1327. For this purpose, the VR-75-30 branch may be omitted and a single resistor of 5000 ohms connected across the output of the pack, with the bias lead connected to the extreme negative end of the resistor.

The transmitter shown in Fig. 1331 should provide sufficient excitation.

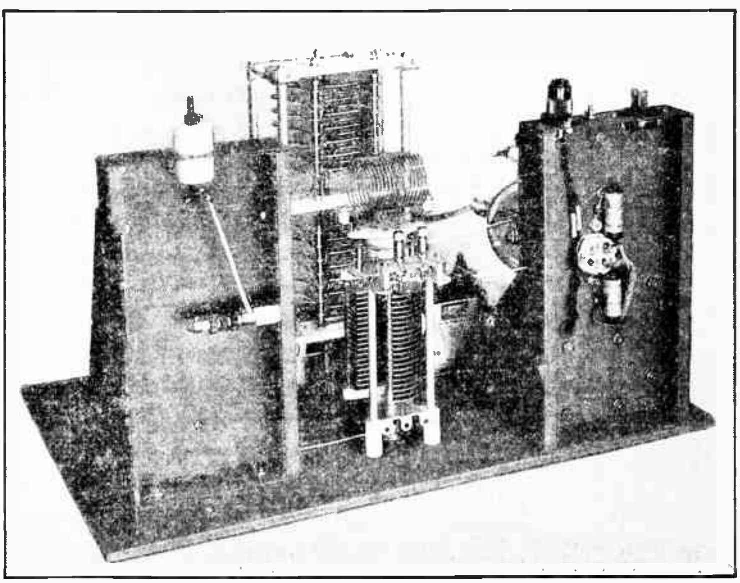
An amplifier operating at high voltage should always, after neutralizing, be tuned up at reduced plate voltage. This may be obtained by connecting a lamp bulb in series with the primary of the plate transformer. Coupling between the exciter and the amplifier should be adjusted so that the grid current does not exceed 40 to 50 ma. with the amplifier tuned and loaded to the rated plate current of 167 ma. Power output of 225 to 300 watts should be obtainable on all bands at plate voltages from 2000 to 3000.

The tube tables in Chapter Twenty should be consulted for data on the operation of other tubes suitable for use in this amplifier.

A 1-Kw. Push-Pull Amplifier

The push-pull amplifier shown in the photographs of Figs. 1375, 1377 and 1378, is built around a pair of Eimac 250TH triodes. It will handle a full kw. input at a plate voltage of 2000 or less, although the plate tank-condenser spacing is sufficient for 3000-volt operation with plate modulation. The driving stage should be capable of delivering approximately 100 watts. The amplifier may be shifted to any

Fig. 1374 — Bottom view of the single-tube 500-watt amplifier. In the lower right-hand corner of the panel is fastened a chassis 9½ × 5 × 1½ inches, on which are mounted, in line, the filament transformer, the tube socket and the neutralizing condenser. A chassis of similar size to the left supports the plate tank coil and the output-link terminals. A large feed-through insulator in the rear edge of this chassis serves as the high-voltage terminal. In wiring the amplifier unit, the importance of well-spaced leads carrying high voltage cannot be stressed too greatly. It must be remembered that the arcing distances and break-down capabilities of voltages as high as 3000 are considerably greater than with the lower plate voltages more commonly used by amateurs.



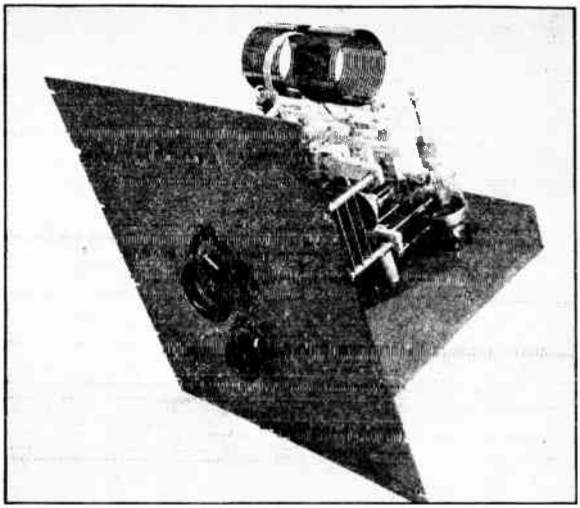


Fig. 1375 — Front view of the kilowatt amplifier. The panel is 21 inches high and of standard 19-inch width.

amateur band by a system of plug-in coils.

The circuit, shown in Fig. 1376, is standard for a push-pull link-coupled neutralized amplifier. The only departure from strict conventionality is the use of the fixed vacuum-type padding condenser (C_9) across the plate tank coil when operating at 3.5 Mc. A filament transformer is included on the chassis to permit short leads which must carry the high heating current.

The components are mounted on a standard 10 × 17 × 3-inch chassis, with the 10-inch side against the panel to provide the necessary depth. The B & W "butterfly"-type plate tank condenser is mounted on heavy 2-inch stand-off insulators, with its shaft along the center line of the chassis, and its front mounting feet centered 2 inches from the panel. Since its rotor is connected to the high-voltage supply, use of a good insulating shaft coupling is of utmost importance as a safety measure. The output tank-coil base assembly, with its adjustable link, is fastened to the two upper-rear stator nuts of the condenser by means of a pair of aluminum angle pieces. Similarly, the clips for the 3.5 Mc. vacuum-type padding condenser are mounted at the front of the condenser. Link output terminals are provided in the form of a pair of large stand-off insulators fastened to the rear of the panel near the top.

The neutralizing condensers are special units designed as accessories to the tank condenser. Each consists of a single disk connected to the grids, the rear stator plates of the plate tank condenser serving as the other side of the neutralizing condenser, for a compact unit.

The by-pass condenser, C_7, is located under the rear end of the tank condenser and is fastened to the chassis with a small metal angle piece which makes the ground connection.

The sockets for the 250THs are submounted. They are spaced 5 inches, center to center, and 4 inches in from the rear edge of the chassis. The grid tank condenser is mounted between the tubes with an extension shaft to the front of the panel. The rotor plates are grounded to the chassis. The high-voltage line to the plate tank condenser and the plate r.f. choke is brought up through the chassis via a large ceramic feed-through insulator.

Underneath, the jack bar for the grid coil is centered between the tube sockets. Connec-

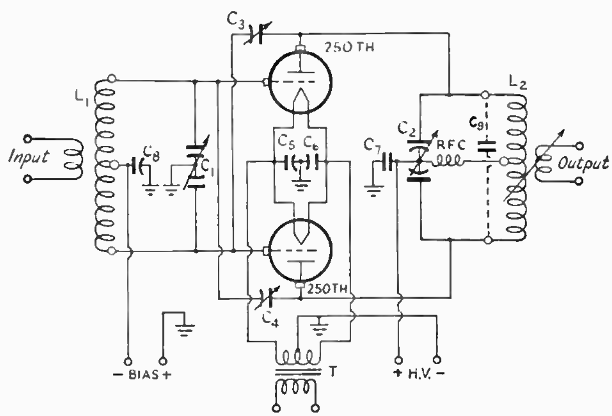


Fig. 1376 — Circuit diagram of the high-power push-pull amplifier.

C_1 — 100 μμfd. per section, 0.05-inch spacing (Hammarlund HFBD-100-C).
C_2 — 60 μμfd. per section, 0.25-inch spacing (B & W CX62-C).
C_3, C_4 — Disk-type neutralizing condenser (B & W N-3).
C_5, C_6, C_8 — 0.01-μfd. paper, 600 volts.
C_7 — 0.001-μfd. mica, 10,000 volts.
C_9 — 25 μμfd., 16,000 volts (GE GL122).
L_1 — B & W BCL coils.
L_2 — B & W HDVL coils.
RFC — 1-mh. r.f. choke (Hammarlund CH-500).
T — 5 volts, 22 amperes (Stancor P6302, see text).

tions between this coil mounting and the condenser on top are made through large clearance holes lined with rubber grommets. Short, direct leads connect the tank circuit to the grid terminals of the tubes.

The filament transformer is mounted directly underneath the plate tank condenser. Since this transformer, as well as the grid coil, protrudes from the underside of the chassis, the chassis is set with its bottom edge 2½ inches above the bottom edge of the panel. The transformer shown in the photographs, and listed under Fig. 1376, is one designed for rectifier service and has high-voltage insulation. If one with 1600- or 2000-volt insulation is available it may be substituted, of course. A Millen safety terminal for the positive high-voltage connection, a three-terminal ceramic strip for bias and ground connections, and a male power plug for the 115-volt connection to the filament transformer are set in the rear edge of the chassis, while a pair of insulated terminals in the right-rear corner are for the excitation input.

Power supply — Fig. 1342 shows the details of a suitable high-voltage plate supply for this amplifier. The biasing unit of Fig. 1327 may be used with an alteration in the voltage-divider resistor in the circuit diagram of Fig. 1326. The total resistance should be 2000 ohms, 100 watts, with the biasing tap taken off at the center. The transmitter of Fig. 1346 will provide adequate excitation.

Adjustment — When the amplifier is completed and ready for operation, the first step in adjustment is the neutralization. This may be done with the amplifier set up with all external connections made, except for the antenna, but with the high voltage turned off.

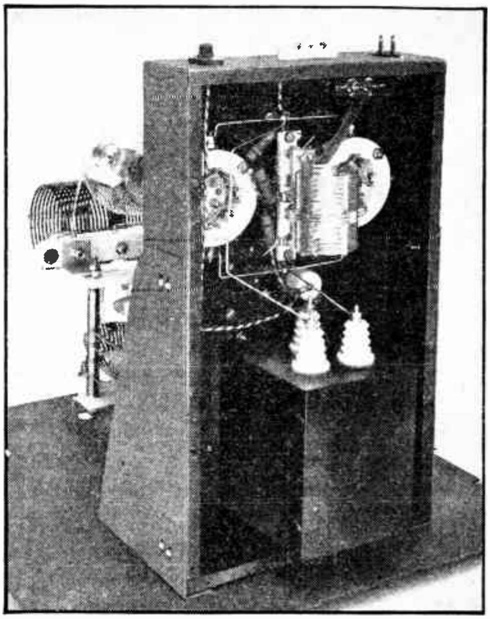


Fig. 1377 — **The filament transformer and grid coil are mounted underneath the chassis.**

With the coils for the desired band plugged in, the tuning of the grid tank circuit should be adjusted until a grid-current reading is obtained. Then the neutralizing condensers should be adjusted simultaneously, bit by bit, keeping the spacing equal. When the amplifier is not neutralized, a dip in grid current will be found as the plate tank condenser is tuned through resonance. The neutralizing condensers should be adjusted until no change in grid current occurs as the plate tank condenser is swung

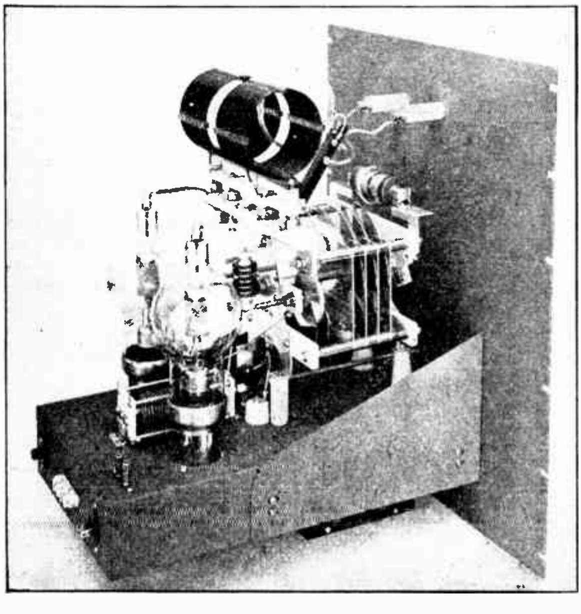


Fig. 1378 — **Rear view of the push-pull 250TH amplifier showing the mounting of the plate tank coil and 3.5-Mc. padding condenser.**

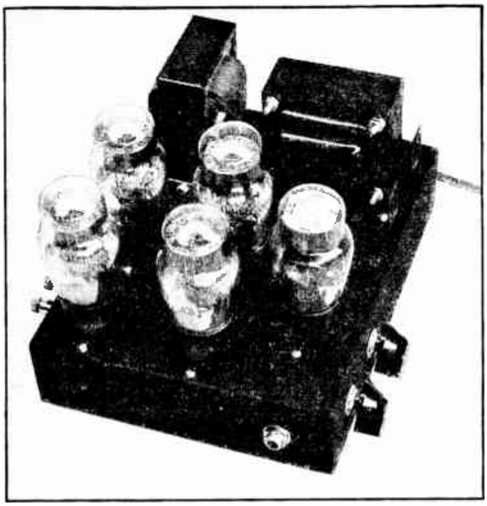


Fig. 1379 — A vacuum-tube keyer, built up on a 7 × 9 × 2-inch chassis with space for four or less keyer tubes and the power-supply rectifier. The resistors and condensers which produce the lag are mounted underneath, controlled by the knobs at the right. The jack is for the key, while terminals at the left are for the keyed circuit.

through its range. This should occur with the adjustable plates of the neutralizing condensers spaced about 1 3/16 inches away from the rear stator plates of the tank condenser.

Although plenty of plate dissipation is available, it is desirable to do the preliminary tuning and loading of the amplifier at reduced plate voltage. Before plate voltage is applied, a grid-current reading of at least 150 to 200 ma. should be possible. The antenna link should be swung out to the minimum-coupling position. As soon as plate voltage and excitation are applied, the plate tank condenser should be adjusted for minimum plate current. Grid current still should be above 150 ma. When the excitation is removed, there should be no indication of oscillation at any setting of the grid- or plate-tank condenser.

The output link may be connected directly to a properly-terminated low-impedance line, or through a link-coupled antenna tuner to the feeders of any antenna system. With excitation and plate power applied, the plate current should increase as the link coupling is tightened and the antenna system tuned to resonance. With each adjustment of coupling or antenna tuning, the plate tank condenser should be retuned for minimum plate current. The minimum reading will increase as the coupling is tightened with the antenna tuned to resonance. The loading may be increased up to the point where the minimum reading is 500 ma., when the input will be 1 kw. at 2000 volts. With the amplifier loaded, the excitation should be adjusted to about 150 ma. for the two tubes.

◖ A Practical Vacuum-Tube Keyer

Fig. 1379 shows a vacuum-tube keyer unit. The diagram is shown in Fig. 1380. T_1, the rectifier, and C_1 and R_1 form the power-supply section for producing the blocking voltage necessary for cutting off the keyer tubes. With only R_2 in the circuit and S_2 in the open position, there will be no lag. As S_2 is turned to introduce more capacitance in the circuit, the keying characteristic is "softened" at both make and break. Adding resistance by turning S_1 to the right affects the "break" only.

As many 45s may be added in parallel as desired. The voltage drop through a single tube varies from 90 volts at 50 ma. to 52 volts at 20 ma. Tubes in parallel will reduce the drop in proportion to the number of tubes. If rated plate voltage is important in the operation of the keyed circuit, the voltage drop through the keyer tubes must be taken into account and the transmitter voltage boosted to compensate for the drop.

If desired, a greater angle of keying lag can be obtained by using a rotary switch with more points and additional resistors and condensers. Suggested values of capacitance in addition to C_2 and C_3, are 0.001 and 0.0022 μfd. From R_2, resistors of 2.2, 3.3 and 4.7 megohms may be added.

When connecting the output terminals of the keyer to the circuit to be keyed, care must be used to connect the grounded output terminal to the negative side of the keyed circuit.

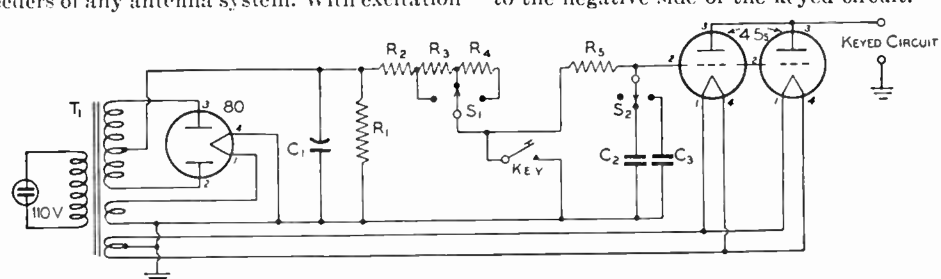


Fig. 1380 — Wiring diagram of the practical vacuum-tube keyer unit and power supply shown in Fig. 1379.

C_1 — 2-μfd. 600-volt paper.
C_2 — 0.0033-μfd. mica.
C_3 — 0.0047-μfd. mica.
R_1 — 0.22 megohm, 1 watt.
R_2 — 50,000 ohms, 10 watts.
R_3, R_4 — 4.7 megohms, 1 watt.
R_5 — 0.47 megohm, 1 watt.
S_1, S_2 — 3-position 1-circuit rotary switch.
T_1 — 325–0–325 volts, 5 volts and 2.5 volts (Thordarson T-13R01).

Rack Construction

Most of the units described in the constructional chapters of this *Handbook* are designed for standard rack mounting. The assembly of a selected group of units to form a complete transmitter is, therefore, a relatively simple matter. While standard metal racks are available on the market, many amateurs prefer to build their own less expensively from wood. With care, an excellent substitute can be made.

The plan of a rack of standard dimensions is shown in Fig. 1381. The rack is constructed entirely of 1 × 2-inch stock of smooth pine, spruce or redwood, with the exception of the trimming strips, *M*, *N*, *O* and *P*. Since the actual size of standard 1 × 2-inch stock runs appreciably below these dimensions, a much sturdier job will result if pieces are obtained cut to the full dimensions.

The main vertical supporting members of the wooden rack each is comprised of two pieces (*A* and *B*, and *I* and *J*) fastened together at right angles. Each pair of these members is fastened together by No. 8 flat-head screws, with heads countersunk.

Before fastening these pairs together, pieces *A* and *J* should be made exactly the same length and drilled in the proper places for the mounting screws, using a No. 30 drill. The length of pieces *A*, *J*, *B* and *I* should equal the total height of all panels required for the transmitter plus *twice* the sum of the thickness and width of the material used. If the dimensions of the stock are exactly 1 × 2 inches, then 6 inches must be added to the sum of the panel heights. An inspection of the top and bottom of the rack in the drawing will reveal the reason for this. The first mounting hole should come at a distance of ¼ inch plus the sum of the thickness and width of the material from either end of pieces *A* and *J*. This distance will be 3¼ inches for stock exactly 1 × 2 inches. The second hole will come 1¼ inches from the first, the third ½ inch from the second, the fourth 1¼ inches from the third and so on, alternating spacings between ½ inch and 1¼ inches (see detail drawing D, Fig. 1381). All holes should be placed ⅜ inch from the inside edges of the vertical members.

The two vertical members are fastened together by cross-member *K* at the top and *L* at the bottom. These should be of such a length that the inside edges of *A* and *J* are exactly 17½ inches apart at all points. This will bring the lines of mounting holes 18¼ inches center to center. Extending back from the bottoms of the vertical members are pieces *G* and *D* connected together by cross-members *L*, *Q* and *E*, forming the base. The length of the pieces *D* and *G* will depend upon space requirements of the largest power-supply unit which will rest upon it. The vertical members are braced against the base by diagonal members *C* and *H*. Rear support for heavy units placed above the

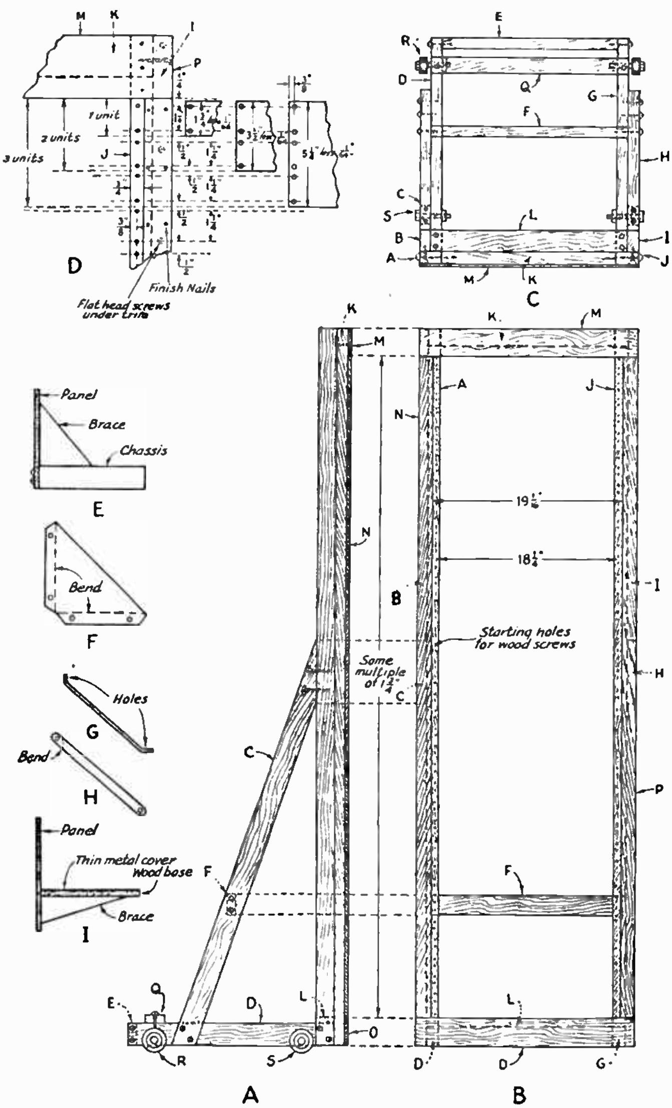


Fig. 1381 — The standard rack. A — Side view. B — Front view. C — Top view. D — Upper right-hand corner detail. E — Panel-and-chassis assembly. F, G, H — Various types of panel brackets. I — Substitute for metal chassis.

base may be provided by mounting angles on *C* and *H* or by connecting these members with cross-braces as shown at *F*.

To finish off the front of the rack pieces of ¼-inch oak strip (*M*, *N*, *O*, *P*) are fastened around the edges with small-head finishing nails. The heads are set below the surface and the holes plugged with putty or plastic wood.

The top and bottom edges of *M* and *O* should be ¼ inch from the first mounting holes, and the distance between the inside edges of the vertical strips, *N* and *P*, 19$\frac{1}{16}$ inches.

To prevent the screw holes from wearing out when panels are changed frequently, ½ × $\frac{1}{16}$ or $\frac{1}{32}$-inch iron or brass strip may be used to back up the vertical members of the frame.

The outside surfaces should be sandpapered thoroughly and given one or two coats of flat black, sandpapering between coats. A finishing surface of two coats of glossy black "Duco" is then applied, again sandpapering between coats. It is very important to allow each coat to dry thoroughly before applying the next, or sandpapering.

Since the combined weights of power supplies, modulator equipment, etc., may total to a surprising figure, the rack should be provided with rollers or wheels so that it may be moved about when necessary after the transmitter has been assembled. Ball-bearing roller-skate wheels are suitable for the purpose.

Standard metal chassis are 17 inches wide. Standard panels are 19 inches wide and multiples of 1¾ inches high. Panel mounting holes start with the first one ¼ inch from the edge of the panel, the second 1¼ inches from the first, the third ½ inch from the second, the fourth 1¼ inches from the third, and the distances between holes from there on alternated between ½ inch and 1¼ inches. (See detail D, Fig. 1381.) In a panel higher than two or three rack units (1¾ inches per unit), it is common practice to drill only sufficient holes to provide a secure mounting. All panel holes should be drilled ⅜ inch in from the edge.

If desired, the rack may be enclosed by completing a framework of one-by-two strip, using ¼-inch plywood for the panels. The panels may be hinged so that three sides are made accessible for servicing. If the transmitter is to be operated in an enclosure, provision should be made for a small amount of forced-air ventilation; otherwise the panels should be open while the transmitter is in operation.

Metering

Various methods of metering are shown in Fig. 1382. A shows the meters placed in the high-voltage plate and bias circuits. MA_1 and MA_2 are for plate current and MA_3 and MA_4 for grid current. When more than one stage operates from the same plate-voltage or bias-voltage supply, each stage may be metered as

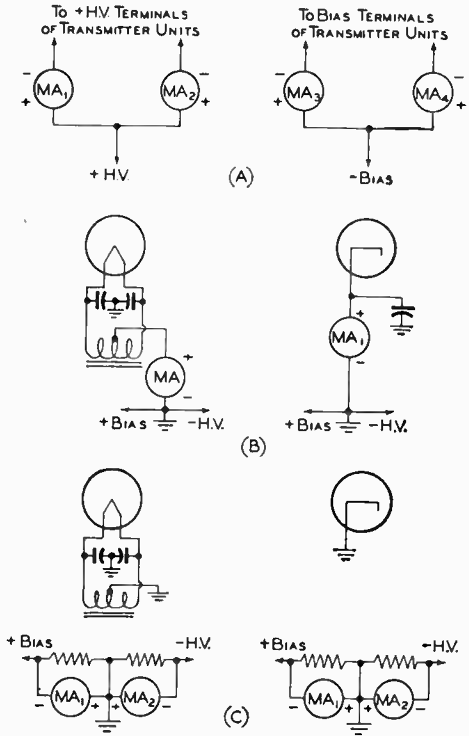


Fig. 1382 — Various methods of connecting milliammeters in grid and plate currents. A — High-voltage metering. B — Cathode metering. C — Shunt metering.

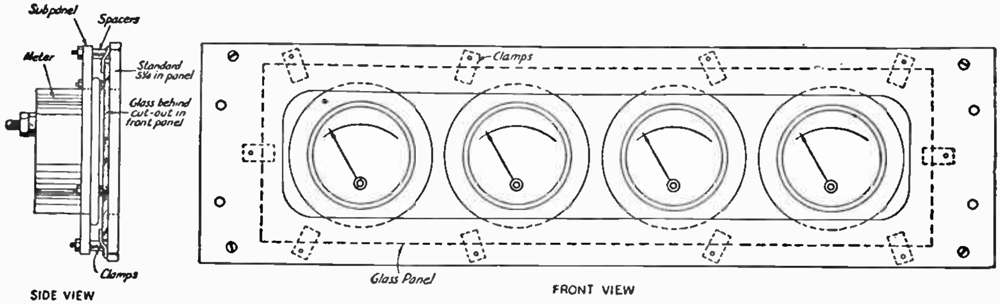


Fig. 1383 — Safety panel for meters. The meters are mounted in the usual manner on an insulating subpanel spaced back of a glass-covered opening in the front panel. The glass is fastened in place with metal clamps or tabs, fastened to the front panel with small screws or pins. The front panel is of standard rack size, 19 × 5¼ inches.

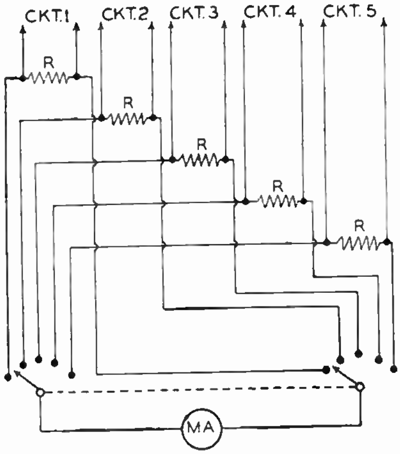


Fig. 1384 — Method of switching a single milliammeter to various circuits with a two-gang switch. The control shaft should be well insulated from the switch contacts, and should be grounded. The resistors, *R*, should have values of resistance ten to twenty times the internal resistance of the meter; 47 ohms will usually be satisfactory.

shown. If this system of metering is used, the meters should be mounted so that the meter dials are not accessible to accidental contact with the adjusting screw. One method of mounting is shown in Fig. 1383, where the meters are mounted behind a glass panel.

When plate milliammeters are to be mounted on metal panels, care must be taken to see that the insulation is sufficient to withstand the plate voltage. Metal-case instruments should not be mounted on a grounded metal panel if the difference in potential between the meter and the panel is to be more than 300 volts; bakelite-case instruments can be used under similar circumstances at voltages up to 1000. At higher voltages than these an insulating panel should be used.

The placing of meters at high-voltage points in the circuit may be overcome by the use of the connections shown in Fig. 1382-B and -C. The disadvantage of the arrangements at B is that the meter reads total cathode current and the grid and plate currents cannot be metered individually. This disadvantage is overcome in C, where the meters are connected across low resistances in the grid- and plate-return circuits. MA_1 reads grid current and MA_2 plate current. The parallel resistors should have a value of not less than 10 to 20 times the resistance of the meter, and should be of sufficient power rating so that there will be no possibility of resistor burn-out. If desired, the resistance values may be adjusted to form a multiplier scale for the meter (see Chapter Nineteen). The same principle is used in the meter-switching system shown in Fig. 1384.

Meters may also be shifted from one stage to another by a plug-and-jack system, but this system should not be used unless it is possible to ground the frame of the jack or unless a suitable guard is provided around the meter jacks to make personal contact with high voltages impossible in normal use of the plug.

Another metering system based upon the use of simple s.p.d.t. toggle switches is shown in the diagram of Fig. 1385. In each case provision is made for metering two circuits with a single milliammeter. Grid returns should be made to filament center-tap or cathode rather than to ground or negative high voltage. If currents included in the meter range are to be measured, the resistors should have a value of about 47 ohms each, otherwise they should be adjusted to give the desired scale multiplication.

Control Circuits

Proper arrangement of controls is important if maximum convenience in operation is to be attained. If the transmitter is to be of fairly high power, it is desirable to provide a special service line leading directly from the public-utility meter board to the operating room. This line should be run in conduit or BX cable, and the conductors should be of ample size to carry the maximum load without undue voltage drop. The line should be terminated with an enclosed entrance switch, properly fused.

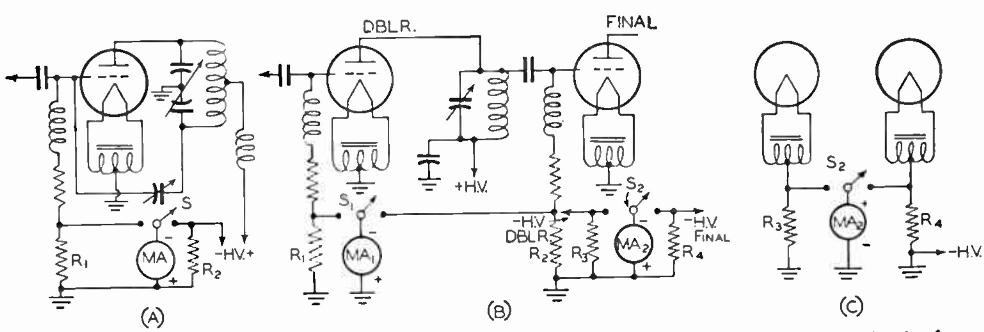


Fig. 1385 — Toggle-switch meter switching. At A is a circuit for switching meter from grid to plate circuit of same stage. At B is a circuit for switching grid meter between two stages and plate meter between two stages. At C is an alternative circuit, similar to the one at B, in which separate filament transformers permit the use of a common plate supply. R_1 and R_2 are grid-circuit meter shunt resistors, while R_3 and R_4 are the plate-circuit shunt resistors.

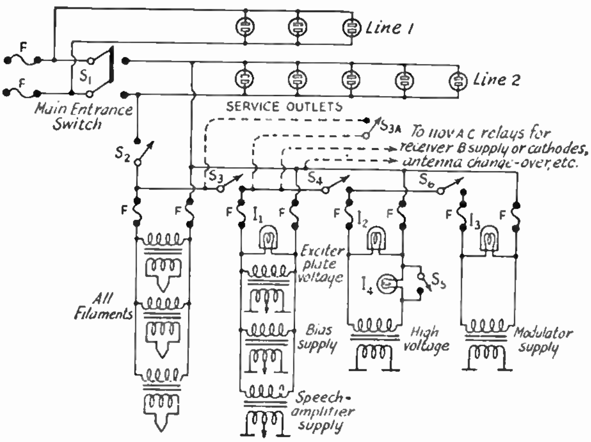


Fig. 1386 — A station control system. No high-voltage supply can be turned on until the filament switch has been closed; the high-power plate supply cannot be turned on until the low-power plate-supply switch has been closed; and modulator power cannot be applied until the final-amplifier plate voltage has been applied. With all switches except S_3 closed, S_3 serves as the main control switch. S_1 — enclosed entrance switch. S_2 — filament switch. S_3 — low plate-voltage and main control switch, preferably of the push-button type which remains closed only so long as pressure is applied. S_4 — high plate-voltage switch. S_5 — low-power and tune-up switch short-circuiting I_4. S_6 — modulator plate-voltage switch. F — fuses. $I_{1\text{-}2\text{-}3}$ — warning lights. I_4 — 100- to 300-watt voltage-reducing lamp.

Fig. 1386 shows the wiring diagram of a simple control system. It will be noticed that, because the control switches are connected in series, none of the high-voltage supplies can be turned on until the filament switch has been closed, and that the high-power plate supply cannot be turned on until the low-power plate-supply switch has been closed. Furthermore, the modulator power cannot be applied until the final-amplifier plate voltage has been applied. S_5 places a 100- to 300-watt lamp, I_4, in series with the primary winding of the high-voltage plate transformer for use during the process of preliminary tuning and for local c.w. work. The final amplifier should first be tuned to resonance at low voltage and S_5 then closed, short-circuiting the lamp. Experience will determine what the low-voltage plate-current reading should be to have it increase to the full-power value when S_5 is closed, so that the proper antenna-coupling and tuning adjustments may be made.

Preferably, S_3 should be of the nonlocking push-button type which remains closed only so long as pressure is applied. A switch of this type provides one of the simplest and most effective means of protection against accidents from high voltage. In the form which is usually considered most convenient, it consists of a switch, located underneath the operating table, which may be operated by pressure of the foot. When used in this manner the operator must be in the operating position, well removed from danger, before high voltage can be applied. If desired, S_{3A} may be wired in parallel on the *front* of the transmitter panel, so that it can be used while tuning the transmitter. S_{3A} also should be of the push-button type.

In more elaborate installations, and in remote-control systems where the transmitter is located some distance from the operating position, similarly arranged switches may be used to control relays whose contacts serve to perform the actual switching at the transmitter.

Two strings of utility outlets, one on each side of the entrance switch, are provided for operation of the receiver and such accessories as the monitor, lights, electric clock, soldering iron, etc. Closing the entrance switch should close those circuits which place the station in readiness for operation. S_2 and S_4 are normally closed and S_3 is normally open. When S_1 is closed upon entering the operating room, the transmitter filaments are turned on as also is the receiver, which should be plugged into line No. 2. With S_4 closed (as well as S_5 and S_6), S_3 performs the job of turning all plate supplies on and off during successive periods of transmission and reception.

All continuously-operating accessories, such as the station clock, should be plugged into line No. 1. This is so that they will not be turned off when S_1 is opened. Line No. 1 is of use also for supplying the soldering iron, lights, etc., when it is desired to remove all voltage from the station apparatus by opening S_1.

Line-Voltage Adjustment

In certain communities trouble is sometimes experienced from fluctuations in line voltage. Usually these fluctuations are caused by a variation in the load on the line and, since most of the variation comes at certain fixed times of the day or night, such as the times when lights are turned on and off for the night, they may be taken care of by the use of a manually-operated compensating device. A

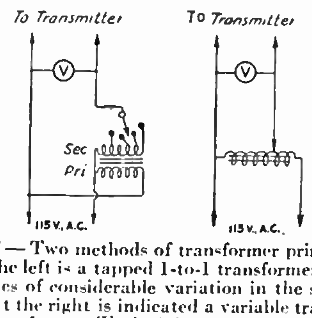


Fig. 1387 — Two methods of transformer primary control. At the left is a tapped 1-to-1 transformer with the possibilities of considerable variation in the secondary output. At the right is indicated a variable transformer or autotransformer (Variac) in series with the transformer primaries.

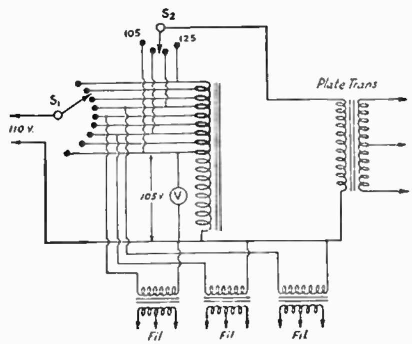


Fig. 1388 — With this circuit, a single adjustment of the tap switch S_1 places the correct primary voltage on all transformers in the transmitter. Information on constructing a suitable autotransformer at negligible cost is contained in the text. The light winding represents the regular primary winding of a revamped transformer, the heavy winding the voltage-adjusting section.

simple arrangement is shown in Fig. 1387. 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.

The secondary is connected in series with the line voltage and, if the phasing of the windings is correct, the voltage applied to the primaries of the transmitter transformers can be brought up to the rated 115 volts by setting the toy-transformer tap switch on the right tap. If the phasing of the two windings of the toy transformer happens to be reversed, the voltage will be reduced instead of increased. This connection may be used in cases where the line voltage may be above 115 volts. This method is preferable to using a resistor in the primary of a power transformer since it does not affect the voltage regulation as seriously.

Another scheme by which the primary voltage of each transformer in the transmitter may be adjusted to deliver the desired secondary voltage, with a master control for compensating for changes in line voltage, is shown in Fig. 1388. This arrangement has the following features:

1) Adjustment of S_1 to make the voltmeter read 105 volts automatically adjusts all primaries to the predetermined correct voltage.

2) The necessity for having all primaries work at the same voltage is eliminated. Thus, 110 volts can be applied to the primary of one transformer, 115 to another, etc.

3) Independent control of the plate transformer is afforded by the tap switch S_2. This permits power-input control and does not require an extra autotransformer.

Grinding Crystals

Crystal blanks, cut to approximate frequency, are available at very reasonable prices. With proper equipment and a little care, these blanks can be ground to the desired frequency. Complete crystal-grinding equipment includes several components. First necessity is a flat piece of plate glass, about 4 inches square or larger. To hold the crystal flat while grinding a flat "button" (shown in Fig. 1389), also of plate glass, either round or square, and slightly larger than the crystal, is required. Both pieces may be obtained at glass stores. Two grades of abrasive, No. 303 emery for surface grinding and No. 600 Carborundum for edge grinding and beveling are obtainable from hardware stores or optician's supply houses. A small paint brush is handy for moistening the abrasive and spreading it around the lapping plate. To facilitate frequent checking of frequency during the grinding process, the quick-change holder shown in Fig. 1390 is desirable. It consists of an FT243 holder with a sliding cover fashioned from sheet metal. Soap, warm water and a toothbrush are used to clean and rinse the crystal. Lintless cloth from an optician's or a clean towel can be used for drying.

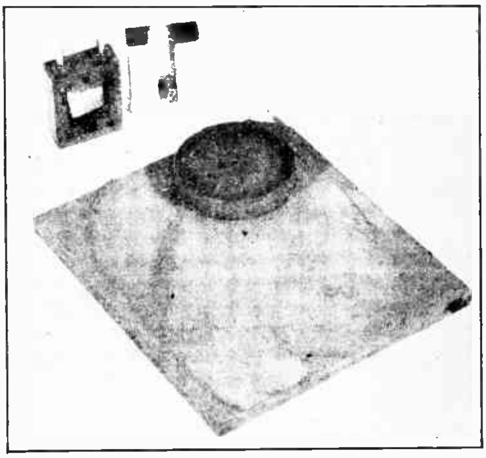


Fig. 1389 — The equipment necessary for grinding a crystal blank to frequency. A piece of plate glass and a "button" of the same material are essential. The "quick-change" adaptation for the crystal holder is a convenience. Not shown, but also convenient, are a small paint brush for spreading abrasive and a toothbrush for scrubbing.

Present-day electrodes have raised lands on each corner, as shown in Fig. 1391, and the crystal should lie at least halfway across these lands and should not be larger than the electrode. The electrodes should be cleaned as carefully as the crystal. Before final assembly both crystal and electrodes should be handled carefully by the corners or edges after their last good scrubbing.

How to grind — The actual grinding is done as follows: Spread the 303 abrasive over an area about a half inch square on the lapping plate, wet the brush, mix water into the spot and spread the abrasive over the lapping plate. Always keep the abrasive moist. Take the button and put a drop of water at its center, and press the dry crystal blank over the drop of water. There should be just enough water in

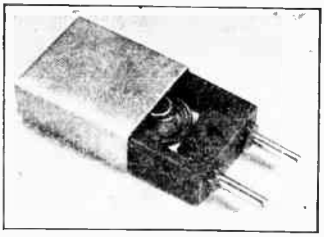


Fig. 1390 — The quick-change crystal holder with sliding cover.

the drop so that it squeezes out under the edges of the blank, where it is wiped away. Place the button, blank down, on the emery and put the index finger in the center of the button. Use just enough pressure to move the button in a figure-8 pattern. This motion is used because it helps keep the blank flat.

After grinding through ten or fifteen "8s" the blank should be rechecked for frequency and activity. The blank's activity is a term used in crystal making to describe how strongly a crystal will oscillate. This may be indicated by the magnitude of the dip in the plate current, grid current to the next stage, or rectified grid current in the crystal oscillator. It is nearly impossible to tell how much change in frequency will occur during the grinding of a crystal, because pressure on the button, the amount of abrasive, and the area of the "8" all will vary the frequency. The frequency change probably will be between 200 and 1000 cycles per "8," using a 7-Mc. crystal. The crystal can be moved along faster as the operator becomes more familiar with the technique, but for the beginner frequent checks of activity are in order so that any drop can be corrected.

To grind a crystal successfully the activity must be good when the crystal is brought to the desired frequency. There are several ways to raise the activity. Assuming that, with careful grinding on a flat plate with a flat button, the two faces of the crystal are parallel, the major cause of low activity will be dirt or moisture on the crystal or electrodes. Before checking activity the crystal should be scrubbed carefully with the toothbrush, using warm water and soap. Wipe the crystal clean and be sure that the electrodes are clean and dry. If the activity is still down the next thing is to bevel all eight edges of the crystal. The beveling can be done with either fine or coarse abrasive, but is usually more effective with the coarse. Beveling, incidentally, will also raise the frequency because of the quartz ground off during the process.

Although beveling will usually improve the activity, another method — and probably the simplest — is to change electrodes. The land heights on the electrodes have a critical effect on activity. If the center of the crystal becomes too high and the lands are so low that the center of the crystal touches the center of the electrodes, the crystal will stop oscillating.

The last step — and the most drastic method of raising activity — is to edge-grind adjacent edges. This grinding is best done with coarse abrasive and should be followed by a slight bevel to remove any chips which may remain. By checking the crystal frequently, a drop in activity can be corrected by the above methods. If the crystal is ground too far and goes completely dead, the frequency may be too high when the crystal is again reactivated.

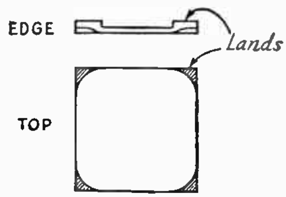


Fig. 1391 — The ½ × ½-inch electrodes used in modern crystal holders, showing the lands at the corners between which the crystal is firmly held.

Building Small Transformers

Power transformers for both filament heating and plate supply for all transmitting and rectifying tubes are available commercially at reasonable prices, but occasionally the amateur wishes to build a transformer for some special purpose or has a core from a burned-out transformer on which he wishes to put new windings.

Most transformers that amateurs build are for use on 110-volt 60-cycle supplies. The number of turns necessary on the 110-volt winding depends on the kind of iron used in the core and on the cross-sectional area of the core. Silicon steel is best, and a flux density of about 50,000 lines per square inch can be used. This is the basis of the table of cross-sections given.

An average value for the number of primary turns to be used is 7.5 turns per volt per square inch of cross-sectional area. This relation may be expressed as follows:

$$\text{No. primary turns} = 7.5\left(\frac{E}{A}\right)$$

where E is the primary voltage and A the number of square inches of cross-sectional area of the core. For 110-volt primary transformers the equation becomes:

$$\text{No. primary turns} = \frac{825}{A}$$

When a small transformer is built to handle a continuous load, the copper wire in the windings should have an area of 1500 circular mils

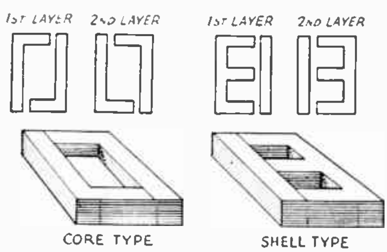


Fig. 1392 — Types of transformer cores and their laminations.

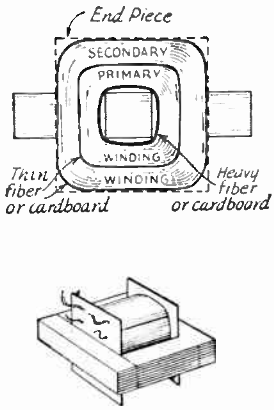


Fig. 1393 — A convenient method of assembling the windings of a shell-type core. Windings can be similarly mounted on core-type cores, in which case the coils are placed on one of the sides. High-voltage core-type transformers sometimes are made with the primary on one core leg and the secondary on the opposite.

for each ampere carried. (See Wire Table in Chapter Twenty.) For intermittent use, 1000 circular mils per ampere is permissible.

The primary wire size is given in the Transformer Design Table; the secondary wire size should be chosen according to the current to be carried, as previously described. The Wire Table in Chapter Twenty shows how many turns of each wire size can be wound into a square inch of window area, assuming that the turns are wound regularly and that no insulation is used between layers. The primary winding of a 200-watt transformer, which has 270 turns of No. 17 wire, would occupy 270/329 or 0.82 square inches if wound with double-cotton-covered wire, for example. This makes no allowance for a layer of insulation between the windings (in general, it is good practice to wind a strip of paper between each layer) so that the winding area allowance should be increased if layer insulation is to be used. The figures also are based on accurate winding such as is done by machines; with hand-winding it is probable that somewhat more area would be required. An increase of 50 per cent should take care of both hand-winding and layer thickness. The area to be taken by the secondary winding should be estimated, as should also the area likely to be occupied by the insulation between the core and windings and between the primary and secondary windings themselves. When the total window area required has been figured — allowing a little extra for contingencies — laminations having the desired leg width and window area should be purchased. It may not be possible to get laminations having exactly the dimensions wanted, in which case the nearest size should be chosen. The cross-section of the core need not be square but can be rectangular in shape so long as the core area is great enough. It is easier to wind coils for a core of square cross-section, however.

Transformer cores are of two types, "core" and "shell." In the core type, the core is simply a hollow rectangle formed from two "L"-shaped laminations, as shown in Fig. 1392. Shell-type laminations are "E"- and "I"-shaped, the transformer windings being placed on the center leg. Since the magnetic path divides between the outer legs of the "E," these legs are each half the width of the center leg. The cross-sectional area of a shell-type core is the cross-sectional area of the center leg. The shell-type core makes a better transformer than the core type, because it tends to prevent leakage of the magnetic flux. The windings are calculated in exactly the same way for both types.

Fig. 1393 shows the method of putting the windings on a shell-type core. The primary is usually wound on the inside — next to the core — on a form made of fiber or several layers of cardboard. This form should be slightly larger than the core leg on which it is to fit so that it will be an easy matter to slip in the laminations after the coils are completed and ready for mounting. The terminals are brought out to the side. After the primary is finished, the secondary is wound over it, several layers of insulating material being put between. If the transformer is for high voltages, the high-voltage winding should be carefully insulated from the primary and core by a few layers of Empire Cloth or tape. A protective covering of heavy cardboard or thin fiber should be put over the outside of the secondary to protect it from damage and to prevent the core from rubbing through the insulation. Square-shaped end pieces of fiber or cardboard usually are provided to protect the sides of the windings and to hold the terminal leads in place. High-voltage terminal leads should be enclosed in Empire Cloth tubing or spaghetti.

After the windings are finished the core should be inserted, one lamination at a time.

TRANSFORMER DESIGN TABLE

Input (Watts)	*Full-load Efficiency*	*Size of Primary Wire*	*No. of Primary Turns*	*Turns Per Volt*	*Cross-Section Through Core*
50	75%	23	528	4.80	1¼″ × 1¼″
75	85%	21	437	3.95	1⅜″ × 1⅜″
100	90%	20	367	3.33	1½″ × 1½″
150	90%	18	313	2.84	1⅝″ × 1⅝″
200	90%	17	270	2.45	1¾″ × 1¾″
250	90%	16	248	2.25	1⅞″ × 1⅞″
300	90%	15	248	2.25	1⅞″ × 1⅞″
400	90%	14	206	1.87	2″ × 2″
500	95%	13	183	1.66	2⅛″ × 2⅛″
750	95%	11	146	1.33	2⅜″ × 2⅜″
1000	95%	10	132	1.20	2½″ × 2½″
1500	95%	9	109	0.99	2¾″ × 2¾″

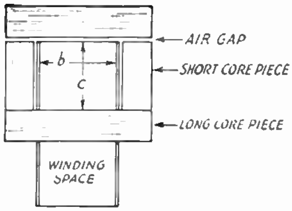


Fig. 1394 — Core arrangement for filter choke coils. The dimensions *b* and *c* refer to the Filter-Choke Design Table.

Fig. 1392 shows the method of building up the core. In the first layer the "E"-shaped laminations are pushed through from one side; the second "E"-shaped lamination is pushed through from the other. The "I"-shaped laminations are used to fill the end spaces. This method of building up the core ensures a good magnetic path of low reluctance. All laminations should be insulated from each other to prevent eddy currents from flowing. If there is iron rust or a scale on the core material, that will serve the purpose very well — otherwise one side of each piece can be coated with thin shellac. It is essential that the joints in the core be well made and be square and even. After the transformer is assembled, the joints can be hammered up tight using a block of wood between the hammer and the core to prevent damaging the laminations. If the winding form does not fit tightly on the core, small wooden wedges may be driven between it and the core to prevent vibration. Transformers built by the amateur can be painted with insulating varnish or waxed to make them rigid and moisture-proof. A mixture of melted beeswax and rosin makes a good impregnating mixture. Melted paraffin should not be used because it has too low a melting point. Double-cotton-covered wire can be coated with shellac as each layer is put on. However, enameled wire should never be treated with shellac as it may dissolve the enamel and hurt the insulation, and it will not dry because the moisture in the shellac will not be absorbed by the insulation. Small transformers can be treated with battery compound after they are wound and assembled. Strips of thin paper between layers of small enameled wire are necessary to keep each layer even and to give added insulation. Thick paper must be avoided since it keeps in the heat generated in the winding so that the temperature may become dangerously high.

Keep watch for shorted turns and layers. If just a single turn should become shorted in the entire winding, the voltage set up in it would cause a heavy current to flow which would burn it up, making the whole transformer useless.

Taps can be taken off as the windings are made if it is desired to have a transformer giving several voltages. Taps should be arranged whenever possible so that they come at the ends of the layers.

After leaving the primary winding connected to the line for several hours it should be only slightly warm. If it draws much current or gets hot there is something wrong. Some short-circuited turns are probably responsible and will continue to cause overheating.

Building Filter Chokes

Filter choke coils may be either of the core or shell type. The laminations should not be interleaved, a butt joint being used instead. An air gap must be provided at some point in the core circuit to prevent magnetic saturation by the d.c. flowing through the winding.

The accompanying table may be used as an approximate guide in winding choke coils. For the same core size, air gap and ampere turns, the inductance will vary approximately as the square of the number of turns. The arrangement of the core is shown in Fig. 1394 and the dimensions *b* and *c* in the table refer to this sketch. The core may be built from straight pieces as shown or with "L"-shaped laminations.

FILTER-CHOKE DESIGN TABLE

L hy.	*Ma.*	*Stack Size In.*	*Core Length*		*Gap In.*	*Winding Form*		*Turns*	*Wire Size*	*Feet*
			Long Piece	*Short Piece*		*b*	*c*			
15	50	½ × ½	½ × 2.2	½ × 0.85	0.035	1	0.68	9500	33	3500
10	100	¾ × ¾	¾ × 2.6	¾ × 0.95	0.03	1	0.67	5000	30	2250
15	100	1 × 1	1 × 3.1	1 × 0.9	0.035	0.96	0.65	4800	30	2550
10	250	2 × 2	2 × 5.2	2 × 1	0.4	1.05	0.68	2000	26	1750
20	250	2 × 2	2 × 5.6	2 × 1.2	0.28	1.43	0.95	4000	26	3820
5	500	2 × 2	2 × 5.5	2 × 1.15	0.17	1.35	0.9	1800	23	1700
10	500	2 × 2	2 × 6.2	2 × 1.5	0.4	2	1.3	3800	23	4100

Chapter Fourteen

Modulation Equipment

In building speech equipment, especially if it is to be used with transmitters operating below 30 megacycles, it should be kept constantly in mind that wide-range audio-frequency response is neither necessary nor desirable. Speech amplifiers should be designed so that the response drops off rapidly above about 3000 cycles; frequencies above this figure are of little help to the receiving operator because the selectivity of the modern superheterodyne is such that they are greatly attenuated when the receiver is tuned to the carrier frequency, but they do cause unnecessary interference to stations working on nearby channels. The speech equipment described in this chapter is adequate for good intelligibility in speech transmission, but is intentionally not designed for "high fidelity." It has been designed to give the required power output as simply and economically as possible while still observing good design principles.

Arrangement of Components

In many respects the arrangement of components is less critical in audio than in r.f. equipment; nevertheless, certain principles must be observed if difficulties are to be avoided. The selection of suitable modulation equipment for any of the transmitters in the preceding chapter is not difficult, if the fundamental principles of modulation described in Chapter Five are understood. If the transmitter is to be plate-modulated and the power input to the modulated stage is to be of the order of 100 watts or higher, a Class B modulator invariably will be selected. A pair of modulator tubes of any type capable of the required power output may be used. The tables in this chapter give the necessary information on the most popular tube types. The driving-power requirements for the modulator stage also are given, so that from this point on the speech-amplifier tube line-up can be selected according to the principles outlined in Chapter Five.

The apparatus to be described is representative of current design practice for speech amplification, with units to provide the various output levels required to drive high- and low-power Class B modulators. In some cases the power output of these amplifier units will be sufficient to modulate low-power transmitters directly, without additional power amplification. Also, practically any of the speech amplifiers shown can be used to grid-modulate transmitters up to the highest power input permitted in amateur transmitters.

Speech-amplifier equipment, especially voltage amplifiers, should be constructed on metal chassis, with all wiring kept below the chassis to take advantage of the shielding afforded. Exposed leads, particularly to the grids of low-level high-gain tubes, are likely to pick up hum from the electrostatic field which 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 audio transformers operating at fairly-high power levels, to prevent magnetic coupling to the grid circuit which might cause hum or audio-frequency feed-back.

If a low-level microphone such as the crystal type is used, the microphone, its connecting cable, and the plug or connector by which it is attached to the speech amplifier, all should be shielded. The microphone and cable usually are constructed with suitable shielding. The cable shield should be connected to the speech-amplifier chassis, and it is advisable — as well as frequently necessary — to connect the

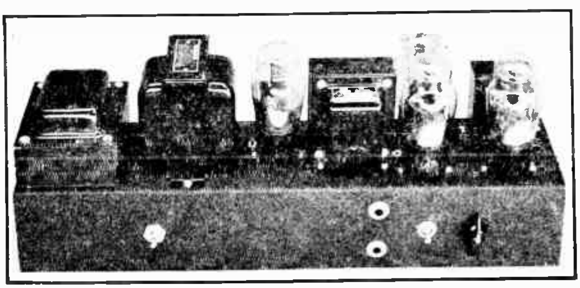


Fig. 1401 — A 10-watt audio unit complete with power supply. Three dual-triode tubes provide a four-stage amplifier with Class B output. Any of the popular types of microphones may be used.

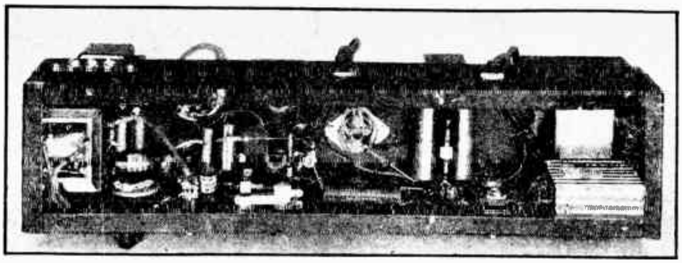


Fig. 1402 — The below-chassis wiring is visible in this view of the 10-watt modulator. The microphone input leads are kept short to reduce hum pick-up.

chassis to a ground such as a water pipe. Heater wiring should be kept as far as possible from grid leads, and either the center-tap or one side of the heater-transformer secondary winding should be connected to the chassis. In a high-gain amplifier the first tube preferably should be of the type having the grid connection brought out to a top cap rather than to a base pin, since in the latter type the grid lead is exposed to the heater leads inside the tube and hence will pick up more hum. With the top-cap tubes, complete shielding of the grid lead and grid cap is a necessity.

A 10-Watt Class-B Modulator for Low-Power Transmitters

A receiving-tube modulator, with a speech amplifier for either crystal or carbon microphones, is shown in Figs. 1401–1403, inclusive. It is suitable for modulating transmitters of 20 watts input or less, such as the low-power equipment frequently used on the very-high frequencies. Type 6N7 or 6A6 tubes are used throughout in the audio circuits. An inexpensive power supply is included, so that the unit is complete and ready for connection to the transmitter.

Fig. 1403 shows the circuit diagram of the speech amplifier-modulator. One section of the first tube is used as the input amplifier for a crystal microphone, the other half being a second speech-amplifier stage. Carbon microphones, which need less gain, are transformer-coupled to the second section of the first 6N7/6A6. The type of jack shown at J_2 in the circuit diagram must be installed if a double-button carbon microphone is to be used. J_2 may be the same as J_1 if a single-button microphone is to be used exclusively.

The gain control is connected in the grid circuit of the second section of the first tube, which is resistance-coupled to the driver. The driver tube has its two sections connected in parallel.

The modulation transformer specified is

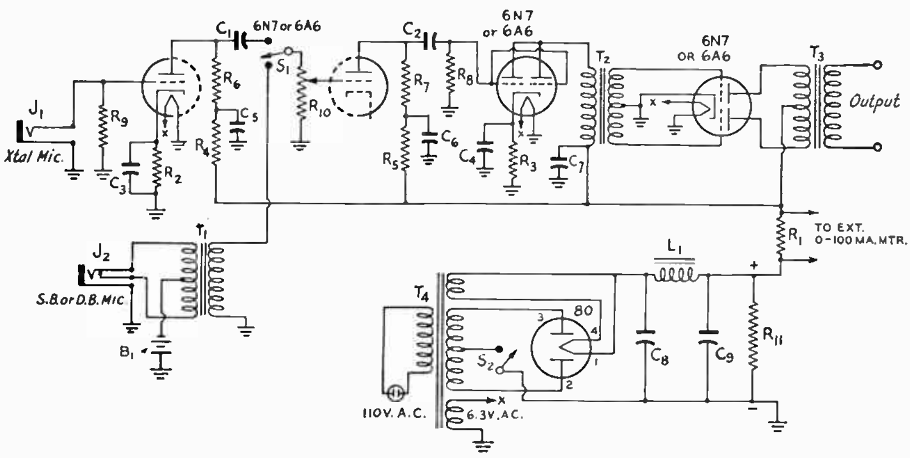


Fig. 1403 — Circuit diagram of the complete 10-watt Class B audio modulator system for low-power transmitters.

C_1, C_2 — 0.1-μfd. 600-volt paper.
C_3, C_4 — 10-μfd. 50-volt electrolytic.
C_5, C_6, C_7, C_8, C_9 — 8-μfd. 450-volt electrolytic.
R_1 — 22 ohms, ½ watt.
R_2, R_3 — 1000 ohms, 1 watt.
R_4, R_5 — 47,000 ohms, ½ watt.
R_6, R_7 — 0.22 megohm, ½ watt.
R_8 — 1 megohm, ½ watt.
R_9 — 4.7 megohms, ½ watt.
R_{10} — 0.5-megohm volume control.
R_{11} — 25,000 ohms, 10 watts.
L_1 — Filter choke, 5 henrys, 200 ma., 80 ohms (Thordarson T-67C49).
B_1 — Microphone battery (see text).
J_1 — Open-circuit jack for crystal microphone.
J_2 — 2- or 3-circuit jack for s.b. or d.b. carbon microphone.
S_1 — S.p.d.t. toggle switch.
S_2 — S.p.s.t. toggle switch (see text).
T_1 — S.b. or d.b. microphone transformer (Stancor A-4351).
T_2 — Driver transformer, parallel 6A6 or 6N7 plates to 6A6/6N7 Class B (Stancor A-4216).
T_3 — Output transformer, 6A6/6N7 Class B to 6500-ohm load (Stancor A-3845).
T_4 — Power transformer, 350-0-350 volts, 90 ma.; 5 volts at 3 amperes; 6.3 volts at 3.5 amperes.

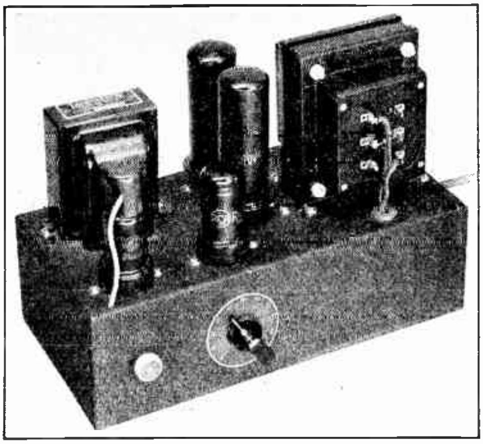


Fig. 1404 — A low-cost speech-amplifier or low-power modulator unit with a maximum audio output of 20 watts. The 6J7 is at the left near corner of the chassis, with the 6J5 to its right, just above the volume control.

designed to work between the plates of a 6N7 or 6A6 and a 6500-ohm load; the impedance ratio used will, of course, depend on the load into which the modulator will work. A milliammeter can be connected across the shunt resistor, R_1, provided to measure the Class B plate current.

The power supply is of the condenser-input type. Using the components specified, it will deliver 350 volts at 90 ma. A switch in the transformer center-tap lead is used for turning the plate voltage on and off without affecting the filament supply.

The power transformer is submounted at the left-hand end of the chassis. Next to it is the filter choke, L_1, followed by the rectifier tube and T_3, the modulation output transformer. The driver tube is at the extreme right-hand end, with T_2, the driver transformer, behind it. The Class B tube is to the rear and in line with the speech-amplifier tube. For convenience in wiring, the audio-tube sockets should be mounted with the filament prongs facing the right-hand end of the chassis.

The plate-voltage switch is on the front of the chassis toward the left in Fig. 1401. The microphone switch, gain control and microphone jacks are grouped at the right. Power input and output terminals are at the rear.

The bottom-view photograph, Fig. 1402, shows the layout for the components mounted below the chassis. T_1 is mounted at the left end. Wiring to the driver-tube socket and the transformer secondary winding should be completed before the transformer is bolted in place, since it is difficult to reach the connecting points with a soldering iron afterward. Short leads between the gain control, the microphone switch and the tube socket can be obtained by making the gain-control contacts face toward the switch, as shown in the photograph.

The compact microphone battery (Burgess Type 3A2) will be held securely in place without brackets or clips if it is wedged in between the bottom of the power transformer and the lips on the bottom of the chassis. A 3-volt battery is sufficient for most carbon microphones, and low current frequently will give better speech quality. The 115-volt a.c. and the meter leads (rubber-covered lamp cord) enter the chassis through rubber grommets. A three-contact terminal strip is located at the right end of the base (left end in the bottom view). One of the contacts on this terminal strip is for an external ground connection and the other two are connected to the modulation-transformer output winding.

The actual measured power output of the unit is 11 watts, as recorded at the point where distortion just begins to be noticeable. This order of audio power output is ample for modulating a low-power transmitter operating with 20 watts or so input to the final stage.

A 20-Watt Speech Amplifier or Modulator

The amplifier shown in Figs. 1404–1406 will deliver audio power outputs up to 20 watts (from the output transformer secondary) with ample gain for ordinary communications-type crystal microphones. Class AB 6L6s are used in the output stage, preceded by a 6J5 and a 6J7 preamplifier.

The unit is built up on a 5 × 10 × 3-inch chassis, with the parts arranged as shown in the photographs. About the only constructional precaution necessary is to use a short lead from the microphone socket (a jack may be used instead of the screw-on type, if desired), and to shield thoroughly the input circuit to the grid of the 6J7. This shielding is necessary to reduce hum. In this amplifier, the 6J7 grid resistor, R_1, is enclosed along with the input jack in a National Type J-1 jack shield, and a shielded lead is run from the jack shield to the grid of the 6J7. A metal slip-on shield covers the grid cap of the tube.

To realize maximum power output, the "B" supply should be capable of delivering about 145 ma. at 360 volts. A condenser-input sup-

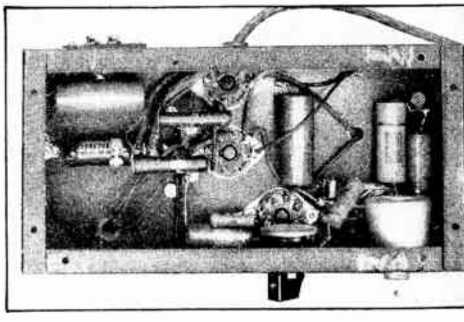


Fig. 1405 — Bottom view of the 20-watt speech amplifier or modulator chassis. The most important constructional point is complete shielding of the microphone input circuit up to the grid of the 6J7 first amplifier.

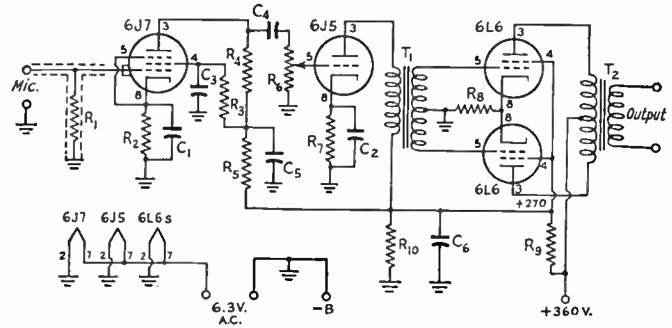


Fig. 1406 — Circuit diagram of the low-cost speech amplifier or modulator capable of power outputs up to 20 watts.

C_1, C_2 — 20-μfd. 50-volt electrolytic.
C_3 — 0.1-μfd. 200-volt paper.
C_4 — 0.01-μfd. 400-volt paper.
C_5, C_6 — 8-μfd. 450-volt electrolytic.
R_1 — 4.7 megohms, ½ watt.
R_2 — 1500 ohms, ½ watt.
R_3 — 1.5 megohms, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_6 — 1-megohm volume control.
R_7 — 1500 ohms, 1 watt.
R_8 — 250 ohms, 10 watts.
R_9 — 2000 ohms, 10 watts.
R_{10} — 20,000 ohms, 25 watts.
T_1 — Interstage audio transformer, single plate to p.p. grids, ratio 3:1 (Thordarson T-57A41).
T_2 — Output transformer, type depending on requirements. A multitap transformer (Thordarson T-19M14) is shown in photos.

ply of ordinary design (Chapter Eight) may be used, since the plate current variation is relatively small. The current is approximately 120 ma. with no input signal and 145 ma. at full output. If an output of 12 or 13 watts will be sufficient, R_9 and R_{10} may be omitted and all tubes fed directly from a "B" supply giving 270 volts at approximately 175 ma.

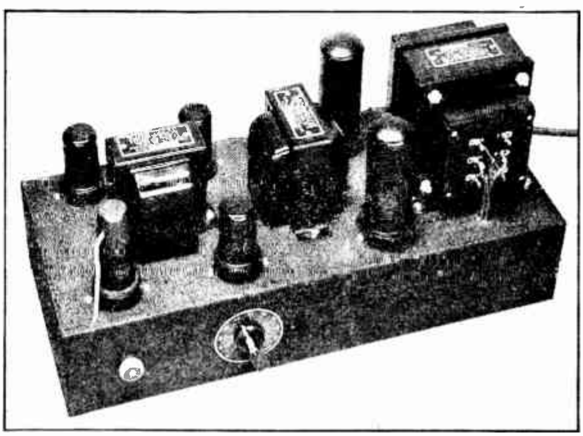
The output transformer shown is a universal modulation type suitable for coupling into the plate circuit of a low-power r.f. amplifier (input 40 watts maximum for 100-per-cent modulation) for plate modulation. For cathode modulation, the r.f. input power that can be modulated can be determined from the data in Chapter Five. The amplifier may also be used for grid-bias modulation with the transformer specified. If the unit is to be used to drive a Class B modulator, it is recommended that the Class B tubes be of the zero-bias type rather than a type requiring fixed bias. A suitable output transformer must be substituted for this purpose; data will be found in catalogues.

◖ A 40-Watt Output Speech Amplifier or Modulator

The 40-watt amplifier shown in Figs 1407–1409 resembles in many respects the 20-watt amplifier just described. The first two stages are, in fact, identical in circuit and construction. To obtain the higher output, however, it is necessary to drive the 6L6s into the grid-current region (Class AB_2 operation), so that a driver stage capable of furnishing sufficient power is required. A pair of transformer-coupled 6J5s in push-pull is used for this purpose, inserted between the single 6J5 stage and the push-pull 6L6s. Decoupling is provided (R_9 and C_5) to prevent motorboating because of the higher over-all gain of the amplifier.

A 6 × 14 × 3-inch chassis is used for the 40-watt amplifier. The photographs show the arrangement of parts. As in the case of the 20-watt unit, complete shielding of the microphone input circuit is essential. The amplifier has ample gain for crystal microphones.

Fig. 1407 — A 40-watt speech amplifier or modulator of inexpensive construction. The 6J7 and first 6J5 are at the front, near the microphone socket and volume control, respectively. T_1 is behind them, and the push-pull 6J5s are at the rear of the chassis behind T_1. T_2, in the center, the push-pull 6L6s, and T_3 follow in order to the right.



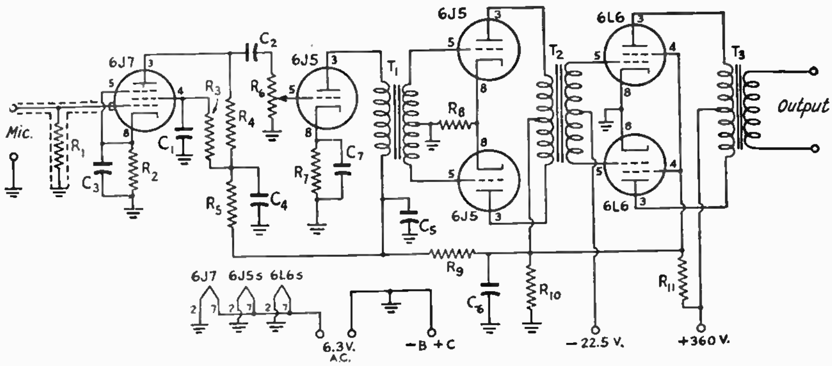


Fig. 1408 — Circuit diagram of the Class AB_2 push-pull 6L6 40-watt output speech amplifier or modulator.

C_1 — 0.1-μfd. 200-volt paper.
C_2 — 0.01-μfd. 400-volt paper.
C_3, C_7 — 20-μfd. 50-volt electrolytic.
C_4, C_5, C_6 — 8-μfd. 450-volt electrolytic.
R_1 — 4.7 megohms, ½ watt.
R_2 — 1500 ohms, ½ watt.
R_3 — 1.5 megohm, ½ watt.
R_4 — 0.22 megohm, ½ watt.
R_5 — 17,000 ohms, ½ watt.
R_6 — 1-megohm volume control.
R_7 — 1500 ohms, 1 watt.
R_8 — 750 ohms, 1 watt.
R_9 — 12,000 ohms, 1 watt.
R_{10} — 20,000 ohms, 25 watts.
R_{11} — 1500 ohms, 10 watts.
T_1 — Inter-stage audio, single plate to p.p. grids, 3:1 ratio (Thordarson T-57A41).
T_2 — Driver transformer, p.p. 6J5s to 6L6s Class AB2 (Thordarson T-84D59).
T_3 — Output transformer, type depending on requirements. A multitap modulation transformer (Thordarson T-19M15) is shown.

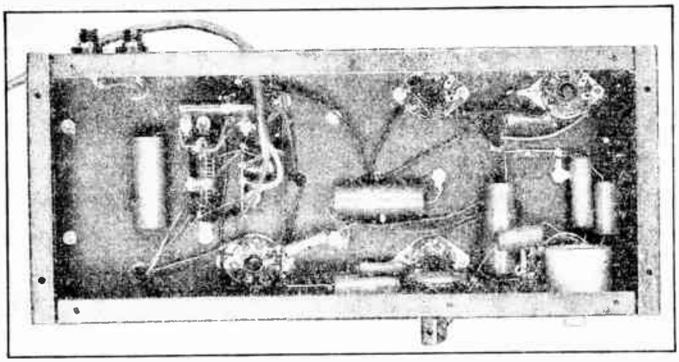


Fig. 1409 — Underneath the chassis of the 40-watt speech amplifier-modulator.

This unit may be used to plate-modulate 80 watts input to an r.f. amplifier. For cathode modulation, the input that can be modulated will depend upon the type of operation chosen, as described in Chapter Five; with 55-per-cent plate efficiency in the r.f. stage, for instance, the input may be of the order of 200 watts, making an allowance for the small amount of audio power taken by the grid circuit.

A high-power Class B modulator can be driven by the unit; data on suitable modulator tubes are given later in this chapter. Zero-bias tubes should be used, because they present a more constant load to the 6L6s than do relatively low amplification-factor tubes which require fixed bias for Class B operation. A suitable Class B driver transformer should be substituted for the universal modulation transformer shown.

The power supply should have good voltage regulation, since the total "B" current varies from approximately 140 ma. with no signal to 265 ma. at full output. A heavy-duty choke-input plate supply should be used; general design data will be found in Chapter Eight. Heater requirements are 6.3 volts at 3 am-

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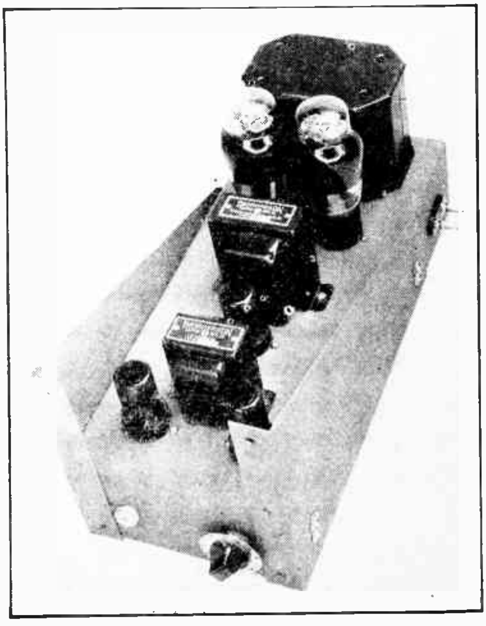

Fig. 1410 — An all-triode speech amplifier with push-pull 6B4G output, for driving a Class B amplifier requiring seven watts or less on the grids. The end-on construction permits mounting another similarly-constructed unit on the same rack panel.

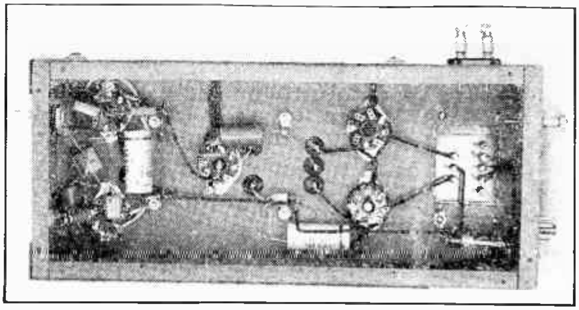

Fig. 1411 — Bottom view of the all-triode speech amplifier. Wiring is simple and the whole unit is easy to construct.

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peres. Bias for the 6L6 stage is most conveniently supplied by a 22.5-volt "B"-battery block; a small-sized unit will be satisfactory, since no current is drawn.

An all-triode speech amplifier — Triodes are preferable to tetrodes as drivers for Class B modulators because their lower plate resistance means better output-voltage regulation and hence less distortion under the varying load presented by the Class B grids. Where an output of 10 watts or less is needed to drive a Class B amplifier, low-μ triodes such as the 2A3, 6A3, and 6B4G can be used. The amplifier shown in Figs. 1410 and 1411 uses a pair of 6B4Gs in push-pull, driven by a three-stage triode amplifier which provides ample gain for communications-type crystal microphones.

The circuit diagram is given in Fig. 1412. The first stage, a 6SF5, is resistance-coupled to a 6J5, which in turn is impedance-coupled to a second 6J5. The latter tube is transformer-coupled to the 6B4G grids. The combination of impedance- and transformer-coupling keeps the stage gain high and restricts the frequency response to the range most useful for voice communication. The volume control is in the grid of the second stage. The circuit is quite straightforward throughout. Bias for the 6B4Gs is obtained from the drop in a resistor (R_{10}) in series with the filament-supply center-tap.

The amplifier is built on a 6 × 14 × 3-inch chassis, arranged for end-mounting from a rack panel. This type of construction uses very little panel space and permits mounting another unit such as a power supply or modulator on the same panel. In Fig. 1410 the tube at the left front, just above the microphone jack, is the 6SF5. The first 6J5 is at the right, with the gain control on the chassis wall below it. The coupling choke, L_1, is behind and between the first two tubes, and is followed, going toward the rear, by the second 6J5, the push-pull coupling transformer, the 6B4Gs, and the output transformer. The wiring underneath the chassis is shown in Fig. 1411.

The type of output transformer to use will depend upon the grid-to-grid impedance of the Class B tubes to be driven, and should have the proper turns ratio to work between that impedance and the 5000 ohms plate-to-plate required for optimum operation of the 6B4Gs. The measured output from the transformer secondary is 7 watts. Power requirements of the amplifier are 3 amperes at 6.3 volts and 100 ma. at 300 volts.

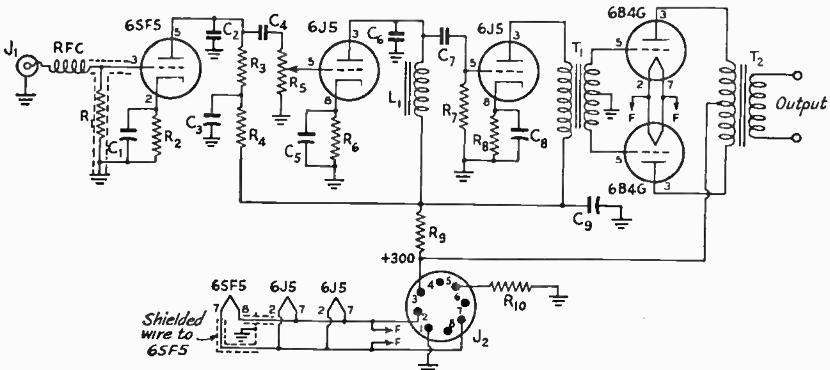


Fig. 1412 — Circuit diagram of the all-triode speech amplifier.

C_1, C_5, C_8 — 10-μfd. 50-volt electrolytic.
C_2 — 470-$\mu\mu$fd. mica.
C_3, C_9 — 8-μfd. 450-volt electrolytic.
C_4, C_6, C_7 — 0.0047-μfd. mica.
R_1 — 2.2 megohms, ½ watt.
R_2 — 4700 ohms, ⅓ watt.
R_3 — 0.47 megohm, ½ watt.
R_4 — 47,000 ohms, ½ watt.
R_5 — 1-megohm potentiometer.
R_6 — 2200 ohms, ½ watt.
R_7 — 0.22 megohm, ½ watt.
R_8 — 1500 ohms, ½ watt.
R_9 — 10,000 ohms, 2 watts.
R_{10} — 1000 ohms, 10 watts.
L_1 — 300 henrys, 5 ma., 6470-ohms d.c. resistance (Thordarson T-37C36).
J_1 — Microphone connector (Amphenol 75-PC1M).
J_2 — Octal socket, male (Amphenol 86-CP8).
RFC — 2.5-mh. r.f. choke.
T_1 — Interstage transformer, single plate to p.p. grids, 3:1 ratio (Thordarson T-57A41).
T_2 — Variable-ratio driver transformer (UTC PA-53AX).

TABLE 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]
6C6, 6J7, 6W7, 7C7, 57 (pentode)	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	350
		2.0	2.95	2300	0.04	4.0	0.0025	100	240
6C8G (one triode unit)	0.1	0.1	—	2120	—	3.93	0.037	55	22
		0.25	—	2840	—	2.01	0.013	73	23
		0.5	—	3250	—	1.79	0.007	80	25
	0.25	0.25	—	4750	—	1.29	0.013	64	25
		0.5	—	6100	—	0.96	0.0065	80	26
		1.0	—	7100	—	0.77	0.004	90	27
	0.5	0.5	—	9000	—	0.67	0.007	67	27
		1.0	—	11,500	—	0.48	0.004	83	27
		2.0	—	14,500	—	0.37	0.002	96	28
6F5, 6SF5, 7B4	0.1	0.1	—	1300	—	5.0	0.025	33	42
		0.25	—	1600	—	3.7	0.01	43	49
		0.5	—	1700	—	3.2	0.006	48	52
	0.25	0.25	—	2600	—	2.5	0.01	41	56
		0.5	—	3200	—	2.1	0.007	54	63
		1.0	—	3500	—	2.0	0.004	63	67
	0.5	0.5	—	4500	—	1.5	0.006	50	65
		1.0	—	5400	—	1.2	0.004	62	70
		2.0	—	6100	—	0.93	0.002	70	70
6F8G (one triode), 6J5, 6J5G, 7A4, 7N7, 6SN7G (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
6R7, 6R7G, 7E6	0.05	0.05	—	1600	—	2.6	0.055	50	9
		0.1	—	2000	—	2.0	0.03	62	9
		0.25	—	2400	—	1.6	0.015	71	10
	0.1	0.1	—	2900	—	1.4	0.03	52	10
		0.25	—	3800	—	1.1	0.015	68	10
		0.5	—	4400	—	1.0	0.007	71	10
	0.25	0.25	—	6300	—	0.7	0.015	54	10
		0.5	—	8400	—	0.5	0.007	62	11
		1.0	—	10,600	—	0.44	0.004	74	11
6SC7 (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
6SJ7	0.1	0.1	0.35	500	0.10	11.6	0.019	72	67
		0.25	0.37	530	0.09	10.9	0.016	96	98
		0.5	0.47	590	0.09	9.9	0.007	101	104
	0.25	0.25	0.89	850	0.07	8.5	0.011	79	139
		0.5	1.10	860	0.06	7.4	0.004	88	167
		1.0	1.18	910	0.06	6.9	0.003	98	185
	0.5	0.5	2.0	1300	0.06	6.0	0.004	64	200
		1.0	2.2	1410	0.05	5.8	0.002	79	238
		2.0	2.5	1530	0.04	5.2	0.0015	89	263
6SQ7, 6B6G, 7B6, 2A6, 75	0.1	0.1	—	1900	—	4.0	0.03	31	31
		0.25	—	2200	—	3.5	0.015	41	39
		0.5	—	2300	—	3.0	0.007	45	42
	0.25	0.25	—	3300	—	2.7	0.015	42	48
		0.5	—	3900	—	2.0	0.007	51	53
		1.0	—	4200	—	1.8	0.004	60	56
	0.5	0.5	—	5300	—	1.6	0.007	47	58
		1.0	—	6100	—	1.3	0.004	62	60
		2.0	—	7000	—	1.2	0.002	67	63

[1] Voltage across next-stage grid resistor at grid-current point.
[2] At 5 volts r.m.s. output.

TABLE II—CLASS-B MODULATOR DATA

Class-B Tubes (2)	Fil. Volts	Plate Volts	Grid Volts App.	Peak A.F. Grid-to-Grid Voltage	Zero-Sig.[1] Plate Current Ma.	Max.-Sig.[1] Plate Current Ma.[2]	Load Res. Plate-to-Plate Ohms	Max.-Sig. Driving Power Watts[3]	Max.-Sig.[1] Power Output Watts[3]
HY65[5]	6.3	450	—	—	—	125	—	0.4	34
HY31Z[4,5]	6.3	300	0	104	20	100	5,000	1.4	18
815[6]	6.3	400 500	−15 −15	60 60	22 20	150 150	8,000 6,200	0.36 0.36	42 54
HY6L6GX[7]	6.3	400 500	−25 −25	80 80	100 100	230 230	3,800 4,550	0.35 0.6	60 75
TZ20	7.5	800	0	160	40	136	12,000	1.8	70
HY61/807[7]	6.3	400	−25	80	100	230	3,800	0.35	60
HY69[5,7]	6.3	300	−25	106	60	150	4,000	0.25	30
HY30Z	6.3	600 750 850	0 0 0	171 167 171	18 22 28	180 180 180	6,000 8,000 10,000	Note 9 " "	75 95 110
807[10]	6.3	400	−25	78	100	240	3,200	0.2	55
HK24	6.3	1000 1250	−29 −42	248 256	30 24	150 136	15,000 21,200	4.5 4.2	105 120
809	6.3	500 750 1000	0 − 4.5 −10	135 140 156	40 40 40	200 200 200	5,200 8,400 11,600	2.4 2.4 3.4	60 100 145
HY40Z	7.5	750 850 1000	0 0 0	171 185 185	32 40 45	225 250 250	6,000 7,000 9,000	Note 9 " "	110 155 185
811	6.3	1250 1500	0 − 9	140 160	48 20	200 200	15,000 18,000	3.8 4.2	175 225
35T	5.0 to 5.1	1000 1250 1500	−22 −30 −40	— — —	— — —	— — —	7,200 9,600 12,800	— — —	150 200 230
TZ40	7.5	1000 1250 1500	0 − 4.5 − 9	220 269 265	— — —	280 280 250	7,350 10,000 12,000	5.5 6.0 6.0	175 225 250
203-A	10	1000 1250	−35 −45	310 330	26 26	320 320	6,900 9,000	10 11	200 260
211	10	1000 1250	−77 −100	380 410	20 20	320 320	6,900 9,000	7.5 8.0	200 260
838	10	1000 1250	0 0	200 200	106 148	320 320	6,900 9,000	7.0 7.5	200 260
HK54	5.0	1500 2000 2500	−45 −70 −85	300 360 360	40 24 20	198 180 150	16,800 36,000 40,000	5.0 6.0 5.0	200 260 275
HY51Z	7.5	850 1000 1250	0 0 0	148 170 155	48 60 90	300 350 300	5,000 6,000 10,000	Note 9 " "	160 260 285
203-Z	10	1000 1250	0 − 4.5	206 215	50 60	350 350	6,200 8,000	6.5 6.75	230 300
ZB120	10	1000 1250 1500	0 0 − 9	190 180 196	70 95 60	310 300 296	6,900 9,000 11,200	5.0 4.0 5.0	200 245 300
8005	10	1250 1500	−55 −80	290 310	40 40	320 310	8,000 2,500	4.0 4.0	250 300
HF100	10 to 11	1500 1750	−52 −62	264 324	50 40	270 270	12,000 16,000	2.0 9.0	260 350
805 RK57	10	1250 1500	0 −16	235 280	148 84	400 400	6,700 8,200	6.0 7.0	300 370
75T	5.0	1000 1500 2000	— — —	— — —	— — —	— — —	6,800 10,000 12,500	— — —	200 300 400
100TH	5.0 to 5.1	2000 2500 3000	Bias adjusted for maximum rated plate dissipation under no-signal conditions Zero bias up to 1250 v. plate				16,000 22,000 30,000	May be driven by push-pull 6L6s	380 460 500
HD203-A	10	1500 1750	−40 −67	— —	36 36	425 425	8,000 9,000	Note 8	400 500
HK254	5.0	2000 2500 3000	−65 −80 −100	400 420 456	50 50 40	260 248 240	16,000 22,000 30,000	7.0 7.0 7.0	328 418 520
810	10	1500	−30	345	80	500	6,600	12	510

[1] Values are for both tubes.
[2] Sinusoidal signal values; speech values are approximately one-half for tubes biased to approximate cut-off and 80 per cent for zero-bias tubes.
[3] Values do not include transformer losses. Somewhat higher power is required of the driver to supply losses and provide good regulation. Input transformer ratios must be chosen to supply required power at specified grid-to-grid voltage with ample reserve for losses and low distortion levels. Driver stage should have good regulation.
[4] Dual tube. Values are for one tube, both sections.
[5] Instant-heating filament type.
[6] Beam tube. Class AB_2. Screen voltage: 125 at 32 ma.
[7] Beam tube. Class AB_2. Screen voltage: 300.
[8] Can be driven by a pair of 2A3s in push-pull Class AB at 300 volts with fixed bias.
[9] Driver: one or two 45s at 275 volts, self-biased (−55 volts).
[10] Beam Tube. Class AB_2. Screen voltage: 300 at 10 ma. Effective grid circuit resistance should not exceed 500 ohms.

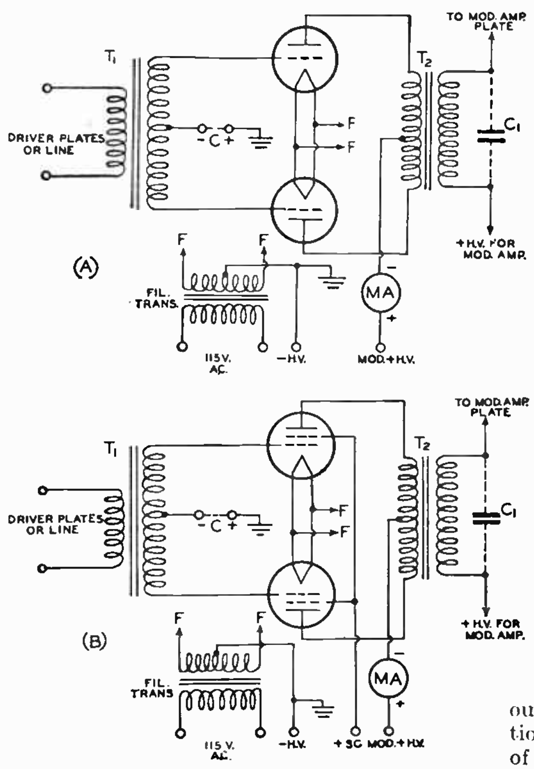


Fig. 1413 — Class B modulator circuit diagrams. Tubes and circuit considerations are discussed in the text.

Class-B Modulators

Class B modulator circuits are practically identical no matter what the power output of the modulator. The diagrams of Fig. 1413 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 cathode should be connected to ground.

Design considerations for Class B stages are discussed in Chapter Five, and data on the performance of various tubes suitable for the purpose are given in the accompanying tables. Once the requisite audio power output has been determined and a pair of tubes capable of giving that output selected, an output transformer should be secured which will permit matching the rated modulator load impedance to the modulating impedance of the r.f. amplifier. Similarly, a driver transformer should be selected which will properly couple the driver stage to the Class B grids.

The plate power supply for the modulator should have good voltage regulation and must be well filtered. 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.

In estimating the output of the modulator, it should be remembered that the figures given in the tables are for the tube output only, and do not include output-transformer losses. The efficiency of the output transformer will vary with its construction, and may be assumed to be in the vicinity of 80 per cent for the less-expensive units and somewhat higher for higher-priced transformers. To be adequate for modulating the transmitter, therefore, the modulator should have a theoretical power capability about 25-per-cent greater than the actual power needed for modulation.

The input transformer, T_1, may couple directly between the driver tube and the modulator grids or may be designed to work from 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

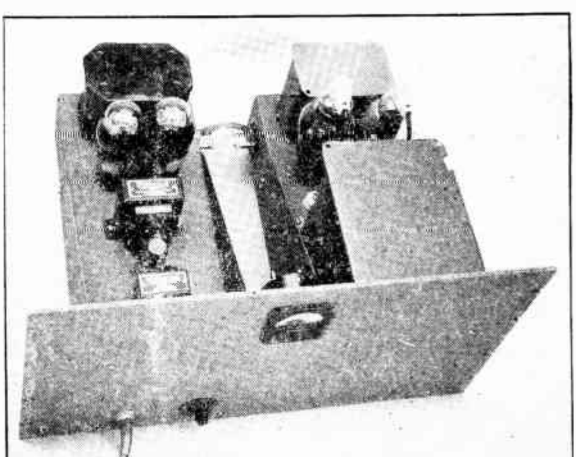


Fig. 1414 — A Class B modulator using 811s or similar tubes (right-hand unit) panel-mounted with its associated speech amplifier. The latter is the all-triode amplifier shown in Fig. 1410. The Class B output transformer is mounted at the panel end of the chassis for good weight distribution. The transformer at the rear is the filament transformer for the Class B tubes.

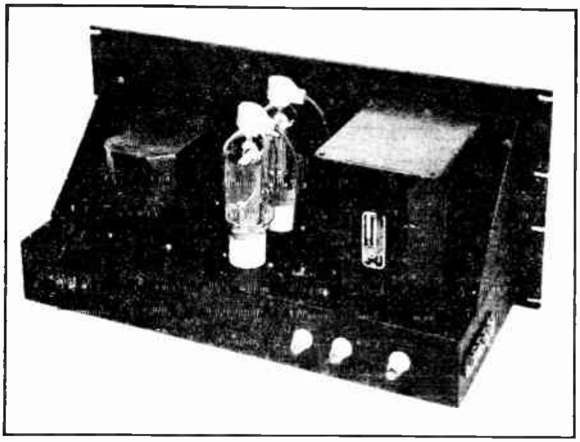


Fig. 1115 — A conventional chassis arrangement for low- and medium-power Class B modulator stages. The mechanical layout in general follows the typical circuit diagrams given in Fig. 1113.

when the driver must be at a considerable distance from the modulator, since the second transformer not only introduces additional losses but also further impairs the voltage regulation.

The bias source for the modulator must have very low resistance. Batteries are the most suitable source. In cases where the voltage values are correct, regulator tubes such as the VR-75-30, VR-105-30, etc., may be connected across a tap on an a.c.-operated bias supply to hold the bias voltage steady under grid-current conditions. Generally, however, zero-bias modulator tubes are preferable, not only because no bias supply is required but also because the loading on the driver stage is less variable and consequently distortion in the driver is reduced.

Condenser C_1 in these diagrams will give a "tone-control" effect and filter out high-frequency sidebands (splatter) caused by distortion in the modulator or preceding speech-amplifier stages. Values in the neighborhood of 0.002 to 0.005 μfd. are suitable. Its voltage rating should be adequate for the peak voltage across the transformer secondary. The plate by-pass condenser in the modulated amplifier will serve the same purpose.

The photographs illustrate different types of construction which may be used for Class B modulators. The actual placement of parts in filling the requirements of any given unit is not critical.

Increasing Modulation Effectiveness

In 'phone transmission communication is carried on by means of the modulation sidebands, not the r.f. carrier. For maximum effectiveness, therefore, the sideband power should be as high as possible. However, modulation in excess of the capability of the transmitter leads to overmodulation "splatter," or spurious sidebands lying outside the normal communication bandwidth. Besides causing unnecessary interference, overmodulation is contrary to the FCC regulations governing amateur 'phone operation.

Methods for increasing the effectiveness of the 'phone transmitter within the limits of modulation capability include restricting the audio-frequency response to those frequencies that contribute most to intelligibility, use of automatic gain control in the speech system, and premodulation clipping of peaks in the voice waveform.

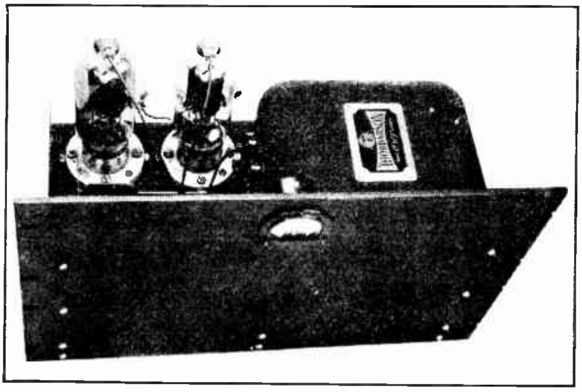


Fig. 1116 — A chassis arrangement for a higher-power Class B modulator. This unit has the filament transformer for the tubes mounted on the chassis. Where the input transformer is included with the speech amplifier, less chassis space will be needed. The tubes are placed near the rear, where the ventilation is good. The plate milliammeter is provided with a small plate over the adjusting screw, to prevent touching the screw accidentally. A Presdwood panel was used for this modulator; with a metal panel, the meter should be mounted behind glass on a well-insulated mount (the meter insulation is not intended for voltages above a few hundred) or connected in the filament center-tap rather than in the high-voltage lead.

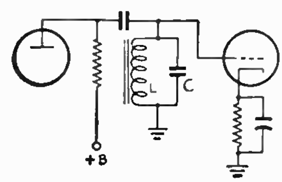


Fig. 1417 — A tuned circuit in the audio amplifier will accentuate the frequencies most useful for voice transmission. This circuit is best adapted to use with a triode amplifier in the preceding stage, but can be used with pentodes if a more "peaked" response is desired. Values are discussed in the text.

Restricting frequency response — Most of the intelligibility in speech is contained in the frequency range from about 500 to 3000 cycles per second. On the other hand, the larger part of the *power* in speech, especially in male voices, is in the frequencies below 500 cycles. With ordinary flat-frequency amplification, therefore, a large part of the modulator power output is devoted to reproducing frequencies that do not contribute materially to understandable speech. By attenuating the frequency response below 400 or 500 cycles the gain can be increased for the higher frequencies without overloading the modulator, thereby considerably increasing the effectiveness of the transmitter for communication purposes.

Fig. 1417 shows a simple tuned circuit that can be installed between two speech-amplifier stages to restrict the frequency response to the most useful frequencies. The *LC* circuit should be adjusted to resonate at approximately 1000 to 1500 cycles. Representative values would be 10 henrys and 0.001 μfd. The resonant frequency can be adjusted either by changing the capacitance of the condenser or by varying the inductance of the coil by varying the width of the air-gap in the core.

In an ordinary resistance-coupled amplifier, the high frequencies can be attenuated by shunting a capacitance from plate to ground or from grid to ground — the common "tone-control" circuit (§7-5). Low-frequency response can be reduced by using a small coupling condenser or low value of grid resistor. If the product of the grid coupling-condenser capacitance (in microfarads) by the grid-leak resistance (in megohms) is made equal to about 0.001 the response will drop off considerably below about 500 cycles.

Volume compression — It is highly desirable to maintain the modulation at as high a level as possible without going into the overmodulation region. Usually the modulation varies over a considerable range as the operator raises or lowers his voice, moves toward or away from the microphone, and so on. If automatic gain control or "volume compression" is incorporated in the speech amplifier the gain may be set at a value that gives full modulation when talking at a low level and

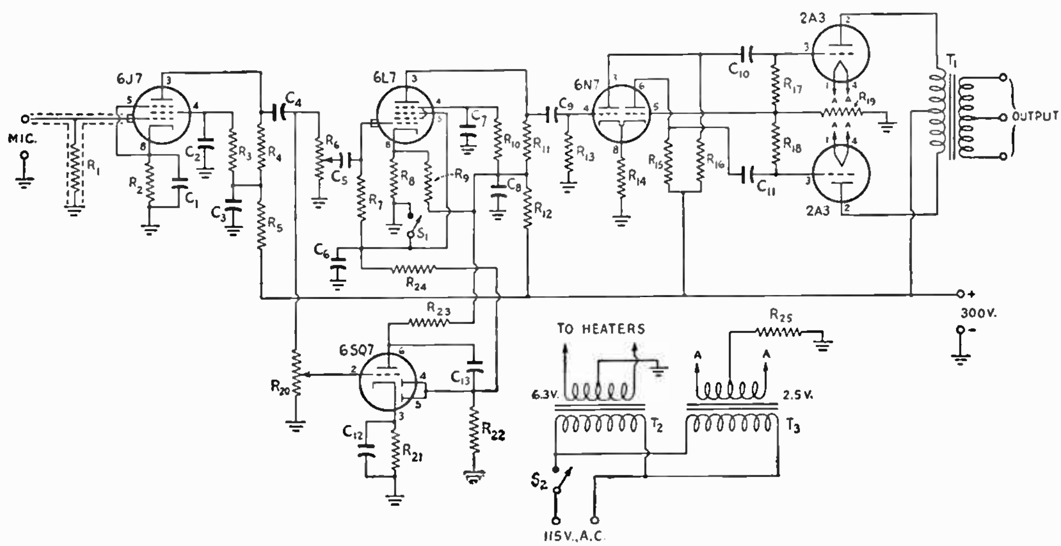


Fig. 1418 — Circuit diagram of the Class-A 2A3 volume-compression speech amplifier.

C_1, C_{12} — 10-μfd. 50-volt electrolytic.
C_2, C_4, C_5, C_6, C_9, C_{10}, C_{11}, C_{13} — 0.1-μfd. 400-volt paper.
C_3, C_8 — 8-μfd. 450-volt electrolytic.
C_7 — 0.47-μfd. 400-volt paper.
R_1 — 4.7 megohms, ½ watt.
R_2, R_8 — 1200 ohms, ½ watt.
R_3, R_7 — 2.2 megohms, ½ watt.
R_4, R_{13}, R_{22}, R_{24} — 0.47 megohm, ½ watt.
R_5 — 47,000 ohms, ½ watt.
R_6, R_{20} — 0.5-megohm variable.
R_9 — 0.22 megohm, 1 watt.
R_{10}, R_{11}, R_{23} — 0.1 megohm, ½ watt.
R_{12} — 10,000 ohms, ½ watt.
R_{14} — 1500 ohms, ½ watt.
R_{15}, R_{16} — 0.1 megohm, 1 watt.
R_{17}, R_{18}, R_{19} — 0.22 megohm, ½ watt.
R_{21} — 4700 ohms, ½ watt.
R_{25} — 750 ohms, 10 watts.
S_1, S_2 — S.p.s.t. switch.
T_1 — Output transformer to match p.p. 2A3s to Class B grids.
T_2 — Filament transformer, 6.3 volts, 2 amperes.
T_3 — Filament transformer, 2.5 volts, 5 amperes.

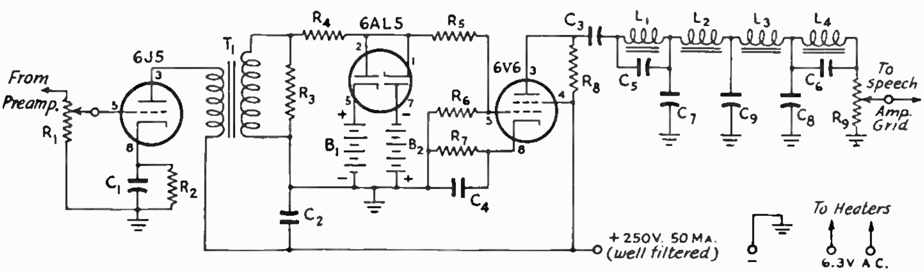


Fig. 1419 — Low-level full-wave clipter system. The gain control R_1 is not required if one of the preceding stages has a gain control.

C_1 — 10-μfd. 25-volt electrolytic.
C_2, C_3 — 8-μfd. 450-volt electrolytic.
C_4 — 25-μfd. 25-volt electrolytic.
C_5, C_6 — 0.023 μfd., ± 5 %.
C_7, C_8 — 0.07 μfd., ± 5 %.
C_9 — 0.08 μfd., ± 5 %.
R_1 — 0.5-meg. pot., a.f. taper.
R_2 — 1000 ohms, 1 watt.
R_3 — 0.15 megohm, 1 watt.
R_4 — 0.22 megohm, 1 watt.
R_5 — 0.47 megohm, 1 watt.
R_6 — 0.22 megohm, 1 watt.
R_7 — 390 ohms, 2 watts.
R_8 — 1000 ohms, ± 10%, 5 watts.
R_9 — 1000-ohm pot., a.f. taper, ± 10% (check with accurate ohmmeter).
L_1 — 30-mh. iron-core choke.
L_2, L_3 — 80-mh. iron-core choke.
L_4 — 30-mh. iron-core choke.
B_1, B_2 — 7½-volt "C" battery.
T_1 — Single-plate-to-p.p.-grids, 2:1 or 3:1 ratio.

the output then will be held at approximately the 100-per-cent modulation level when louder sounds strike the microphone. Automatic gain control is simple in principle; some of the audio output is rectified and filtered to produce a d.c. voltage that varies with the speech amplitude, and this voltage is used to bias a tube in the early speech-amplifier stages so that the louder the sound the greater the reduction in over-all gain.

Fig. 1418 is the circuit diagram of a speech amplifier with volume compression, suitable for working from a crystal microphone and having a power output (6 watts or more, depending upon the efficiency of the output transformer) sufficient to drive a Class B modulator to an output of about 250 watts. The automatic gain-control circuit uses a separate amplifier and rectifier combined in one tube, a 6SQ7. The rectified output of this circuit is filtered and applied to the No. 1 and 3 grids of a pentagrid amplifier tube, thereby varying its gain in inverse proportion to the signal strength. With proper adjustment, an average increase in modulation level of about 7 db. can be secured without exceeding 100-per-cent modulation on peaks.

The amplifier proper consists of a 6J7 first stage followed by a 6L7 amplifier-compressor. The 2A3 grids are driven by a 6N7 self-balancing phase inverter. The operation of the 2A3s is purely Class A, without grid current.

The amount of compression is controlled by the potentiometer, R_{20}, in the grid circuit of the 6SQ7. A switch, S_1, is provided to short-circuit the rectified output of the compressor when normal amplification is required.

Adjustment of the compressor control is rather critical. First set R_{20} at zero and adjust the gain control, R_6, for full modulation with the particular microphone used. Then advance the compressor control until the amplifier just "cuts off" (output decreasing to a low value) on peaks; when this point is reached, back off the compressor control until the cut-off effect is gone but an obvious decrease in gain follows each peak.

Because of the necessity for filtering out the audio-frequency component in the rectifier output, there will be a slight delay (amounting to a fraction of a second) before the decrease in gain "catches up" with the peak. This is caused by the time constant of the circuit, and so is unavoidable.

When a satisfactory setting is secured, as indicated by good speech quality with a definite reduction in gain on peaks, the gain control, R_6, should be advanced to give full output with normal operation. Too much volume compression, indicated by the cut-off effect following each peak, is definitely undesirable, and the object of adjustment of the compressor control should be to use as much compression as possible without danger of overcompression.

Clipter circuit — Sideband power can be greatly increased by cutting off those components in the speech wave that have high *peak* but low *average* amplitude. For distortionless amplification the presence of such peaks requires considerable power *capability* on the part of the modulator, but this capability cannot be utilized because the ratio of peak to average amplitude is high. Cutting off the peaks decreases this ratio to such an extent that much more effective communication is possible, at some sacrifice in naturalness. The intelligibility is greatly improved when the signal is weak at the receiving station because the greater sideband power "cuts through" noise and interference. As much as 25 db. of clipping is advantageous under such conditions.

The clipping must be done in the speech

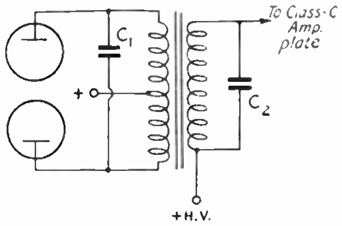


Fig. 1420 — "Building out" the modulation transformer to form a low-pass filter to cut off high frequencies resulting from distortion in the Class B modulator. Values of condensers C_1 and C_2 must be determined by trial as explained in the text. Mica condensers generally are required in view of the audio voltages present across the transformer windings. Capacitances from 0.001 to 0.006 μfd. usually are in the proper range.

amplifier and the clipped output must be passed through a filter to eliminate the high-frequency harmonics that result from clipping. The filter should be of the low-pass type designed to have a cut-off frequency in the vicinity of 3000 to 4000 cycles.

Fig. 1419 shows a premodulation clipping and filter circuit, or "clipter," that may be inserted between two stages in any ordinary speech amplifier at a point where the level is about 6 volts peak. At this level the clipter will provide about 25 db. of clipping at the maximum setting. It consists of a 6J5 amplifier transformer-coupled to a pair of oppositely-connected diodes (6AL5) which short-circuit the output of the 6J5 above a predetermined level. Both positive and negative peaks are clipped. The resultant signal is fed to the grid of a 6V6 amplifier and thence through a low-pass filter consisting of L_1, L_2, L_3, L_4, C_5, C_6, C_7, C_8 and C_9. The output at R_9 is about 4 volts peak for all degrees of clipping. The filter shown has a cut-off frequency of approximately 4000 cycles.

Under conditions of maximum clipping, the peak voltage across the secondary of T_1 will reach about 200 volts. A husky interstage transformer with a well-clamped core is necessary in order to avoid acoustical lamination chatter.

The shunt diode clipper shown has a negligible time constant and holds the peak output voltage to a negligible rise as clipping is increased from threshold to 25 db. It is important that leads between R_4, R_5, R_6, the 6AL5, and the control grid of the 6V6 be kept short and not cabled with other leads or run against the chassis, in order to minimize the time constant of the chopper circuit.

The filter illustrated was designed to use standard values of commercially-available chokes. The filter capacitance values can best be obtained by checking with an accurate capacitance meter or bridge, paralleling two or more capacitors to get the desired value when necessary. Tubular paper capacitors have sufficiently high Q for the purpose, and the better grades will be found to run within 5 per cent of their marked values.

To take full advantage of the clipping feature the transmitter must be capable of 100-per-cent *sine-wave* modulation with low distortion. To adjust the system, turn the gain control (R_1 or other preamplifier control) full on and the clipping level control, R_9, full off. Then, using ordinary speech, advance R_9 until the transmitter shows signs of being modulated at a low level. Listening on a 'phone monitor or the station receiver, adjust R_1 to the highest setting that gives good intelligibility.

Now advance R_9 to a position just below the point where splatter is heard when the station receiver (assuming it is a superhet with antenna terminals shorted to ground) is tuned just off the signal. Have another station, preferably nearby, check for splatter just to be sure. Potentiometer R_9 then need not be touched unless the adjustments to the modulated r.f. stage, particularly loading, are altered appreciably.

If an oscilloscope is available it may be used to check the waveform out of the modulator to ascertain whether the tops of the clipped waves are flat. It may also be used to check the modulation envelope of the r.f. carrier and determine whether the negative peaks are being clipped *in the Class C stage* (negative modulation in excess of 100 per cent). The latter condition is the worst offender so far as splatter is concerned, particularly with plate modulation. If the condition exists, it will be necessary to back off on R_9 until it is corrected. For further details see article by W. W. Smith in February, 1946, *QST*.

Reduction of high-frequency sidebands — Even though means may be incorporated in the speech amplifier to attenuate frequencies above those necessary for intelligible speech, it is still possible for high-frequency sidebands to be radiated if distortion occurs in the modulator, or if the transmitter is overmodulated. High frequencies arising as the result of modulator distortion can be attenuated by the circuit arrangement shown in Fig. 1420. The condensers across the primary and secondary act in conjunction with the leakage reactance of the transformer windings to form a low-pass filter having a cut-off frequency determined by the capacitance and the leakage inductance. Since the latter will vary with different transformers it will be necessary to determine the proper values of capacitance by trial. The voltage ratings of the condensers should be at least as high as the d.c. plate voltage applied to the tube or tubes with which the transformer winding is associated.

The condenser values can be found with the aid of an audio-frequency oscillator and output-voltage measuring device such as a copper-oxide-rectifier voltmeter. For test purposes the audio-frequency input voltage can be kept low so that the meter range will not be exceeded. With the Class C r.f. amplifier disconnected, the meter should be connected across the Class B output-transformer sec-

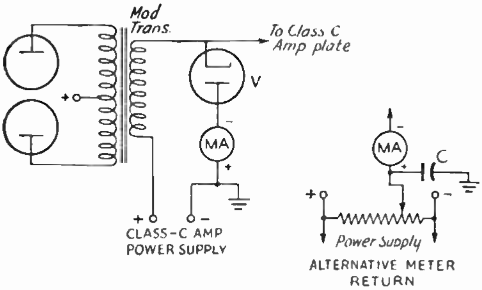


Fig. 1421 — A negative-peak overmodulation indicator. Milliammeter *MA* may be any low-range instrument (up to 0–50 ma. or so). The inverse peak-voltage rating of the rectifier, *V*, must be at least equal to the d.c. voltage applied to the plate of the r.f. amplifier. The alternative meter-return circuit can be used to indicate modulation in excess of any desired value below 100 per cent.

ondary and the audio oscillator should be connected in place of the microphone. (If the oscillator output voltage is too high to permit this, it may be cut in at a later speech stage.) With constant input voltage, vary the frequency while trying different values of capacitance across the primary and secondary until values are found that result in a pronounced drop in output above about 3000 cycles. The same meter may be used for checking both input and output voltages if it is of the multi-range type and the oscillator output is applied to a speech-amplifier stage where a level of a few volts is permissible.

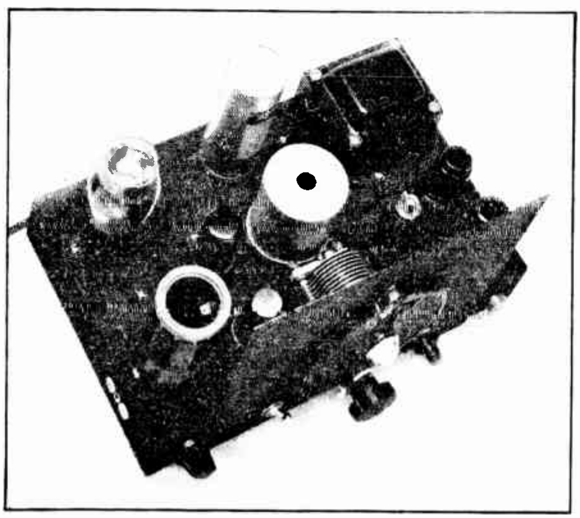
The spurious sidebands set up by overmodulation will not be prevented by the system above. The only way to prevent overmodulation is to monitor the transmissions continuously, with a device such as a simple cathode-ray oscilloscope (by far the most satisfactory type of 'phone monitor) or the negative-peak indicator shown in Fig. 1421. Overmodulation on negative peaks is more likely to result in spurious sidebands than positive overmodulation because of the sharp break that occurs when the carrier is suddenly cut off and on. The milliammeter in the negative-peak indicator of Fig. 1421 will show a reading on each overmodulation peak that carries the instantaneous voltage on the plate of the Class C modulated amplifier "below zero" — that is, negative. The rectifier, *V*, cannot conduct so long as the negative half cycle of audio output voltage is less than the d.c. voltage applied to the r.f. tube. The rectifier tube must be of a type suitable for the Class C plate voltage employed, and its filament transformer must have similarly-rated insulation.

The effectiveness of the monitor is improved if it indicates at somewhat less than 100-per-cent modulation, as it will then warn of the danger of overmodulation before it actually occurs. It can be adjusted to indicate at any desired modulation percentage by making the meter return to a point on the power-supply bleeder as shown in the alternative diagram. The by-pass condenser, *C*, insures that the full audio voltage appears across the indicator circuit. The modulation percentage at which the system indicates is determined by the ratio of the d.c. voltage between the meter tap and the positive terminal to the total d.c. voltage.

Frequency Modulation

At the present time the common method of frequency modulation is to vary the frequency of the controlling oscillator in the transmitter by means of a reactance modulator. This type of modulator may be applied either to a self-controlled or crystal-controlled oscillator. In the former case it can produce fairly wide-band frequency modulation in the v.h.f. region, and of course may be designed so that it gives narrow-band f.m. (in which the deviation ratio is limited to about 0.5, thus giving an f.m.

Fig. 1422 — This modulator-oscillator unit is used with normally crystal-controlled v.h.f. transmitters for frequency-modulated output. It contains a speech amplifier and power supply, so that no additional equipment is needed. The oscillator coil is in the round shield can in the center. The coil in the left foreground is the buffer output circuit. The speech amplifier and modulator are at the right, with the power supply along the rear. A 7 × 11-inch chassis is used.



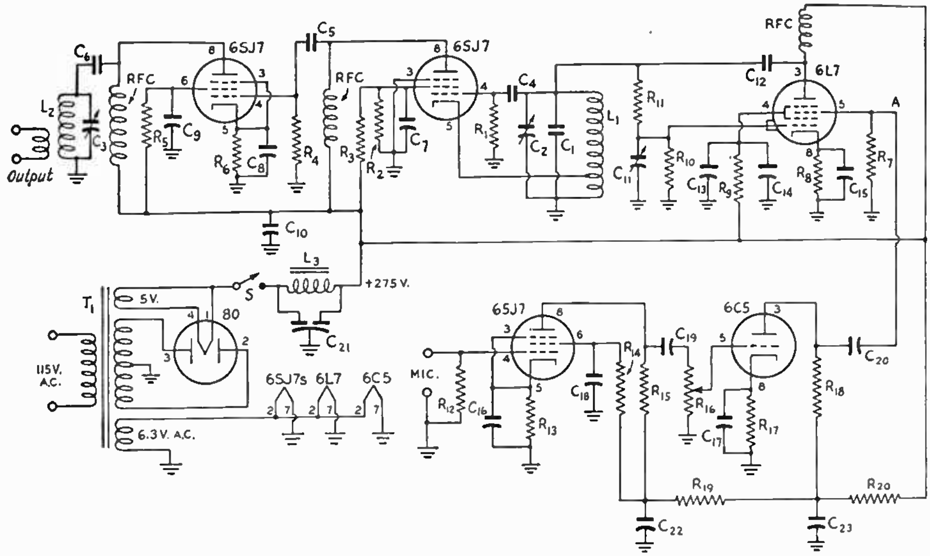


Fig. 1423 — Circuit diagram of the f.m. control unit for use with normally crystal-controlled v.h.f. transmitters.

- C_1 — 150-μμfd. silvered (mica for 7 Mc.)
- C_2 — 100-μμfd. variable (National SE-100).
- C_3 — 50-μμfd. variable (Hammarlund HF-50).
- C_4 — 100-μμfd. mica.
- C_5, C_{12} — 220-μμfd. mica.
- C_6 — 0.001-μfd. mica.
- C_7, C_8, C_9, C_{10}, C_{13}, C_{15}, C_{19}, C_{20} — 0.01-μfd. paper.
- C_{11} — 3-30-μμfd. mica trimmer.
- C_{14}, C_{22}, C_{23} — 8-μfd. 450-volt electrolytic.
- C_{16}, C_{17} — 10-μfd. 25-volt electrolytic.
- C_{18} — 0.1-μfd. 200-volt paper.
- C_{21} — Dual 450-volt 8-μfd. electrolytic.
- R_1 — 0.1 megohm, 1 watt.
- R_2 — 22,000 ohms, 1 watt.
- R_3, R_4, R_5, R_{11} — 47,000 ohms, 1 watt.
- R_6, R_8 — 330 ohms, ½ watt.
- R_7, R_{10} — 0.47 megohm, ½ watt.
- R_9 — 33,000 ohms, 1 watt.
- R_{12} — 4.7 megohms, ½ watt.
- R_{13} — 1000 ohms, ½ watt.
- R_{14} — 1 megohm, ½ watt.
- R_{15}, R_{19} — 0.22 megohm, ½ watt.
- R_{16} — 0.5-megohm volume control.
- R_{17} — 2200 ohms, ½ watt.
- R_{18} — 47,000 ohms, ½ watt.
- R_{20} — 0.15 megohm, 1 watt.
- L_1 — 7 Mc.: 11 turns No. 18 e., length ¾ inch, 1-inch diameter, tapped 3rd turn from ground.
- L_2 — 14 Mc.: 10 turns No. 18 wire on 1½-inch diameter form (Hammarlund SWF-4). Link 3 to 5 turns (not critical).
- L_3 — Filter choke, 10 henrys, 40 ma.
- RFC — 2.5-mh. r.f. choke.
- S — S.p.s.t. toggle switch.
- T_1 — 250-0-250 volts, 40 ma.; 6.3 volts at 2 amperes; 5 volts at 2 amperes. (Thordarson T-13R11).

signal that occupies substantially no more channel width than an a.m. signal) on the 28-Mc. band. With a crystal oscillator, the reactance modulator is capable of giving sufficient deviation for narrow-band f.m. at 28 Mc.

The unit shown in Figs. 1422, 1423 and 1424 is a complete VFO-modulator designed to work into a normally crystal-controlled transmitter using either 7- or 14-Mc. crystals. The r.f. output of the unit is intended to be fed through a link to a tuned-circuit coil wound on a coil form which substitutes for the crystal holder in the crystal oscillator. This tuned circuit is resonant at the same frequency as the output tank of the control unit. L_2C_3 in Fig. 1423 and can, in fact, be identical in construction.

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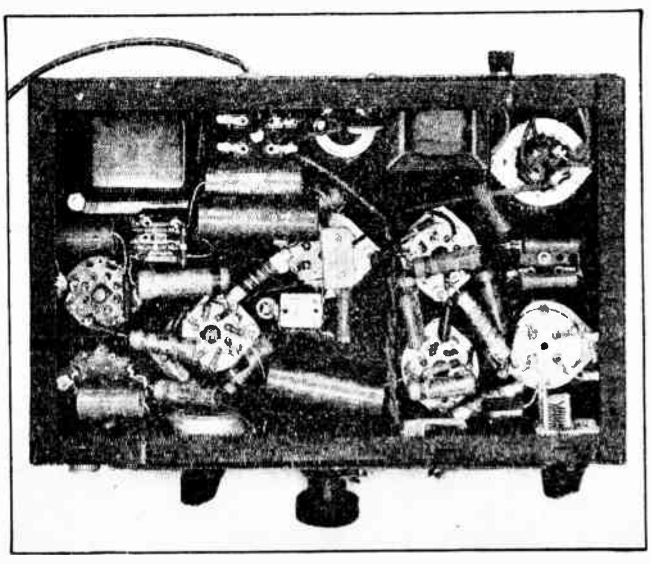
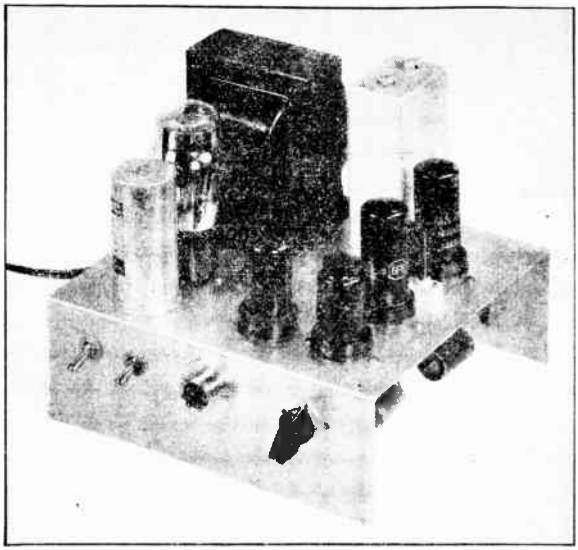


Fig. 1424 — In this bottom view of the f.m. modulator unit, the r.f. section is at the right and the audio at the left. The oscillator socket is to the right of the coil socket in the center.

Fig. 1425 — **An f.m. frequency-control and modulator unit using reactance modulation of a crystal oscillator.**



In transmitters using triode oscillators, or pentode crystal oscillators in which the tubes are not well screened, it is advisable to use the crystal oscillator tube as a doubler rather than as a straight amplifier. If the transmitter uses a crystal oscillator operating in the vicinity of 14 Mc., for example, the output of the unit may be on 7 Mc. and the grid circuit of the ex-crystal tube also tuned to 7 Mc. This will avoid difficulty with self-oscillation in the ex-crystal stage. With a pentode oscillator it is possible to work straight through, provided the grid tank substituted for the crystal is tuned well on the high-frequency side of resonance, but this procedure is not advisable since it may make the modulation nonlinear. It is rather important that all circuits in the transmitter be tuned "on the nose" for best performance. Of course, if the crystal tube is a well-screened transmitting type it can be used as a straight amplifier.

With harmonic-type crystal oscillators the input frequency can be the same as that of the crystal, since the output frequency of the crystal tube is already a harmonic. In the Tri-tet oscillator, the cathode tank should be short-circuited; in the types using a cathode impedance to provide feed-back, this impedance

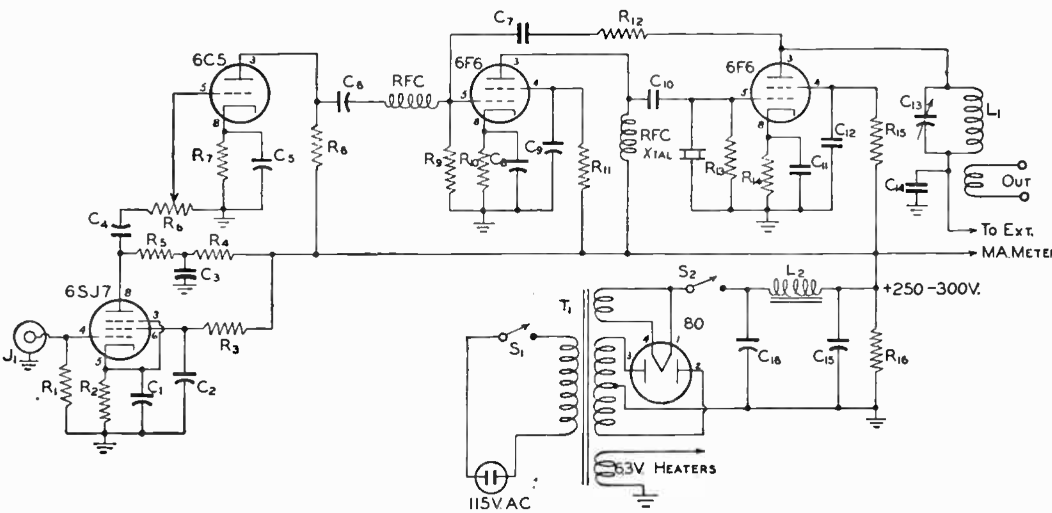


Fig. 1426 — **Circuit diagram of the narrow-band modulator unit with crystal-controlled oscillator.**

C_1, C_5, C_8 — 5-μfd. 50-volt electrolytic.
C_2, C_3, C_4, C_6, C_9 — 0.01-μfd. paper.
C_7 — 5.5-μμfd. ceramic (3-30-μμfd. trimmer adjusted to same capacitance may be used).
C_{10}, C_{11}, C_{12}, C_{14} — 0.001-μfd. mica.
C_{13} — 50-μμfd. variable.
C_{15} — 20-μfd. 450-volt electrolytic.
C_{16} — 10-μfd. 450-volt electrolytic.
R_1 — 4.7 megohms, ½ watt.
R_2 — 1000 ohms, ½ watt.
R_3 — 0.47 megohm, ½ watt.
R_4 — 22,000 ohms, ½ watt.
R_5 — 0.22 megohm, ½ watt.
R_6 — 1-megohm volume control.
R_7 — 1500 ohms, ½ watt.
R_8 — 0.1 megohm, ½ watt.
R_9, R_{12} — 0.47 megohm, ½ watt.
R_{10} — 390 ohms, ½ watt.
R_{11} — 0.1 megohm, 1 watt.
R_{13} — 4700 ohms, 1 watt.
R_{14} — 470 ohms, ½ watt.
R_{15} — 22,000 ohms, 1 watt.
R_{16} — 10,000 ohms, 25 watts.
L_1 — 56 turns No. 26 enam., ⅞-inch diam., 1⅜ inches long. Link, 8 turns.
L_2 — 15 henrys, 70 ma.
J_1 — Microphone-cable socket.
RFC — 2.5-mh. r.f. choke.
S_1, S_2 — S.p.s.t. toggle.
T_1 — Receiver-type power transformer; 250 to 30 volts each side c.t. at 70 ma.

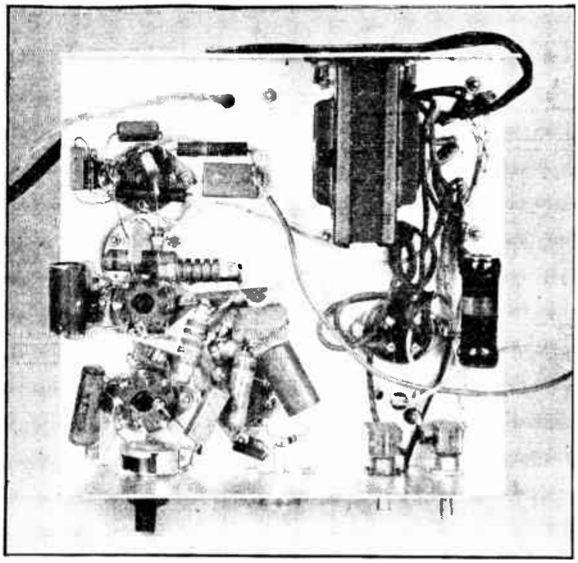


Fig. 1427 — Bottom view of the crystal-controlled f.m. unit.

The sensitivity of the modulator is controlled by the setting of C_{11}. The higher the capacity of this condenser the smaller the frequency deviation for a given audio input voltage to the modulator. At maximum sensitivity, with C_{11} at minimum capacity, the linear deviation is approximately 1.5 kc. with an a.f. input to the modulator grid of 2 volts peak. The actual deviation at the output frequency of the transmitter depends upon the amount of frequency multiplication following the modulated oscillator. The maximum linear deviation is approximately 6 kc. at 28 Mc., 12 kc. at 50 Mc., and 36 kc. at 144 Mc.

also should be shorted. Care should be taken to avoid short-circuiting the grid bias, whether from a cathode resistor or grid leak. In the latter case this usually will mean that a blocking condenser (470 μμfd. or larger) should be connected between the "hot" end of the grid tank and the grid of the ex-crystal tube, with the grid lead (and choke) connected on the grid side of the condenser. Such a blocking condenser may be incorporated in the plug-in tank. The grid-tank tuning condenser may be a small air padder mounted in the coil form.

Where a suitable power supply and speech amplifier are already available, the lower part of Fig. 1423 can be omitted and only the oscillator, buffer and modulator units need be built. With transformer input, the transformer and gain control should be connected between ground and point *A* of Fig. 1423, R_7 being omitted. Any of the conventional methods may be used to couple the modulator to an available speech amplifier, with one precaution — if a high-impedance connection is used, the "hot" lead should be shielded to prevent hum pick-up.

If the transmitter to be used has a self-excited oscillator, electron-coupled or otherwise, a separate oscillator need not be built. The reactance modulator can be connected directly across the tank circuit of the oscillator.

The circuit constants of the oscillator in the unit pictured are adjusted to cover the frequency range 6000-7425 kc. so that the output can be multiplied into the 28-, 50- and 144-Mc. bands. For 28-Mc. operation a multiplication of 4 is required; for 50 Mc., a multiplication of 8; and for 144 Mc., a multiplication of 24. The output circuit, L_2C_3, is tunable over the range 12-15 Mc., and thus is adapted to feeding into a transmitter using crystals operating in this range. For replacing crystals operating at half this frequency, L_2 should have 20 turns with all other coil dimensions remaining the same.

Crystal-controlled F.M. — The frequency-control unit shown in Figs. 1425 and 1427 provides reactance-tube modulation of a crystal oscillator, and thus meets the needs of those who want narrow-band frequency modulation with high carrier-frequency stability. The circuit is given in Fig. 1426. With AT-cut crystals it is possible to obtain a deviation of 200 cycles at 3.5 Mc., which when multiplied to 28 Mc. gives a deviation of 1600 cycles — or a deviation ratio of approximately 0.5, based on an upper audio-frequency limit of 3000 to 4000 cycles. This order of deviation is sufficient for reception on ordinary communications-type superheterodyne receivers when the receiver is detuned slightly from the carrier frequency. With X- and Y-cut crystals, deviations at the fundamental frequency of approximately 1000 and 2500 cycles, respectively, are obtainable.

The circuit values are rather critical and should be followed closely for optimum results. The plate tank condenser of the crystal oscillator, C_{13}, should be set slightly on the high-frequency side of the minimum plate-current point to obtain optimum modulation. If the condenser is set too near the point of minimum plate current it is possible that the crystal will be swung out of oscillation on voice peaks. The setting that gives maximum stable modulation with good voice quality can be determined by listening to the 28-Mc. harmonic from the unit on a regular receiver.

The r.f. output of the unit readily can be coupled into a transmitter having a 3.5-Mc. crystal oscillator by winding a link of a few turns around the plate coil of the transmitter oscillator and removing the regular oscillator tube, the link being connected to the output terminals of the unit shown. The crystal-controlled f.m. unit is built on a U-shaped chassis measuring 7½ by 8 by 3 inches.

Chapter Fifteen

V.H.F. Receivers

IN ITS essentials most modern receiving equipment for the 28- and 50-Mc. bands differs very little from that used on lower frequencies. The 28-Mc. band serves as the meeting ground between what are ordinarily termed "communications frequencies" and the very-highs, and it will be found that most of the receivers described in Chapter Twelve are capable of working on 28 Mc. In this chapter are described receivers and converters capable of good performance on 50 Mc. and higher.

Federal regulations impose identical requirements on all frequencies below 54 Mc. respecting stability of frequency and, when amplitude modulation is used, freedom from frequency modulation. Thus receivers for 50-Mc. a.m. reception may have the same selectivity as those designed for the lower frequencies. This order of selectivity is not only possible but desirable, since it permits a considerable increase in the number of transmitters which can work in the band without undue interference. High selectivity also aids greatly in improving the signal-to-noise ratio, both as concerns noise originating in the receiver itself and in its response to external noise. The effective sensitivity of such a receiver can be made considerably higher than is possible with nonselective receivers.

Receivers for f.m. signals usually are designed with less selectivity, so that they can accommodate the full swing of the transmitter. At least for 28- and 50-Mc. f.m. reception, however, the h.f. oscillator must be as stable as in a narrow-band a.m. receiver.

The superheterodyne system of reception is used almost universally on frequencies below 54 Mc. because it is the only type that fulfills the stability requirements. A.m. superheterodynes and those for f.m. reception differ only in the i.f. amplifier and second detector, so that a single high-frequency converter may be used for either a.m. or f.m.

Superheterodynes for 50 Mc. should have fairly high intermediate frequencies to reduce both image response and oscillator "pulling." For example, a difference between signal and image frequencies of 900 kc. (the difference when the i.f. is 450 kc.) is a very small percentage of the signal frequency; consequently, the response of the r.f. circuits to the image frequency is nearly as great as to the desired signal frequency. To obtain discrimination against the image equal to that obtainable at 3.5 Mc. would require an i.f. 16 times as high, or about 7 Mc. However, the Q of tuned circuits is less at 50 Mc. than it is at the lower frequencies, chiefly because the tube loading is considerably greater, and thus still higher i.f.s are desirable. A practical compromise is reached at about 10 Mc.

To obtain high selectivity with a reasonable number of i.f. stages, the double-superheterodyne principle is often employed. A 10-Mc. intermediate frequency, for example, is changed to a second i.f. of perhaps 450 kc. by an additional oscillator-mixer combination.

Few amateurs build complete 50-Mc. superheterodyne receivers. General practice in this band has been to use a conventional communications receiver to handle the i.f. output of a simple 50-Mc. frequency converter. Even an all-wave broadcast receiver may be used with excellent results on 50 Mc. by the addition of a relatively simple converter.

The superheterodyne type of receiver is finding increased favor for 144-Mc. work also, as the occupancy of that band increases. Especially in heavily-populated areas, stabilization of transmitters and an improvement in the selectivity of receivers are becoming almost mandatory, particularly for those operators who are interested in exploiting the full possibilities of this band. The ideal receiver for present conditions is one having a pass-band of around 100 kc. Greater selectivity is hardly desirable, not only because it will discriminate against transmitters having the slightest instability, but because the receivers themselves are inclined to be somewhat less stable at this frequency. This approach has been used in most of the recent pioneering efforts by amateurs working in the microwave field.

A converter working into an f.m. receiver, or into a broad-band i.f. channel designed for either a.m. or f.m. reception, provides a quite satisfactory means of reception of signals, not only at 144 Mc., but on up through the microwave range.

The simplest type of v.h.f. receiver is the superregenerator, long favored in amateur work. It affords good sensitivity with few tubes and elementary circuits. Its disadvantages are lack of selectivity and, if the oscillating detector is coupled to an antenna, a tendency to radiate a signal which may cause interference to other receivers. To some extent the lack of selectivity is advantageous, since it makes for easy tuning, and permits reception of all signals within its tuning range, however unstable they may be. To reduce radiation, a superregenerative detector should be preceded by an r.f. stage, or, if the detector is coupled directly to the antenna, it should be operated at the lowest plate voltage which will permit superregeneration.

From a practical aspect, superregenerative receivers may be divided into two general types. In the first the quenching voltage is developed by the detector tube functioning as a "self-quenched" oscillator. In the second, a separate oscillator tube is used to generate the quench voltage. Self-quenched superregenerators have found wide favor in amateur work. The simpler types are particularly suited for portable equipment, which must be kept as simple as possible. Many amateurs have "pet" circuits claimed to be superior to all others, but the probability is that the arrangement of a particular circuit has led to correct operating conditions. Time spent in minor adjustments will result in a smooth-working receiver.

A 144/235-Mc. Superregenerator

The receiver in Figs. 1501, 1502 and 1503 affords excellent sensitivity on both 144 and 235 Mc. For the amateur who wishes to experiment on these bands, it will provide satisfactory reception at minimum expense. The circuit is the familiar self-quenched superregenerative detector, followed by two stages of audio amplification.

The receiver is built on a 7 × 7 × 2-inch chassis. The tuning condenser is mounted on a metal bracket, cut in the shape of a U to clear the stator connections. The dial is connected to the condenser by a flexible bakelite coupling.

The improvised socket for the plug-in coils utilizes contacts obtained from an Amphenol miniature-type tube socket, by the process of squeezing the socket in a vise until the bakelite cracks. One contact is soldered to each of the tuning-condenser connections and a third to a lug supported by one of the extra holes in the Isolantite base of the condenser. The contacts must all be placed at exactly the same height, so that the plug-in coil will seat properly. The band-set condenser, C_2, is mounted by soldering short strips of wire between the lugs and the tuning-condenser terminals.

The polystyrene tube socket for the 9002 is mounted on a metal bracket, placed near the tuning condenser so as to allow a very short lead from the condenser to the plate terminal and just enough room between the rotor connection and the grid terminal for the grid condenser. Heater and cathode leads are brought through the chassis in a rubber grommet.

The variable antenna coupling coil, L_1, is mounted on a polystyrene rod supported by a shaft bearing. The rod is prevented from moving axially in the bearing by cementing a fiber washer to the shaft and tightening the knob on the other side. The antenna coupling loop should be adjusted so that, when rotated, it will just clear the coils plugged into the socket.

The coils are mounted on small strips of ⅛-inch polystyrene (Millen QuartzQ) which have three small holes drilled in them corresponding exactly with the coil socket. Each coil is cemented to the strip with Duco cement at the points where the wire passes through the base. The No. 18 wire used for the coils will fit snugly in the sockets if the contacts are pinched slightly. The coils are trimmed to fit the bands by spreading or squeezing the turns slightly. The mica-trimmer band-set condenser

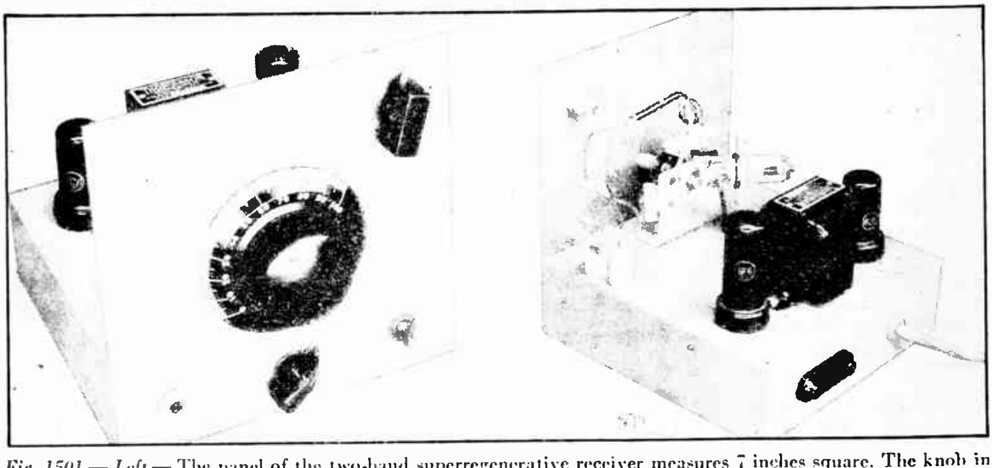


Fig. 1501 — Left — The panel of the two-band superregenerative receiver measures 7 inches square. The knob in the upper right-hand corner adjusts antenna coupling, while the knob below the tuning dial controls regeneration. *Right* — A rear view of the two-band superregenerative receiver. The 235-Mc. plug-in coil is in the foreground.

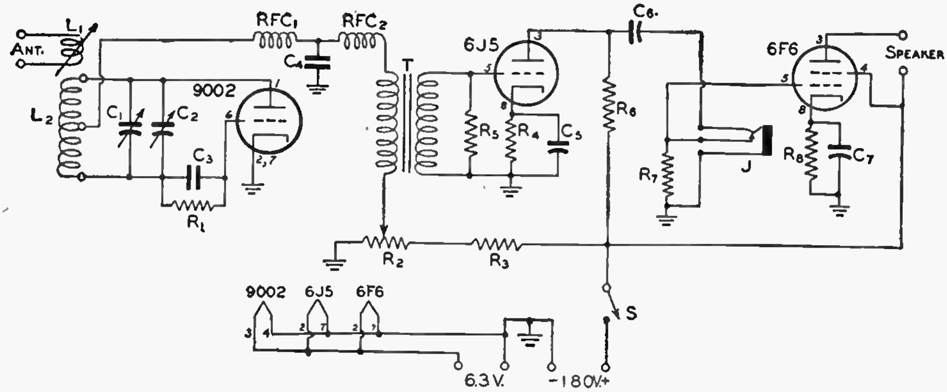
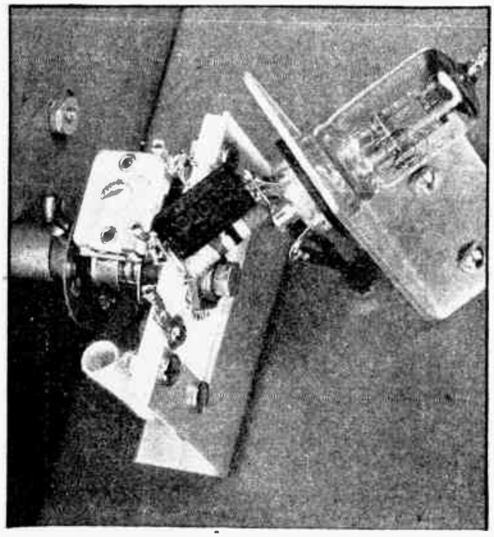


Fig. 1502 — Wiring diagram of the superregenerative receiver for 144 and 235 Mc.

C1 — 5-μμfd. midget variable (National UM-15, 4 plates removed).
C2 — 3-30-μμfd. mica trimmer.
C3 — 50-μμfd. mica.
C4 — 0.0033-μfd. mica.
C5, C7 — 10-μfd. 25-volt electrolytic.
C6 — 0.01-μfd. 400-volt paper.
R1 — 10 megohms, ½ watt.
R2 — 50,000-ohm wire-wound variable.
R3, R5, R6, R7 — 0.1 megohm, ½ watt.
R4 — 2200 ohms, ½ watt.
R8 — 470 ohms, 1 watt.
L1 — 1 turn No. 14 e., ⅜-inch inside diameter.
L2 — 144 Mc.: 3 turns No. 18 e., ½-inch diameter, spaced over ¼ inch. Tapped 1¼ turns from plate end.
235 Mc.: 2 turns No. 18 e., ¼-inch diameter, spaced over ½ inch. Tapped at center of coil.
J — Closed-circuit jack.
RFC1 — 25 turns No. 24 d.c.c., close-wound, ¼-inch diameter.
RFC2 — 8-mh. r.f. choke.
S — S.p.s.t. toggle switch.
T — Plate-to-grid interstage audio transformer (Thordarson T-57A36).

gives some further range of adjustment. In the receiver as described, it is screwed down fairly tightly for the 144-Mc. band and loosened about four revolutions for 235 Mc. In the absence of good marker stations, an absorption frequency meter or a Lecher wire system (described in Chapter Nineteen) may be used for spotting the band limits.

Two factors which will be found to influence sensitivity are the value of C_4 and the degree of antenna coupling. Values of C_4 from 0.001 to 0.0047 μfd. should be tried. The antenna coupling will vary greatly with the setting of L_1 and the type of antenna used; it is well worth while to tune the antenna circuit and then vary the coupling with the panel control.



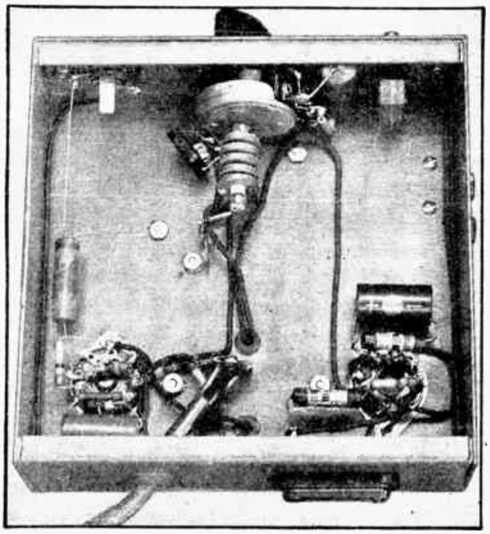


Fig. 1503 — *Left* — A close-up view of the tuning assembly, showing how the leads from the tuning condenser to the tube socket have been kept short and how the coil socket is mounted on the tuning condenser. Hidden by the grid condenser (the 50-μμfd. condenser so prominent in the picture), the plate terminal of the tube socket goes to a lug which has been added to the stator of the tuning condenser. *Right* — The arrangement of parts under the chassis may be seen in this photograph. The 6J5 socket is at the left and the 6F6 socket is at the right, near the 'speaker terminals. The 8-mh. r.f. choke, seen just under the regeneration control at the top center, is supported by tie-strips.

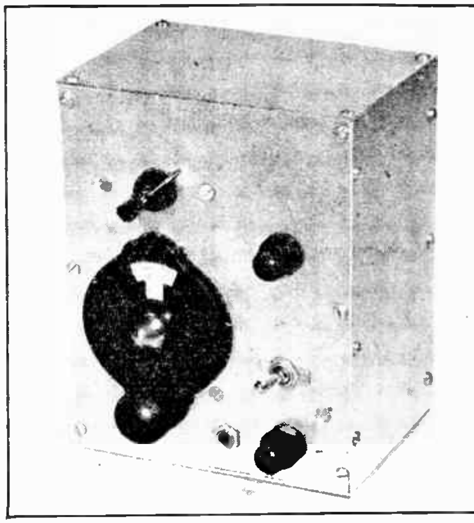


Fig. 1504 — Front view of the 144-Mc. t.r.f. receiver. The pointer knob above the vernier dial tunes the r.f. stage. The small round knobs are for audio volume (lower right) and detector plate-voltage variation. Outside dimensions of the handmade case are 7 × 5½ × 4 inches.

Tight coupling usually will give better results than loose coupling. The coupling can be increased almost up to the point where the detector no longer oscillates, with no ill effects except increased radiation.

An audio volume control could be installed in place of the fixed grid resistor, R_7, if desired. In the original model of this receiver, the value of R_7 was adjusted until normal loudspeaker output was obtained; this value may be varied to meet any particular requirements.

Though a 9002 detector is shown, a 6C4 will work equally well. Socket connections are similar, but some experimentation with different values for R_1 and C_1 may be necessary.

T.R.F. Superregenerative Receiver

The 144-Mc. receiver in Figs. 1504–1508 uses miniature tubes throughout and is intended for either home or portable/mobile use. The r.f. amplifier stage furnishes some additional gain over a straight superregenerative detector, affords freedom from antenna effects, and — most important of all — prevents radiation from the receiver. Although the r.f. and detector circuits are individually tuned, the broad tuning of the r.f. stage makes the receiver essentially a single-dial affair — important in mobile work — and the low-priced miniature tubes permit compact assembly and low current consumption. Heater current is 625 ma. at 6.3 volts, and the total plate drain from 135 volts of "B" battery is less than 10 ma.

The tuned r.f.-amplifier stage uses a 6AK5 pentode which is coupled through C_5 to the 6C4 superregenerative triode detector. This in turn is transformer-coupled to a 6C4 audio stage which drives the 6AK6 output stage. A plate coupling choke, L_4, and the coupling condenser C_{12} remove d.c. from the output jack, J_2, and eliminate the possibility of short-circuiting the plate supply at this point.

The receiver chassis and partitions are built from pieces of 1/16-inch aluminum held together at the corners with machine screws and strips of 1/4-inch square brass rod. The over-all dimensions are 7 × 5 × 4 inches — the chassis that mounts the audio components is 4 × 5 inches with a 1¾-inch folded lip. To eliminate oscillation in the r.f. stage and radiation from the detector, completely-separate compartments are used for the r.f. and detector stages. These compartments consist of identical boxes that measure 1⅞ inches square and 3 inches long. The tube sockets are mounted on the end plates, and all of the connections to the sockets are made before the boxes are completely assembled. The wire between C_5 and L_3 runs through two Millen 32150 bushings in the walls of the two shield compartments. This interconnection, the only one except for the power circuits, is made by running separate leads from the condenser and coil through the bushings and then soldering the two ends together after the two units are mounted on the front panel.

The detector tuning condenser, C_8, is a regular Cardwell ZV-5-TS modified by adding a single

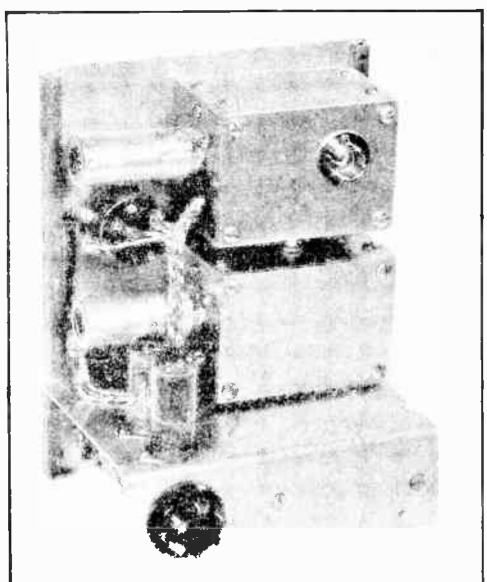


Fig. 1505 — Rear view of the complete receiver. Note that the r.f. stage and superregenerative-detector circuit components are in separate completely-enclosed compartments, for elimination of radiation. Miniature tubes are used throughout, for compactness and low current consumption.

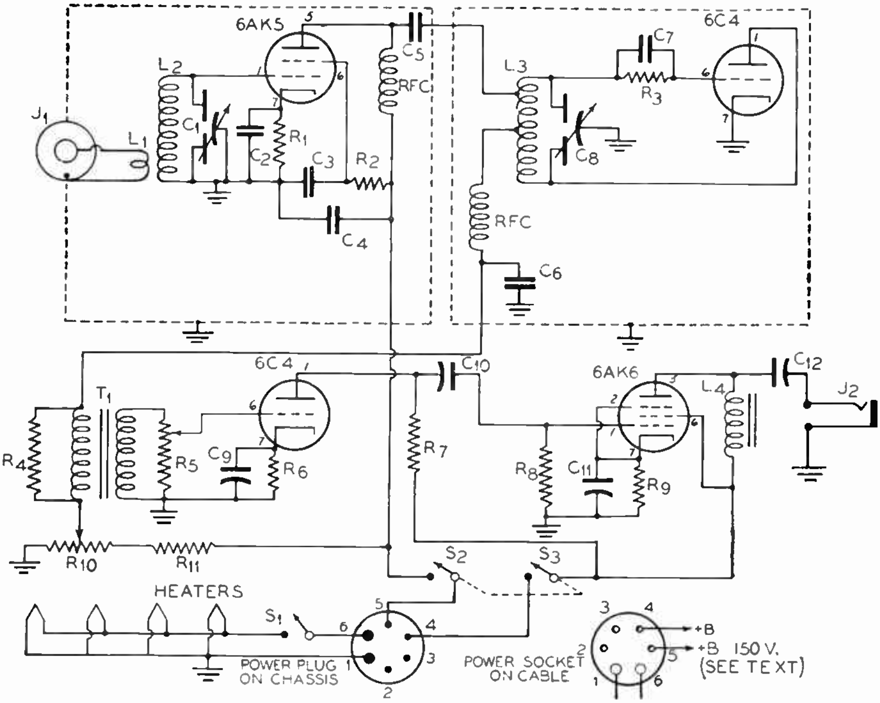


Fig. 1506 — Wiring diagram of the four-tube t.r.f. superregenerative receiver. Boundaries of shield compartments housing r.f. and detector stages are shown by dotted lines.

C_1, C_8 — Split-stator condenser (Cardwell ZV-5-TS). See text.
C_2, C_3, C_4 — 500-μμfd. midget mica.
C_5, C_7 — 17-μμfd. midget mica.
C_6 — 0.0022-μfd. midget mica.
C_9, C_{11} — 10-μfd. 25-volt midget electrolytic.
C_{10}, C_{12} — 0.1-μfd. paper.
R_1 — 1500 ohms, ½ watt.
R_2, R_7, R_8 — 0.1 megohm, ½ watt.
R_3 — 3.3 megohms, ½ watt.
R_4 — 39,000 ohms, ½ watt. See text.
R_5 — 0.5-megohm potentiometer.
R_6 — 2200 ohms, ½ watt.
R_9 — 680 ohms, ½ watt.
R_{10} — 50,000-ohm potentiometer.
R_{11} — 22,000 ohms, 1 watt.
L_1 — 2 t. ⅜-inch i.d. No. 18 enam. inserted between turns of L_2 at cold end.
L_2 — 1 t. ⅜-inch i.d. ¾ inch long. No. 18 tinned.
L_3 — 5 t. center-tapped, ½ inch long, No. 18 tinned. R.f. coupling tap, 1 t. from grid end.
L_4 — Midget audio or filter choke (Inca D-92).
J_1 — Coaxial socket (Jones S-201). Matching plug for antenna is P-101 or P-201.
J_2 — Headphone or speaker jack
RFC — See text
S_1 — S.p.s.t. switch on R_{10}.
S_2, S_3 — D.p.s.t. toggle switch.
T_1 — Midget audio transformer (Thordarson T-13A34).

circular plate to the regular one-plate rotor. This additional constant capacitance across the circuit increases the bandspread and, because it decreases the *L*-to-*C* ratio, smooths out the regeneration so that the regeneration control, R_{10}, does not have to be readjusted within the 144-Mc. band.

The two r.f. chokes, *RFC*, are homemade affairs wound on 1-watt IRC composition resistors — 0.22 megohm or higher — the insulated type that is ¼ inch in diameter and 2⁵⁄₃₂ inch long. The ends are notched with a small file or saw, to prevent the ends of the

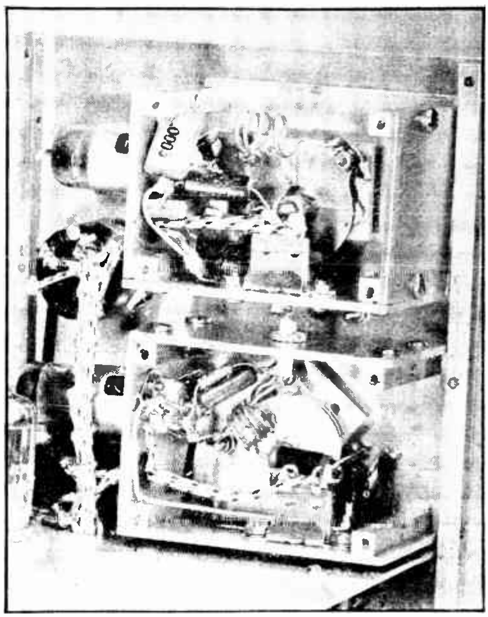


Fig. 1507 — Close-up view of the r.f. and superregenerative-detector compartments, with back plates removed to show details. Top, back, and right side may be removed from either assembly, providing accessibility despite compact design.

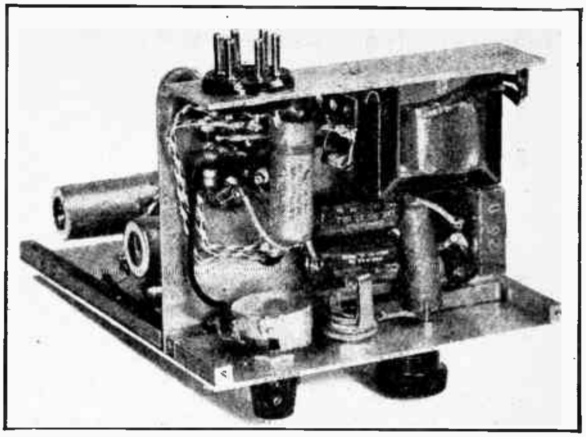


Fig. 1508 — Bottom view, showing audio-component arrangement.

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coil wire from slipping after they have been soldered to the pigtail leads of the resistor, and then a single layer of No. 30 d.s.c. is wound on for a length of $1\frac{7}{32}$ inch. No lacquer or dope should be used on the winding because of the increased distributed capacity that will result.

When the receiver is completely wired the first move should be to check detector operation. With the 6AK5 in its socket, but with no plate or screen voltage applied to it, apply the plate voltage to the detector and check for the customary hiss. Try the regeneration control, R_{10}, to determine whether the detector goes in and out of superregeneration smoothly. Some variation in values of R_3, R_4 and C_6 may be necessary to attain this end, and some 6C4s work better than others in this respect.

Next, the tuning range should be checked by means of Lecher wires or an absorption-type wavemeter. With the values given, 144 Mc. should fall at about 80 on the dial, with 148 Mc. at around 60. The position of the r.f. coupling tap on L_3 will have considerable effect on the resonant frequency of the combination. Its position is not critical, except for its effect on the tuning range of the detector circuit, but the spacing of the turns in the coil will have to be changed if the position of the tap is materially different from that given.

When the detector is found to be in the band, the r.f. stage may be put into operation. With any of the shields removed, or with no antenna connected, the 6AK5 will probably oscillate, blocking the detector, but this effect will disappear when the two compartments are completely assembled and an antenna attached by means of the coaxial connector. If the r.f. stage is operating properly there will be slight change in the character of the hiss when the stage is tuned through resonance. Using a signal generator (the harmonic of any oscillator which falls in the 144-Mc. band will do) or the signal of a 144-Mc. station, there will be a pronounced drop in background noise and a slight change in dial setting of the detector when the r.f. stage is tuned "on the nose." Once the r.f. tuning is adjusted for maximum response, preferably on a weak signal near the middle of the band, it may be left at that setting for all except the very weakest signals at either end.

Power-supply filtering and regulation are important factors in attaining smooth and efficient performance with superregenerative detectors. The power plug mounted on the back of the chassis provides a separate connection (Pin 5) for the detector and r.f. +B, in order that this may be drawn from a regulated source, such as a VR-150. The other pin marked "+B" (Pin 4) supplies the audio tubes, and the voltage used here need not be regulated. If "B" batteries are used — and they are highly recommended for mobile operation — Pins 4 and 5 may be connected together in the power socket on the cable. The use of "B" batteries in mobile work will result in better sensitivity and more quiet operation than will be available with any sort of mobile power supply, vibrator or dynamotor, and the drain from the car battery will be negligible during receiving periods. A set of medium-size "B" batteries (135 volts is sufficient for good 'speaker volume) will last through a year or more of normal operation. When batteries are used, the on-off switch, S_2-S_3, should be thrown to the "off" position when the receiver is not in use, otherwise there will be a small continuous drain on the batteries through the R_{10}-R_{11} bleeder.

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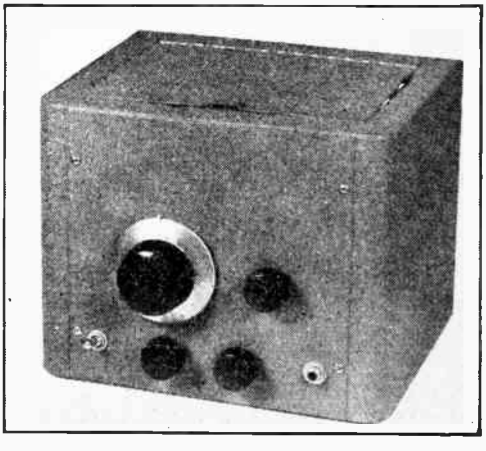


Fig. 1509 — The four-tube 144-Mc. superheterodyne, dressed up in a modern cabinet. The large dial is oscillator tuning, and the small dial is for mixer tuning. The two knobs control regeneration (right) and volume (left).

ℂ A 144-Mc. Superregenerative Super-heterodyne Receiver

A superheterodyne, using a superregenerative second detector is shown in Figs. 1509, 1510, 1511 and 1512. A 6J6 miniature twin triode is used as local oscillator and mixer, and its high transconductance (5300 μmhos) and small size make for good performance in the 2-meter band. The superregenerative second detector is a triode-connected 6V6 working at 33 Mc., and this is followed by a 6J5 for headphones and a 6F6 for loudspeaker operation. The wiring diagram, Fig. 1510, shows no coupling condenser between the oscillator and mixer because stray coupling between grid pins at the socket gives adequate injection to the grid-leak biased mixer. A small coil, L_4, is used in the plate circuit of the mixer to resonate in series with C_5 to the signal frequency, and the resistor, R_{13}, is included to make this resonance a broad one. The condenser C_5 also tunes the primary, L_5, of the i.f. transformer. The i.f. transformer is adjustable only in the secondary circuit, since with just one stage there is no tuning requirement other than that the primary and secondary be tuned to the same frequency. A switch, S_1, removes the plate voltage from the second detector and following stages during transmission periods, but plate voltage is left on the oscillator (and mixer) to avoid drift. This is an unnecessary refinement, however, since the oscillator drift is considerably less than the bandwidth of the i.f. amplifier.

Inductive tuning of the oscillator circuit is used, by moving a copper vane which acts as a low-resistance shorted turn in the field of the coil. As the vane is moved into the field of the coil, the inductance is reduced. No current flows through the insulated shaft supporting the vane, and consequently there is no "jumping" of frequency such as is caused by erratic contact to a condenser rotor

The mixer coil, L_2, is wound on a National XR-50 form in which the iron slug has been replaced by a brass one from an AR-2 form. The coil is peaked for the center of the band — the tuning is broad — and additional trimming is done with the antenna condenser, C_1. Three antenna binding posts are available, so that either series or parallel tuning of L_1 can be used.

The receiver is designed to be mounted in a commercial-type 8 × 10 × 8-inch cabinet. The panel, part of the standard cabinet, measures

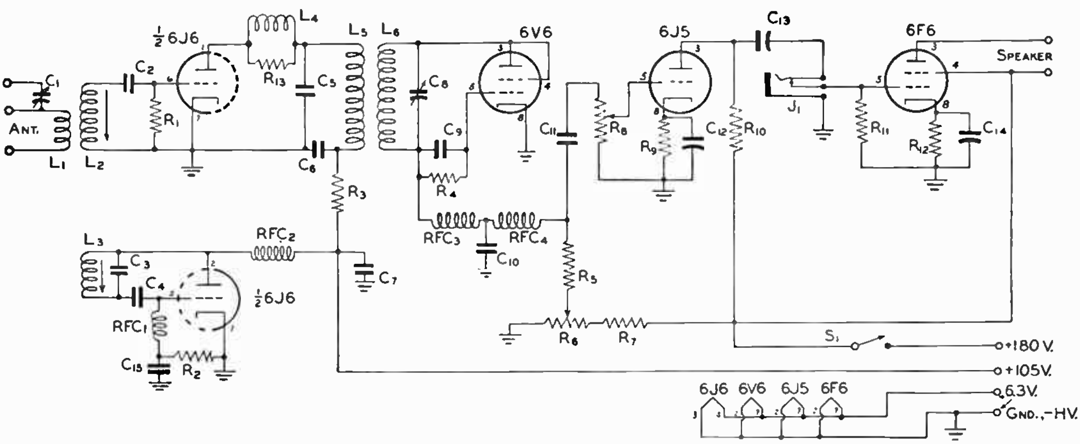


Fig. 1510 — Wiring diagram of the 144-Mc superheterodyne.

C_1 — 35-$\mu\mu$fd. variable (National UM-35).
C_2 — 27-$\mu\mu$fd. ceramic.
C_3, C_5 — 10-$\mu\mu$fd. ceramic.
C_4 — 10-$\mu\mu$fd. mica or ceramic.
C_6 — 470-$\mu\mu$fd. mica.
C_7, C_9 — 100-$\mu\mu$fd. mica.
C_8 — 4–20-$\mu\mu$fd. adjustable ceramic (Erie or Centralab).
C_{10} — 0.0047-μfd. mica.
C_{11} — 0.01-μfd. 400-volt paper.
C_{12}, C_{14} — 25-μfd. 25-volt electrolytic.
C_{13} — 0.1-μfd. 400-volt paper.
C_{15} — 4.7-$\mu\mu$fd. ceramic.
R_1 — 1.8 megohms, ½ watt.
R_2 — 8200 ohms, ½ watt.
R_3 — 1000 ohms, ½ watt.
R_4 — 10 megohms, ½ watt.
R_5 — 68,000 ohms, ½ watt.
R_6 — 50,000-ohm 2-watt potentiometer, preferably wire-wound.
R_7 — 47,000 ohms, 1 watt.
R_8 — 0.5-megohm volume control.
R_9 — 2500 ohms, ½ watt.
R_{10}, R_{11} — 0.1 megohm, ½ watt.
R_{12} — 470 ohms, 1 watt.
R_{13} — 1000 ohms, 1 watt (¼-inch diam.).
L_1 — 2 turns No. 22 d.c.c. wound over ground end of L_2.
L_2 — 4 turns No. 18 enam. wound on National XR-50 form and spaced to occupy ½ inch.
L_3 — 2 turns No. 14 enam., 5/16-inch i.d., spaced 2 × wire diam.
L_4 — 5 turns No. 18 enam., spaced to occupy ⅜ inch, wound on R_{13}.
L_5 — 9 turns No. 22 enam., close-wound.
L_6 — 8 turns No. 22 enam., close-wound, on same form as L_5 and spaced 11/16 inch from L_5.
J_1 — Closed-circuit telephone jack.
RFC_1 — 28 turns No. 30 d.c.c. close-wound on 3/16-inch diam. form. See text.
RFC_2 — 48 turns No. 22 enam., close-wound on ¼-inch diam. form. See text.
RFC_3 — One pie from 4-pie 2.5-mh. choke. See text.
RFC_4 — 80-mh. iron-core r.f. choke (Meissner 19–6846).
S_1 — S.p.s.t. toggle switch.

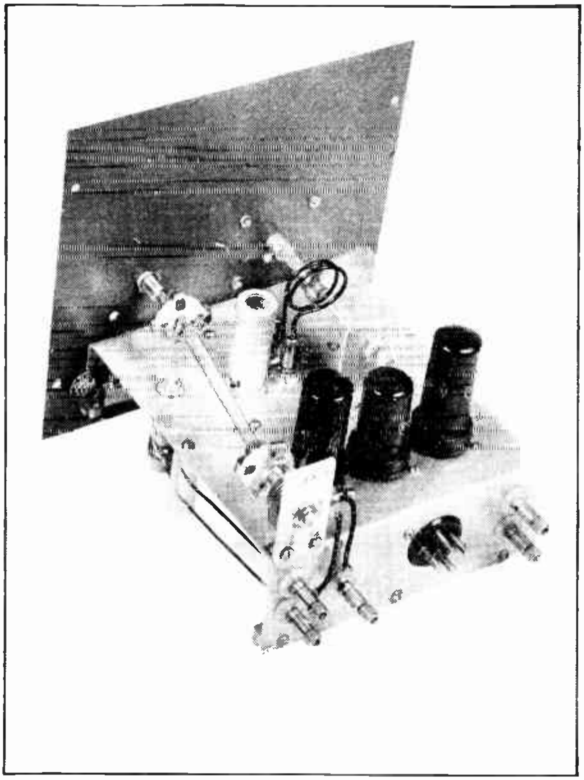


Fig. 1511 — A top view of the receiver shows the construction of the inductive-tuning device used in the oscillator circuit. The tubes along the back, from left to right, are superregenerative second detector, audio and output.

8 × 8 inches. The chassis was bent out of 1/16-inch aluminum and is 6¼ inches wide and 7 inches deep. A 2½-inch lip is bent down at the rear and a 1¾-inch lip is formed at the front. The front bend is made shorter to avoid the lip at the bottom of the cabinet. The chassis is held to the panel by the two potentiometers (regeneration and volume controls) while a ⅜-inch square dural bar bolted to the edge of the 2½-inch lip picks up two screws through the bottom of the cabinet to give a rigid structure.

Bakelite sockets (Amphenol MIP) are used for the octal tubes, and the miniature tube socket is the ceramic one made by Eby. A metal shield to match the socket also acts as a tube lock. The socket is mounted with Pin 5 toward the panel. National FWA binding posts mounted on National XP-6 polystyrene buttons support the oscillator coil, and allow the coil to be changed readily for experimental purposes. The antenna and loudspeaker leads are brought out to similar posts at the rear of the chassis.

The ¼-inch diameter polystyrene rod used for the oscillator tuning vane shaft is supported at the panel end by the National AM dial and at the other by a panel bushing mounted in an aluminum bracket. The vane is made of a piece of thin copper soldered to a brass shaft coupling. After soldering the vane to the coupling, the copper is cut roughly in the form of a straight-line-wave-length condenser rotor plate. It can be trimmed up later to give something resembling straight-line frequency tuning, but this is hardly essential. By moving the vane closer to the coil the tuning range can be increased, and vice versa. The antenna condenser, C_1, is mounted at the rear of the chassis on the bracket furnished with the condenser, and a shaft bearing in the front panel connects to the condenser through an extension shaft and two flexible couplings.

RFC_1 and RFC_2 are wound on 1-megohm resistors. A small notch is filed at each end of the resistor to keep the wire in place, and the wires for the chokes are soldered to the leads of the resistor. A 1-watt size is used for RFC_1 and a 2-watt size for RFC_2. RFC_3 is made by mounting a single pie from a 2.5-mh 4-pie r.f. choke on a 1-megohm 1-watt resistor similar to that used for RFC_1. The easiest way to remove the pies from the ceramic form on which they come is to melt the metal from one end of the choke with a hot soldering iron and then force a sharp ice pick or nail down the hole in the center of the ceramic form until the ceramic splits. The pies can then be removed and one mounted on the resistor with Duco cement.

The i.f. transformer is wound on a National PRE-3 polystyrene form. Two additional small holes, 90 degrees apart, are drilled in the form between the two windings, and one lead of C_8 is snaked through to furnish a support for one end of the condenser as well as a tie-point for one end of L_5 and the isolating resistor R_3. Another hole in the form, below L_5, is used to support one end of R_{13} and serve as a tie-point for C_5 and L_5.

In wiring the receiver, it is convenient to wire the heater circuits first. On the metal tubes, Pins 1 and 2 are grounded to lugs fastened under the screws holding the sockets to the chassis. On the miniature socket a jumper goes from Pin 3 to the central shield of the socket and thence to a lug under one of the screws fastening the socket to the chassis, on the Pin 7 side. Some care should be taken in wiring the r.f. components on the miniature socket, to insure short leads. One connection of R_1, R_2, C_5 and C_6 goes to Pin 7. C_3 mounts between Pin 2 and the binding post supporting the grid side of L_3, and C_4 is mounted from this post to Pin 5. C_7 and C_{10} return to the ground lug for the 6J6 heater circuit mentioned above. A tie-point joins RFC_1 and R_2.

Checking of the receiver is best done by starting at the output and working toward the input. Connect heater voltage and high voltage to check the superregenerative-detector operation. With a speaker or headset connected, advancing the regeneration control should result in the familiar superregenerative hiss. At this point the 105 volts for the mixer and oscillator can be connected, because the adjustment of C_8 should be made with plate voltage on the mixer. With the regeneration control only slightly beyond the point where the hiss starts to be heard, adjust C_8 for the point which requires maximum advancing of R_6 for oscillation. This brings L_6C_8 into resonance with L_5C_5. If it is found that the second detector won't oscillate at one very sharp setting of C_8, the coupling between L_5 and L_6 is too tight. In this event the coils should be backed away from each other, if possible, or else C_8 can be detuned slightly. The former procedure is preferable. The setting of C_8 where the primary circuit pulls the detector out of oscillation should be quite sharp — if it isn't, the setting isn't right. When the detector is oscillating and C_8 is not set properly, it is quite likely that the hiss will also contain some unpleasant high-frequency whistles. The exact frequency of the i.f. can be checked on a calibrated communications-frequency receiver, if desired, but a frequency check is not essential. With the constants given the i.f. will be around 33 Mc.

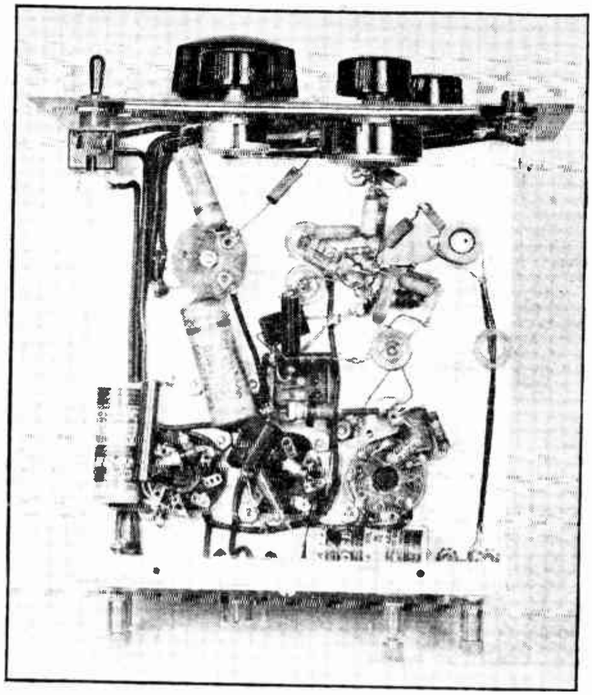


Fig. 1512 — A view underneath the chassis, showing the arrangement of parts. Note the ceramic trimmer condenser between the second-detector socket and the i.f. transformer. This trimmer condenser is adjustable from above the chassis. To the left of the ceramic condenser can be seen RFC_3, the single-pie r.f. choke.

Knowing the i.f. makes it a bit easier to adjust the oscillator portion of the 6J6, because an absorption wavemeter or Lecher wires can be used to put the oscillator on the right frequency. If one knows the i.f. and has some means of checking the oscillator frequency, the oscillator can be adjusted to give a tuning range from 143 Mc. minus the i.f. to 149 Mc. minus the i.f. The tuning range is adjusted by spacing the turns of L_3 and by moving the vane on the shaft. Moving the vane closer to the coil will increase the tuning range but increases the minimum frequency a trifle, and vice versa. If a calibrated 144-Mc. superregenerative receiver or transmitter is available, it can be used as a signal source and the oscillator tuning range can be adjusted without knowing the i.f.

The mixer coil and antenna coupling can be checked by listening to a weak signal (whose weakness is under your control, however), or to ignition noises, and it will be found that best sensitivity will be obtained with fairly tight coupling.

V.H.F. Converters

For the amateur who already possesses a communications-type high-frequency receiver or a good all-wave broadcast receiver capable of tuning to either 5 or 10 Mc., there is no necessity for building a separate v.h.f. receiver, particularly for operation on the 50-Mc. band. It is not only easier but often more satisfactory to build a v.h.f. converter which, in conjunction with the already existing receiver, can be used as a double superheterodyne. This arrangement is particularly successful if the receiver has controllable or broad-band selectivity to permit reception of the less-stable signals on the higher-frequency bands.

The output transformer for such a converter should be designed to tune to an i.f. of either 5 or 10 Mc. (the higher frequency being preferable for operation on bands above 50 Mc.), with a low-impedance secondary. The output from the converter may be coupled through a low-impedance shielded line to the input circuit of the communications receiver, in much the same manner as link coupling is used between stages in a transmitter. The r.f. and mixer circuits of the receiver must be tuned to the same frequency as the output transformer — 5 or 10 Mc. — which then becomes the first i.f. Thereafter the receiver dial remains untouched, all tuning being done with the converter. The volume control, however, will be the gain control on the receiver into which the converter works. A converter may have its own built-in power supply, but with simpler designs it is often possible to draw the filament and plate voltages from the receiver with which the converter is to be used.

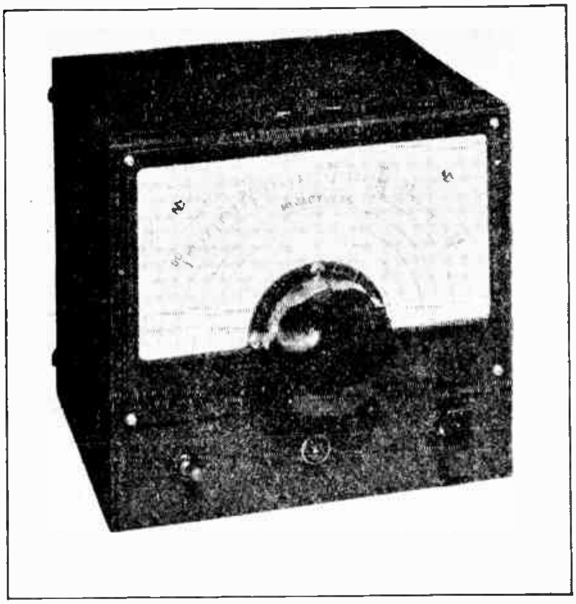


Fig. 1513 — This two-tube 50-Mc. converter incorporates new miniature tubes and obtains its power from the communications receiver with which it is used. The toggle switch at the left cuts the filament circuit when the unit is not in use. The control at the lower right transfers the antenna from the converter to the receiver for normal reception.

◖ A Simple Two-Tube Converter for 50 Mc.

When a high intermediate frequency is used, image rejection is not a problem, and r.f. selectivity in the converter is not particularly important, especially when the converter is used in conjunction with a highly-selective communications receiver. Thus quite satisfactory performance can be obtained without the use of an r.f. amplifier stage. The new high-transconductance miniature pentodes, such as the 6AK5, are excellent as mixers, and a two-tube converter incorporating the 6AK5 in an appropriate circuit will give a degree of performance formerly obtainable only with more complex designs. Such a converter is shown in Figs. 1513–1517. It was designed by Richard W. Houghton, W1NKE, and was described in detail in *QST* for June, 1946. Though it was laid out particularly for use with an HRO it may be used effectively with any communications receiver capable of tuning to 10.5 Mc.

As shown in the schematic diagram, Fig. 1514, the oscillator voltage is injected at the screen grid of the mixer tube. The coupling

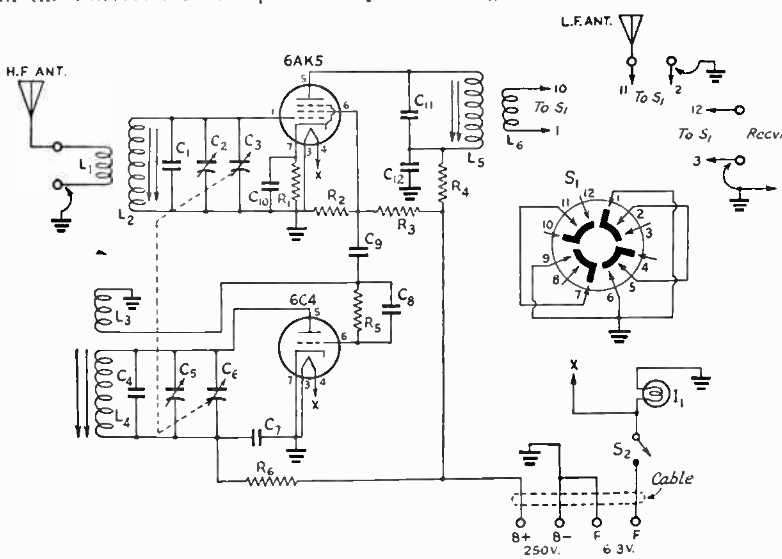


Fig. 1514 — Circuit diagram of the 50-Mc. converter.

C_1 — 15-$\mu\mu$fd. fixed ceramic, zero temp.-coëf. (Erie NPOA).
C_2, C_5 — 2-6-$\mu\mu$fd. ceramic trimmer (Centralab 820-A).
C_3 — 11-$\mu\mu$fd. variable (National UMA-10 with 1 stator plate removed).
C_4 — 12-$\mu\mu$fd. fixed ceramic, zero temp.-coëf. (Erie NPOA).
C_6 — 9-$\mu\mu$fd. variable (National UMA-10 with 1 stator and 1 rotor plate removed).
C_7, C_8, C_9 — 100-$\mu\mu$fd. mica or ceramic.
C_{10}, C_{12} — 47-$\mu\mu$fd. mica or ceramic.
C_{11} — 35-$\mu\mu$fd. fixed ceramic, zero temp.-coëf. (Erie NPOA).
R_1 — 6800 ohms, ½ watt.
R_2 — 1.5 megohms, ½ watt.
R_3 — 0.47 megohm, ½ watt.
R_4 — 0.1 megohm, ½ watt.
R_5 — 22,000 ohms, ½ watt.
R_6 — 10,000 ohms, 1 watt.
L_1 to L_6, inc. — See Fig. 1516.
I_1 — 6.3-volt pilot lamp.
S_1 — 4-pole double-throw switch, preferably with ceramic wafers (Oak Type HC).
S_2 — S.p.s.t. toggle.

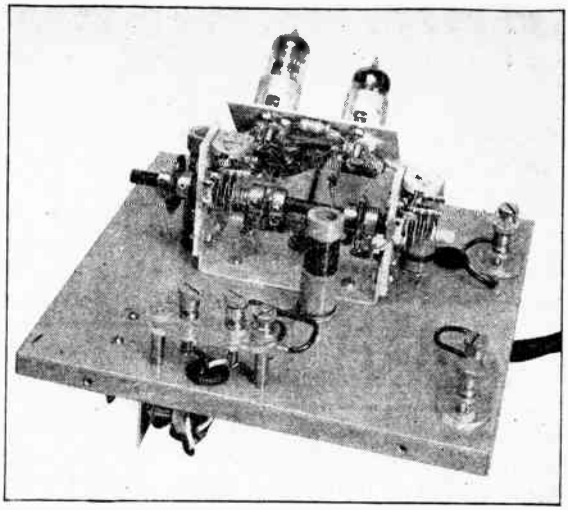


Fig. 1515 — The r.f. construction of the 50-Mc. converter is shown in this above-chassis view. The 6C4 oscillator is at the left and 6AK5 mixer at the right on the subchassis. The 10.5-Mc. i.f. output coil is in the foreground. Flexible ground leads are shown connected to their binding posts in the position normally used for grounded antenna systems.

condenser, C_9, has sufficient capacitance to act as the 6AK5 screen by-pass condenser as well. The grid tank circuit, comprised of L_2 in parallel with C_1, C_2, and C_3, resonates over the operating frequency range, 49.5 to 54.8 megacycles. C_3 is ganged with the oscillator tuning condenser, C_6.

The oscillator operates over a range 10.5 Mc. higher than that of the mixer, and the mixer plate circuit is tuned to this intermediate frequency. With this i.f., the fifth harmonic of the receiver's local oscillator ($10.955 \times 5 = 54.775$ Mc.) appears just outside the high end of the tuning range, sufficiently far from the calibrated band so that it does not interfere with normal operation.

Tracking is easily accomplished over the frequency range under consideration because the percentage of frequency change is small. Starting with two identical tuning condensers (National Type UMA-10), two plates are removed from the one used in the oscillator and one plate from the one in the mixer. Sufficient fixed padding capacitance, using a zero-temperature-coefficient ceramic for low over-all temperature drift, is added to give the required range. The coil forms used are provided with adjustable cores of high-frequency powdered iron, providing an easily-accessible inductance adjustment. Figs. 1515 and 1517 show the layout of these coils.

The wafer-type switch S_1 provides a convenient means of channeling either the converter output or a low-frequency antenna into the antenna terminals of the receiver. When the converter is in use both low-frequency antenna terminals are switched to ground, thus minimizing direct receiver pick-up at the intermediate frequency. Single-wire or doublet antennas may be used at either high- or low-frequency inputs.

When operating the receiver over its normal frequency range, the converter filaments may be turned off by means of switch S_2. This function also could be accomplished by means of an additional wafer on S_1.

A four-prong-to-four-prong adapter, of the sort used for making tube substitutions, is used

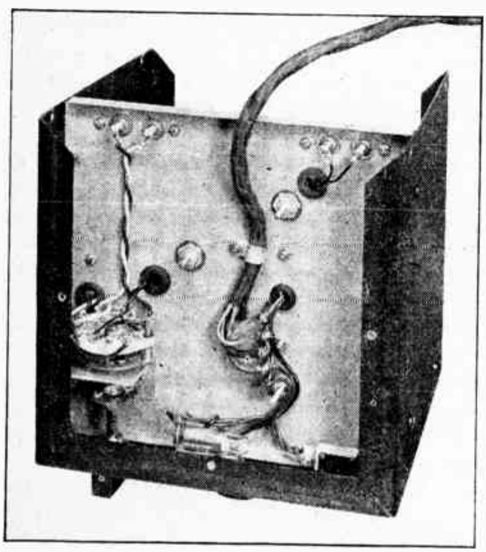


Fig. 1517 — A bottom view of the converter. S_1, the antenna-transfer switch, is at the lower left. Low-impedance antenna leads should be twisted loosely as shown. The three adjusting screws for the iron-core inductances protrude from the chassis on either side of the power cord.

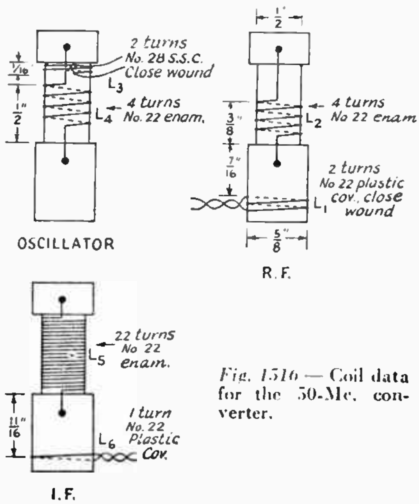


Fig. 1516 — Coil data for the 50-Mc. converter.

on the power cord to enable both it and the receiver cord to be plugged into the HRO power pack simultaneously. With receivers having integral power packs a different arrangement would be required, one possibility being to use a similar plug adapter under one of the power tubes in the receiver, picking up the "B" voltage at the screen-grid pin.

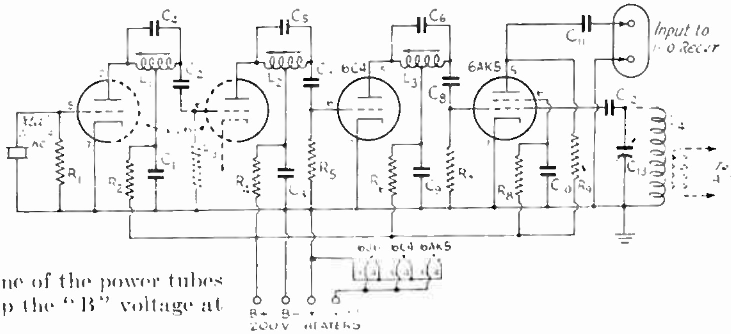


Fig. 1519 — Schematic of the 3-tube 2-meter converter, using a 28-Mc. crystal.

C_1, C_3, C_9 — 470-μμfd. mica.
C_2, C_7 — 100-μμfd. mica.
C_4, C_6 — 15-μμfd. (10 to 20) ceramic or mica.*
C_5 — 22-μμfd. (15 to 25) ceramic or mica.*
C_8 — 2.2-μμfd. ceramic or mica.
C_{10}, C_{12} — 47-μμfd. mica.
C_{11} — 100-μμfd. mica.
C_{13} — 15-μμfd. variable, National UMA-15.
R_1 — 22,000 ohms, ½ watt.
R_2 — 4700 ohms, ½ watt.
R_3 — 0.1 megohm, ½ watt.
R_4 — 4700 ohms, ½ watt.
R_5 — 0.1 megohm, ½ watt.
R_6 — 4700 ohms, ½ watt.
R_7 — 0.25 megohm, ½ watt.
R_8 — 0.75 megohm, ½ watt.
R_9 — 4700 ohms, ½ watt.
L_1 — XR-50 coil form, ungrooved, 11 turns No. 22 enam., close-wound, center-tapped.
L_2 — XR-50 coil form, ungrooved, 5 turns No. 16 enam., spaced ½ dia. of wire, center-tapped.
L_3 — XR-50 coil form, ungrooved, 3 turns copper strip, 3/32 inch wide, spaced 3/32 inch, center-tapped.
L_4 — 1½ turns of No. 14 copper wire, ½ inch in diameter.
* C_4, C_5 and C_6 should be selected in value so that plugs are fairly well out from center of coil at resonance.

ℂ A Crystal-Controlled Converter for 144 Mc.

While most converters are used in the manner described above (by leaving the communications receiver set at a given intermediate frequency and tuning the converter over the desired frequency range), it is quite possible to reverse the procedure, using a fixed-frequency oscillator in the converter and tuning the receiver. This approach is particularly advantageous at 144 Mc. and higher, where the selectivity of the tuned circuits is such that no adjustment of the converter circuits is required when the i.f. (in this case usually a broad-band receiver) is varied over a four-megacycle range.

Several converters employing this principle were described by Calvin F. Hadlock, W1CTW, in the May, 1946, issue of *QST*. The simplest is shown in Figs. 1518, 1519, and 1520. It uses a 6J6 oscillator-doubler, operating with a 28-Mc. crystal, followed by a 6C4 doubler and a 6AK5 mixer, the grid circuit of which is tuned to 146 Mc. and coupled to the antenna. The plate circuit of the mixer is the input circuit of a receiver (see Fig. 1519) which tunes the range between 30 and 34 Mc. The converter was designed for use with the National One-Ten, a superregenerative receiver, but it should provide excellent results when used with any of several a.m.-f.m. receivers which are capable of tuning this range.

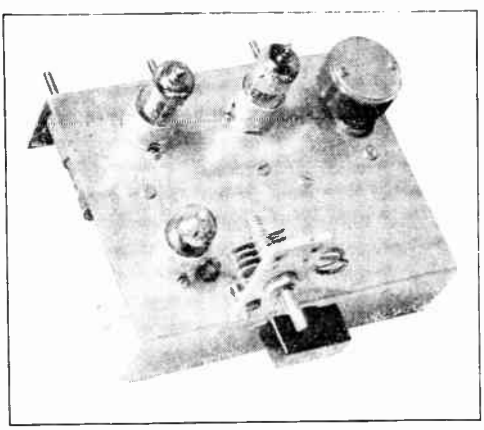


Fig. 1518 — Top view of the three-tube 144-Mc. converter using a 10-meter crystal. Space is provided at the right of the mixer for addition of an r.f. stage.

The model shown in Figs. 1518, 1519 and 1520 is built on a chassis of folded aluminum 6 × 4½ × 1½ inches in size. Space is left on the chassis for addition of an r.f. stage, if desired. The first half of the 6J6 is a conventional triode crystal oscillator, the second half acting as a doubler, driving a 6C4 doubler. With the values shown, the second 6J6 grid will have about 20 volts of excitation, as measured with a high-resistance voltmeter across R_3. The voltage developed across R_5 will be about 25 to 30 volts. The 6C4 doubler provides about 10 volts on the mixer grid before the r.f. input circuit is connected. With the input circuit connected and adjusted to approximately the middle of the 2-meter band the excitation voltage drops to about 1 volt, which is sufficient for good conversion with the grid-leak injection shown. A very high-resistance voltmeter should be used for these measurements. A 100-microampere meter with a 0.5 megohm resistor in series is suitable.

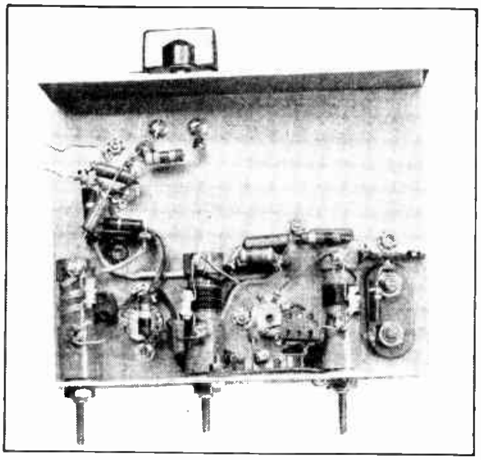


Fig. 1520 — Bottom view of the three-tube 2-meter converter. Note the fixed-tuned tank circuits mounted along the back edge of the chassis. The two short leads at the upper left connect to the antenna terminals of a One-Ten receiver.

This type of converter, used in conjunction with a superregenerative receiver on 30 Mc., will give a degree of performance on 144 Mc. roughly comparable to that of the receiver alone on 30 Mc. With the One-Ten, it discriminates against the poorer oscillator signals, but anything which does not swing more than 200 kc. or so will be received with good quality. The added selectivity afforded by such an arrangement will add greatly to the effectiveness of any station in a locality where there is appreciable activity on 144 Mc.

Mobile Receiving Equipment for 2, 6 and 10 Meters

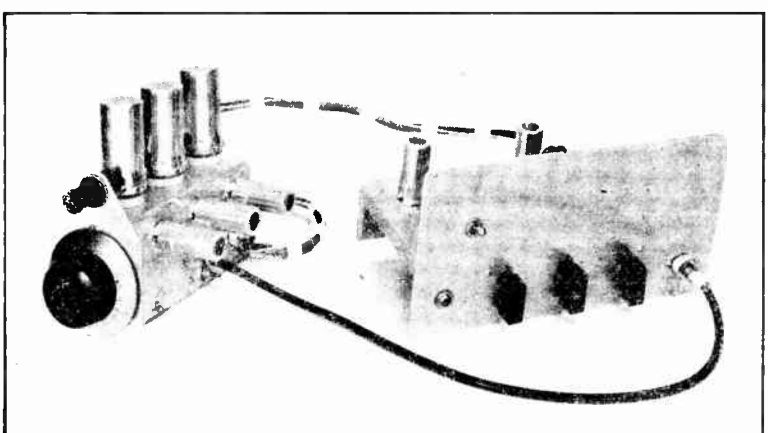
The high sensitivity, noise rejection, and a.v.c. characteristics of the superregenerative detector make it useful in mobile operation. The chief difficulties inherent in this type of receiver, broadness of tuning and radiation of an interfering signal, can be overcome by using a superregenerative stage as the second detector in a superheterodyne receiver. The i.f.-amplifier-and-audio unit shown in Figs. 1521 1524 was designed especially for mobile operation. Two converters, shown in Figs. 1521, 1525–1529, working with this unit, provide mobile reception on 2, 6, 10, and 11 meters. The space available in a particular make of car will influence the form factor of the units, but these are representative designs. The two converters, one for 6–11 meters and one for 2 meters, are intended for steering-post mounting, while the i.f.-audio unit is shaped to fit into a glove or radio compartment.

Little need be said about the i.f. unit, as there are few critical factors, and mechanical layout is relatively unimportant. Only four tubes are used: a 6AG5 11-Mc. i.f. amplifier, a 6C4 superregenerative second detector, a 6C4 first audio amplifier, and 6AK6 second audio. Note that both audio stages are transformer-coupled, this method having been used in preference to resistance coupling, as experience has shown that the former makes for smooth, quiet operation when superregenerative detectors are employed.

The input stage of the unit should be well shielded, not only to prevent oscillation, but to reduce pick-up on 11 Mc. When the unit is installed in a car this is not troublesome, but in home-station work, 11-Mc. interference can become quite severe, especially during evening hours.

The tuned circuits used in the 11-Mc. amplifier, the superregenerative detector, and as output coupling units in the two converters, are all similar. The coils are wound of No. 22 enameled wire on National XR-50 core-tuned forms, the secondary winding occupying the entire winding space. A simple way of securing the primary is to wrap a layer of Scotch Tape, sticky side *out*, around the ground end of the secondary. The primary winding will then stick as it is wound on, and holding it in place will be no problem. A small tab of tape, or

Fig. 1521 — The three-tube converter for 6 and 10 meters connected to the 11-Mc. i.f. amplifier and audio system. The converter is mounted on the steering post, while the i.f. unit is designed for glove-compartment mounting. The object above the converter dial is an adjustable-beam dial light.



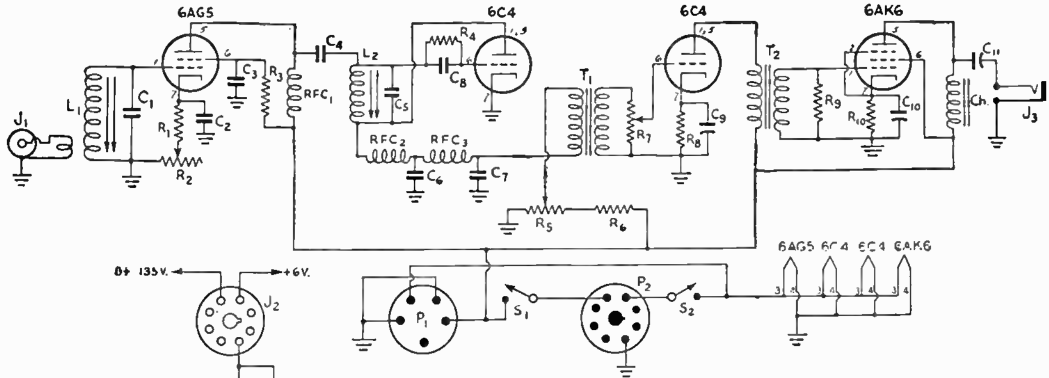


Fig. 1522 — Wiring diagram of the i.f. unit using a super-regenerative second detector and two audio stages.

C_1, C_5 — 47-μμfd. ceramic.
C_2, C_3 — 470-μμfd. midget mica.
C_4, C_8 — 100-μμfd. midget mica.
C_6, C_7 — 0.0068-μfd. mica.
C_9, C_{10} — 25-μfd. 50-volt electrolytic.
C_{11} — 0.1-μfd. 600-volt tubular.
R_1 — 270 ohms, carbon.
R_2 — 10,000-ohm potentiometer.
R_3 — 1000 ohms.
R_4 — 4.7 megohms.
R_5 — 50,000-ohm potentiometer.
R_6 — 47,000 ohms, 1 watt.
R_7 — 0.25-megohm potentiometer.
R_8 — 2200 ohms.
R_9 — 0.22 megohm.
R_{10} — 680 ohms.
All resistors ½-watt type unless otherwise indicated.

L_1, L_2 — 22 turns No. 22 enam., close-wound on National XR-50 form. Primary: 3 turns No. 22 enam. close-wound on layer Scotch Tape over ground end of L_1.
Ch. — Midget filter or audio choke.
J_1 — Coaxial socket (Jones S-201).
J_2 — Octal socket on power cable.
J_3 — Speaker or headphone jack.
P_1 — 5-prong plug for converter power, mounted on back of chassis.
P_2 — Octal plug, mounted on back of chassis.
RFC_1 — 2.5-mh. r.f. choke (National R-100).
RFC_2 — One "pie" from National R-100, mounted on 1-watt resistor.
RFC_3 — 80-mh. r.f. choke.
S_1 — S.p.s.t. toggle switch, bat-handle type.
S_2 — S.p.s.t. switch, mounted on R_7.
T_1, T_2 — Midget interstage transformers.

household cement, will suffice.

The three-tube converter shown in Figs. 1521, 1525, and 1526 covers the 50–54-Mc. and 27–30-Mc. ranges by means of plug-in coils. Using the 11-Mc. intermediate frequency, it is possible to cover the two bands with a common oscillator coil, the oscillator running on the low side of the signal frequency for 50–54 Mc. and on the high side for 27–30 Mc. It is thus merely necessary to change the mixer and r.f. coils when changing bands. Three tubes are used: a 6AK5 r.f. amplifier, a 6AK5 mixer, and a 6C4 oscillator.

The converter layout, shown in Fig. 1525, makes some sacrifices in accessibility for the sake of compactness; however, by planning the construction carefully, the builder should have no trouble in assembling or adjusting the converter. Parts are mounted on an "L"-shaped aluminum chassis, with a cover of the same general shape, making a case which is 2 inches wide, 3 inches high, and 6½ inches long.

Octal sockets for the plug-in coils (Millen 74001 shielded core-tuned forms) are mounted along the top edge, with the corresponding tube sockets projecting from the right side. The oscillator compartment is at the front, nearest the dial — a "must" when flexible couplings are used for ganging. The middle compartment houses the mixer-stage components,

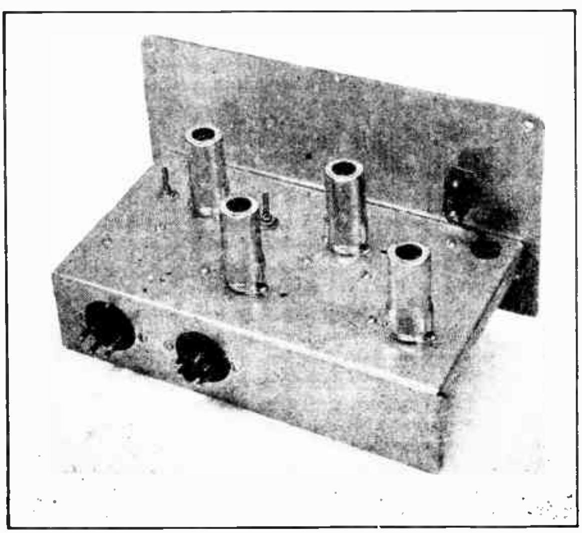


Fig. 1523 — Rear view of the 11-Mc. i.f.-audio unit. The tubes nearest the panel are the i.f. amplifier, left, and the superregenerative detector. The octal plug on the back of the chassis is for the power cable, while the 5-prong plug connects through another cable to the converter. The toggle switch is the B+ stand-by switch.

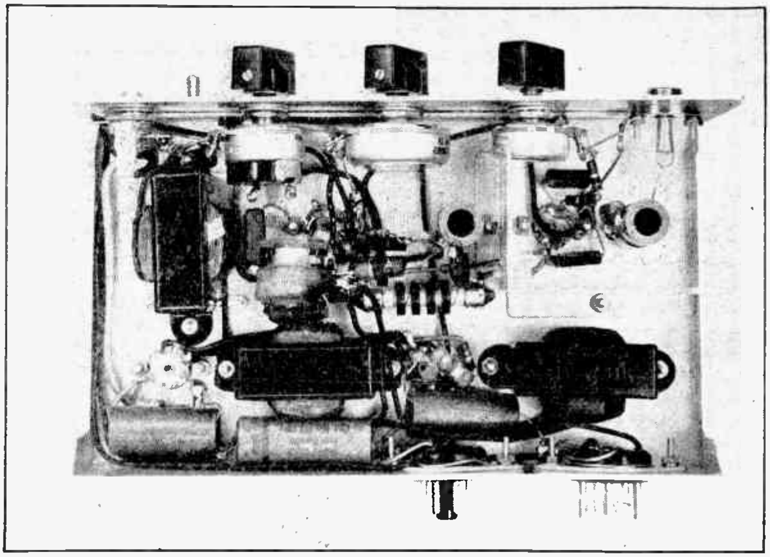


Fig. 1524 — Bottom view of the i.f.-audio unit, showing arrangement of parts. At the upper right, in a partially-shielded compartment, are the parts comprising the i.f.-amplifier input circuit. In the center are the detector socket and associated parts. At the left and rear are the audio components.

including the core-tuned i.f. output coupling transformer. Coupling between the oscillator and mixer is obtained by means of a piece of "push-back" wire which is soldered to the oscillator tuned circuit and then wrapped around the r.f.-plate or mixer-grid lead. The coupling should be set at the lowest value which will provide maximum signal strength. At the back is the r.f. section, which is provided with a coaxial input jack for antenna connection.

As this converter may be used with conventional i.f. systems, provision was made for incorporating a.v.c. Instead of grounding the grid returns from the r.f. and mixer tubes, these returns are brought out, through resistors R_1 and R_5, to a separate pin on the power-cable socket. The corresponding pin in the i.f. unit is connected to ground.

The oscillator circuit is high-C, for maximum stability, the capacity other than that of the variable condenser being supplied by a fixed ceramic padder, consisting of 20-μμfd. and 27-μμfd. units in parallel with the tuning condenser. Adjustable padders are used on the mixer and r.f. circuits to facilitate tracking. These are mica trimmers, to which some may raise the objection of instability, but the coil inductance is adjusted so that the trimmers tune nearly wide open, so that small changes in plate spacing have a negligible effect on the capacity. Tracking is made easy by the adjustable-inductance feature of the coil forms used.

In putting the converter into operation it is best to start by establishing the tuning range of the oscillator, which may be checked with an absorption wavemeter or monitored by a receiver which is capable of tuning from 37 to 43 Mc. It is useful to have the receiver capable of tuning in the high end of the old f.m. band, so the oscillator may be made to hit 37 Mc. or

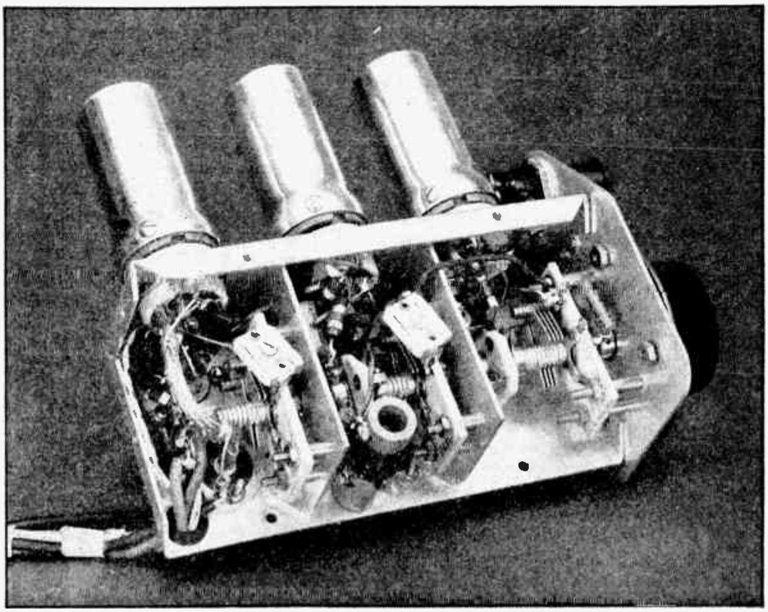


Fig. 1525 — Interior view of the 28- and 50-Mc. converter, with cover removed. The mica trimmers are adjusted through small holes in the chassis cover. The oscillator compartment is at the front (right), the mixer in the middle, and the r.f. amplifier at the left.

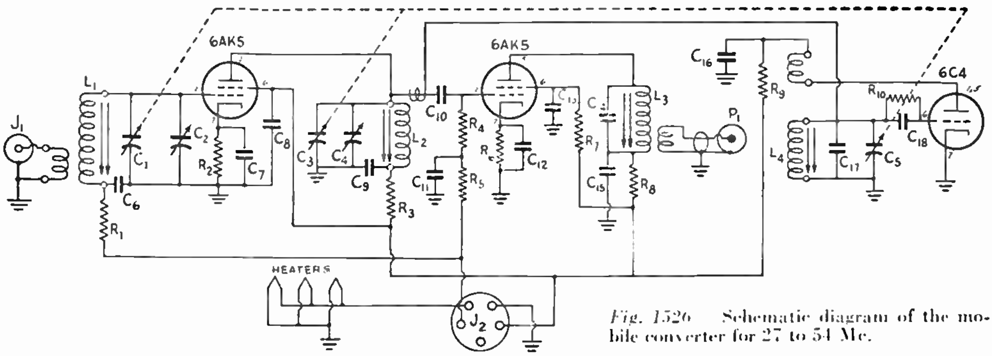


Fig. 1526 Schematic diagram of the mobile converter for 27 to 54 Mc.

C_1, C_3 — R.f. and mixer tuning condensers (National UM-15 reduced to 2 stator and 2 rotor plates).
C_2, C_4 — 3–30-μμfd. mica trimmer.
C_5 — Oscillator tuning condenser (National UM-35 reduced to 1 stator and 4 rotor plates).
C_6, C_7, C_8, C_9, C_{11}, C_{12}, C_{13}, C_{15}, C_{16} — 470-μμfd. midget mica.
C_{10}, C_{18} — 100-μμfd. mica.
C_{14} — 47-μμfd. ceramic.
C_{17} — 47-μμfd. ceramic (20 μμfd. and 27 μμfd. in parallel).
R_1, R_5 — 0.22 megohm.
R_2, R_3, R_8, R_9 — 270 ohms, carbon.
R_4, R_7 — 1.0 megohm.
R_6 — 6800 ohms.
R_{10} — 47,000 ohms. (All resistors ½-watt rating.)

L_1 — R.f. coil. 28 Mc.: 10 turns No. 22 enam., ¾ inch long. Primary: 2 turns No. 28 d.s.c. interwound in cold end of L_1. 50 Mc.: 5 turns No. 22 enam., ⅜ inch long. Primary similar to 28-Mc. coil.
L_2 — Mixer coil. 28 Mc.: 9 turns No. 22 enam., ¾ inch long. 50 Mc.: 4 turns No. 22 enam., ⅜ inch long.
L_3 — I.f. output transformer. 22 turns No. 22 enam., close-wound on National XR-50 form. Coupling winding: 2 turns No. 20 "push-back," wound at cold end of L_3.
L_4 — Oscillator coil. 2¼ turns No. 22 enam., 3/16 inch long. Feed-back winding: 2 turns No. 28 d.s.c. interwound between turns of L_4.
J_1 — Coaxial socket (Jones S-201).
J_2 — 5-prong socket on power cable.
P_1 — Coaxial plug (Jones P-201).

so at the low-frequency end of its range. If the inductance of the coil is properly adjusted, 43 Mc. (oscillator frequency) will come at the high end. This gives a spread of about 70 divisions for the 50-Mc. band, and about 50 divisions for 27 to 30 Mc. If more spread is desired for the 10-meter band, a separate oscillator coil for that band may be made, and additional padder capacitance built into the r.f. and mixer coils for 10 meters.

Once the oscillator is tuning the desired range, the mixer should be put into operation. For test purposes, a temporary primary may be wound on the mixer coil, using two of the spare pins on the coil and socket for bringing out the leads thereto. From here on, a signal generator which tunes the desired frequency ranges is useful, but it is not absolutely necessary. A signal from a VFO, or the harmonics of several crystals, can be made to serve the same purpose. The signal from the oscillator in a communications receiver can be used also. The signal source should be fed into the converter, by direct connection to the temporary primary, or by means of a pick-up antenna, and the output of the converter fed into a communications receiver tuned to 11 Mc. If the converter is working there will be an appreciable increase in receiver noise as the plate voltage is applied to the mixer, and this will increase as the mixer grid and plate circuits are resonated.

Tracking is accomplished in the usual way, except that no squeezing of turns is required for inductance adjustment. With a signal near the high end of the band, adjust the trimmer, C_4, for maximum signal or noise. Tune to near the low end, and recheck the setting of C_4. If the trimmer capacity has to be increased, the coil inductance is low; if the capacity has to be decreased the inductance is too high. Adjust the coil inductance by moving the core (moving the core into the coil increases the inductance) and repeat the trimmer-setting process until the band can be tuned without any readjustment of C_4. When the mixer is functioning properly the same procedure should be followed with the r.f. coil. It is well to note the performance of the mixer alone, as this will serve to determine whether the r.f. stage is performing as it should. There should be a noticeable increase in sensitivity when the r.f. stage is added, but if the mixer is functioning correctly it should be possible to get quite good performance with the mixer alone.

It is well to make all converter adjustments with a communications receiver serving as the i.f., as it is difficult to observe minor changes when the superregenerative detector is used, because of its strong a.v.c. characteristics. The i.f. system should be peaked at 11 Mc. with a signal generator, and then the converter connected to it for an over-all check. The performance, using the superregenerative i.f. unit will be somewhat lower than that of the converter-receiver combination, but it should be possible to copy any signal on the mobile set-up which is solidly readable when the communications receiver is used for an i.f. system.

The circuit of the two-tube 144-Mc. converter, shown in Figs. 1527–1529, is similar to the lower-frequency unit, except that the r.f.

stage is omitted for the sake of simplicity. Even without the r.f. stage, performance well above that of the better superregenerative receivers is obtainable. The 2-meter converter uses a 6AK5 mixer and a 955 oscillator. Because the mixer tuning is fairly broad, no attempt was made to gang the tuned circuits, and only the oscillator is tuned by the vernier dial. The mixer tuning is provided with a front-panel knob, but once set for maximum signal at 146 Mc., it can be left in the same position for both ends of the band with a negligible sacrifice in sensitivity.

From the schematic diagram, Fig. 1528, it may be seen that the circuits of the converters are somewhat similar except for the elimination of the r.f. stage, and the use of a cathode-tapped coil in the oscillator circuit of the 2-meter unit. The converter was originally laid out using a 6J6 push-push mixer, but due to the difficulty of obtaining satisfactory performance with this arrangement, it was changed to the 6AK5. The "butterfly" tuning condenser used is a hangover from the 6J6 set-up — an ordinary Trim-Aire, with its stator sawed in half, would do.

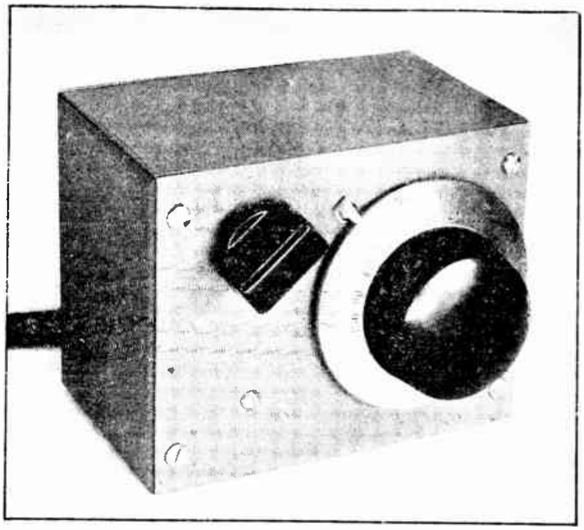


Fig. 1527 — **Front view of the 144-Mc. converter. The entire unit is contained in a standard 3×4×5-inch case.**

All the parts are mounted on the front panel, so that the complete unit can be removed from the case intact. Sections of the folded-over edge of the case were sawed out at several points to provide space for easy removal. The oscillator and mixer assemblies are mounted on individual subpanels of folded aluminum, and most of the wiring can be done before these assemblies are fastened to the front panel. The coaxial socket for the antenna connection is mounted on a separate bracket, and projects through a hole in the back of the case.

Injection of oscillator voltage is accomplished in a manner similar to that used in the other converter, except that a smaller capacity must be used, otherwise the oscillator will "pull out" when the mixer circuit is tuned to resonance. A 4.7-μμfd. ceramic condenser is connected to the hot end of the oscillator tuned circuit, and the coupling lead is run from this condenser to the mixer grid lead. By bringing the two tuned circuits closer together, it would be unnecessary to provide any coupling other than that between the two coils.

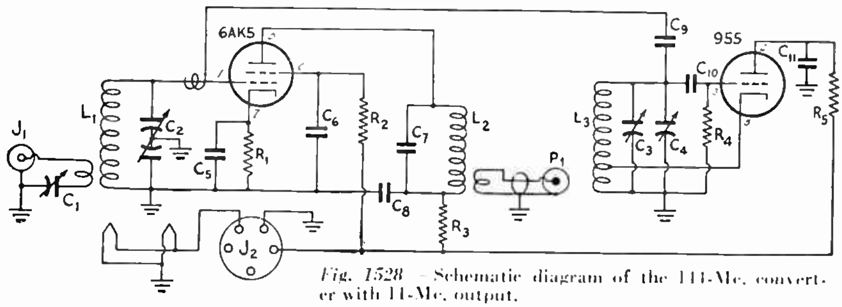


Fig. 1528 — Schematic diagram of the 144-Mc. converter with 14-Mc. output.

C1 — 3–30-μμfd. mica trimmer.
C2 — Cardwell "butterfly" condenser, 1 rotor plate with 1 stator plate on each side. See text.
C3 — 25-μμfd. trimmer with screwdriver adjustment (Millen 26025).
C4 — Oscillator tuning condenser (Millen 20015 reduced to 1 stator and 1 rotor plate).
C5, C6, C8, C11 — 470-μμfd. mica midget.
C7 — 47-μμfd. ceramic.
C9 — 4.7-μμfd. ceramic.
C10 — 100-μμfd. mica midget.
R1 — 10,000 ohms.
R2 — 1.0 megohm.
R3 — 270 ohms.
R4 — 22,000 ohms.
R5 — 10,000 ohms.
All resistors ½-watt carbon.
L1 — 3 turns No. 12 tinned, 3/8 inch long, 3/8-inch inside diameter. Primary: 2 turns No. 20 "push-back" interwound at cold end of *L1*.
L2 — 22 turns No. 22 enam., closewound on National XR-50 form. Coupling winding: 3 turns No. 22 enam. wound on layer of Scotch Tape over cold end of *L2*.
L3 — 3 turns No. 12 tinned, ½ inch long, ¼-inch inside diameter, tapped 1 turn from cold end.
J1 — Coaxial socket (Jones S-201).
J2 — 5-prong socket on power cable.
P1 — Coaxial plug (Jones P-201).

The oscillator tuning condensers C_3 and C_4, are similar mechanically, except that one has a shaft to which is affixed the vernier dial, and the other a screwdriver adjustment. It is important that two similar condensers be used in this arrangement, where the two are mounted at right angles, in order that the stators and rotors line up for direct connection without leads. With the condensers and coil used here, the 144-Mc. band covers about 50 divisions on the dial, permitting coverage up to 150 Mc. This is useful, as commercial signals are available in this range in many locations, and they are quite helpful in making receiver adjustments and in judging the condition of the band.

To do a completely effective job of mobile operation requires considerable attention to noise reduction. With this sort of receiver, the worst interference comes, not from the car's ignition system, but from the generator. The superregenerative detector provides effective silencing for noise pulses of short duration, such as ignition interference, but its inherent a.v.c. characteristics make it respond to a continuous noise such as the whine of the generator, to the exclusion of any weaker signal. It is for this reason that the use of "B" batteries for receiver plate supply is recommended. There is almost certain to be enough noise from any vibrator or generator plate supply to effect at least a slight reduction in the over-all sensitivity of a receiver of this type.

Several types of reception are possible through variation in the setting of the regeneration control. With the plate voltage on the detector near maximum, the loudest "shush" and widest bandwidth are obtained. This is the setting normally used for 144-Mc. reception. Backing off the regeneration control reduces the hiss level and sharpens the response, and best all-around reception on 28 or 50 Mc. is usually obtained in this position. Further reduction of the plate voltage results in a whistle being heard as carriers are tuned in, and quite satisfactory c.w. reception is possible at this setting. From here down, the detector is operating in a condition in between superregeneration and straight regeneration for a considerable variation in the plate voltage. It goes into straight oscillation and then out of oscillation entirely as the voltage is reduced nearly to zero. Reception of modulated signals is possible when the detector is operated in a manner similar to that used with regenerative detectors, and "hiss-less" reception is possible at this point. Sensitivity is considerably lower, however, giving striking proof of the value of superregeneration as a means of attaining high performance with a few tubes.

F.M. I.F. Amplifiers

As was pointed out earlier in this chapter, an f.m. superheterodyne receiver differs from an a.m. receiver mainly in that the pass-band of the intermediate-frequency amplifier must be wider, and in that a limiter and discriminator are used instead of a second detector. The front end of an f.m. receiver usually follows the conventional pattern, and any v.h.f. converter can be used for the purpose if its output frequency is that of the i.f. amplifier.

The f.m. i.f. amplifier employed with the converter may be either the i.f. amplifier of a standard f.m. broadcast receiver or one built especially for the purpose by the amateur himself.

If the i.f. system of an f.m. broadcast receiver is used, the intermediate frequency should first be determined so that the output of the converter can be designed to tune to this frequency and coupled to the grid of the mixer tube of the receiver. If the output transformer in an existing converter does not tune to the required frequency, it is usually feasible to add or remove enough turns from the coil to enable it to be tuned to the receiver i.f. A change in the h.f. oscillator tuning will also be required.

The use of an f.m. i.f. amplifier of this type, in conjunction with a suitable converter, is highly recommended for reception of modulated-oscillator signals such as are common on the 144-Mc. and higher-frequency bands. If the received station holds down its modulation to the point where the signal just fills the pass-band of the i.f. amplifier, best quality and signal-to-noise ratio will be obtained. Under these conditions weaker signals can be received more intelligibly than with the simpler types of receiving systems, and one's receiving range can be extended considerably.

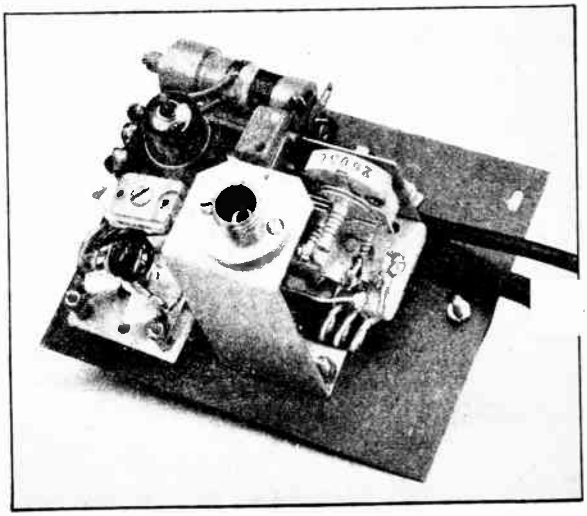


Fig. 1529 — **Back view of the 2-meter converter. Two similar condensers mounted at right angles comprise the tuning assembly for the oscillator in the 2-meter converter.**

V.H.F. Transmitters

Beginning with the v.h.f. region, frequency assignments are no longer in direct harmonic relationship. This fact, coupled with the necessity for extreme care in selection and arrangement of components for low circuit capacitance and minimum lead inductance, makes it highly desirable to construct separate r.f. equipment for v.h.f. work, rather than attempt to adapt for v.h.f. use a transmitter designed for the lower frequencies.

Transmitter stability requirements for 50 Mc. are the same as for the lower-frequency bands, and, by careful attention to component placement, a rig may be made to serve well on 50, 28, and even 14 Mc., but incorporation of 50 Mc. and higher in the usual "all-band" transmitter is not generally feasible.

At 144 Mc. and higher, no restrictions are imposed on transmitter stability, except that the whole emission must be kept within the band limits. This permits the use of modulated-oscillator transmitters, and a large proportion of the stations now working on 144 Mc. and above employ this simple and economical type of gear. By proper choice of tubes and circuits, crystal control is applicable to 144 Mc. however, and the greatly-increased occupancy of the band in metropolitan areas makes stabilization of at least the higher-powered stations almost mandatory, if the full possibilities of the band are to be realized. Crystal control, or its equivalent, may even be employed on 235 and 420 Mc., but the use of these frequencies has not reached the point where stabilization is particularly important.

Throughout the v.h.f. and u.h.f. regions, frequency modulation as well as amplitude modulation is permitted by the amateur regulations. The 300-watt transmitter for 50 and 144 Mc. described in this chapter makes provision for the use of f.m., and any crystal-controlled transmitter can be adapted for f.m. through the addition of a frequency-modulated oscillator to replace the crystal, in the manner described in Chapter Fourteen.

At 420 Mc. and higher, most standard transmitting tubes cannot be employed with any degree of success. Instead, special tubes designed for these frequencies must be employed. Such tubes have extremely close electrode spacing, to reduce transit-time effects, and are constructed with leads having virtually no inductance. Several more-or-less-conventional triode tubes are now available which will operate with fair efficiency up to above 500 Mc., and the disk-seal or "lighthouse" variety will function up to about 3000 Mc.

Above about 2000 Mc. the most useful types of tubes are the klystron and magnetron. These are essentially one-band devices, the frequency-determining circuits being an integral part of the tube itself. Tuning over a small frequency range, such as an amateur band, is possible, usually by warping the cavity employed, but the tubes are not independent of frequency in the conventional sense.

Practically all the recently-opened bands in the ultrahigh and superhigh regions have already seen some pioneering activity, and they offer interesting possibilities to the experimentally-inclined.

A 40-Watt A.M.–F.M. 50-Mc. Transmitter

The transmitter shown in Figs. 1601–1603, inclusive, has an output of approximately 40 watts in the 50-Mc. band and is so designed that either frequency or amplitude modulation may be used. Aside from power supplies, no auxiliary apparatus is needed for f.m. transmission, since the primary frequency control is a variable-frequency oscillator and a reactance modulator is included in the unit. For amplitude modulation, a modulator having an audio power output of about 30 watts is required.

As an alternative to electron-coupled VFO control, provision also is made for crystal control, using a Tri-tet oscillator. As shown in the circuit diagram, Fig. 1602, the crystal oscillator and e.c. oscillator have a common plate circuit, the frequency being doubled in this circuit in both cases. The oscillators are followed by a 6V6 doubler, and this in turn drives the final amplifier, an 815.

The tuned circuits are designed to cover a little more than the range required for the 50-Mc. band so that the transmitter as shown can be used to drive a power frequency multiplier tripling into the 144-Mc. band. The VFO grid circuit tunes from 12 to 13.5 Mc., the range from 12.5 to 13.5 Mc. being used for the 50-Mc. band, and the range from 12 to 12.35 Mc. being available for the 144-Mc. band. When crystal control is to be used, frequencies within the appropriate ranges should be selected, since the oscillator portion of the Tri-tet circuit works over the same frequency range as the grid circuit of the VFO. Appropriate crystals in the 8-Mc. range may also be used, as the 6AG7 triples effectively.

The common oscillator-plate circuit tunes from 24 to 27 Mc., with the 6V6 doubling to 48 to 54 Mc. Either oscillator may be selected by means of a switch, S_{1A-B-C}, which closes the cathode circuit of the desired oscillator. To prevent any possibility of accidental frequency modulation when amplitude modulation is being used, a three-position switch is employed, giving a front-panel choice of crystal or VFO control (for a.m. or c.w.) and VFO control with f.m.

Stability under changes in supply voltage is attained by supplying the VFO screen from a VR-150. This holds the screen voltage at 150 when the plate potential is varied from 150 to 600 volts. The cathode current to the oscillator, measured in J_2, remains practically constant when the plate voltage is varied over this wide range, and the total frequency shift is only a few hundred cycles. With variations in plate voltage which would result from even the most severe line-voltage fluctuations, the frequency shift in the oscillator is only a few cycles.

Other sources of VFO instability are excessive tube and component heating, variations in circuit capacity due to nonrigid mechanical design, and interaction because of improper placement of components. In this design, oscillator input is held to less than half the rated plate dissipation of the tube, keeping drift because of heating to a minimum. All circuit components are mounted below the chassis, away from the heat given off by the metal tubes, and in such position as to prevent interaction so far as possible without extensive shielding. A silvered-mica fixed condenser is used in parallel with the grid coil, and rigid components are used throughout. The result of these precautions is a VFO whose stability compares favorably with that of the associated crystal oscillator.

The transmitter is built on a 10 × 17 × 3-inch chassis, with all components except tubes, crystal and the final-stage output circuit mounted below the deck. Viewing the unit from the top front, the microphone transformer and 6SA7 reactance modulator are at the right front, with the VR-150 at the rear, adjacent to the antenna coupling assembly. The crystal, crystal oscillator, and VFO are grouped near the middle of the chassis, with the doubler and final tubes at the left.

The front panel is a standard 8¾ × 19-inch crackle-finished Masonite unit. The VFO tuning dial is centrally placed, with the oscillator and doubler tuning condensers at the left, and the a.m./f.m. switch and deviation control at the right. The final plate tuning knob is above the VFO dial, at the left, and the swinging-link adjustment is at the right. Jacks, from left to right, are J_4, J_3, J_2 and J_1.

The two wires protruding through the chassis close to the 815 are neutralizing "condensers," labeled C_{X1} and C_{X2} on the schematic diagram. They consist of two pieces of No. 14 enameled wire soldered to the grid prongs of the 815 socket, crossed under the chassis, and brought through the chassis and held in position by two small Isolantite feed-through bushings (Millen 32150).

Adjustment is simple and straightforward. The tuning range of the VFO should be checked first. This may be done with only the two oscillator tubes in place, and the a.m./f.m. switch in the VFO position. The oscillator plate condenser should be tuned for maximum r.f. indication in a neon bulb adjacent to L_2, and the frequency checked in a receiver having a fairly accurate calibration for the region around 12, 24, or 48 Mc.

The size of the VFO grid coil, L_1, is extremely critical, and if some pruning of this coil is to be avoided it would be advisable to make the 50-μμfd. section of C_{10} an adjustable

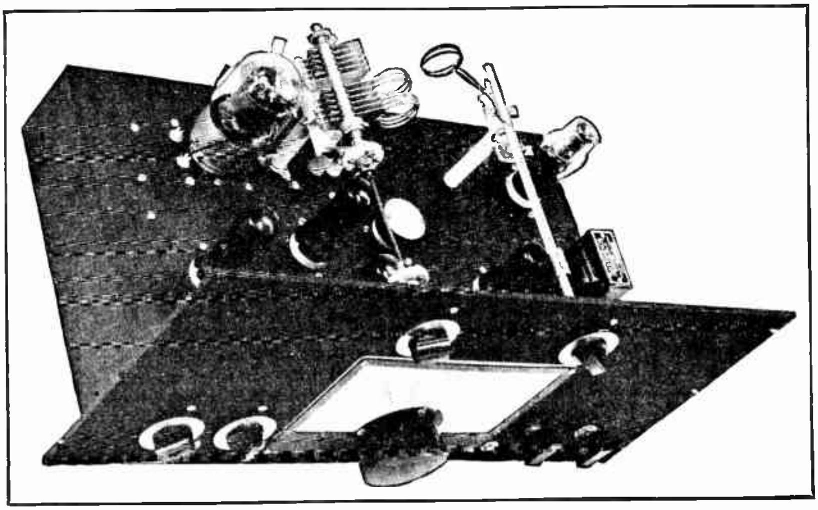


Fig. 1601 — Front view of the 50-Mc. a.m./f.m. transmitter. The r.f. section of the unit occupies the left-hand portion of the chassis. The VR-150, 6SA7 reactance modulator, and microphone transformer are at the right. Note the neutralizing-capacity wires at the left of the 815.

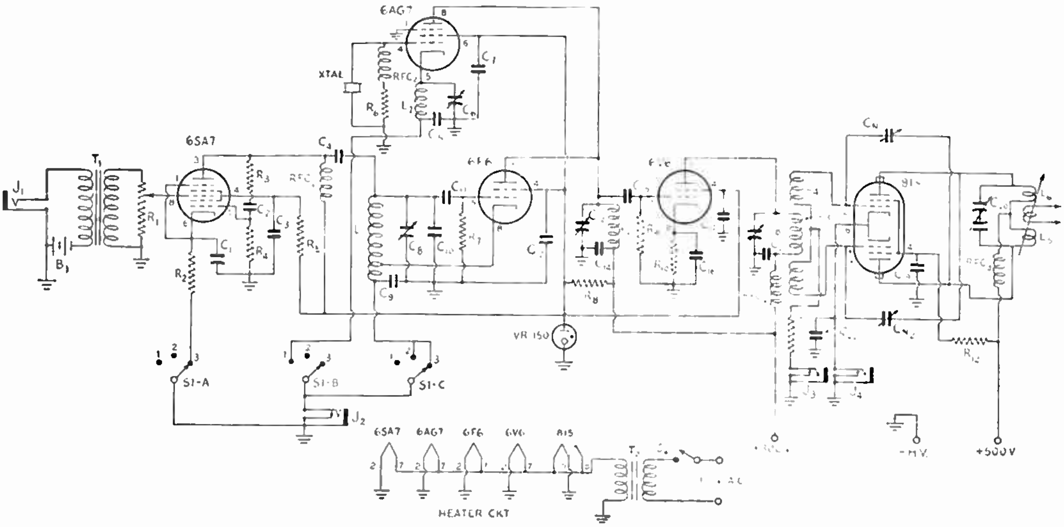


Fig. 1602 — Wiring diagram of a 50-Mc. a.m./f.m. transmitter.

C_1 — 0.01-μfd. 400-volt paper tubular.
C_2 — 0.001-μfd. mica.
C_3 — 8-μfd. 450-volt electrolytic and 0.005-μfd. mica in parallel.
C_4, C_{19} — 470-$\mu\mu$fd. mica.
C_5, C_7, C_9, C_{12}, C_{14}, C_{16}, C_{17}, C_{21}, C_{22} — 0.0022-μfd. mica.
C_6 — 100-$\mu\mu$fd. midget variable, screwdriver adjustment (Hammarlund APC-100).
C_8 — 50-$\mu\mu$fd. variable, "straight-line-frequency" type (Hammarlund MC-50-M).
C_{10} — 100-$\mu\mu$fd. and 50-$\mu\mu$fd. in parallel (Sickles Silver-cap). See text.
C_{11} — 100-$\mu\mu$fd. mica.
C_{13}, C_{18} — 50-$\mu\mu$fd. variable (Hammarlund MC-50-S).
C_{15} — 47-$\mu\mu$fd. mica.
C_{20} — 35-$\mu\mu$fd. per section, split stator (Hammarlund MCD-35-MX).
C_{N1}, C_{N2} — Neutralizing capacity. See text.
R_1 — 0.5-megohm volume control, switch type.
R_2 — 680 ohms, ½ watt.
R_3 — 47,000 ohms, ½ watt.
R_4, R_6 — 0.22 megohm, ½ watt.
R_5 — 4700 ohms, ½ watt.
R_7, R_9 — 0.1 megohm, ½ watt.
R_8 — 5000 ohms, 5 watts.
R_{10} — 220 ohms, 1 watt.
R_{11} — 15,000 ohms, 1 watt.
R_{12} — 15,000 ohms, 5 watts.

L_1 — 8 turns No. 18 tinned, ¾-inch diameter, 1-inch length, on National PRF-2 form. Tapped 2 t. from ground end.
L_2 — 10 turns No. 14 e., ½-inch diameter, spaced one diameter, air-wound.
L_3 — 4 turns No. 14 e., ½-inch diameter, spaced one diameter, air-wound.
L_4 — 5 turns each section, No. 14 e., ½-inch diameter. Adjust spacing for best coupling. See text.
L_5 — 3 turns each section, No. 12, tinned, 1⅛-inch diameter, spaced one diameter.
L_6 — 2 turns No. 14 e., 1-inch diameter, swinging link. See photos and text.
L_7 — 35 turns No. 24 d.c.c., close-wound on 9/16-inch diameter form (National PRE-3).
B_1 — Microphone battery (Burgess).
J_1 — Open-circuit jack.
J_2, J_3, J_4 — Closed-circuit jack.
RFC_1, RFC_2, RFC_4 — 2.5-mh. r.f. choke (National R-100).
RFC_3 — 2.5-mh. r.f. choke, end-mounting (National R-100-U).
$S_{1A\,B\,C}$ — 3-position 3-contact rotary switch (Mallory).
S_2 — Switch on deviation control, R_1.
T_1 — Single-button microphone transformer (Thordarson T-83A78).
T_2 — 6.3-volt 4-amp. filament transformer.

padder condenser, such as a Hammarlund APC-50, which can then be adjusted until 12 Mc. appears at about 90 on the VFO vernier dial. The high-frequency limit, 13.5 Mc., should then come at approximately 10, giving a spread of about 18 divisions for the 144-Mc. band and 54 divisions for the 50-Mc. band. Without such a variable condenser, the number of turns on L_1 must be adjusted by cut-and-try until the proper tuning range is secured. In either case, the final adjustment of band coverage should be made with the 6SA7 reactance modulator in its socket so that its plate-to-ground capacity will be across the tuned circuit.

Operation of the crystal oscillator may next be checked. With a 100-ma. meter connected through J_2, and the a.m./f.m. switch in the "crystal" position, adjust the crystal-oscillator cathode tuning, C_6, until the current dips sharply, indicating oscillation. This control should be set at the point which gives the lowest cathode current consistent with easy crystal starting. Cathode current should be similar for both oscillators — about 20 ma.

The doubler stage may next be tested by installing the 6V6 and 815 tubes, leaving the plate power off the 815. A meter having a 10-ma. range should be used to measure the grid current in the 815, at J_3. The current should come up to about 6 ma. when the spacing between L_3 and L_4 is optimum, though this is more than is actually needed for satisfactory operation of the 815.

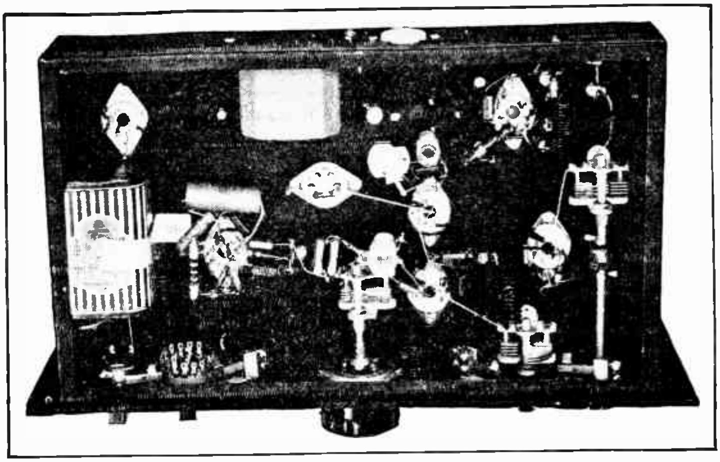


Fig. 1603 — Under-chassis view of the 50-Mc. a.m./f.m. transmitter. At the lower center are the VFO grid coil and associated components. Over these are the crystal and cathode circuit for the 6AG7 crystal oscillator. At the upper right are the inductively-coupled doubler plate coil and final grid coil. The coil and condenser at the lower right comprise the plate circuit which is common to both oscillators. The doubler plate tuning condenser is at the far right.

Next the position of the neutralizing wires can be adjusted. The 815 plate tuning condenser, C_{20}, should be rotated slowly, meanwhile watching the grid current for any variation. The position of the neutralizing wires should be adjusted until there is no sign of fluctuation in grid current as the tuning condenser is rotated. A length of wire extending about one inch above the metal ring on the 815, at a position about ⅛ inch from the glass envelope, should be sufficient. If this should be inadequate, small tabs of copper or brass can be soldered to the ends of the wires to make additional capacity to the tube plates. The neutralizing capacity is necessary in order to ensure completely stable operation.

After neutralization, power may be applied to the 815 plates, while noting the cathode current as indicated on a 200-ma. meter plugged into J_4. The dip at resonance should bring the current to about 50 ma. with no load. A 25-watt lamp connected across the swinging link terminals should then give a full-brilliancy indication when the link is adjusted for maximum coupling. This is with 500 volts applied, which should be used only after it has been determined that everything is functioning properly. If trouble is encountered, further tests should be made with reduced voltage to avoid damaging the tube.

When the transmitter is put on the air, the full 500 volts at 150 ma. may be used for f.m. or c.w. operation. For plate modulation, the voltage should be reduced to about 400 for maximum tube life, even though the tube plates may show no color at the higher voltage.

For frequency modulation, the 6SA7 reactance modulator provides the simplest possible means of obtaining the desired swing in frequency. It may be operated with a single-button microphone plugged into J_1, or the modulator may be driven from a speech amplifier and crystal or dynamic microphone. The output of the speech amplifier should then be connected across potentiometer R_1, and T_1 may be omitted. In either case, R_1 serves as a deviation control, the swing being adjusted to suit the receiver at the station being worked.

In addition to the filament transformer, T_2, indicated in the circuit diagram, the transmitter requires two plate power supplies. One, for the 815, should have an output of 400 to 500 volts at 175 ma.; the other, for the remaining tubes, should deliver 300 volts at approximately 100 milliamperes.

300-Watt Driver-Amplifier for 50 and 144 Mc.

A companion high-power driver-amplifier for the 50-Mc. transmitter described in the preceding section is shown in Figs. 1604 to 1607, inclusive. The amplifier uses a pair of 35-TG tubes in push-pull while the driver, a frequency tripler used for 144 Mc. only, is a single 35-TG. If operation on 144 Mc. is not desired the driver may be omitted, in which case everything to the left of terminals *B-B* in the circuit diagram, Fig. 1606, may be ignored.

Looking at the front-panel view, the two large dials are the plate tuning controls for both stages. The small dial at the left controls the swinging link, the center dial is the grid tuning control for the final stage, and the one at the far right is the tripler grid tuning control. All parts are mounted well back from the panel, and Lucite rods are used for extension shafts.

The rear view shows the general placement of parts. At the left, attached to the back of the 7 × 17 × 3-inch chassis, is the jack bar containing terminals *A-A* and *C-C*, into which the link from the exciter is plugged to furnish drive for either the tripler or final. The tripler grid coil, L_1, is just above the link socket, with the plate condenser, C_5, and coil, L_2, for this stage between the tube and the front panel. The link between L_3 and L_2 is a plug-in affair, and its socket (which is a mechanical mounting only) is between the tripler plate and final grid condensers. Between the grid tuning condenser and the final tubes are the ganged neutralizing

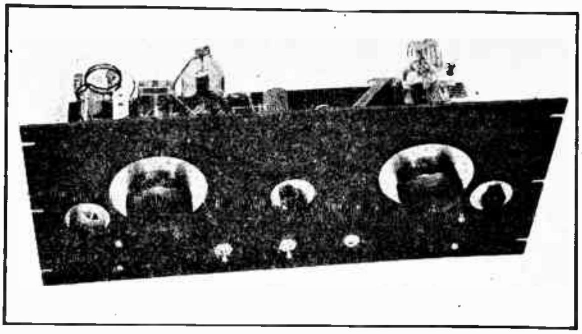


Fig. 1604 — Front view of the 300-watt driver-amplifier for 50 and 144 Mc. The two large dials are the plate tuning controls. The small dial at the left adjusts the position of the output coupling link, the center dial is the grid tuning control for the final, and the third small dial is the tripler-grid tuning control. Across the lower center are the filament switches and grid-current meter jack.

condensers. These are triple-spaced midget condensers mounted back-to-back with coupled shafts. The final tank condenser is mounted as closely as possible to the two tubes, at the right. The jack bar for the final plate coil and the homemade swinging link assembly are at the far right. All components are mounted as close together as possible without being so crowded that tubes cannot be removed from the sockets.

When the amplifier is to be used on 50 Mc. the switch S_1 is left open so that the filament of the tripler will not light when S_2 is closed. The link from the exciter is plugged into terminals *C-C* in the jack bar, which is a Millen Type 40205 coil socket. The output of the exciter is thus connected to the link terminals on the final grid-coil socket, L_3, which is a National Type XB-16. The plug-in link is left out of its socket, *B-B*, which is a Millen Type 33002 crystal socket mounted on a small cone stand-off.

For operation on 144 Mc., switch S_1 is closed, lighting the filament of the tripler tube. The exciter link is inserted at terminals *A-A* on the link jack bar, coupling the exciter to the tripler grid coil, L_1. The plug-in link which transfers the energy from L_2 to L_3 is inserted in its socket, and 144-Mc. coils are inserted in the sockets for L_3 and L_4.

In order to eliminate the stray capacitance and inductance usually encountered in any plug-in base, the 144-Mc. coils for L_3 and L_4 are made to plug directly into their respective sockets. The grid coil, being of No. 12 wire, fits the socket contacts; the plate coil is fitted with pins removed from an old tube base or plug-in coil form. For the same reason, the plug-in link terminals on the L_3 coil socket are not used for 144 Mc.

The final-stage plate tank condenser is made from a Cardwell dual neutralizing condenser, which originally had an insulated flexible coupling between the two rotor sections. This was removed and a section of ¼-inch brass rod, tapped for 6/32 thread, was inserted in its place. A piece of ⅛-inch thick Lucite was fitted to the bottom of the condenser assembly and serves as a mounting base. The result is a split-stator condenser which has sufficiently wide spacing to eliminate the danger of flash-over, yet is extremely compact.

There is really no necessity for a plug-in coil at L_1, inasmuch as it is never changed, but it was employed to permit the use of a standard commercial unit. Two turns were removed from one end, making it essentially an end-linked coil. The same type of coil (National AR-16, 10-C) assembly is used for the 50-Mc. coil for L_3. One turn was removed from each end in this case, a center-linked assembly being needed at this point.

Meters should be provided for reading the tripler plate, final grid, and final plate currents,

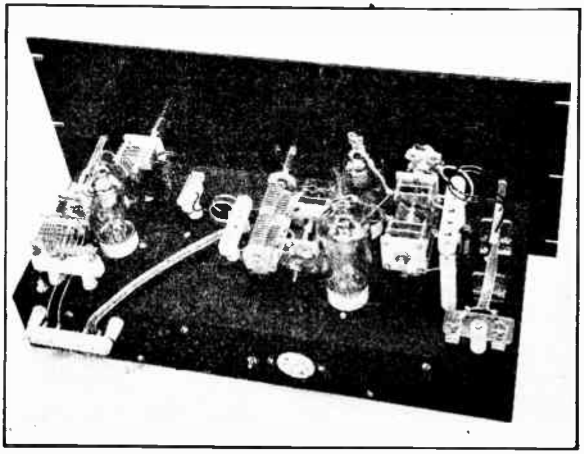


Fig. 1605 — Rear view of the v.h.f. amplifier unit with 144-Mc. coils in place. All components are grouped for minimum lead length. Lucite rods are used for extension shafts on all tuning controls. Note the plug-in link between the tripler plate coil and the final grid circuit. Flexible links, for the final grid- and output-coupling circuits, are low-loss 300-ohm line (Amphenol 21-056).

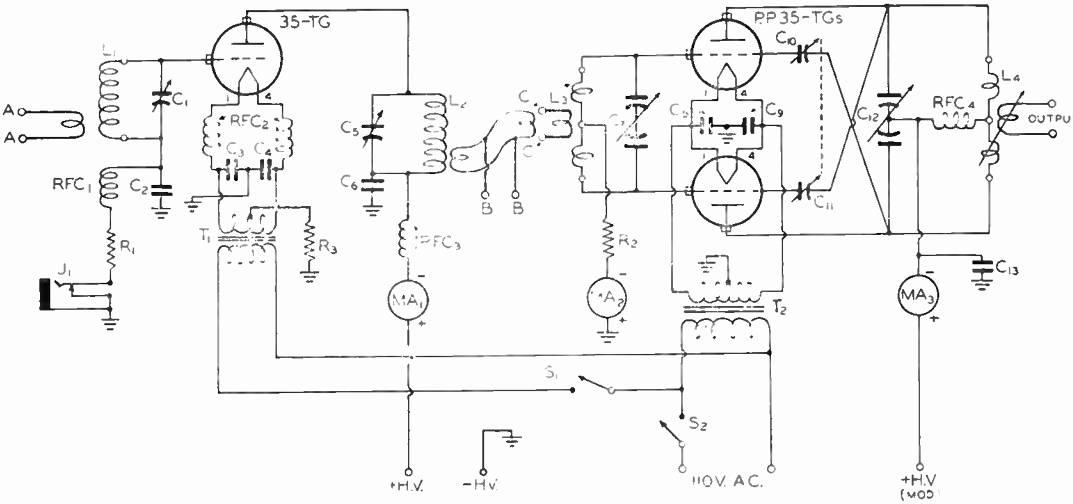


Fig. 1606 — Schematic diagram of the 50–144-Mc. driver-amplifier using 35-TGs.

C_1, C_5 — 15-μμfd. variable (Hammarlund HFA-15-E.)
C_2, C_3, C_4, C_8, C_9 — 0.001-μfd. mica.
C_6, C_{13} — 0.0005-μfd. 5000-volt mica.
C_7 — 15-μμfd. per section, split stator (Hammarlund HF-15-X).
C_{10}, C_{11} — Neutralizing condensers (Cardwell Trim-Aire, 2 plates, triple spacing).
C_{12} — 4-μμfd. per section, split stator (Cardwell ED-4-DI). See text.
R_1 — 50,000 ohms, 10 watts.
R_2 — 3000 ohms, 10 watts.
R_3 — 250 ohms, 10 watts.
L_1 — 6 turns No. 18, 1¼-inch diameter, 1 3/16 inches long, 3-turn end link (National AR-16, 10-C, with two turns removed from one end).
L_2 — 2 turns No. 14 e., 1-inch diameter, spaced ⅛ inch. Link, L_2, L_3 — 2 turns No. 14 e., each end. Plug-in device is for mechanical mounting only.
L_3 — 50–60 Mc. — Same as L_1, but with one turn removed from each end of the original unit. 144 Mc. — 2 turns, No. 12 tinned, ¾-inch diameter, spaced ½ inch. No plug-in base is used — coil leads plug directly into socket.
L_4 — 50–60 Mc. — 3 turns each side of center, No. 12 tinned, 2-inch diameter. Adjust turns spacing so that low-frequency end of range comes with tuning condenser at maximum capacity. Base is a Millen Type 40205 midget plug. 144 Mc. — 1 turn each side of center, No. 12 tinned, spaced to fit holes in jack bar (Millen Type 41205 midget socket). Pins for this coil may be removed from an old tube base or plug-in coil form.
J_1 — Closed-circuit jack.
MA_1 — 0–150 ma.
MA_2 — 0–50 ma.
MA_3 — 0–300 ma.
RFC_1, RFC_4 — V.h.f. r.f. choke (Ohmite Z-1).
RFC_2 — 10 turns No. 14 e., self-supporting, close-wound on ⅜-inch diameter.
RFC_3 — V.h.f. r.f. choke (Ohmite Z-0).
S_1, S_2 — S.p.s.t. toggle switch.
T_1 — Filament transformer, 5 volts, 4 amperes.
T_2 — Filament transformer, 5 volts, 8 amperes.

as indicated in the circuit diagram, although these meters are not included in the unit itself. The jack on the front panel is for a meter for measuring the tripler grid current, and is normally used only during initial tuning operations.

The final stage should be tuned up on 50 Mc. first. The exciter link should be plugged into terminals *C-C* on the jack bar, and the 50-Mc. coils inserted at L_3 and L_4. With power on the exciter but no plate voltage on the amplifier, rotate C_7 for maximum grid current. Set the neutralizing condensers at maximum capacity and rotate C_{12}. If the final-stage plate circuit is capable of being tuned to resonance there will be a pronounced dip in the grid current.

The neutralizing condensers, C_{10} and C_{11}, should then be adjusted a small amount at a time until the dip in grid current disappears. Power may then be applied to the plate circuit. If everything is in order, the dip in plate current at resonance should bring the plate current down to less

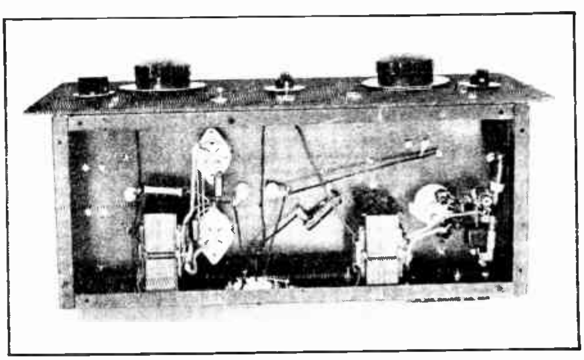


Fig. 1607 — Underchassis view of the 35-TG driver-amplifier. Separate filament transformers are used for the two stages. The driver-tube socket and the two filament r.f. chokes are at the right.

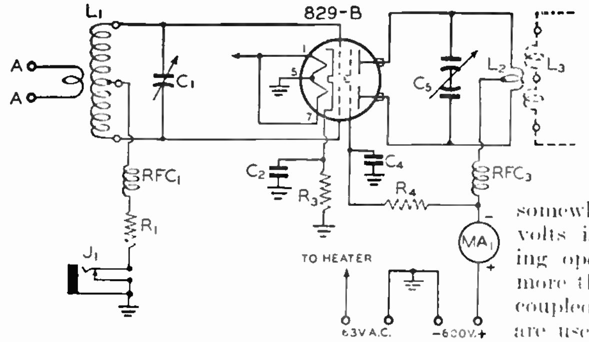


Fig. 1608 — Alternate tripler stage to replace the 35-TG shown in Fig. 1606. Components are the same with the following exceptions:

C_5 — Split-stator midget (Hammarlund HFD-15X).
R_4 — 10,000 ohms, 10 watts.
L_1 — 6 turns No. 18, 1¼-inch diam., 1⅜ inches long, 3-turn center link. (National AR-16, 10-C, with 1 turn removed from each end.)
L_2 — 2 turns No. 12 enam., ⅝-inch diam., center-tapped, coupled inductively to L_3.

than 50 ma. The amplifier may be loaded up to nearly 300 ma., at a plate voltage of 1500 — an input of 425 watts or more — before the plates of the 35-TGs show more than their normal bright-orange color.

Next, tripler operation should be checked. With the exciter on 48 Mc. and the link inserted in the terminals *A-A*, adjust C_1 for maximum grid current. This should be around 20 ma. when no plate voltage is applied to the tripler. For initial tests 750 volts is sufficient — the maximum voltage should not be used until everything is in order. Apply the plate voltage and tune C_5 for resonance, which should occur near minimum capacity.

When it has been determined that the output is actually the third harmonic, or 144 Mc., insert the plug-in link at *B-B* and the coils for 144 Mc. at L_3 and L_4. Repeat the process of checking the final stage as outlined above for 50 Mc. Some change in the setting of the neutralizing condensers may be required for complete neutralization at 144 Mc. (the setting for this band is much more critical than for 50 Mc.), but the adjustment for 144 will usually be found to be satisfactory for the lower frequency as well.

Tests on 144 Mc. should be conducted at a lower voltage than is used for 50 Mc. Up to 2000 volts may be used at the lower frequency after everything is tuned up, but with the somewhat lower efficiency at 144 Mc., 1300 volts is the recommended maximum. Tuning operations should be conducted at not more than 1000 volts. A load should be kept coupled to the final stage when high voltages are used, otherwise the circuit losses at this frequency will cause sufficient tank-circuit heating to melt soldered connections.

Circuit losses make the dip in plate current high (about 100 ma. at 1000 volts) at 144 Mc., but the resonance dip is not a true indication of performance. Lamp loads, too, are unreliable at this frequency. The best test is the color of the tube plates. If the color does not indicate greater heat than is shown when 150 watts input is run with no excitation, then there is no cause to worry about harming the tubes.

◖ Alternate Tripler Stage Using an 829

A more efficient tripler stage, for use in driving the amplifier on 144 Mc., is shown in Fig. 1608. It may be used in place of the 35-TG tripler shown in Fig. 1606, or as a source of excitation for any 144-Mc. amplifier in the medium-power class. It employs an 829 or 829-B, and uses most of the components of the 35-TG stage it replaces. Best transfer of energy to the amplifier stage is obtained with direct inductive coupling of L_2 and L_3, dispensing with the plug-in link shown in Fig. 1606.

By driving the tripler stage very hard it is made to operate at quite good efficiency, the output, with 600 volts on the plates, being nearly 40 watts. Grid current, under load, is about 10 ma. through the 50,000-ohm grid resistor. At 125-ma. plate current, the 829 tripler will provide a satisfactory amount of grid drive for the push-pull 35-TG final stage.

Fig. 1609 — Rear view of the 100-watt 144-Mc. transmitter. The 815 tripler is at the center of the chassis, with the two preceding tripler stages at the right. The final stage is assembled on a separate "L"-shaped chassis, to permit substitution of an alternate arrangement. Note the very close coupling between the tripler-plate and final-grid tanks.

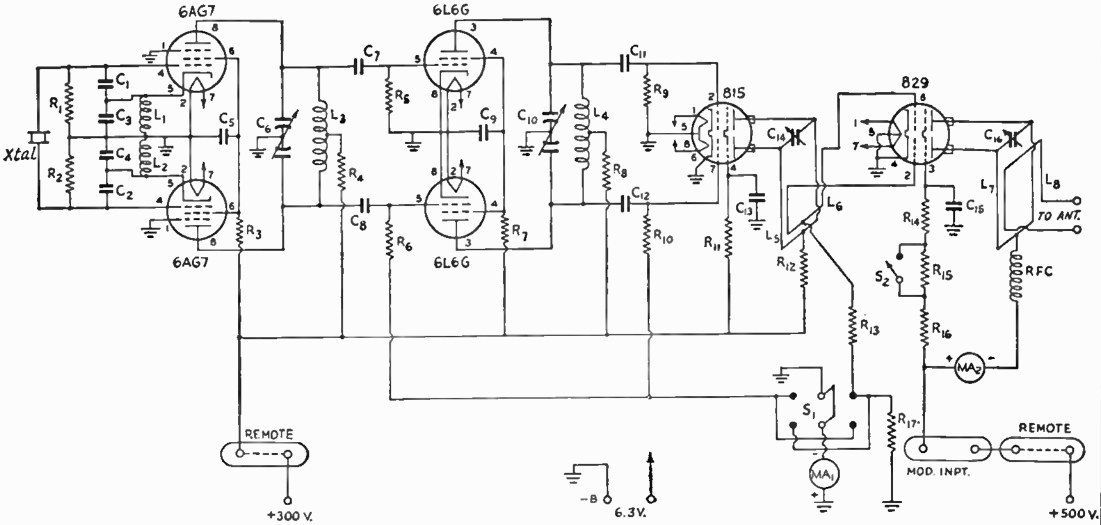


Fig. 1610—Circuit diagram of the 144-Mc. crystal-controlled transmitter.

C_1, C_2—4.7-μμfd. ceramic, 500 volts.
C_3, C_4—100-μμfd. mica, 500 volts.
C_5, C_9, C_{13}, C_{15}—0.001-μfd. mica, 500 volts.
C_6, C_{10}—50-μμfd.-per-section variable (National STD-50).
C_7, C_8, C_{11}, C_{12}—220-μμfd. mica, 500 volts.
C_{14}, C_{16}—10-μμfd. variable (National UMA-10; See text).
R_1, R_2—0.22 megohm, ½ watt.
R_3—15,000 ohms, 2 watts.
R_4, R_8, R_{12}—220 ohms, 2 watts.
R_5, R_6, R_9, R_{10}—0.1 megohm, ½ watt.
R_7—33,000 ohms, 2 watts.
R_{11}—10,000 to 20,000 ohms, 4 watts.
R_{13}—4700 ohms, 1 watt.
R_{14}—5000 ohms, 5 watts.
R_{15}—0.1 megohm, 2 watts.
R_{16}—5000 to 10,000 ohms, 5 or 10 watts.
R_{17}—Resistance equal to grid-meter resistance, if 0–10 milliammeter is used.
L_1, L_2—750 μh. (See text).
L_3—20 turns No. 22 on ½-inch dia. form (National XR-50), close-wound, center-tapped.
L_4—7 turns No. 16 on ½-inch dia. form, length ⅝ inch, center-tapped.
L_5, L_7—See Figs. 1612 and 1613.
L_6—See text.
L_8—See text.
MA_1—0-10 milliammeter (0-20 ma. may be used, in which case R_{17} is not required).
MA_2—0-300 milliammeter.
RFC—40 turns No. 26 close-wound on ¼-inch dia. form.
S_1—D.p.d.t. toggle switch.
S_2—S.p.s.t. toggle switch.

ℂ A 100-Watt 144-Mc. Transmitter

A crystal-controlled transmitter for 144 Mc. need not be especially complicated, particularly if the rig is designed for one-band operation. Figs. 1609–1614 show a simple easily-constructed transmitter which is capable of delivering up to 100 watts of power at 144 Mc., with only slightly more apparatus than would be required for similar output on a lower frequency. It was designed by Calvin F. Hadlock, W1CTW.

The oscillator stage uses two 6AG7s in push-pull, with a crystal in the range between 5.33 and 5.48 Mc. The plate circuit is tuned to the third harmonic of the crystal frequency, driving a pair of 6L6Gs operating as frequency triplers. This stage, in turn, drives an 815 tripler, the output of which is on 144 Mc. The final stage may be either an 815 or an 829, and examples of construction are shown for both tubes. Output and efficiency will be considerably higher with the 829, and it is somewhat more stable in operation and easier to drive.

By making provision for more than adequate driving power, capacity coupling is permitted in the exciter stages, and a nonresonant grid circuit can be used in the final stage, which makes for completely stable operation without neutralization. Similar tank circuits of novel design (see Figs. 1612 and 1613) are used in the two 144-Mc. plate circuits. The grid circuit of the final stage consists of an untuned "U"-shaped loop which is tightly coupled to the plate circuit of the 815 tripler stage. As the resonant frequency of this grid circuit is much higher than 144 Mc., there is no tendency to oscillation.

The coil and condenser values for the crystal-oscillator grid circuit are not particularly critical, but should be adjusted roughly for optimum oscillation. Pie-wound 2.5-mh. chokes might be used for the coils by removing one or two of the pies. The use of the two 4.7-μμfd. condensers connected between the cathode and grid of each 6AG7 assures strong oscillation, even with sluggish crystals. With the constants given, the crystal will always oscillate, regardless of the setting of the various controls.

The plate circuits of the first two stages are tuned with receiving-type split-stator condensers, the rotors of which are grounded. The

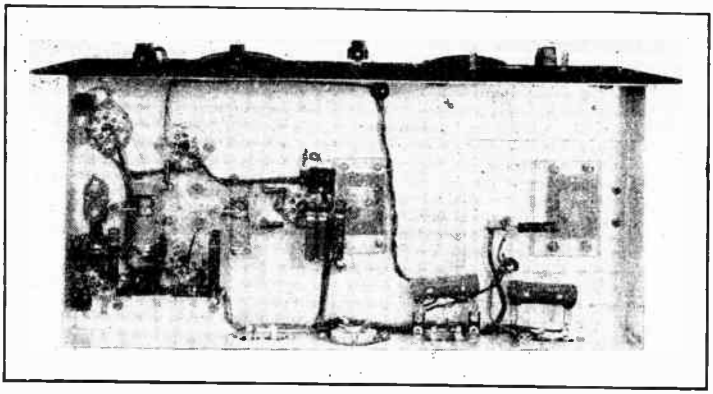


Fig. 1611 — Bottom view of the 144-Mc. crystal-controlled transmitter, showing the simplicity of the layout. The two "U"-shaped plate tanks are mounted on blocks of polystyrene. Oscillator and tripler plate coils are at the left.

coils are wound on ungrooved coil forms (National XR-50, with the core removed). Any ½-inch diameter form could, of course, be substituted. Coils are mounted below the chassis, with the tuning condensers above.

The 815 and the 829 plate circuits are made of 1/16-inch copper strip, ¾ inch wide and 11 inches long, folded into "U" shape, as shown in Fig. 1612. The National UMA-10 condensers used for tuning these circuits are removed from their Isolantite mounting plates and remounted on the ends of the copper-strip inductances. A National GS-1 Isolantite insulator is fastened just below the condenser to make the assembly rigid. Connection to the tube plates is made with short lengths of flexible copper ribbon ¼ inch wide.

The final grid circuit is made of 1/16-inch copper strip, ¼ inch wide, bent into a "U" which is the same width as the plate tank to which it is coupled. The grid "U" shown is about three inches long. It should be as short as possible and yet provide adequate grid drive when coupled as closely as possible to the tripler plate circuit. Making several grid "coils" in order to attain this end is preferable to using longer tanks and looser coupling, as the smaller the grid tank is, the less tendency there will be for instability in the final stage.

The transmitter is mounted on a 17 × 8 × 2-inch chassis, and is designed for rack mounting, using a standard 8¾-inch panel. The tripler plate circuit is in the exact center of the chassis, and is mounted on a small rectangle of polystyrene which is bolted to the bottom of the chassis. In order to permit the use of alternate amplifier units, a square hole was cut from the chassis top at the point where the amplifier units are mounted, and the two final stages were assembled on chassis of folded aluminum 6 inches high by 7½ inches wide. If only one amplifier stage is to be built, it can, of course, be mounted directly on the chassis. An advantage of the separate-chassis method is that the holes by which it is mounted can be made into slots, permitting the whole assembly to be moved back and forth slightly for adjustment of coupling to the tripler stage.

The screen-supply switch, S_2, provides a safe method for tuning up the transmitter without damaging the 829. With the meter measuring a branch of the exciter grid current (left-hand position of S_1, Fig. 1610), tuning the oscillator plate circuit for maximum indication should produce about 1¼ to 1½ ma. Tuning the second stage should raise the total reading to about 3 ma. With the grid-meter switch in the right-hand position, tuning the tripler plate circuit to resonance should produce about 12-ma. final grid current under load, when the coupling between the tripler plate and final grid circuits

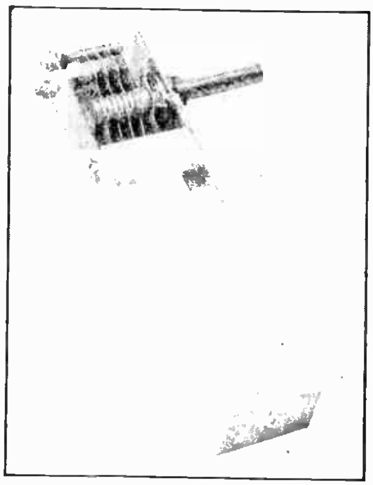


Fig. 1612 — Detail photo of the tank circuit used in the 815 and 829 plate circuits. Dimensions are given in Fig. 1613.

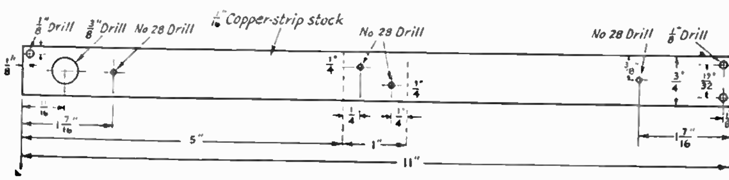


Fig. 1613 — Dimensional drawing of the 144-Mc. tank inductance before bending. The material is 1/16-inch copper strip.

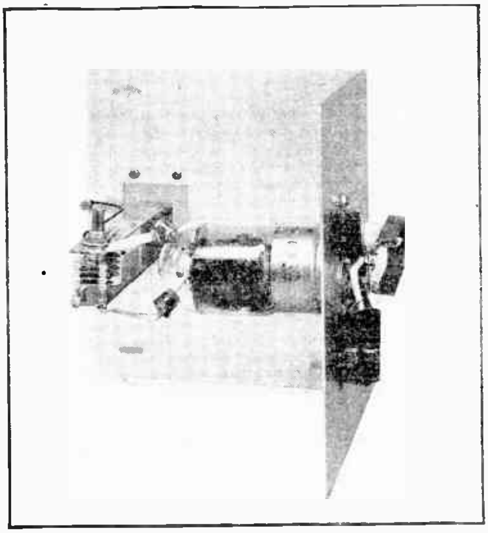


Fig. 1614 — An alternate amplifier unit using an 815, which may be substituted for the 829 unit shown in the complete assembly, Fig. 1609.

is adjusted correctly. The final-stage plate circuit may then be adjusted and the full screen voltage applied. With antenna or dummy load, the output coupling (L_8 is approximately size of L_6, but of No. 12 enameled wire) may be adjusted to raise the final plate current to 200 to 250 ma., depending upon the plate voltage used. The transmitter is then ready for modulation, which may be supplied by any audio unit having 75 watts output, or the unit may be used as an exciter, in which case it is capable of driving a final stage of 500 watts or higher rating.

A Mobile Transmitter for 50 and 28 Mc.

Low over-all battery drain in mobile operation is best obtained through the use of filament-type tubes which are lighted only during transmission periods. The mobile unit for 6, 10, and 11 meters, shown in Figs. 1615–1619, employs filament-type beam tetrodes throughout. Five 2E30s are used, as crystal oscillator, frequency multiplier, Class A driver, and push-pull Class AB modulators. The final stage is a 2E25, a tube of somewhat larger design, having its plate connection at the top of the envelope. Total filament current is only 4.3 amperes, and there is no drain whatever when the rig is not actually on the air.

The transmitter is housed in a crackle-finished cabinet of modern design (Par-Metal CA-202) which may be mounted in back of the seat in coupe-type vehicles or in the trunk compartment of sedans.

Special attention is paid to ruggedness of construction, all leads being made as short and direct as possible. Small components are supported with terminal strips at each end where possible, and tuning controls are equipped with dial locks (National ODL). The meter (a Marion 0–10-ma. sealed unit) is back-of-panel mounted, with a sheet of Lucite serving as a protecting window. This method of mounting the meter, about ½ inch in back of the panel, also provides a convenient method for illuminating the meter face. Dial lights are mounted at either side of the meter, as shown in Figs. 1617 and 1619.

By using 100-$\mu\mu$fd. variable condensers for C_2 and C_3, the range of the oscillator and multiplier plate circuits is extended, so that it is unnecessary to change these coils in changing

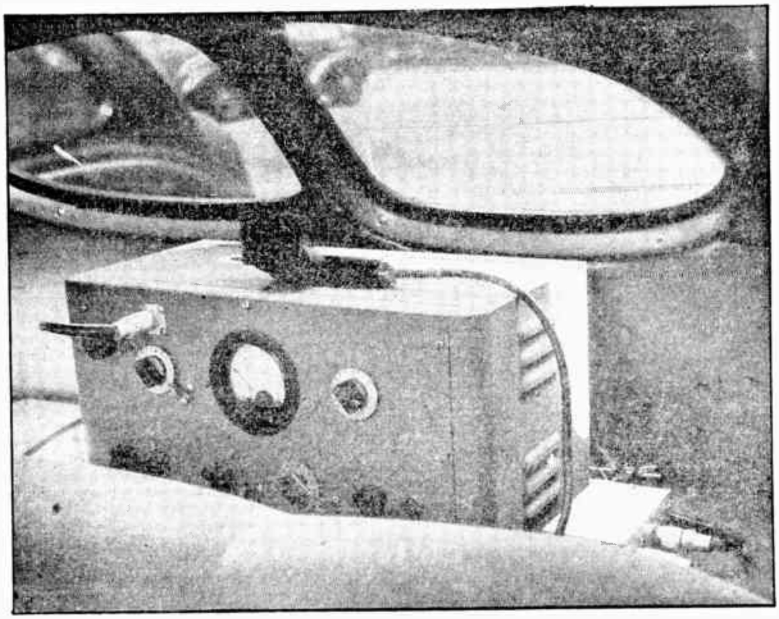


Fig. 1615 — A typical installation of the 6- and 10-meter mobile transmitter. The small aluminum box at the right of the unit houses the antenna change-over relay. The genemotor and its starting relay are mounted under the hood, adjacent to the car battery. Operation of the transmitter is controlled entirely by the push-to-talk switch on the microphone.

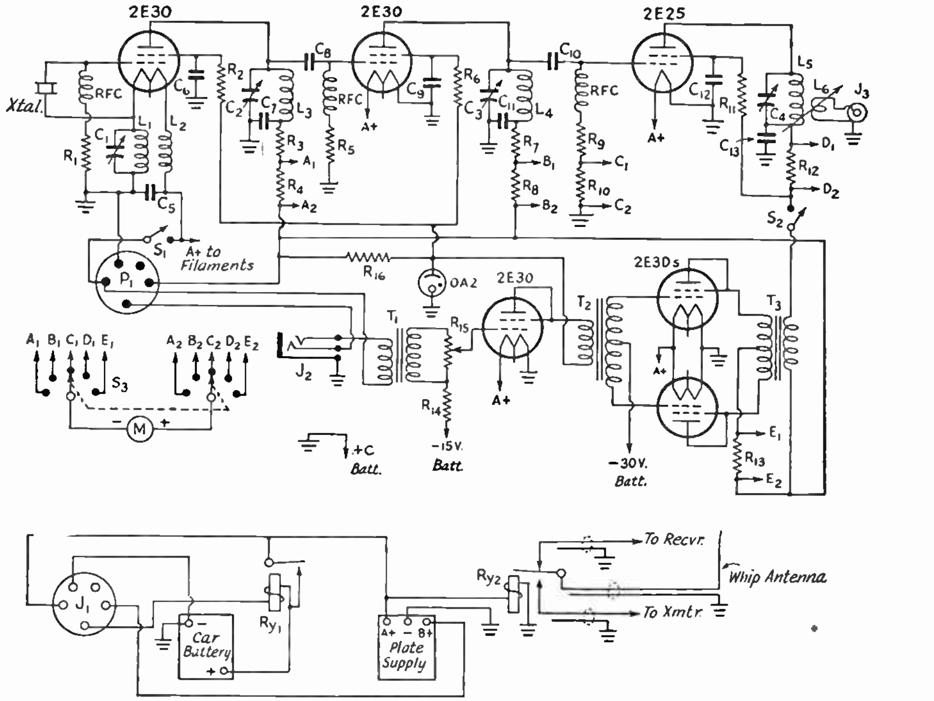


Fig. 1616 — Wiring diagram of the mobile rig for 6 and 10 meters.

C_1 — 100-μμfd. midget, screwdriver-adjustment type (Hammarlund APC-100).
C_2, C_3 — 100-μμfd. midget, shaft type (Hammarlund HF-100).
C_4 — 15 μμfd., double spaced (Hammarlund HFA-15-E).
C_5 — 0.001-μfd. mica.
C_6, C_7, C_9, C_{11}, C_{12}, C_{13} — 170-μμfd. midget mica.
C_8, C_{10} — 100-μμfd. midget mica.
R_1 — 82,000 ohms, 1 watt.
R_2, R_6 — 1000 ohms, ½ watt.
R_3, R_7, R_{10} — 100 ohms, ½ watt.
R_4, R_8, R_{12}, R_{13} — Special shunts. (See text.)
R_5 — 150,000 ohms, 1 watt.
R_9 — 33,000 ohms, 1 watt.
R_{11}, R_{16} — 5000 ohms, 10 watts.
R_{14} — 10,000 ohms, ½ watt.
R_{15} — 0.5-megohm potentiometer.
L_1, L_2 — 7 turns each, No. 20 d.c.c., ⁵⁄₁₆ inch long on 1-inch dia. form, windings interwound.
L_3 — 10 turns No. 12 enam., close-wound on 1-inch dia. form.
L_4 — 6 turns No. 12 enam., ¾ inch long, ½-inch inside dia., self-supporting.
L_5 — 28 Mc.: 10 turns No. 12 enam., 1½ inches long, 1-inch inside dia., self-supporting.
50 Mc.: 5 turns No. 12 enam., 1 inch long, 1-inch inside dia., self-supporting.
L_6 — 3 turns on ½-inch polystyrene rod — see text and detail photo.
J_1 — Socket on power cable, 5 prong.
J_2 — Double-button microphone jack. If T-17-B microphone is used, a special jack designed for this microphone must be obtained.
J_3 — Coaxial fitting (Amphenol 83-1R. Matching plug is 83-1SPN).
M — 0–10-ma. sealed unit (Marion).
P_1 — Power plug on transmitter chassis.
RFC — 2.5-mh. r.f. choke, National R-100.
Ry_1, Ry_2 — See text.
S_1, S_2 — S.p.s.t. snap switch.
S_3 — 2-section 5-position wafer-type switch.
T_1 — Single-button microphone transformer.
T_2 — Driver transformer (Stancor A-4752).
T_3 — Modulation transformer (UTC S-18).

bands. Only the crystal and the final plate coil, L_5, need be changed. Complete push-to-talk operation is made possible through the use of two relays. Ry_1 starts the genemotor and applies the filament voltage to the transmitter. Ry_2 transfers the antenna from receiver to transmitter. Both are controlled by the switch on the microphone, which may be any single-button type which has a control switch. The Army T-17-B, now currently available as government surplus, is shown with the rig.

The crystal oscillator is a Tri-tet, modified for filament-type tubes. Interwound coils are inserted in the filament leads, and one of these is tuned. The setting of this adjustment is not critical and may be left near maximum capacity for 6.8-, 7-, and 8.4-Mc. crystals. The oscillator doubles in, its plate circuit at all times.

The stage following the oscillator is operated as a doubler for 27- and 28-Mc. work, and as a tripler for 50 Mc. The 2E30 is an effective frequency multiplier, and there is adequate excitation for the final in either case. Screen voltage on the exciter stages is stabilized with a miniature voltage-regulator tube, an 0A2. With a screen voltage of 150, the plate input to both 2E30s is held to about 6 watts per tube.

The final stage uses a 2E25, whose top-cap plate connection permits the mounting of the plate circuit above the chassis, well isolated from the other tuned circuits. A small shield,

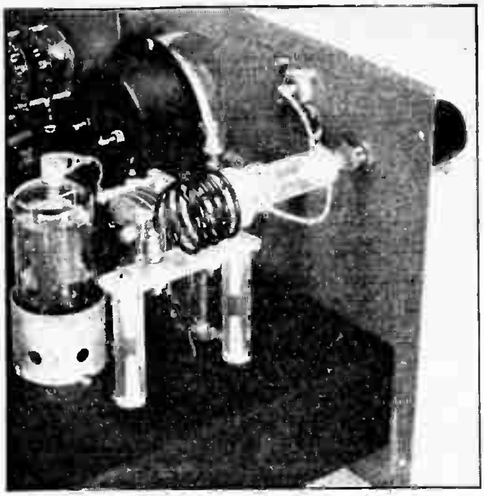


Fig. 1617 — Detail photo of the 2E25 final stage, showing method of coupling to the antenna. The coupling coil, wound on a polystyrene rod, is adjustable from the front panel. The plate coil is mounted by means of G.R. plugs.

cut from an old-style tube shield to a length of about one inch, comes up to the bottom of the 2E25 plate assembly. These precautions are sufficient to provide completely stable operation without neutralization.

The antenna coupling coil, L_6, is wound on a short length of polystyrene rod ½ inch in diameter, into which is inserted a ¼-inch rod of the same material. This shaft projects through the front panel, where a shaft-locking panel bushing (Bud PB-532 bushing, Millen 10061 shaft lock) holds it in the desired position. Coupling is adjusted by pushing or pulling the knob affixed to the shaft, following which the bushing may be tightened for permanent setting. The bushing may also be set finger-tight, allowing the coupling to be adjusted, yet holding it with sufficient firmness to prevent its being jarred out of position.

Three 2E30s are used for the modulator, one as a Class A driver, and two in push-pull as Class AB modulators. All three are triode-connected. Bias is supplied by a 30-volt hearing-aid battery, which can be tapped at 15 volts by opening up the cardboard case and soldering on a lead at the point where the two 15-volt sections are joined together. This lead is brought out to the unused terminal on the battery socket and plug.

Metering of all circuits is provided by a 10-ma. meter, a 2-section 5-position switch, and a set of shunts. The shunts are made from small 100-ohm resistors, on which is wound about 7 feet of No. 30 enameled wire. The shunts should be wound with an excess of wire, the length of which may be reduced until the multiplication of the meter scale is just right. The resistor R_{10} in the final grid circuit is left without a shunt, giving direct reading on the 10-ma. scale for measuring the final grid current.

Except for the speech stages, the unit may be tested using 6.3 volts a.c. on the filaments and an a.c. power supply. A storage battery must be used for filament supply when the speech equipment is to be tested, as a.c. on the filaments will produce excessive hum. Initial testing should be carried on with about 200 volts on the tube plates. When operation has been found to be satisfactory, this may be raised to 300.

To place the unit in operation, set S_1 to the "on" position, leaving S_2 "off." With the meter switch in position "A," apply plate voltage and note meter reading, which is the oscillator plate current. This will be about 20 ma., dipping slightly at resonance as C_2 is adjusted. Next switch to position "B" and adjust C_3. The dip here may not be as pronounced as in the oscillator, and the final grid current, position "C," 10-ma. scale, is the best indication of resonance in the preceding adjustments. This reading should be about 4 ma., dropping to 3 ma. under load. With S_2 turned on, the final plate current, position "D," should drop to below 10 ma. at resonance, and coupling of the antenna should raise it to 50 to 60 ma. Modulator plate current will be about 20 ma., rising to 60 ma. or more on audio peaks. No metering position is provided for the Class A driver current, but this should

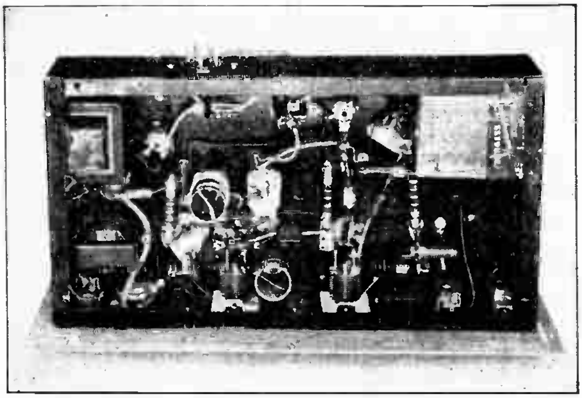


Fig. 1618 — Bottom view of the mobile rig. At the left center are the interwound coil and tuning condenser which are part of the oscillator filament circuit. Audio components are at the left, with oscillator and multiplier plate circuits near the front panel.

Fig. 1619 — The plate circuit of the final stage of the mobile transmitter is the only r.f. circuit above the chassis. The three tubes at the left are the driver and audio stages, with the oscillator and multiplier tubes directly in back of the meter. The tube to the right of the modulation transformer is the 0A2 voltage regulator. Chassis size is 7 × 13 × 2 inches.

be approximately 10 ma.

With the coil and condenser values given, it is impossible to get output from the final stage on a wrong frequency, but excitation to the final may be obtained on incorrect harmonics; hence it is advisable to check the frequency of each stage with a calibrated absorption-type wavemeter.

For maximum convenience, the same antenna should be used for both transmission and reception. Antenna change-over is handled with a conventional 6-volt antenna relay which was mounted in a small box made up for the purpose from folded sheet aluminum. Amphenol coaxial fittings, mounted on the sides of the relay box as close to the relay contacts as possible, provide for connection to the transmitter, the receiver, and the antenna by means of coaxial line. The relay case is grounded and only the inner conductor of the coaxial line is switched.

A headlight relay for genemotor starting may be purchased from any auto-accessory store, and this and the genemotor should be mounted as close to the car battery as possible, in order to minimize voltage drop. Battery wiring and filament cables should be as heavy wire as possible, with No. 10 as the minimum.

For actual mobile operation, the quarter-wave telescoping "whip" antenna, operating as a Marconi in the manner shown in Fig. 1616, is convenient. Much greater range in stationary operation from high locations may be had with half-wave radiators or multielement arrays, either of which may be arranged for easy on-the-spot assembly. An example of such a portable array for 50 Mc. is shown in Fig. 1711, Chapter Seventeen.

A 6C4 Oscillator for 144 and 235 Mc.

Figs. 1620 to 1623, inclusive, show the details of construction of a low-power oscillator using a 6C4, a miniature triode power tube having a plate dissipation rating of 5 watts and designed for use as an oscillator in the v.h.f. range. At the rated plate input of 300 volts at 25 milliamperes the oscillator develops an r.f. output of about 2 watts in the 144-Mc. band. With minor modifications, to be described later, the oscillator may also be made to work on 235 Mc., with somewhat lower efficiency.

As shown by the diagram, Fig. 1622, the circuit is the ultraudion with an adjustable feedback condenser, C_3, connected between grid and cathode. To reduce frequency modulation when the oscillator is amplitude-modulated, the tuned circuit has a fairly high C/L ratio,

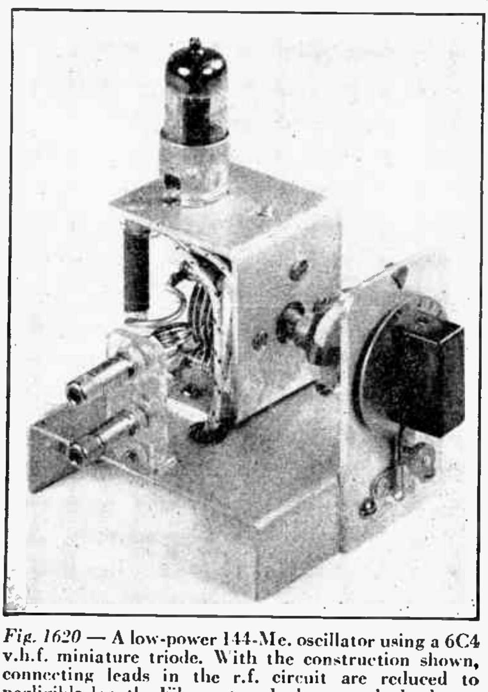


Fig. 1620 — A low-power 144-Mc. oscillator using a 6C4 v.h.f. miniature triode. With the construction shown, connecting leads in the r.f. circuit are reduced to negligible length. Filament and plate-supply leads are brought through the bottom chassis to a connection strip on the rear lip. The excitation control is adjusted through a hole in the top of the supporting member.

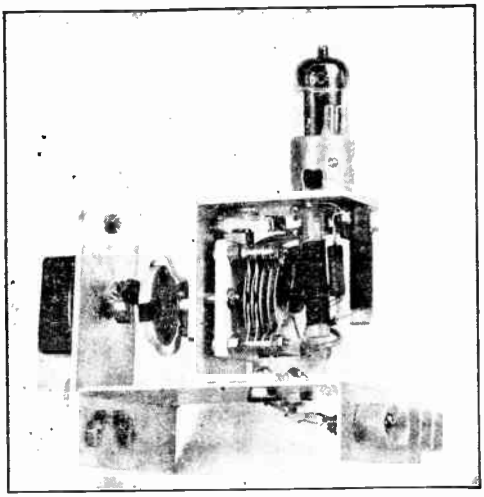


Fig. 1621 — A view showing the assembly of components of the 6C4 144-Mc. oscillator. The r.f. chokes are mounted by drilling and tapping the ends of the polystyrene rod. The grid choke is held in place by one of the socket mounting screws.

using a tuning condenser having a fixed as well as a variable section. The condenser rotor consists of three circular plates and two "butterfly" plates. The circular plates rotate between two sets of stators having plates of regular shape and thus provide a fixed capacity. The butterfly plates rotate between two sets of opposed 90-degree stator plates, each set consisting of two plates. The assembly (now available as Cardwell Type ER-14-BF/SL) is made from a Cardwell ER double condenser, with only the front Isolantite plate used for a mounting. This method of construction results in a split-stator condenser having a minimum of inductance, since the r.f. current flows over the rotor plates without having to travel along the shaft. The plate shapes and details of assembly are shown in Fig. 1623.

Lead lengths in the circuit are reduced to a minimum by the construction shown in Figs. 1620 and 1621. The entire oscillator assembly is mounted on a piece of 3/32-inch-thick aluminum bent in the general shape of a "U." The mounting is 1⅞ inches wide and the bent-over top portion is 1⅞ inches deep. The over-all height is 2¼ inches. The bottom lip dimension can be anything convenient so long as enough area is provided to make a solid mechanical mounting. The tuning condenser, C_1, is centered on the vertical portion and is mounted on the screws and spacers provided with the condenser. The hole for the shaft is made amply large so that the condenser rotor is not grounded. The condenser is mounted so that the two sets of stator plates are at top and bottom.

The tube socket is mounted so that the plate lead can drop in as straight a line as possible to the terminal at the right on the upper stator plates of C_1. The grid condenser, C_2, is supported at one end by the grid prong on the tube socket and at the other by the left-hand terminal on the lower stator plates. The excitation control, C_3, has its movable-plate tab bent at a right angle so it can be bolted to the vertical support, and the stationary-plate tab is soldered directly to the grid prong on the tube socket. The grid choke, grid leak, and plate choke are supported as shown in the photograph. The condenser along the rear edge of the assembly is the heater by-pass condenser, C_4.

The oscillator assembly is mounted on a 3¼ by 3¼-inch aluminum channel ¾ of an inch deep. A small panel at the front provides a place for a tuning dial which drives the condenser shaft through an insulated coupling. A dial lock is provided so the condenser can be locked at a given frequency setting.

A polystyrene-insulated double binding-post assembly mounted vertically from a small bracket provides output terminals and a support for the antenna coupling coil, L_2. The coupling can be varied by bending the soldering lugs that support the coil so that L_2 is moved nearer to or farther away from L_1.

The condenser construction provides just enough capacity variation to cover the 144–148-Mc. band adequately. Because of slight differences in the construction of similar units, it may be necessary to vary the inductance of L_1 slightly to bring the band on the dial: this can be done by squeezing the turns together or pulling them apart. The frequency range can be checked with Lecher wires or a calibrated absorption wavemeter. (See chapter on frequency measurement.) Final adjustment of L_1 should be made after C_3 has been adjusted for optimum output from the oscillator, since the setting of this condenser has some effect on the frequency of oscillation.

To adjust C_3, solder two pieces of wire about

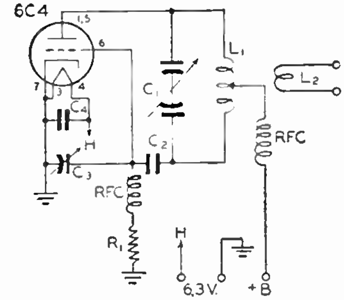


Fig. 1622 — Circuit diagram of the 6C4 oscillator.

C_1 — Tuning condenser: see text and Fig. 1623.
C_2 — 17-μμfd. midget mica.
C_3 — 3–30-μμfd. ceramic trimmer (National M-30).
C_4 — 170-μμfd. midget mica.
R_1 — 22,000 ohms, ½ watt.
L_1 — 2 turns No. 12 bare wire; inside diameter 9/16 inch, length 1 inch: plate-supply tap at center.
L_2 — 2 turns No. 14 enameled; inside diameter ⅜ inch, slight spacing between turns.
RFC — 1-inch winding of No. 24 d.s.c. or s.c.c. on ¼-inch diameter polystyrene rod.

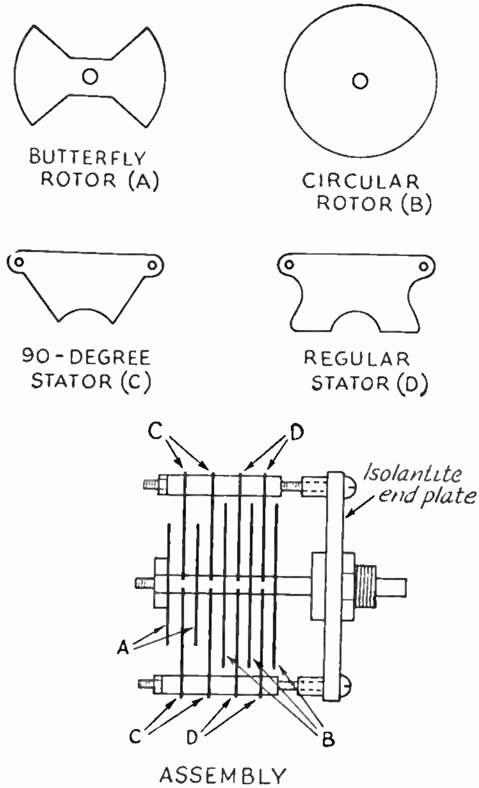


Fig. 1623 — Plate shapes and assembly of C_1, the tuning condenser used in the 6C4 oscillator.

¾ inch long to the terminals of a small flashlight lamp or dial light and connect them to the output terminals. A milliammeter of 0–50 or 0–100 range should be connected in the plate-supply lead. Adjust the coupling between L_2 and L_1 for maximum glow in the lamp and then vary the capacity of C_3 until the best output is obtained. C_3 need not be touched again after the proper setting is determined.

In using the oscillator for transmitting, the coupling between L_1 and L_2 should be kept as loose as possible, particularly if the antenna or feeders can swing in a breeze, because any change in the antenna circuit will be reflected as a change in the oscillator frequency. In any event, the coupling should not be increased beyond the setting that makes the oscillator plate current 25 milliamperes. At 300 volts the plate current should be about 20 ma. without any r.f. load.

For operation on 235 Mc. it is merely necessary to remove one rotor disc and one set of stators and replace the coil with a "U"-shaped inductance similar to that used in the 235-Mc. transceiver, Fig. 1638.

A High-C 144-Mc. Oscillator

The inherent instability of a modulated oscillator — that is, the change in frequency with the change in plate voltage under modulation — can be markedly reduced if the oscillator tank circuit is made to have as high a C/L ratio as possible. Although this usually entails some sacrifice of power output, the overall effectiveness of the transmitter is increased because the radiated energy is more nearly on one frequency. This is a particularly important consideration when selective receivers are used. In addition, the fact that there is less frequency modulation also means that there is less interference to other stations operating in the same band.

A high-C 144-Mc. oscillator is shown in Figs. 1624, 1625, and 1626. It uses an HY75 tube and a tank circuit consisting of a low-inductance v.h.f. condenser and a one-turn tank coil of heavy conductor. The circuit, shown in Fig. 1625, is the ultraudion with a tuned filament circuit to provide control of excitation. The oscillator is mounted on a 3 × 4 × 5-inch box, with the tube socket mounted below the top by means of pillars so that only the glass bulb is protruding. To bring the condenser terminals on the same level as the grid and plate terminals of the tube the condenser is mounted on ⅝-inch-high blocks.

The tube socket is positioned so that the plate cap of the tube is near one set of the stator plates of C_1. This leaves room to mount the grid condenser, C_2, between the grid cap and the other stator terminal, thus making

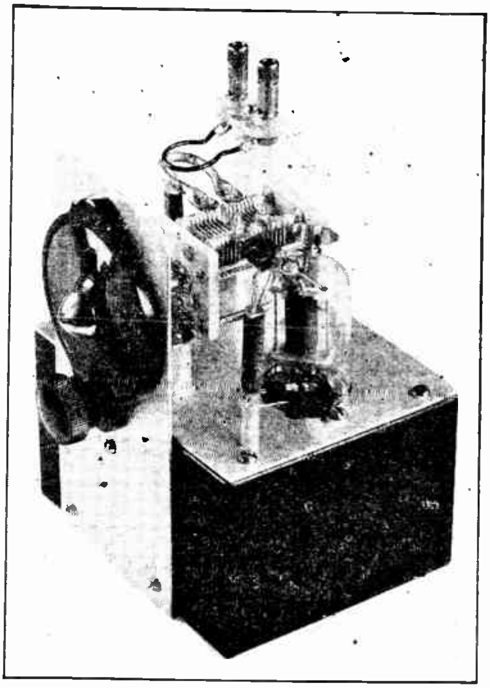


Fig. 1624 — A high-C 144-Mc. oscillator using an HY75. This type of oscillator has considerably less frequency modulation than those using low-C circuits, consequently causes less interference and can be more effectively received on selective receivers.

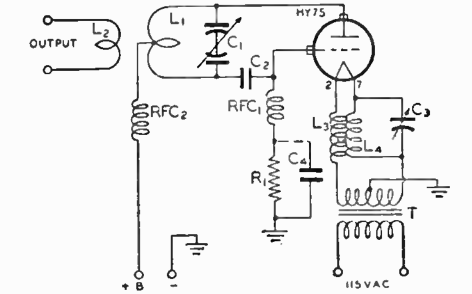


Fig. 1625 — Circuit diagram of the high-*C* 144-Mc. oscillator.

C_1 — Split-stator condenser. 31.5 μμfd. total (Hammarlund VU-30).
C_2 — 47-μμfd. midget mica.
C_3 — 3-30-μμfd. ceramic trimmer.
C_4 — 100-μμfd. midget mica.
R_1 — 4700 ohms, 1 watt.
L_1 — 1 turn of 5/32-inch copper tubing, approximately horseshoe shape; over-all length from mounting holes in lugs, 1¾ inches; outside diameter at widest point, 13/16 inches; plate tap at center.
L_2 — 1 turn No. 14 enameled; diameter ¾ inch.
L_3, L_4 — 6 turns No. 18 d.c.c. on ½-inch form; L_3 interwound with L_4, no spacing between turns.
RFC_1, RFC_2 — 1-inch winding of No. 24 d.s.c. or s.c.c. on ¼-inch diameter polystyrene rod.
T — 6.3-volt filament transformer.

the leads between the tank circuit and the tube as short as possible. The output coupling coil, L_2, is soldered to lugs under the binding posts of a two-post assembly mounted on a 2½-inch Isolantite stand-off insulator. A friction-type vernier dial is used to tune the circuit, because the tuning is rather critical with the high-C circuit and because the type of condenser used requires this or a similar type of dial to hold the setting, since the shaft turns on ball bearings. The dial mounts on a small supporting panel with rounded corners, as shown in the photographs.

The tuned filament circuit consists of L_3, L_4 and C_3. L_3 is wound between the turns of L_4 so that the coupling is very tight; thus both filament leads can be tuned by one condenser, C_3. C_3 is adjusted for maximum output as judged by the brilliance of a lamp connected to the output terminals; it has relatively little effect on the frequency of oscillation. Once this adjustment has been made its setting may be considered permanent except for occasional rechecking.

The inductance of L_1 should be adjusted so that the low-frequency end of the 144-Mc. band is reached with C_1 set as close as possible to maximum capacity. It is advisable to start with the coil a little larger than necessary and cut a little at a time off the ends until the proper inductance is found. The connections between the coil and the condenser are made by means of lugs fashioned from tubing just enough larger in diameter than the coil so that the ends of the coil will fit inside. One end of each lug is flattened and drilled to fit the condenser terminals, and the coil is soldered in the unflattened ends.

With a plate input of 350 volts at about 60 milliamperes the power output of the oscillator is approximately 4 watts. When received on a superheterodyne-type receiver with a beat-frequency oscillator, the carrier will be quite clean and stable provided the mechanical construction is rigid. Under modulation, the frequency band occupied is only about a fifth as much as that taken up by a low-C oscillator operated at the same plate voltage. In these days of crowded v.h.f. bands a high-C transmitter of this type will go a long way toward improving conditions, though it is far from the ultimate.

A Stabilized 144-Mc. Transmitter

In general, a modulated oscillator is not a desirable type of transmitter for use in a band such as 144 Mc. where there is considerable activity. Even when stabilized by the use of a high-C tank circuit this type of transmitter leaves much to be desired, because there is still a great deal more frequency modulation than is present in a master-oscillator power-amplifier transmitter. In addition, an oscillator coupled to an antenna is subject to frequency change whenever the antenna constants change slightly, as they will with changes in weather and with any vibration or swinging of the feeder wires. Besides, an oscillator cannot be modulated 100 per cent without considerable distortion because in most cases oscillation cannot be sustained at plate voltages below 50 to 100 volts. Finally, the efficiency of an oscillator is quite low compared to the efficiency of a properly-driven amplifier, so that considerably more power output can be obtained from the same tube when it is used as an amplifier than when it is used as a self-controlled oscillator.

An amplifier driven by an oscillator, al-

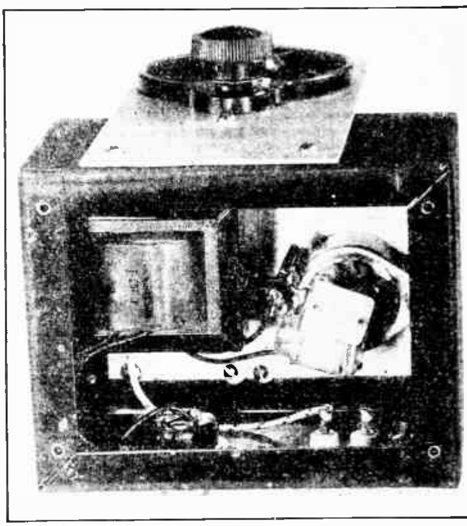


Fig. 1626 — Below-chassis view of the high-*C* 144-Mc. oscillator. The filament transformer and filament tuned circuit are mounted inside the box.

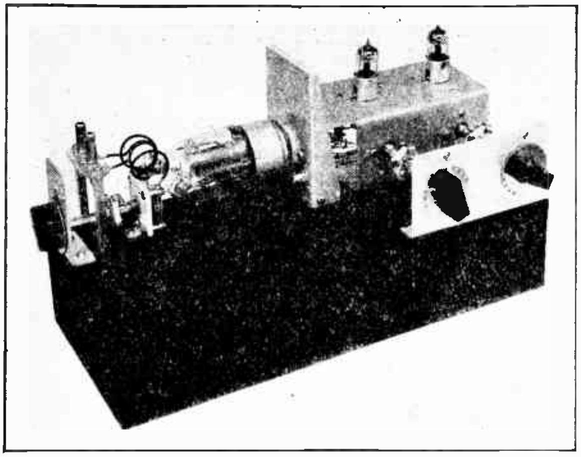


Fig. 1627 — A three-stage transmitter using a 6C4 master oscillator, 6C4 buffer amplifier, and 815 final amplifier for stabilized transmission in the 144-Mc. band. The oscillator and buffer are built as a unit on the folded aluminum chassis at the right. The transmitter develops a carrier output of about 40 watts.

though more stable than an oscillator alone, is still subject to frequency-modulation effects because the change in power input to the amplifier with modulation causes a change in the grid impedance of the amplifier, and this in turn reacts on the oscillator to change the frequency. Hence it is desirable to use at least one buffer-amplifier stage between the oscillator and amplifier. If this is done it is quite possible to get satisfactory performance with inexpensive low-power tubes in both oscillator and buffer stages, while if the buffer is omitted it is necessary to use a fairly high-power oscillator. This is because the coupling between the oscillator and modulated amplifier must be very loose if the oscillator frequency is not to be affected by whatever happens in the amplifier plate circuit; consequently the oscillator must develop much greater power than actually is needed to drive the amplifier since only a small part of the power can be utilized with the loose coupling required.

A three-stage transmitter in which frequency-modulation effects are quite small is shown in Figs. 1627–1630, inclusive. It includes a 6C4 oscillator, 6C4 neutralized buffer amplifier, and 815 final amplifier, as shown in the circuit diagram, Fig. 1628. The oscillator and buffer are built as a unit on a "U"-shaped piece of aluminum 6½ inches long on top, 2⅜ inches high, and 2⅞ inches deep on the top. The 815 is mounted on a vertical aluminum piece measuring 4¼ inches high and 3 inches wide, reinforced by bending side lips as shown in the photographs. The two sections are assembled on a 6 × 14 × 3-inch chassis.

The oscillator circuit and components are identical with those already described in *QST*

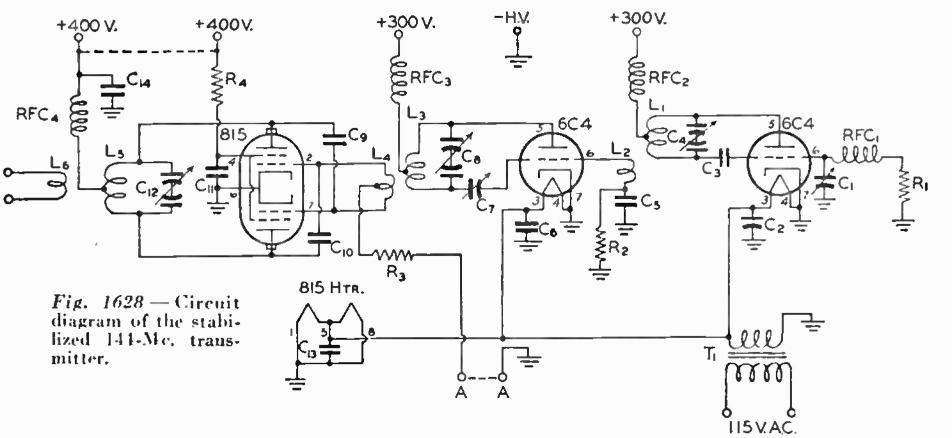


Fig. 1628 — Circuit diagram of the stabilized 144-Mc. transmitter.

C1 — 3–30-μμfd. trimmer.
C2, C6, C11, C13 — 470-μμfd. midget mica.
C3, C5 — 47-μμfd. midget mica.
C4 — Oscillator tuning: see text.
C7 — Neutralizing; see text.
C8 — Buffer tuning; see text.
C9, C10 — Amplifier neutralizing; see text.
C12 — Amplifier tuning; see text.
C14 — 100 μμfd., 2500 volts.
R1 — 22,000 ohms, ½ watt.
R2 — 22,000 ohms, ½ watt.
R3 — 15,000 ohms, 1 watt.
R4 — 15,000 ohms, 10 watts.
L1 — 2 turns No. 12 bare wire; inside diameter 9/16 inch, length 1 inch; plate-supply tap at center.
L2 — 2 turns No. 14, inside diameter ½ inch; turns spaced wire diameter.
L3 — 4 turns No. 14, inside diameter ¾ inch, length 1 inch; plate-supply tap at center.
L4 — 2 turns No. 14, inside diameter ½ inch; turns spaced diameter of wire; tapped at center.
L5 — 2 turns No. 12, inside diameter 1 inch, length 1 inch; plate supply tap at center.
L6 — 2 turns No. 12, inside diameter ¾ inch.
RFC1, RFC2, RFC3, RFC4 — 1-inch winding No. 24 d.s.c. on ¼-inch diam. polystyrene rod.
T1 — 6.3-volt 2-amp. filament transformer.

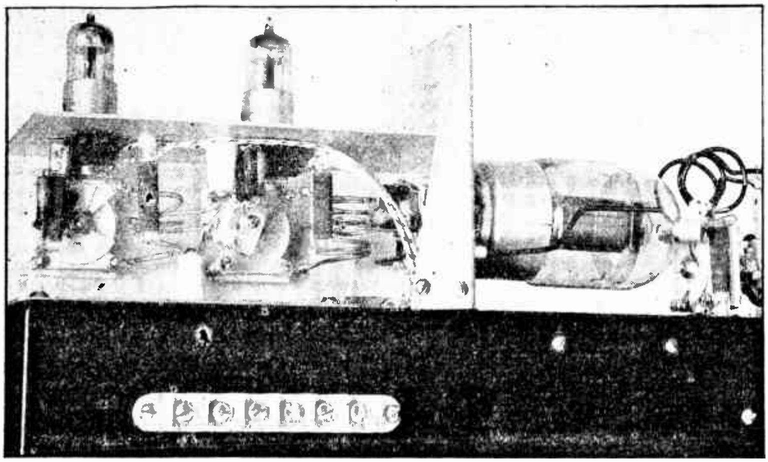


Fig. 1629 — A rear view of the three-stage 144-Mc. transmitter. The oscillator is at the left, with the buffer amplifier in the center. The 815 final is at the right.

for April, 1946. The construction of the buffer amplifier is quite similar to that of the oscillator. The buffer tuning condenser consists of a rotor having three butterfly plates and two stators each having two 90-degree plates. The grid circuit of the buffer is self-resonant, the tuning being adjusted by squeezing the turns of the grid coil L_2 together, or prying them apart. The buffer neutralizing condenser, C_7, mounted directly between the grid of the 6C4 and the lower set of stator plates of C_8, is a 3–30-$\mu\mu$fd. trimmer with the movable plate removed and a washer soldered under the head of the adjusting screw. The washer, by replacing the movable plate, reduces the capacity of the condenser to a value suitable for neutralizing the 6C4. This capacitor may be conveniently adjusted through the open end of the chassis. Its location is clearly shown in Fig. 1629.

The grid coil of the final amplifier also is resonant with the input capacity of the 815, just as the buffer grid circuit is self-resonant. For best operation, the 815 requires neutralization at this frequency. The neutralizing "condensers," C_9 and C_{10} in the circuit diagram, are simply pieces of No. 14 wire extending from the grid of one section of the 815 to the vicinity of the plate of the other section. The wires are crossed at the bottom of the tube socket and go through Millen 32150 bushings in the metal partition. The screen and filament by-pass condensers are mounted so that the leads between the socket prongs and the nearest ground point are as short as possible. This wiring should be done before mounting the partition.

The amplifier plate tank circuit uses a condenser of the same construction as that used in the buffer tank. It is mounted as closely as possible to the plate caps on the 815, and to preserve circuit symmetry the condenser is tuned from the left-hand edge of the chassis. If the transmitter is to be equipped with a regular panel this condenser may be operated by a right-angle drive from the front.

The output terminals are a standard binding-post assembly on polystyrene, mounted on metal posts 2¾ inches high to bring the coupling coil in proper relation to the amplifier plate tank coil, L_5. Coupling is adjusted by bending L_6 toward or away from L_5.

The plate by-pass condenser and screen dropping resistor are mounted underneath the chassis, as shown in Fig. 1630, together with the filament transformer. Separate power-supply terminals are provided for the oscillator plate, buffer plate, amplifier grid (terminals A-A), amplifier screen, and amplifier plate so that the currents can be measured separately. An external 0–200-ma. milliammeter will serve in making all adjustments. However, if a meter of lower range is available, it may be used profitably in the low-current circuits.

In putting the transmitter into operation, the first step is to adjust the frequency range of the oscillator, using Lecher wires or a calibrated absorption-type wavemeter. This should be done after C_1 has been adjusted for maximum output. Then, using loose coupling between the buffer grid coil, L_2, and the oscillator tank coil, L_1 (the coupling may be adjusted by bending L_2 away from L_1 on its mounting lugs), adjust L_2 by changing the turn spacing until the grid circuit is resonant. Resonance will be indicated by maximum oscillator plate current; it can also be checked by measuring the voltage across the buffer grid leak, R_2, with a high-resistance voltmeter. The maximum voltmeter reading

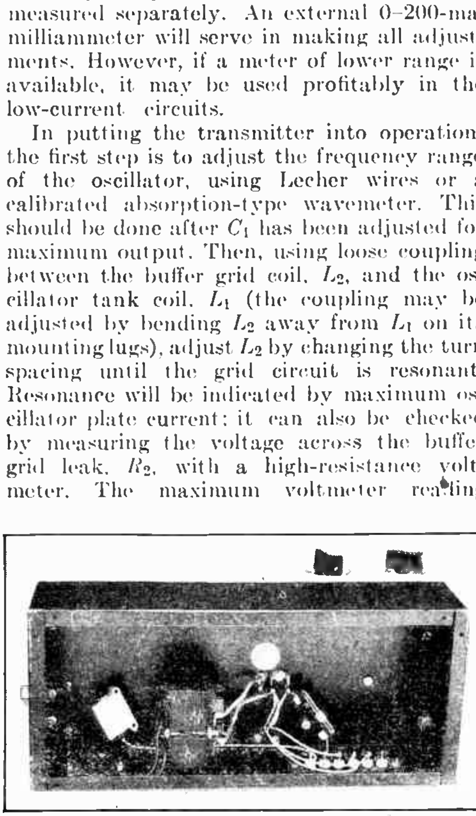


Fig. 1630 — Underneath the chassis of the 144-Mc. MOPA transmitter. The filament transformer, amplifier plate by-pass condenser, and screen dropping resistor are mounted here.

(about 40 volts) indicates resonance. The buffer should next be neutralized by varying the capacity of C_7 until there is no change in the voltage across R_2 when the buffer tank condenser, C_8, is tuned through resonance. The point of correct neutralization also can be determined by coupling a sensitive absorption wavemeter such as is described in the chapter on frequency measurement to the buffer plate coil, and adjusting C_7 for minimum reading. With this method, care must be used to avoid coupling between the wavemeter and the oscillator: link coupling between L_3 and the wavemeter, with the latter far enough away so that it does not give a reading from the oscillator alone, should be used. Another method of checking neutralization is to adjust the turn spacing of the amplifier grid coil, L_1, to resonance and measure the 815 grid current (with no plate or screen voltage on the tube) and adjust C_7 for zero grid current.

After the buffer is neutralized, plate voltage may be applied and C_8 adjusted to resonance, as indicated by minimum plate current. If the coupling to the final amplifier is quite loose, the minimum plate current should be approximately 17 ma. The amplifier grid coil may next be resonated (by adjusting the spacing between turns) and the coupling increased until the maximum grid current is secured. The grid current should be 4 milliamperes or more and the buffer plate current should rise to about 28 ma.

Neutralization of the 815 is the next step. If the grid current changes when the plate condenser, C_{12}, is tuned through resonance, the neutralizing wires should be moved closer to or farther away from the tube plates until tuning C_{12} has no effect on the grid current. When this condition is reached the amplifier is neutralized. Plate and screen voltage may then be applied. With no load on the amplifier the plate current should dip to approximately 65 ma. at resonance. Loading the amplifier to a plate current of 150 ma. should not cause the grid current to drop below about 3.5 ma. A 40-watt lamp used as a dummy load should light to practically normal brightness at this input, using a plate-supply voltage of 400.

For greatest stability, the coupling between the oscillator and buffer should be as loose as possible. It is better to obtain the rated 815 grid current of 3 milliamperes by using tight coupling between the buffer and amplifier and loose coupling between the oscillator and buffer than vice versa. With normal operation the oscillator plate current should be approximately 25 ma. and the buffer plate current 28 ma., at 300 volts.

A modulator for the transmitter should have an audio output of 35 watts, using a coupling transformer designed to work into a 2500-ohm load.

A 144-Mc. Double Beam-Tetrode Power Amplifier

An amplifier set-up suitable for use with double-beam-tetrode tubes is shown in Figs. 1631, 1632 and 1633. The tube in the photographs is an 829, but an 815 or 832 can be used in the same layout. The only change that might be required would be in the inductances of the grid and plate coils, L_2 and L_3; these may have to be made slightly smaller or larger in diameter to compensate for the differences in input and output capacitances in the various types. The physical arrangement of the components is similar to that used for the 815 amplifier incorporated in the three-stage transmitter described in a preceding section.

The amplifier of Fig. 1633 is built on an aluminum chassis formed by bending the long edges of a 5 × 10-inch piece of aluminum to form vertical lips ¾ inch high, so that the top-of-chassis dimensions are 3½ by 10 inches. The tube socket is mounted on a vertical aluminum partition measuring 3½ inches high by 3¼ inches wide on the flat face, with the sides bent as shown in the photographs to provide bracing. The partition is mounted to the chassis by right-angle brackets fastened to the sides. The socket is mounted with the cathode connection at the top, the cathode prong being directly grounded to the nearest mounting screw for the socket. The heater by-pass condenser, C_6, is mounted directly over the center of the tube socket, extending between the paralleled heater prongs at the bottom and the cathode prong at the top. The screen by-pass is connected with as short leads as possible between the screen prong and the nearest socket mounting screw.

The grid coil, L_2, is supported by the grid prongs on the socket. The two turns of the coil

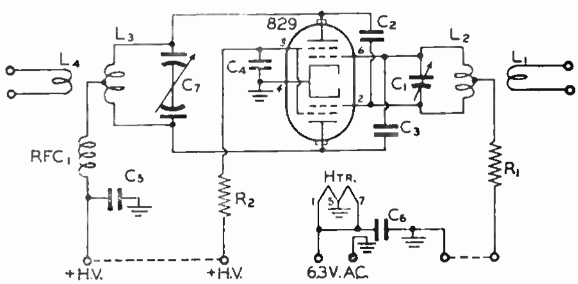


Fig. 1631 — Circuit of the 829 amplifier for 144 Mc.

C_1 — 3–30-μμfd. ceramic trimmer.
C_2, C_3 — Neutralizing condensers; see text.
C_4 — 500-μμfd. mica, 1000 volts.
C_5 — 500-μμfd. mica, 2500 volts.
C_6 — 170-μμfd. mica.
C_7 — Split stator, 15 μμfd. per section (Cardwell ER-15-AD).
R_1 — 4700 ohms, 1 watt.
R_2 — 10,000 ohms, 10 watts.
L_1 — 2 turns No. 12, diameter ½ inch.
L_2 — 2 turns No. 12, diameter ½ inch, length ½ inch.
L_3 — 2 turns No. 12, diameter 1⅛ inches, length 1 inch.
L_4 — 2 turns No. 12, diameter 1 inch
RFC_1 — 1-inch winding of No. 24 d.s.c. or s.c.c. on ¼-inch diameter polystyrene rod.

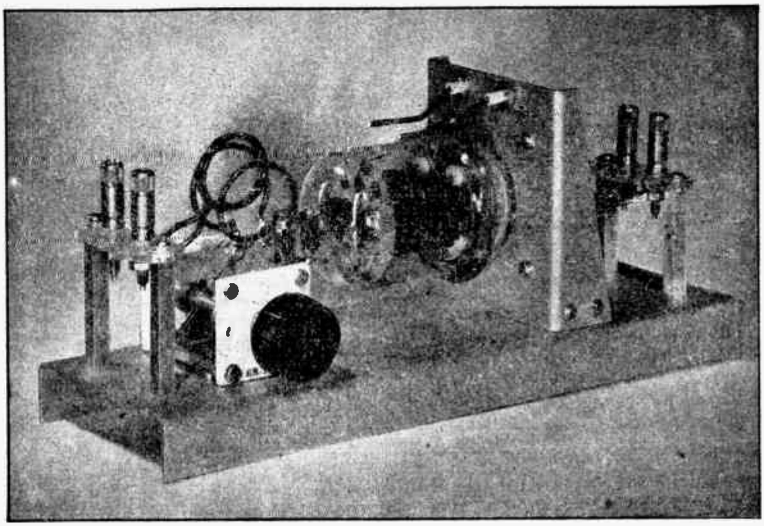


Fig. 1632 — A 144-Mc. amplifier using a double beam tetrode. This type of construction is suitable for the 815 and 832 as well as the 829 shown. The vertical partition provides support for the tube as well as shielding between the input and output circuits. Note the neutralizing "condensers" formed by the wires near the tube plates.

are spaced about one-half inch to allow room for the input coupling coil, L_1, to be inserted between them. The coupling is adjusted by bending L_1 into or out of L_2. The grid tuning condenser, C_1, is mounted between the socket prongs; although the condenser has mica insulation it is used essentially as an air-dielectric condenser since the movable plate does not actually contact the mica at any setting inside the band. The coupling link is soldered to lugs under binding posts on a National FWG strip, the strip being mounted on metal pillars 1½ inches high to bring the link to the same height as the grid coil.

Although the shielding between the input and output circuits of the tube is sufficiently good so that the circuit will not self-oscillate, tuning of the plate circuit will react on the grid circuit to some extent because the grid-plate capacity, although small, is not zero. To eliminate this reaction it is necessary to neutralize the tube. The neutralizing "condensers" are lengths of No. 12 wire soldered to the grid prongs on the socket. The wires are crossed over the socket and then go through small ceramic feed-throughs at the top of the vertical shield, projecting over the tube plates on the other side as shown in Fig. 1633.

Connections between the plate tank condenser, C_7, and the tube plate terminals are made by means of small Fahnestock clips soldered to short lengths of flexible wire. The tank coil, L_3, is mounted on the same condenser terminals to which the plate clips make connection. The output link, L_4, is mounted similarly to the grid link except that the posts are 1⅞ inches high. The plate choke, RFC_1, is mounted vertically on the chassis midway between the plate prongs of the tube, the mounting means being a short machine screw threaded into the end of the polystyrene rod. The "cold" lead of the choke is by-passed by C_5 underneath the chassis.

Supply connections are made through a 5-post strip on the rear edge of the chassis. The dotted lines between connections in Fig. 1631 indicate that these connections are normally short-circuited; leads are brought out so that the grid and screen currents can be measured separately.

In adjusting the amplifier, the plate and screen voltages should be left off and the d.c. grid circuit closed through a milliammeter of 0–25 or 0–50 range. The driver should be coupled to the amplifier input circuit through a link (Amphenol Twin-Lead is suitable, because of its constant impedance and low r.f. losses). Use loose coupling between L_1 and L_2 at first, and adjust C_1 to make the grid circuit resonate at the driver frequency, as indicated by maximum grid current. The coupling between L_1 and L_2 may then be increased to make the grid current slightly higher than the rated load value for the tube used — approximately 12 ma. for the 829. If the driver is an oscillator, the coupling between L_1-L_2 should be as loose as possible with proper grid current.

Neutralization can be checked by rotating C_7 through resonance. A flicker in grid current as C_7 is rotated indicates that the neutralizing capacity is not correct. The neutralizing wires should be bent in relation to the tube plates until the grid current remains constant when C_7 is tuned through resonance. Care should be used to keep the wires symmetrical with respect to the two sections of the tube.

After neutralization, plate and screen voltage may be applied. If possible, the plate voltage should be low at first trial so there will be no danger of overloading the tube. Adjust C_7 to resonance, as indicated by minimum plate current (this should be measured independently of the screen); with the 829, the minimum plate current should be in the neighborhood of 80 milliamperes with 400 volts on the plate and no load on the circuit. A dummy load such as a 60-watt lamp should light to something near full brilliance when the coupling between L_3 and L_4 is adjusted to make the tube draw a plate current of 200 ma. When the loading is set, the grid current should be checked to make sure it is up to the rating for the tube. If it has decreased, the coupling between L_1 and L_2 should be increased to bring it back to normal.

Power-supply and modulator requirements will depend upon the particular tube used. For the 829, the plate supply should have an output voltage of 400 to 500 with a current capacity of 250 milliamperes. With a 400-volt supply the modulator power required is 50 watts, with an output transformer designed to work into a 1600-ohm load; with a 500-volt supply slightly over 60 watts of audio power is needed, the load being 2000 ohms.

Transceivers

The transceiver is a combination transmitter-receiver in which, by suitable switching of d.c. and audio circuits, the same tube and r.f. circuit functions either as a modulated transmitting oscillator or as a superregenerative detector. This makes for extreme compactness and light weight, making the transceiver popular for hand-carried portable equipment. It is a compromise with respect to other features, however. The transceiver can be a source of serious interference, and its efficiency is not equal to that of other types of gear wherein separate tubes and circuits are used for transmission and reception.

As a matter of good amateur practice the use of transceivers should be confined to very low-power operation — as in "walkie-talkie" or "handie-talkie" equipment — in the 144-Mc. band, and to experimental low-power operation in the higher-frequency bands. The use of transceiver-type equipment should be avoided entirely for regular operation on the 144-Mc. band.

A Complete 144-Mc. Portable-Mobile Station

The transmitter shown in Figs. 1633 to 1637, inclusive, is designed to be used for portable or mobile operation in conjunction with a vibrator power supply giving 100 ma. at 300 volts. Separate tubes are used for the r.f. sections of the receiver-transmitter combination. The oscillator is operated at 15 watts input and delivers approximately four watts of power output.

As shown in Fig. 1637, the oscillator uses an HY-75 tube in the ultraudion circuit, using high C to improve the carrier stability and reduce frequency modulation. Coupling between the oscillator and the antenna is by means of a variable link and a short length of coaxial cable connected between the link and the antenna switch, S_{1B}. In general, the oscillator circuit is similar to the one shown in Fig. 1625, and additional information on adjustment will be found in the description of that oscillator.

A 6C4 triode tube is used in the superregenerative detector. Fairly high C is used in this circuit because the 6C4 superregenerates more smoothly with this arrangement. A variable link, mounted on a polystyrene pillar, is connected to the antenna switch by means of coaxial cable. The circuit goes into superregeneration with 50 to 60 volts on the plate element, and the resulting low power input causes relatively low receiver radiation.

For transmission, the audio section of the unit employs a single-button carbon microphone working into a 6C5 Class A driver stage, transformer-coupled to a 6Y7G Class B modulator. With a 6-volt battery, the microphone output is more than adequate for full power output from the speech system. The Class B modulator gives higher power efficiency and lower average plate current than a Class A modulator and, as a result, the proportion of

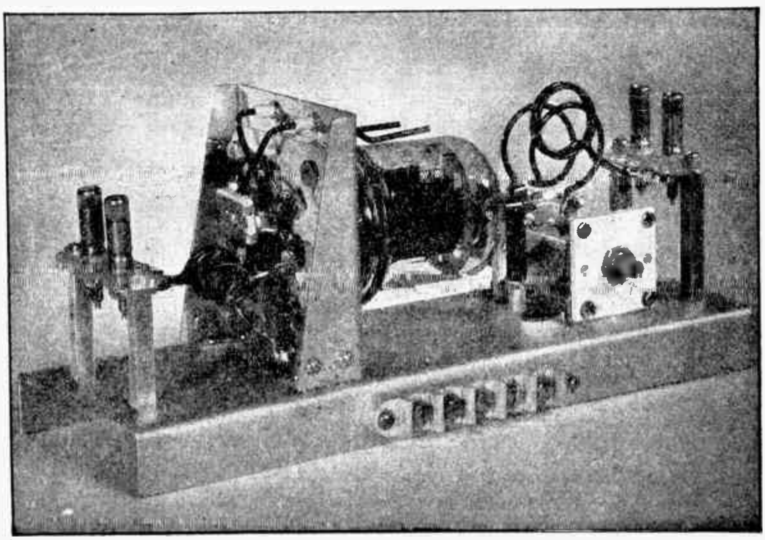


Fig. 1633 — Another view of the 144-Mc. amplifier. The neutralizing wires are crossed over the socket before going through the feed-through insulators. The input circuit is designed for link coupling to the driver stage.

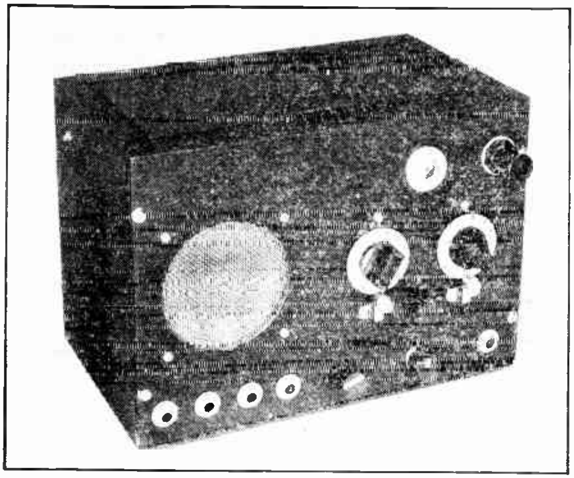


Fig. 1634 — A front view of the complete 144-Mc. portable-mobile station.

the limited power-supply output current which must be reserved for the audio section is relatively low. The 6Y7G requires a plate-to-plate load resistance of about 14,000 ohms. The oscillator, operating with 300 volts at normal current, represents a load impedance of 6000 ohms, so that the primary-to-secondary impedance ratio required in the coupling transformer is 2.3 to 1. With the transformer specified a close approximation to this ratio is secured when the taps specified for matching 10,000 ohms to 4500 ohms are used. The 6Y7G output is transferred from the headphone-'speaker circuit to the oscillator plate circuit by means of one section of the send-receive switch.

The microphone winding of the transceiver transformer, T_1, is opened when the unit is changed over to the receiving position and a second winding is cut in to complete the circuit between the detector and the 6C5 driver stage. With switch S_{1C} in the receive position the audio output is fed into the 'speaker. Headphones, plugged into the 'phone jack, J_5, will disconnect the 'speaker. R_6 is the regeneration control. The small padding condenser, C_3, increases the C in the circuit and serves for band-setting. The loading resistor, R_3, may or may not be required. It is used to prevent the howling which frequently is encountered at certain settings of the regeneration control. If necessary, the value of this resistor may be decreased to 22,000 ohms.

The transmitter is enclosed in a 6 × 7 × 10-inch metal cabinet. Most of the parts are mounted on a chassis measuring 1½ × 5½ × 9½ inches. Unfortunately, a 9-inch-long "U"-shaped chassis was not available and, as a result, the rolled-over edges at the front of the cabinet must be cut away to allow clearance for the chassis. The panel and chassis are fastened together by the mounting sleeves for the regeneration and gain controls and the jacks, as shown in Fig. 1636; the microphone jack, J_1, is the one at the right. The regeneration control is to the left of J_1 and the gain control is next in line. The three metering jacks, J_2, J_3 and J_4 are at the left of the panel while the 'phone jack, J_5, is next to the gain control. From left to right along the center of the panel are the 'speaker, the oscillator frequency-control dial and the detector tuning dial. Both dials are equipped with locks to prevent frequency shift during mobile operation. The send-receive switch knob is centered below the two tuning controls. A coaxial antenna fitting is located at the top of the panel and to the left side of the dial light.

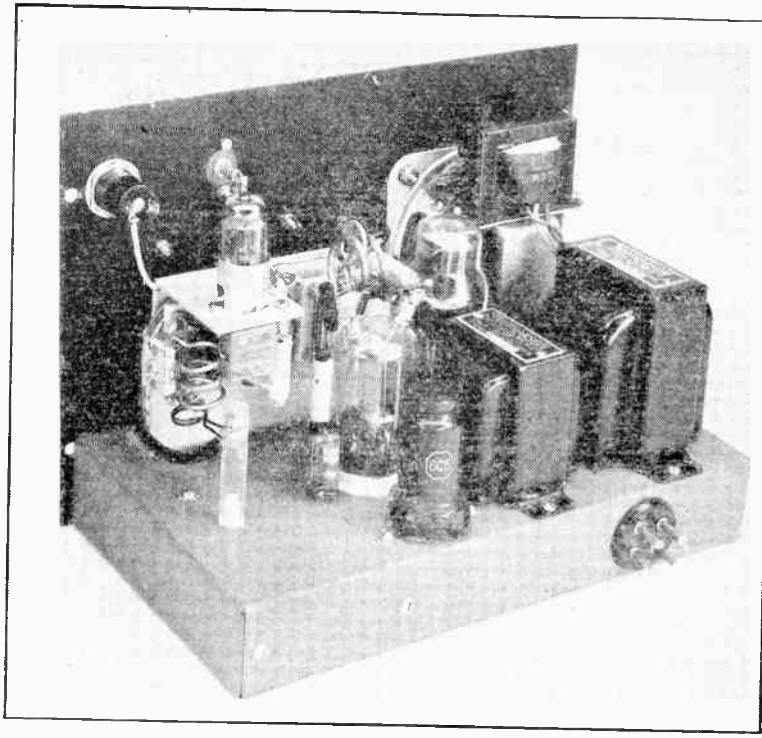
Fig. 1635 shows the chassis arrangement of the main components of the unit. The 6C5, T_2, and T_3 appear from left to right along the rear edge of the chassis. The 6Y7G is directly in back of T_2 and the 'speaker transformer, T_4, is mounted on the lugs provided in the 'speaker design. Power leads are brought to the four-prong plug mounted on the rear wall of the chassis.

The r.f. section is kept as compact as possible so short leads can be maintained. The tuning condensers, C_1 and C_2, are mounted on a chassis formed from 3/32-inch aluminum stock. This chassis is 4 inches wide, has a ¾-inch mounting surface at the bottom, and is 3 inches high. A section of the stock, measuring 1½ inches wide by 2 inches deep, is bent over at the top to form a mounting plate for the 6C4.

A small terminal block, made from sheet polystyrene, is mounted on spacers between the two variable condensers; the coaxial feed-line for the oscillator is moored at this point along with the hot end of the antenna link, L_2. The oscillator-plate r.f. choke, RFC_1, is mounted at the upper right-hand corner of the chassis and the tank coil, L_1, has its ends drilled to fit the condenser terminals on which it mounts. The grid r.f. choke, by-pass condenser, and grid leak are all mounted on the ¼-inch diameter rod shown at the left of the HY-75. The rod is drilled and tapped so that it can be mounted by means of the tube-socket mounting screw. The HY-75 socket is mounted below the chassis by means of ⅝-inch metal spacers.

The detector coil is mounted on the upper and lower stator-plate terminals at the left of the condenser. C_3 is connected across the same two points. The feed-back condenser, C_7, and the grid-leak resistor, R_2, are tied in parallel and mounted between the lower stator plate terminal of C_2 and the tube socket. The r.f. choke is at the rear-right corner of the shelf. C_8, the quench by-pass, is located between the r.f. choke and ground. The antenna coupling

Fig. 1635 — Rear view of the 144-Mc. transmitter-receiver.



coil, L_4, is supported by a polystyrene rod which has been drilled and tapped for chassis mounting. The detector end of the coaxial cable is soldered to the open ends of the link at the points where they protrude through the rod.

A bottom view of the chassis is shown in Fig. 1636. T_1, the microphone transformer, is mounted on the rear wall at the right end of the chassis. The 6C5 plate decoupling condenser, C_{10}, rests against the base to the front of T_1. C_9 and R_5, the 6C5 cathode condenser and resistor, are to the left of the tube socket at the rear of the chassis and the blocking condenser, C_{11}, is in front of the HY-75 socket. One of the spare pins of the HY-75 socket is used as a tie-point for the connection between C_{11} and switch, S_{1C}. The filament tuning condenser, C_4, is connected between Pin 2 of the oscillator socket and a soldering lug which is held in place by one of the tube-socket mounting nuts.

For mobile work only, the microphone voltage may be secured by connecting the hot end of the microphone transformer winding to any one of the 6-volt points inside the chassis. However, if operation with an a.c. supply is contemplated, it is necessary to bring out the transformer lead so that a microphone battery may be connected externally. This connection can be made by connecting the transformer lead through the input plug.

Plate current can be measured by using a 0–100 milliammeter fitted with a plug for the plate jacks, J_2, J_3 and J_4. The oscillator plate-current reading should be approximately 40 ma. with no antenna load connected and with 300 volts on the plate. The antenna coupling and tuning should be adjusted to obtain a full-load current of approximately 50 ma., using the loosest antenna coupling which will give the desired plate current.

The modulator plate-current reading should be about 25 ma. without speech and should rise to about 100 ma. on modulation peaks. Under full modulation the plate current of the oscillator will kick downward slightly, because of the lowered oscillator plate voltage caused by the power-supply regulation, as the modulator current increases.

The preliminary testing might well be carried on with a dummy load coupled to the oscillator. This procedure is recommended unless the transmitter frequency has been set

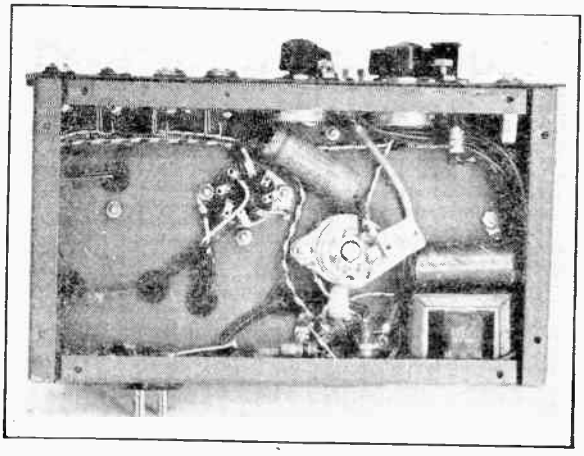


Fig. 1636 — A bottom view of the complete 144-Mc. portable-mobile unit.

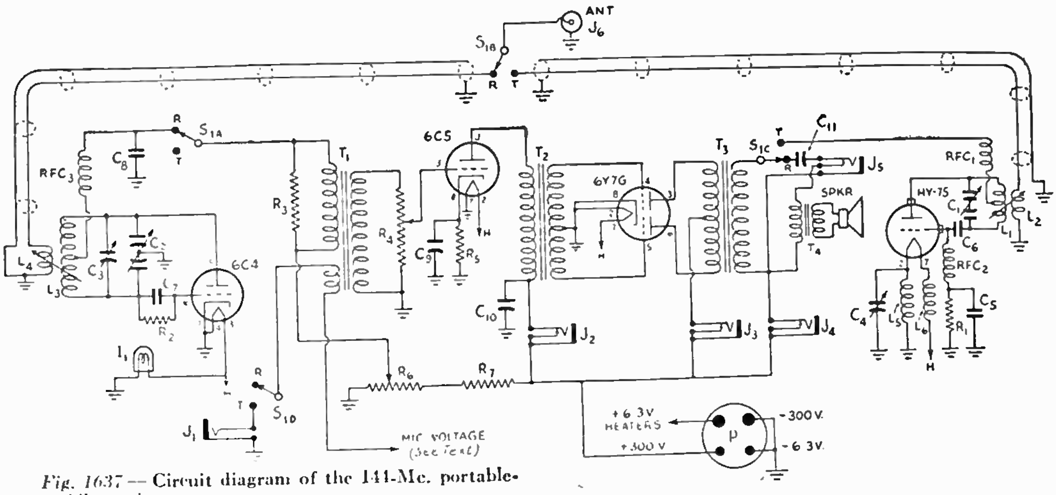


Fig. 1637 — Circuit diagram of the 144-Mc. portable-mobile station.

C_1 — Split-stator "butterfly" condenser, 14 μμfd. total (Cardwell ER-14-BF/SL).
C_2 — Split-stator condenser, 6 μμfd. total (Cardwell ER-6-BF/S).
C_3, C_4 — 3–30-μμfd. ceramic trimmer.
C_5 — 100-μμfd. midget mica.
C_6, C_7 — 47-μμfd. midget mica.
C_8 — 0.0047-μfd. mica.
C_9 — 8-μfd. 25-volt electrolytic.
C_{10} — 8-μfd. 150-volt electrolytic.
C_{11} — 0.1-μfd. 600-volt paper.
R_1 — 4700 ohms, 1 watt.
R_2 — 3.3 megohms, 1/2 watt.
R_3 — 47,000 ohms, 1 watt.
R_4 — 0.5-megohm volume control.
R_5 — 1000 ohms, 1 watt.
R_6 — 0.1-megohm potentiometer.
R_7 — 0.1 megohm, 1 watt.
L_1 — 2 turns of 1/8-inch copper tubing; inside diameter 5/8 inch; turns spaced approximately 5/16 inch between centers; plate tap at center.
L_2 — 1 turn No. 12 bare wire; inside diameter 1/2 inch.
L_3 — 3 turns No. 12 bare wire; inside diameter 1/2 inch, length 7/8 inch; plate tap at center.
L_4 — 2 turns No. 18 enameled wire; inside diameter 1/2 inch; turns spaced diameter of wire.
L_5, L_6 — 3 turns No. 18 d.c.c.; inside diameter 1/2 inch. L_5 interwound with L_6, no spacing between turns.
I_1 — 6-volt dial light.
J_1 — Midget open-circuit jack.
J_2, J_3, J_4, J_5 — Midget closed-circuit jack.
J_6 — Coaxial-cable connector.
P — 4-prong male plug.
RFC_1, RFC_2, RFC_3 — 1-inch winding of No. 24 d.s.c. or s.c.e. on 1/4-inch rod; rods drilled and tapped for mounting.
Spkr — 3-inch permanent-magnet dynamic 'speaker.
$S_{1A-B-C-D}$ — 4-pole double-throw switch.
T_1 — Transceiver transformer (UTC R-53).
T_2 — Interstage audio, single plate to push-pull grids (Thordarson T-19D06).
T_3 — Output transformer. 10,000-ohm primary to 4500-ohm secondary (Thordarson T-17M59).
T_4 — Output transformer, 4500-ohm primary to voice coil (UTC R-59).

inside the 144-Mc. band before the actual installation in the automobile is started. In any event, always check the frequency carefully each time before starting regular operation because the antenna loading will affect the frequency. Also, because the oscillator has a high-C tuned circuit, a small variation in the setting of C_1 will cause a considerable jump in frequency. It is wise to check the frequency whenever an adjustment of any kind is made. Frequency checking can be done with an absorption-type frequency meter, with Lecher wires, or by listening on a calibrated receiver.

Testing the detector for superregeneration is a simple matter inasmuch as the superregenerative hiss becomes plainly audible when the circuit goes into operation. It is possible that a component layout slightly different than that of the original model will necessitate some experimentation with the values of R_2, C_7 and C_8. Values which provide the smoothest regeneration and the greatest sensitivity should be selected. The padding condenser, C_3, should be adjusted so as to allow the tuned circuit, C_2L_3, to cover the 144-Mc. band. An accurate check on the frequency coverage can be made by employing any one of the instruments or methods suggested for calibration of the transmitter frequency.

A 300-volt 100-ma. vibrator-type power supply is recommended for mobile operation. The self-rectifying type is the least expensive and places the smallest load on the car battery. However, any supply that will deliver the necessary voltage and current will be quite satisfactory. An a.c. supply for testing purposes may also be provided; it should have the same output capabilities as the vibrator supply, and should include a filament transformer designed to deliver 6.3 volts at 3 to 3.5 amperes.

Under normal conditions the plate voltage applied to the transmitter and audio tubes will be the full power-supply output voltage minus the small *IR* drop caused by the audio-transformer windings. The 6C5 draws approximately 10 ma. and has a cathode bias of 8 to 9 volts. The 6C4, when superregenerating with 50 to 60 volts applied to the plate, will draw less than 1 ma. of plate current.

The antenna can be either a quarter-wave (19-inch) or a half-wave (39-inch) rod. Coaxial feed can be used with the short antenna and a two-wire tuned transmission line should be used with the half-wave radiator.

A Simple 235-Mc. Transceiver

The transceiver shown in Figs. 1639, 1640 and 1641 can be used either as a piece of fixed-station equipment or for portable-mobile work. The circuit diagram of the transceiver is shown in Fig. 1638. The detector-oscillator section of the unit employs a 6C4 triode in a high-C circuit similar to the one shown in Fig. 1622.

The audio section of the transceiver consists of a 6J5 driving a 6V6. With the send-receive switch in the "send" position the microphone circuit is closed while the audio input winding of T_1 and the 'speaker winding of T_2 are disconnected. The primary winding of T_2 becomes the modulation choke during transmission. Voltage for the single-button carbon microphone is developed across R_8, the 220-ohm resistor in the 6V6 cathode circuit.

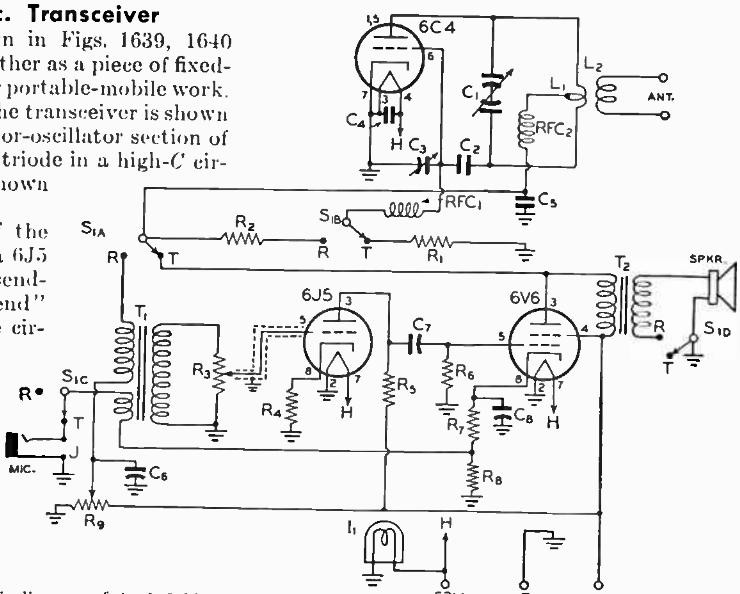


Fig. 1638 — Circuit diagram of the 235-Mc. transceiver.

C_1 — Tuning condenser; see text (Cardwell Type ER-14-BF/SL).
C_2 — 47-$\mu\mu$fd. midget mica.
C_3 — 3–30-$\mu\mu$fd. ceramic trimmer.
C_4, C_5 — 470-$\mu\mu$fd. midget mica.
C_6, C_7 — 0.1-μfd. paper, 400 volts.
C_8 — 25 to 50 μfd., electrolytic, 50 volts.
R_1 — 22,000 ohms, ½ watt.
R_2 — 4.7 megohms, ½ watt.
R_3 — 0.1-megohm volume control.
R_4 — 1000 ohms, ½ watt.
R_5 — 0.1 megohm, 1 watt.
R_6 — 0.1 megohm, ½ watt.
R_7 — 470 ohms, 1 watt.
R_8 — 220 ohms, 1 watt.
R_9 — 50,000-ohm volume control.
L_1 — 1 turn of 5/32-inch copper tubing, approximately horseshoe shape; over-all length from mounting holes, 15/16 inch; outside diameter at open end, 15/16 inch; place tap at center.
L_2 — 2 turns No. 14 or 16; diameter 7/16 inch.
I_1 — 6–8-volt pilot light.
J — Open-circuit jack.
RFC_1, RFC_2 — 1-inch winding of No. 24 d.s.c. or s.c.e. on ¼-inch diameter polystyrene rod.
$S_{1A\text{-}B\text{-}C\text{-}D}$ — 4-pole 2-position switch.
Spkr. — 3-inch permanent-magnet dynamic speaker.
T_1 — Transceiver transformer.
T_2 — Output transformer, pentode to voice coil.

The transceiver is housed in a utility cabinet of 5 × 6 × 9 inches. The front cover of the box is used as the panel. Fig. 1639, a front view of the unit, shows the location of the variable controls, antenna terminals, pilot light, microphone jack and 'speaker. All of these components, with exception given to the tuning condenser and dial, are mounted on the panel.

As shown in the rear view, Fig. 1641, the r.f. assembly is mounted on a small "L"-shaped chassis at the left end of the panel. This chassis, formed from 1/16-inch aluminum, has a width of 2 inches and is 2¼ inches high by 2 inches deep. The layout of this chassis is identical to the arrangement shown in Fig. 1620. C_1 is modified somewhat. One set of stator plates and one circular rotor plate are removed to make the total capacitance appropriate for 235 Mc.

The swinging antenna link shown at the

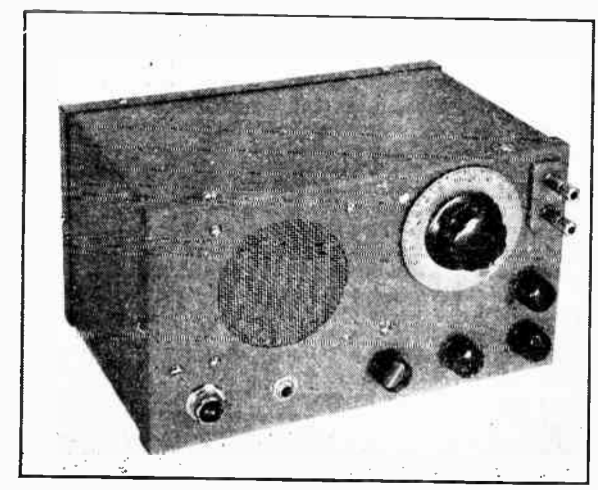


Fig. 1639 — A simple 235-Mc. transceiver. The 'speaker, tuning dial and antenna terminals are shown at the top of the panel. The pilot light, microphone jack, audio gain control, send-receive switch and regeneration controls run from left to right across the bottom of the panel. The swinging-link control is above the regeneration potentiometer. Ventilation holes are drilled in the rear panel.

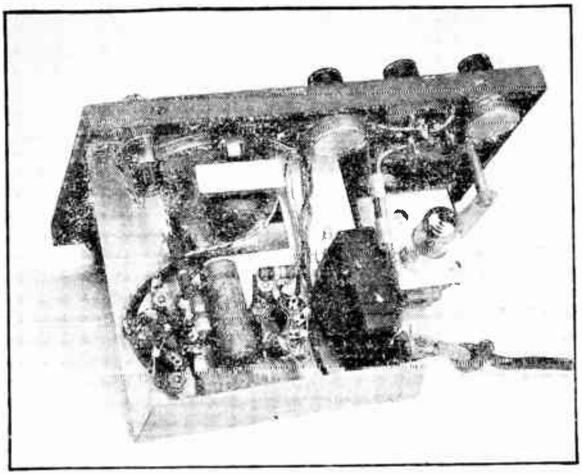


Fig. 1640 — A bottom view of the simple transceiver.

left of the r.f. section proved to be more of a refinement than a necessity, and may be replaced by a tightly-coupled fixed link. Grid leaks for the 6C4 are mounted directly on the send-receive switch contacts and all input and output leads for the circuit are cabled and fed to the switch and the audio sections.

The audio section, shown at the right in Fig. 1641, is mounted on an aluminum chassis cut from 1/16-inch stock five inches wide. It has a 1/2-inch lip at the front for securing to the panel and a 2-inch vertical member at the rear. A cut-out in the chassis provides clearance space for the speaker. The 6J5 audio-input tube is at the left of the audio section and the 6V6 output tube is on line with and to the right of the 6J5.

Looking at the bottom of the chassis, Fig. 1640, the output-tube socket is at the left with the 6J5 socket mounted at the rear center. The transceiver transformer, T_1, is at the right and the output transformer, T_2, can be seen protruding through the chassis at the back and center. The coupling condenser, C_7, rests on the base between the tube sockets and the 6V6 cathode by-pass condenser, C_8, lies against and parallel to the rear wall of the chassis. Spare tube-socket pins are used to secure resistor tie-points wherever convenient. Heater, ground and plate-voltage leads are terminated at a lug strip mounted at the right end of the chassis wall. An ordinary battery cable completes the circuits from these points to the external power supply.

It is possible that in a particular layout some experimentation with r.f. component values will be necessary. Effective superregeneration depends considerably on the grid choke, and the number of turns used should be adjusted so that the cold end can be touched with the finger without disturbing the operation of the oscillator. A grid-leak value allowing the smoothest operation should be selected and plate by-pass condenser values between 0.0022 and 0.0047 μfd. should be tried.

The inductance of the tuned-circuit coil, L_1, should be adjusted to bring the band on the dial by increasing or decreasing the length of the closed and open ends. The frequency may be checked as described in Chapter Nineteen. Coupling is adjusted by bending the leads of the antenna coil, L_2, to bring the coil nearer to or farther from L_1. The coupling should be adjusted so that with the switch in the "receive" position the oscillator goes into superregeneration smoothly; if the coupling is too tight it may not be possible to obtain superregeneration at all. The feed-back condenser, C_3, will have some slight effect on the frequency and this condenser should be adjusted for maximum oscillator output before the final frequency check is made. A 60-ma. dial light should be used to provide a means of judging r.f. output. A table listing approximate voltages at suggested test points is given below:

	Transmit	*Receive*
Input Terminal	335 volts	350 volts
Osc. plate	300 "	150 "
6J5 plate	75 "	80 "
6J5 cathode	2.2 "	2.3 "
6V6 plate	300 "	325 "
6V6 screen	335 "	350 "
6V6 cathode	0 "	0 "
Microphone circuit open	21 "	22 "
Microphone circuit closed	18.5 "	
Microphone voltage	1.9 "	

Filament: 6.3 volts at 0.9 amp.

Maximum total current drain is 55 ma. in "transmit" position and 45 ma. in "receive" position.

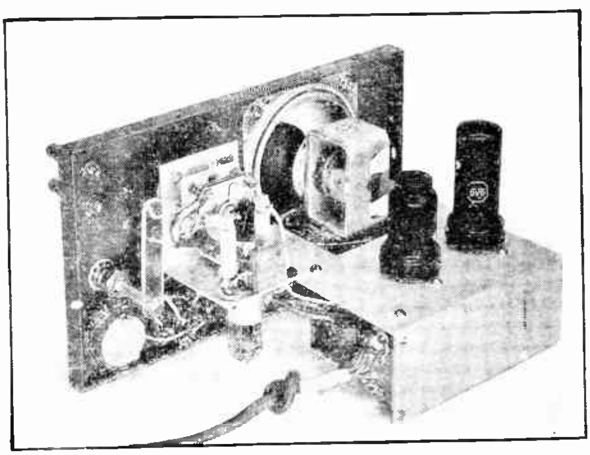


Fig. 1641 — An inside view of the 235-Mc. transceiver.

A Disk-Seal Tube Oscillator for 144, 235 and 420 Mc.

At frequencies above 300 Mc. or so tubes of conventional construction will not operate, for the reasons outlined at the beginning of this chapter. The disk-seal or "lighthouse" tubes will function nicely, however, in ordinary linear circuits at frequencies up to several hundred megacycles. No special types of circuit construction (such as cavity resonators) are required, therefore, when disk-seal tubes are used in the 420-Mc. band.

Details of construction of a transmitter using the 2C44, for operation in the 144-, 235-, and 420-Mc. bands, are given in Figs. 1642–1645, inclusive. Using parallel lines, it is only necessary to change the position of the shorting bar to obtain output on any of the three bands. The shorting bar is moved to a previously-calibrated point on the lines and locked, and then any frequency within the amateur band is obtained by proper setting of a tuning condenser connected across the lines at the point where they connect to the tube. The antenna coupling loop is connected to the shorting bar so that the two are moved simultaneously.

The circuit is shown in Fig. 1642. It will be recognized as the conventional circuit used in most 144-Mc. gear. The only critical component in the unit is RFC_2, the grid choke. There is an optimum value of choke for any one frequency, with which maximum output will be obtained at that frequency, but the value shown is a good compromise for the three-band range of this transmitter. The cathode is above ground by RFC_3 and RFC_4, but these inductors are not critical.

The transmitter is built on a 6 × 28 × 1-inch board. The "cold" ends of the ¼-inch rods used in the line are supported by two panel bushings mounted in an aluminum bracket which is fastened to the baseboard. These two panel bushings are of the locking type and make it a simple matter to position the rods properly. The plate rod is terminated at the plate and in a hole in the plate cap. The plate cap consists of a ½-inch length of ¾-inch diameter brass rod with a ⅜-inch hole drilled in the center and a ¼-inch hole drilled in the side. Holes are drilled at right angles to the large holes and tapped for 6–32 setscrews. The ⅜-inch hole fits over the plate cap of the tube, and the ¼-inch hole slides over the end of the plate rod. The grid half of the parallel line is approximately one inch shorter than the plate rod, to provide room for the grid condenser, C_2. The grid end of the line is supported by a small polystyrene post, and the grid socket is made by forming a narrow band of copper around the grid disk of the lighthouse tube and tightening it with a 2-56 machine screw and nut.

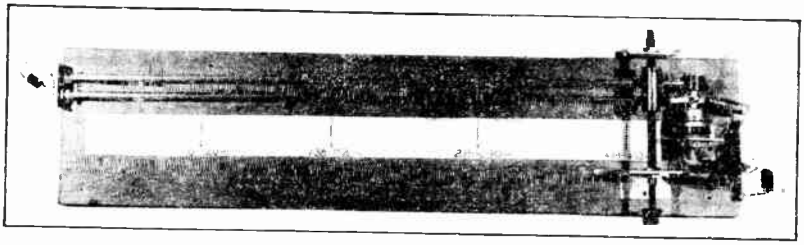


Fig. 1643 — A three-band oscillator (144, 235 and 420 Mc.) using the 2C44. The shorting bar on the parallel lines is moved to the proper point and locked, and tuning over the band is accomplished by the homemade variable condenser mounted at the ends of the lines near the tube.

The shorting bar for the parallel lines is made of two locking-type panel bushings set in a copper strap. These bushings are tightened just enough to insure good contact and still allow the bar to slide without too much effort. It is imperative, therefore, that the two rods be smooth and straight, although they can be either brass rod or brass tubing. The coaxial-cable connector for the antenna feed line and the antenna loop are mounted to a piece of $\frac{1}{16}$-inch bakelite bolted to the shorting bar. The antenna loop rides under the lines so that it will not hit the tuning condenser when the shorting bar is near the condenser. The size of the loop may vary with different antennas, but it should be about 2 inches long and spaced the same as the lines. The coupling can be increased by bending the loop closer to the lines.

The tuning condenser is of the split-stator type with the rotor floating. The stator plates consist of two strips of copper, $\frac{3}{16}$ inch wide by 1 inch long, formed in two arcs and soldered to the tuning rods (see Fig. 1645). The rotor

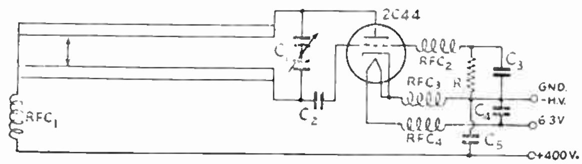


Fig. 1642 — Circuit diagram of the three-band oscillator.

C_1 — See text and Fig. 1645.
C_2 — 10-μμfd. ceramic.
C_3, C_4, C_5 — 100-μμfd. mica.
R_1 — 3300 ohms, 1-watt composition.
RFC_1 — 13 turns No. 18 enam., ¼-inch diam., close-wound.
RFC_2 — 25 turns No. 18 enam., ⅜-inch diam., spaced wire diam.
RFC_3, RFC_4 — R.f. choke (Ohmite Z-1).

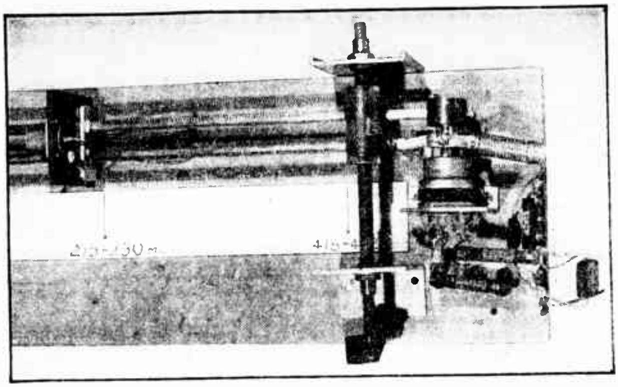


Fig. 1644 — A close-up view of the tuning condenser of the three-band oscillator also shows the details of the socket mounting and tube connections (W1DBM).

Oscillation can be determined by using a small neon bulb or a flashlight lamp and loop of wire held close to the lines. Grid current is also an excellent oscillation indicator. If no oscillation is obtained, it probably means an incorrect grid choke, and its construction should be checked or modified slightly. To get the best efficiency, particularly on any one band, may require some slight revision in the inductance of the grid choke or in the value of the r.f. by-pass capacitors, the effect of such changes being checked by watching the output as indicated by a lamp load and the input as indicated by a plate milliammeter. Tuning up should be done at reduced plate voltage, say around 250 or 300, at which value the loaded plate current should run around 15 to 20 ma., after which the maximum input of 40 ma. at 500 volts can be applied if considered necessary.

uses a piece of ¾-inch diameter polystyrene rod through which is drilled a ¼-inch diameter hole for a bakelite or polystyrene shaft. If desired, the solid polystyrene can be replaced by a ¾-inch diameter coil form by cementing a disk of polystyrene to the open end of the coil form.

The rotor plate, a "U"-shaped strip of copper one inch square, is formed and then cemented to the polystyrene form. A "U"-shaped piece is necessary because it was found that at 450 Mc. a cylindrical rotor acted as a capacitor plate as it was first brought near the stator plates, but as rotation continued the rotor began to act as a shorted turn in the field of the lines, thus counteracting the effect of the additional capacity and limiting the tuning range to only a small frequency variation. Two metal brackets with panel bushings are used to support the rotor shaft. It is a good idea to mount the panel bushings in slots rather than the usual clearance holes, so that the shaft can be moved toward the stator plates until the desired capacity range is obtained.

The tube socket is mounted on an aluminum bracket which is screwed to the baseboard. No connection is made to the r.f. cathode connection because the oscillator was found to work better over the entire range that way.

Forced ventilation must be used on the tube if anything like the rated maximum input of 20 watts is to be used. As much of the plate heat as possible must be conducted away by the plate rod, and for this reason the connection between plate and rod must be as good as possible from a heat as well as an electrical standpoint. The forced ventilation of the plate can best be obtained by the use of a small electric fan whose blast is directed at the plate connection whenever the plate power is applied. A small blower tube can be rigged up from stiff cardboard and attached to the fan.

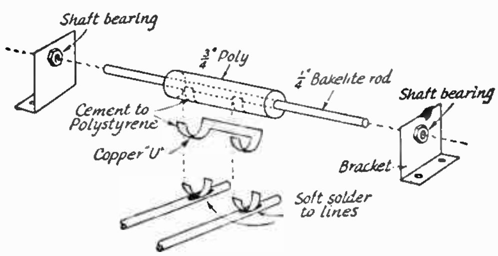


Fig. 1645 — Constructional and assembly details of the tuning capacitor for the 144/235/420-Mc. oscillator.

A good set of Lecher wires or an accurately-calibrated absorption wavemeter is essential for finding the different amateur bands. Although a wire line is probably the most convenient for the 144- and 235-Mc. bands, a more rigid line for the 420-Mc. band can be made by using ¼-inch rod or tubing, supporting it in the same manner that the tuned circuit is supported for the oscillator. After the oscillator has been calibrated, a cardboard scale can be added to the baseboard and the positions marked for the three amateur bands. The approximate settings of the shorting bar follow:

Distance from Center of Plate of 2C44 to Shorting Bar	*Frequency Range*
14 inches	138–152 Mc.
8¼ "	215–245 "
2½ "	418–452 "

Considerable care must be exercised in moving the shorting bar (and in removing the tube from its socket) because of the possibility of breaking the tube seals.

Chapter Seventeen

V.H.F. Antennas

Design Factors

While the basic principles of antenna operation are essentially the same for all frequencies, certain factors peculiar to v.h.f. work call for changes in antenna technique for the frequencies above 50 megacycles. Here the physical size of multielement arrays is reduced to the point where an antenna system having some gain over a simple dipole is possible in nearly every location, and experimentation with various types of arrays is an important part of the program of most progressive workers. The importance of high-gain antennas in v.h.f. work cannot be overemphasized. A good antenna system is often the sole difference between routine operation and outstanding success in this field. By no other means can so large a return be obtained from a small investment as results from the erection of a good directional array.

Beginning with the 50-Mc. band, the frequency range over which antenna arrays should operate effectively is often wider in percentage than that required of lower-frequency systems; thus greater attention must be paid to designing arrays for maximum frequency response, possibly to the extent of sacrificing other factors such as high front-to-back ratio. As the frequency of operation is increased, losses in the transmission line rise sharply; hence it becomes more important that the line be matched to the antenna system correctly. Because any v.h.f. transmission line is long, in terms of wavelength, it is often more effective to use a high-gain array at relatively low height, rather than to employ a low-gain system at great height above ground, particularly if the antenna location is not completely shielded by heavy foliage, buildings, or other obstructions in the *immediate* vicinity. This concept is in direct contrast to early notions of what was most desirable in a v.h.f. antenna system. An appreciable clearance above surrounding terrain is desirable, but great height is by no means so all-important as it was once thought to be. Outstanding results have been obtained by many v.h.f. workers, especially on 50 and 144 Mc., with antennas not more than 25 to 40 feet above ground.

Polarization

Practically all the early work on frequencies above 30 Mc. was done with vertical antennas, probably because of the somewhat stronger field in the immediate vicinity of a vertical system. When v.h.f. work was confined to almost pure line-of-sight distances, the vertical dipole produced a stronger signal at the edge of the working range than did the same antenna turned over to a horizontal position. With the advent of high-gain antennas and extended operating ranges, horizontal systems began to assume importance in v.h.f. work, especially in parts of the country where a considerable degree of activity had not already been established with verticals.

Numerous tests have shown that there is very little difference in the effective working range with either polarization, *if* the most effective element arrangements are used, and the same polarization is employed at both ends of the path. Vertical polarization still has its adherents among 50-Mc. enthusiasts and much fine work has been done with vertical antennas, but an effective horizontal array is somewhat easier to build and rotate. Simple 2-, 3- or 4-element horizontal arrays have proven extremely effective in 50-Mc. work, and the postwar era has seen an increase in the use of such arrays which has amounted to standardization on horizontal polarization.

The picture is somewhat different when one goes to 144 Mc. and higher. At these frequencies, the most effective vertical systems (those having two or more half-wave elements, vertically stacked) are more easily erected than on 50 Mc. Important, in considering the polarization question, is the existence of countless 144-Mc. mobile stations, whose antenna systems must, of necessity, be vertical. While horizontal polarization will undoubtedly find increased favor at 144 Mc. and higher, particularly for point-to-point work in rural areas, it is probable that vertical polarization will continue to dominate this field for some years to come. Under certain conditions, notably a station directly in the shadow of a hill, there may be a considerable degree of polarization shift, but ordinarily it may be assumed that best results in 144-Mc. work will be obtained by matching the polarization of the stations one desires to contact.

Impedance Matching

Because line losses tend to be much higher in v.h.f. antenna systems, it becomes increasingly important that feed lines be made as nearly "flat" as possible. Transmission lines commonly used in v.h.f. work include the open-wire line of 500 to 600 ohms impedance, usually spaced about two inches; the polyethylene-insulated flexible lines, available in

300-, 150-, and 72-ohm impedances; and coaxial lines of 50 to 90 ohms impedance. These may be matched to dipole or multielement antennas by any of several arrangements detailed below.

The "J" — Used principally as a means of feeding a stationary vertical radiator, around which parasitic elements are rotated, the "J" consists of a half-wave vertical radiator fed by a quarter-wave matching section, as shown at A, Fig. 1701. The spacing between the two

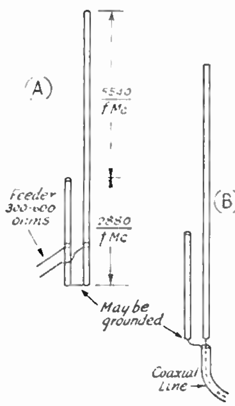


Fig. 1701 — Two versions of the "J" antenna, often used in mobile installations, or in vertical arrays where parasitic elements may rotate around a fixed radiator.

sides of the matching section should be two inches or less, and the point of attachment of the feed line will depend on the impedance of the line used. The feeder should be slid along the matching section until the point is found which gives the best operation. The bottom of the matching section may be grounded for lightning protection. A variation of the "J" for use with coaxial-line feed is shown at B in Fig. 1701. The "J" is also useful in mobile applications.

The delta or "Y" match — Probably the simplest arrangement for feeding a dipole or parasitic array is the familiar delta, or "Y" match, in which the feeder system is fanned out and attached to the radiator at a point where the impedance along the element is the same as that of the line used. Information on figuring the dimensions of the delta may be found in Chapter Ten. Chief weakness of the delta is the likelihood of radiation from the matching section, which may interfere with the effectiveness of a multielement array. It is also somewhat unstable mechanically, and quite critical in adjustment.

The "Q" section — An effective arrangement for matching an open-wire line to a dipole, or to the driven element in a 2- or 3-element array having wide (0.25 wavelength or greater) spacing, is the "Q" section (Chapter Ten). This consists of a quarter-wave line, usually of ½-inch or larger tubing, the spacing of which is determined by the impedance at the center of the array. The parallel-pipe "Q" section is not practical for matching multielement arrays to lines of lower impedances than about 600 ohms, nor can it be used effectively with close-spaced parasitic arrays. The impedance of the "Q" section required in these cases is lower than can be obtained using parallel sections of tubing, but a concentric line may be used for this purpose. A quarter-wave section of flexible coaxial line of 72 ohms impedance is a convenient arrangement for matching a 300 to 600-ohm line to the low center impedance of a 3- or 4-element array. The length of such a line will be approximately 65 per cent of a quarter wavelength or

$$L = \frac{1920}{f_{Mc}}$$

where L is the length of the line required, in inches. This figure takes into account the propagation factor of the solid-dielectric coaxial line. For the line made of parallel pipes, length in inches is determined by

$$L = \frac{2880}{f_{Mc}}$$

The "T" match — The principal disadvantages of the delta system can be overcome through the use of the arrangement shown in Fig. 1712, commonly called the "T" match. It has the advantage of providing a means of adjustment (by sliding the clips along the parallel conductors), yet the radiation from the matching arrangement is lower than with the delta, and its rigid construction is more suitable for rotatable arrays. It may be used with coaxial lines of any impedance, or with the various other forms of transmission lines up to 300 ohms. The position of the clips should, of course, be adjusted for maximum loading and minimum standing-wave ratio. The "T" system is particularly well suited for use in all-metal "plumbing" arrays.

The folded dipole — Probably the most effective means of matching a wide range of line impedances to almost any sort of parasitic array is the folded dipole (Figs. 1702, 1703, and 1704), described in Chapter Ten. A 300-ohm line may be used to feed a 4-element array (Fig. 1703) by using ¼-inch rod or tubing for the fed section of the folded dipole and 1-inch tubing for the parallel section. A 3-element array of the same general construction may be matched by using ¾-inch tubing for the parallel section of the dipole.

The impedance at the center of the system may also be increased by using three or more elements in parallel, the center impedance being increased approximately as the *square* of the number of elements used. This applies only if

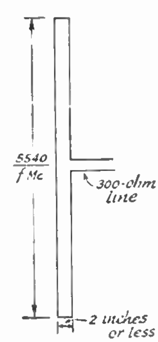


Fig. 1702 — Details of the folded dipole.

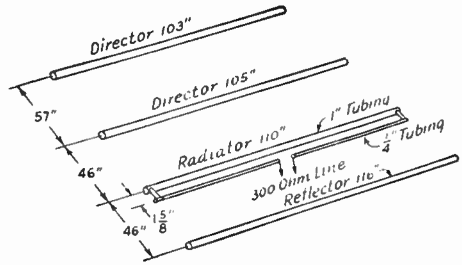


Fig. 1703 — Dimensional drawing of a 4-element 50-Mc. array. Element length and spacing were derived experimentally for maximum forward gain at 50.5 Mc.

the elements are all the same conductor size.

◖ Arrays for 50 Mc.

Since the same basic principles apply to all antennas regardless of frequency, little discussion is given here of the various simple dipoles which may be used when nondirectional systems are desired. Details of such antennas may be found in Chapter Ten, and the only modification necessary for adaptation to use on 50 Mc. or higher is the reduction in length necessary for increased conductor diameter at these frequencies.

A simple but effective array which requires no matching arrangement is shown in Fig. 1705. Its design takes into account the drop in center impedance of a half-wave radiator when a parasitic element is placed a quarter wavelength away. A director element is shown, as the drop in impedance using a slightly-shortened parasitic element is just about right to provide a good match to a 50-ohm coaxial line. The element lengths are not extremely critical in such a simple system, and the figures shown may be used with satisfactory results.

The importance of broad frequency response in any antenna designed for v.h.f. work cannot be overlooked. The disadvantage of all parasitic systems is that they tend to tune quite sharply, and thus are often effective over only a small portion of a given band. One way in which the response of a system can be broadened out is to increase the spacing between the parasitic elements to somewhat more than the 0.1 or 0.15 wavelength normally considered to provide optimum forward gain and highest front-to-back ratio. Some broadening may also be obtained by making the directors slightly shorter and the reflector slightly longer than the optimum value. The folded dipole is useful as the radiator in such an array, as its frequency response is somewhat broader than other types of driven elements.

A 4-element array for 50 Mc. having an effective operating range of about 2 Mc. is shown in Figs. 1703 and 1704. It employs a

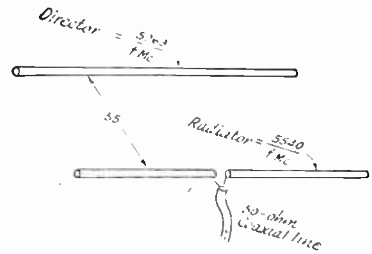
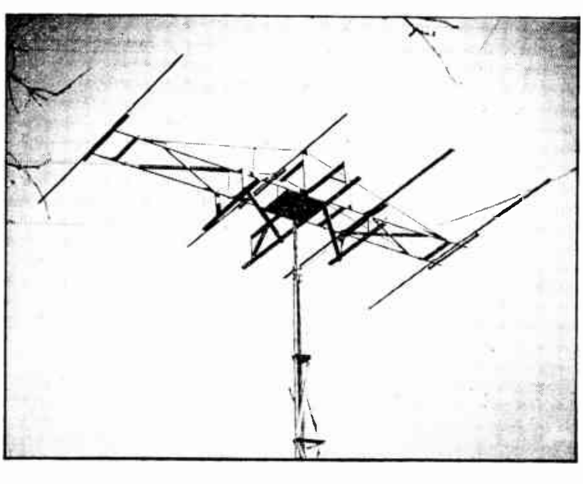


Fig. 1705 — A simple 2-element array for 50 Mc. No matching devices are needed with this arrangement.

folded dipole having nonuniform conductor size. Reflector and first director are spaced 0.2 wavelength from the driven element, while the forward director is spaced 0.25 wavelength. The spacing and element lengths given were derived experimentally, and are those which give optimum forward gain at the expense of some front-to-back ratio. As the latter quality is not of great value in 50-Mc work, it can be neglected entirely in the tuning procedure for such an array.

The dimensions given are for peak performance at 50.5 Mc. For other frequencies, the length of the folded dipole should be figured as recommended in Chapter Ten. The reflector will then be 5 per cent longer than the driven ele-

Fig. 1704 — The 4-element rotary array for 50 Mc. installed atop a steel tower. The frame extending below the main framework serves as a rotating device. The array frame is mounted on a pipe flange, to which is fitted a length of pipe which serves as a vertical support.



ment, the first director 5 per cent shorter, and the second director 6 per cent shorter. A broadening of the response may be obtained, at a slight sacrifice in forward gain, by adding to the reflector length and subtracting from the director lengths. For those interested in experimenting with element lengths, slotted extensions may be inserted in the ends of the various elements, other than the dipole, as shown in Fig. 1706. A 3-element array may be built, using the same general dimensions, except that the driven element, in this case, should have a ¾-inch diameter element in place of the 1-inch tubing used in the 4-element system.

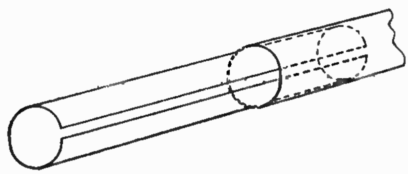


Fig. 1706 — Detail drawing of inserts which may be used in the ends of the elements of a parasitic array to permit accurate adjustment of element length.

Excellent results in work over distances up to 400 miles are being obtained by 50-Mc. workers using various more-complex directional arrays than the ones described above. The most important factor in such work is the attainment of the lowest possible radiation angle, and this purpose is well served by stacking of elements, in either vertical or horizontal systems. The use of two parasitic arrays, one a half wavelength above the other, fed in phase, provides a gain of 3 db. or more over that of a single array. The system shown in Fig. 1707 is excellent for either vertical or horizontal polarization, as is the "H" array, using four half-wave elements, with or without parasitic elements.

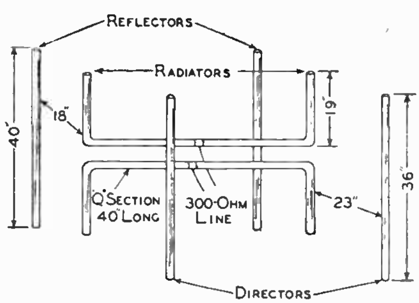


Fig. 1707 — A double-"Q" array for 144 Mc. The horizontal portion of the half-wave "H" acts as a "Q" section, matching the antenna impedance to the 300-ohm line attached at the center of the array. This array works well in either vertical or horizontal positions.

ℭ Arrays for 144 Mc.

Any of the above arrangements may, of course, be used for 144 as well as for 50 Mc., but, as two of them are designed for maximum effectiveness in a horizontal position, other designs may be used more effectively where vertical polarization is employed. To obtain the lowest radiation angle with vertical systems, those comprising a number of half-wave elements fed in phase are most useful. An important feature of such systems is that they are not so sharp in frequency response as are arrays having two or more parasitic elements in the same plane; consequently, adjustment of even quite complex systems such as the 16-element array shown in Fig. 1708 is not at all critical.

Plane reflectors are usable at 144 Mc., their size at this frequency being within reason. An interesting possibility in connection with this type of reflector is its use with two different sets of driven elements, one on each side of the reflecting screen. A set of elements arranged for vertical polarization may be used on one side, and a set of horizontally-polarized elements on the other, or the plane reflector may be made to serve on two different bands by a similar arrangement of elements for two frequencies, on opposite sides of the reflector. The screen need not be a solid sheet of metal, or even a close-mesh screen. A set of wires or rods arranged in back of the driven elements will work almost equally well. The dimensions of the reflector are not critical. For maximum effectiveness, the plane reflector should extend at least ¼ wavelength beyond the area occupied by the elements, but reflecting curtains no larger than the space occupied by the reflectors shown in Fig. 1708 have been used with good results.

In designing directional arrays having more than one driven element it is advisable to arrange for feeding the array at a central point. A simple 6-element array of high performance, incorporating this principle, is shown in Fig. 1707. All the elements may be made of soft-copper tubing, ¼ inch in diameter. The driven elements are comprised of two pieces which are bent into two "U"-shaped sections and arranged in the form of a half-wave "H." The horizontal portion of the "H" is then a double quarter-wave "Q" section, matching the impedance of the two radiators to that of the feed line. With the wide spacing used, the position of the parasitic elements is not particularly critical, except as it affects the impedance of the system, and the spacing of the elements may be varied to provide the best match. The spacing of the horizontal section may be varied for the same purpose. With the dimensions given, a spacing of one inch between centers is about right for feeding with a 300-ohm line. The radiation pattern of this array is similar in both horizontal and vertical planes; thus it will work with equal effectiveness in either position, provided the polarization is the same as that of the stations to be contacted.

By using a curtain of eight half-wave elements, arranged as shown in Figs. 1708 and 1709, backed up by eight reflectors, a degree of performance can be obtained which is truly

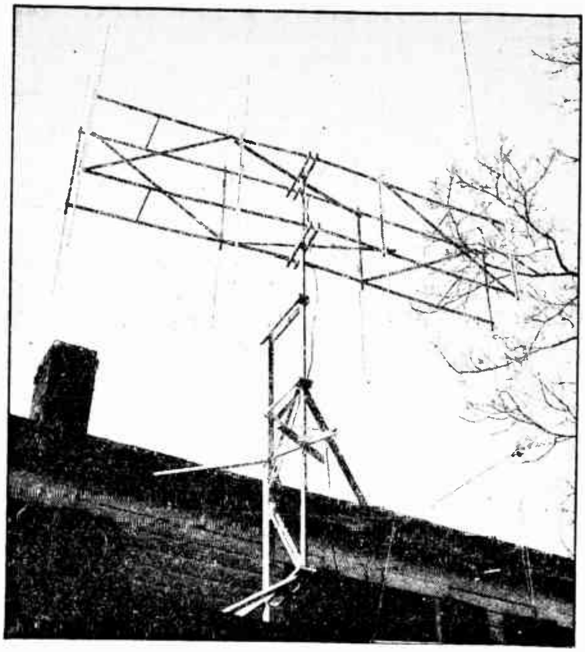


Fig. 1708 — A 16-element array for 144 Mc., showing supporting structure and "rotating mechanism." Sash cord wrapped three times around the criss-cross pulley permits 360-degree rotation.

outstanding. A gain of as much as 15 db. can be realized with such an arrangement, effecting an improvement in operating range which could never be obtained by any other means. Such an array is neither difficult nor expensive to construct, and its performance will more than repay the builder for the trouble involved in its construction.

The cumbersome nature of the structure required to support such an array would make its construction out of the question for a lower frequency, but for 144 Mc. the outside dimensions are only 1½ × 7½ × 10 feet, and the supporting frame can be made quite light.

The center pole (a 1½-inch rug pole 10 feet long) turns in three bearings which are mounted on braced arms extending out about two feet from a "two by three," which is braced in a vertical position. An improvised pulley made of two pieces of 1 × 2-inch "furring" notched in the ends and fastened criss-cross fashion near the bottom of the center pole serves as a "rotating mechanism." Sash cord wrapped three times around this "pulley" and run over to the window on small pulleys allows the beam to be rotated more than 360 degrees before reversal is required. To keep the array from twisting in high winds light sash cords are attached near each end of the supporting structure. These cords are brought through the window near the rotating ropes and are pulled up tight and fastened when the antenna is not in use.

The elements are of 7/16-inch soft-aluminum tubing for light weight. To stiffen the structure, and to help to maintain alignment, inserts were turned down from ½-inch polystyrene rod to fit tightly into the elements at the point where the cross-over or phasing wires are connected. Similar inserts are used for the reflector elements also. The interconnecting phasing sections are of No. 16 wire, spaced about 1½ inches. The feed line, connected at the center of the system, is Amphenol 21-056 Twin-Lead, 300 ohms impedance. The impedance at the center of the array is about right for direct connection of the 300-ohm line, without the necessity for a matching section of any kind. It is probably somewhat lower than 300 ohms, actually, and if a perfect match is desired, a "Q" section may be used. The performance is not greatly affected by such a change, as the standing-wave ratio is relatively low with the connection as shown.

The center section of the array may be used without the outside 8 elements, if space is limited, and a simpler array of good performance is desired. The simple "H" with reflectors may also be fed with 300-ohm line without the necessity for special matching devices.

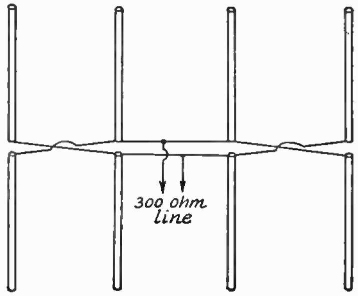


Fig. 1709 — Schematic of the radiating portion of the 16-element 144-Mc. array. Reflectors are omitted for clarity. Radiators are 38 inches long, reflectors 40.5 inches. Cross-over or phasing sections are also 40.5 inches long. Reflectors are mounted 17 inches in back of each radiator.

Mobile and Portable Antennas

A common type of antenna employed for mobile operation on 50 and 144 Mc. is the quarter-wave radiator which is fed with a coaxial line. The antenna, which may be a flexible telescoping "fish-pole," is mounted in any of several places on the car. The inner conductor of the coaxial line is connected to the antenna, and the outer conductor is grounded to the frame of the car. Quite a good match may be obtained by this method with the 50-ohm coaxial line now available; however, it is well to provide some means of tuning the system, so that all variables can be taken care of. The simplest tuning arrangement consists of a variable condenser connected between the low side of the transmitter coupling coil and ground, as shown in Fig. 1710. This condenser should have a maximum

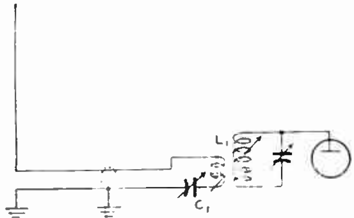


Fig. 1710 — Method of feeding quarter-wave mobile antennas with coaxial line. C_1 should have a maximum capacity of 75 to 100 μμfd. for 28- and 50-Mc. work. L_1 is an adjustable link.

capacity 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.

The short antenna required for 144 Mc. (approximately 19 inches) permits mounting the antenna on the top of the car. Such an arrangement provides good coverage in all directions, the car body acting as a ground plane. When the antenna is mounted elsewhere on the car, it is apt to show quite marked directional characteristics. Because of this it is desirable to make provisions for the use of the same antenna for both transmitting and receiving.

The best antenna possible for operation under mobile conditions is not particularly effective, as compared with antenna systems normally used in fixed-station work. To make the most of the fine opportunities for DX work afforded by countless high-altitude locations which are accessible by car, it is helpful to have some sort of collapsible antenna array which can be assembled "on the spot." Even a sim-

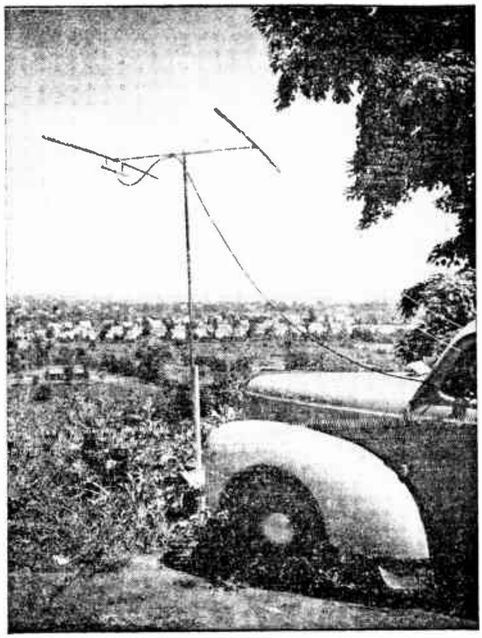


Fig. 1711 — A 2-element collapsible array for 50-Mc. portable use.

ple array like the one shown in Figs. 1711 and 1712 will effect a great improvement in the operating range of the low-powered gear normally used for mobile operation. This one is designed for 50-Mc. use, but similar arrangements can be made for other frequencies.

The array shown is a 2-element system, comprised of a radiator which is fed with coaxial line by means of a "T" match, and a reflector which is spaced 0.15 wavelength in back of the driven element. It is made entirely of ¾-inch dural tubing, except for the vertical support, which is 1-inch tubing of the same material. A suggested method of mounting is shown in Fig. 1711. A short length of 1 × 2-inch or larger wood is bolted to the car bumper. A piece of ¾-inch dural tubing is bolted to this upright, and the 1-inch vertical section of the array slips over the top of the ¾-inch section. The array is turned by means of ropes attached

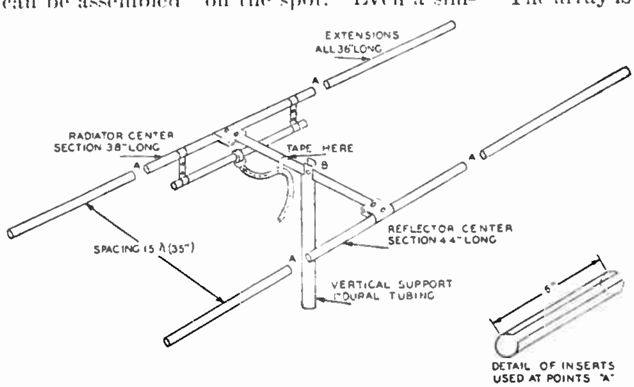


Fig. 1712 — Detail drawing of the collapsible 50-Mc. array, shown in Fig. 1711. All parts except the vertical support, which is 1 inch in diameter, are made of ¾-inch duralumin tubing. For carrying purposes, it is taken apart at Points "A" and "B," inserts of slotted dural tubing being used at Point "A" to hold the sections together. All extensions are the same length, the difference in element length being provided by the length of the center sections.

to the reflector element. Height of the array may be increased over that shown by using a longer wooden support, in which case it is desirable to use a 2 × 2 for greater strength. An anchoring pin made from a spike inserted in the bottom end of the wooden support is helpful to prevent tilting of the array. With such a device embedded in the ground, the whole assembly will remain rigid, which is helpful in the high winds usually encountered in mountain-top locations. Portability is provided by making the elements in three sections, with the end sections all the same length. The center section of the radiator is 6 inches shorter than that of the reflector.

The fed section of the "T" matching device is composed of two pieces of ¾-inch dural tubing about 14 inches long. The two sections are held together mechanically, but insulated electrically, by a piece of polystyrene rod which is turned down just enough to make a tight fit in the tubing. The inner and outer conductors of the coaxial line are fastened to the two inside ends of the matching section. Clips made of spring bronze are used for connection between the radiator and the "T." The position of these should be adjusted for maximum loading and minimum standing-wave ratio on the line. The idea for this array was suggested by W7OWX in "Hints and Kinks," April, 1946, *QST*, page 148.

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. 1713 was developed to eliminate feeder radiation. The center conductor of a 70-ohm concentric transmission line is extended one-quarter wave beyond the end of the line, to act as the upper half of a half-wave antenna. The lower half is provided by the quarter-wave sleeve, the upper end of which is connected to the outer conductor of the concentric line. The sleeve acts as a shield about the transmission line and very little current is induced on the outside of the line by the antenna field. The line is nonresonant, since its characteristic impedance is the same as the center impedance of the half-wave antenna. The sleeve may be made of copper or brass tubing of suitable diameter to clear the transmission line. The coaxial antenna is somewhat difficult to construct, but is superior to simpler systems in its performance at low radiation angles.

Cylindrical antennas — Radiators such as are used for television and broad-band f.m. are of interest in amateur v.h.f. operation because they work at high efficiency without adjustment throughout the width of an amateur band.

At the very-high frequencies an ordinary dipole or equivalent antenna made of small wire is purely resistive only over a very small frequency range. Its Q, and therefore its selectivity, is sufficient to limit its optimum performance to a narrow frequency range, and readjustment of the length or tuning is required for each narrow slice of the spectrum. With tuned transmission lines, the effective length of the antenna can be shifted by retuning the whole system. However, in the case of antennas fed by matched-impedance lines, any appreciable frequency change requires an actual mechanical adjustment of the system. Otherwise, the resulting mismatch with the line will be sufficient to cause significant reduction in power input to the antenna.

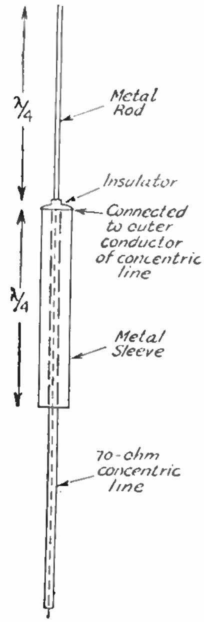


Fig. 1713 — Coaxial antenna. The insulated inner conductor of the 70-ohm concentric line is connected to the quarter-wave metal rod which forms the upper half of the antenna.

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 res-

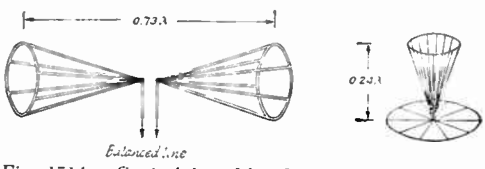


Fig. 1714 — Conical broad-band antennas have relatively constant impedance over a wide frequency range. The three-quarter wavelength dipole at left and the quarter-wave vertical with ground plane at right have the same input impedance — approximately 65 ohms. Sheet-metal or spine-type construction may be used.

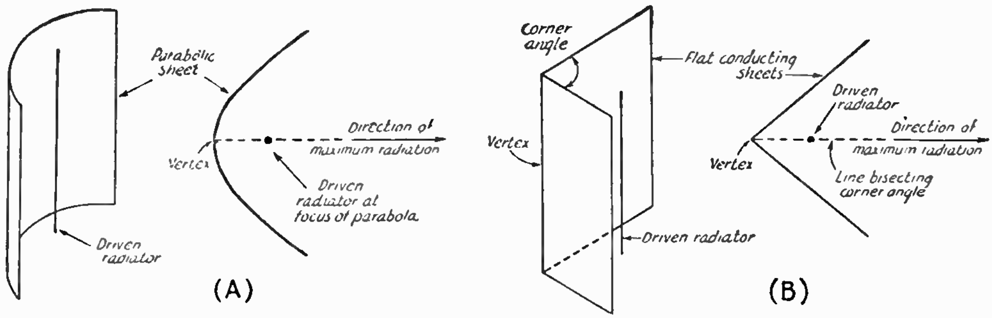


Fig. 1715 — Plane sheet reflectors for v.h.f. and u.h.f. A shows a parabolic sheet and B a square-corner reflector.

onant 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 characteristic impedance can be reduced to a very low value suitable for extremely wide-band operation. The cone may be made up either of sheet metal or of multiple wire spines, as in Fig. 1714.

Plane sheet reflectors — The small physical size of v.h.f. antennas makes practical many methods not feasible on lower frequencies. For example, a plane flat-sheet reflector may be used with a half-wave dipole, obtaining gains of 5 to 7 db. Much higher gains are attainable with a number of stacked dipoles, spaced ¼ or ¾ wavelength apart, and a larger reflecting sheet; such an arrangement is called a "billboard" array.

Plane reflectors need not be constructed of solid sheets. Wire mesh, or a grid of closely-spaced parallel-wire spines, is more easily erected and offers lower wind resistance.

Parabolic reflectors — A plane sheet may be formed into the shape of a parabolic curve and used with a driven radiator situated at its focus, to provide a highly-directive antenna system. If the parabolic reflector is sufficiently large so that the distance to the focal point is a number of wavelengths, optical conditions are approached and the wave across the mouth of the reflector is a plane wave. However, if the reflector is of the same order of dimensions as the operating wavelength, or less, the driven radiator is appreciably coupled to the reflecting sheet and minor lobes occur in the pattern. With an aperture of the order of 10 or 20 wavelengths, a beam width of 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 portion of the radiation.

Corner reflectors — The "corner" reflector consists of two flat conducting sheets which intersect at a designated angle. The corner reflector antenna is particularly useful at v.h.f. where structures one or two wavelengths in maximum dimensions are more practical to build than larger systems.

The plane surfaces are set at an angle of 90 degrees, with the antenna set on a line bisecting this angle. For maximum performance, the distance of the antenna from the vertex should be 0.5 wavelength, but compromise designs can be built with closer spacings. The plane surfaces need not be solid sheets; spines spaced about 0.1 wavelength apart will serve as well. The spines do not have to be connected together electrically.

If the driven radiator is situated on a line bisecting the corner angle, as shown in Fig. 1715, maximum radiation is in the direction of this line. There is no focus point for the driven radiator, as with a parabolic reflector, and the radiator can be placed at a variety of positions along the bisecting line.

Corner angles larger than 90 degrees can be used, with some decrease in gain. A 180-degree "corner" is equivalent to a single flat-sheet reflector. With angles smaller than 90 degrees, the gain theoretically increases as the corner angle is decreased. However, to realize this gain the size of the reflecting sheets must also be increased.

At a spacing of 0.5 wavelength from the driven dipole to the vertex, the radiation resistance of the driven dipole is approximately twice the radiation resistance of the same dipole in free space. Smaller spacings of driven dipole and vertex are practical, but at a slight sacrifice in efficiency. The alternative design for the 144- and 50-Mc. square-corner reflector has a dipole-to-vertex spacing of 0.4 wavelength. At this spacing the driven dipole radiation resistance is still somewhat higher than its free-space value, but is considerably less than when the spacing is 0.5 wavelength.

Chapter Eighteen

Emergency and Portable

EMERGENCY self-powered equipment is no longer a nice toy to play with when regular amateur activities pall; it has become the moral obligation of every amateur to be prepared in case of any communications emergency. Large-scale disasters in the past have demonstrated the tremendous value of amateur emergency stations in relaying relief messages when all other communication channels are closed. Aside from the all-important emergency phase, the use of portable equipment has been extended through organized activity in the annual ARRL "Field Days," and the problem of providing equipment suitable for use in rural districts, where commercial power is not available, has always been with us.

The most vital need for self-powered equipment occurs in connection with emergency activity, and the basic design of all such equipment should be predicated on emergency use. Every amateur, no matter where he may be located, can reasonably expect that sometime he may be called upon to perform emergency communications duty, and it is his responsibility to the public welfare, to himself, and to amateur radio as a whole to see that he is in some measure prepared.

It is not to be expected that every amateur will prepare himself for an emergency by having available a complete and separate self-powered station, although a large number of individuals and club groups do so. There is, however, no reason why every amateur cannot prepare his station for an emergency by having an emergency power supply ready and a quick means for utilizing all or part of his regular station equipment as an emergency-powered station. The emergency power supply can be anything from a small vibrator supply and/or batteries to a large gasoline-driven generator.

Battery and Vibrator Data

The use of dry batteries, storage batteries and vibrator-transformer packs or genemotors is discussed in Chapter Eight. Table I shows the service which may be expected from standard-brand dry batteries under various load conditions. Various types of manufactured vibrator-transformer units are listed in Table II, while Table III is a listing of available dynamotors which are suitable for emergency and portable work.

Construction of Vibrator Supplies

Vibrator-type power supplies are not difficult to construct. The transformer usually is a special type designed for the purpose, although a heavy-duty receiver or low-power transmitter transformer may be pressed into service if it has suitable filament windings which may be connected as the 6-volt vibrator primary. A supply may be designed to operate from a 6-volt storage battery only, or a dual-primary transformer or separate transformers may be changeably on either 115-volt a.c. or 6-volt d.c.

Typical circuit diagrams are shown in Fig. 1801. The one shown at A is the simplest, although it operates from a 6-volt d.c. source only. S_1 turns the high voltage on and off.

The circuit of B provides for either 6-volt d.c. or 115-volt a.c. operation with a dual-primary transformer. S_2 is the a.c. on-off switch while S_3 switches the heater of the 6X5 rectifier from the storage battery to the 6.3-volt winding on the transformer. Filament supply for the transmitter or receiver is switched by shifting the power plug to the correct output socket, X when operating from a 6-volt d.c. source and Y when 115-volt a.c. input is used.

The circuit of Fig. 1801-C may be used when a dual-primary transformer is not available. The filter is switched from one rectifier output to the other by means of the d.p.d.t. switch, S_4, which also shifts filament connections from a.c. to d.c. The filter section of the switch could be eliminated if desired by connecting the filtering circuit permanently to the output terminals of both rectifiers and removing the unused rectifier tube from its socket. Similarly, the filament section of S_4 could be dispensed with by providing two output sockets as in the circuit at B. If a separate rectifier filament winding is available on T_3, directly-heated rectifier types may be substituted for the 6X5 in the a.c. supply. In some cases where the required filament windings are not available, a rectifier of the cold-cathode type, such as the 0Z4, which requires no heater voltage, may be used to advantage.

If suitable filament windings are available, a regular a.c. transformer will make an acceptable substitute for a vibrator transformer. If the a.c. transformer has two 6.3-volt windings, they may be connected in series, their junction

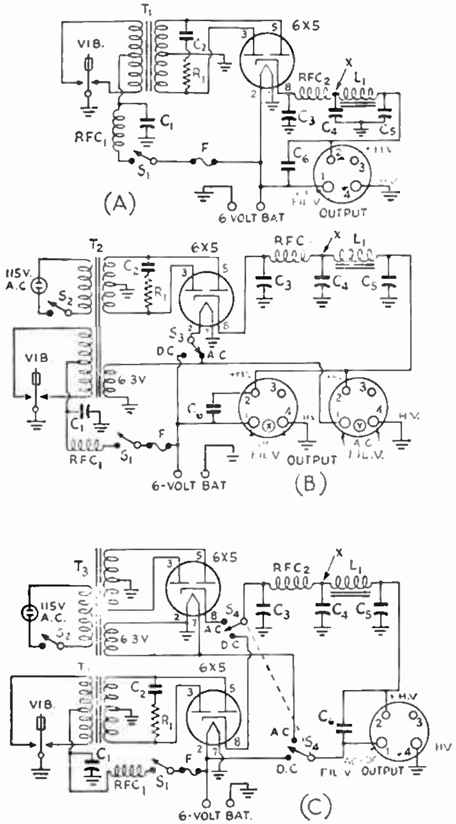


Fig. 1801 — Typical vibrator-transformer power-supply circuits. The circuit at A shows a simple arrangement for 6-volt d.c. input: the one at B illustrates the use of a combination transformer for operation from either 6 volts d.c. or 115 volts. a.c. The circuit of C is similar to that of B but uses separate transformers.

C_1 — 0.5-μfd. paper, 50-volt rating or higher.
C_2 — 0.005 to 0.01 μfd., 1600 volts.
C_3 — 0.01-μfd. 600-volt paper.
C_4 — 8-μfd. 150-volt electrolytic.
C_5 — 32-μfd. 150-volt electrolytic.
C_6 — 100-$\mu\mu$fd. mica.
R_1 — 4700 ohms, ½ or 1 watt.
L_1 — 10–12-henry 100-ma. filter choke, not over 100 ohms (Stancor C-2303 or equivalent).
F — 15-ampere fuse.
RFC_1 — 55 turns No. 12 on 1-inch form, close-wound.
RFC_2 — 2.5-mh. r.f. choke.
S_1 — S.p.s.t. toggle — battery switch.
S_2 — S.p.s.t. toggle — a.c. power switch.
S_3 — S.p.d.t. toggle — rectifier-heater change-over switch.
S_4 — D.p.d.t. toggle — a.c.-d.c. switch.
T_1 — Vibrator transformer.
T_2 — Special vibrator transformer with 115-volt and 6-volt primaries, to give approximately 300 volts at 100-ma. d.c. (Stancor P-6166 or equivalent).
T_3 — A.c. transformer, 275 to 300 volts each side of center-tap, 100 to 150 ma.; 6.3-volt filament.
VIB — Vibrator unit (Mallory 500P, 294, etc.)
X — Insert a series resistor of suitable value to drop the output voltage to 300 at 100-ma. load, if necessary. If transformer gives over 300 volts d.c., a second filter choke may be used to give additional voltage drop as well as more smoothing.
NOTE — All ground connections should be made to a single point on the chassis.

forming the required center-tap. A 6.3-volt and a 5-volt winding may be used in a similar manner even though the junction of the two windings does not provide an accurate center-tap. A better center-tap may be obtained, if a 2.5-volt winding also is available, since half of this winding may be connected in series with the 5-volt winding to give 6.25 volts.

R.f. filters for reducing hash are incorporated in both primary and secondary circuits. The secondary filter consists of a 0.01-μfd. paper condenser directly across the rectifier output, with a 2.5-mh. r.f. choke in series ahead of the smoothing filter. In the primary circuit a low-inductance choke and high-capacitance condenser are needed because of the low impedance of the circuit. A choke of the specifications given should be adequate, but if there is trouble with hash it may be beneficial to experiment with other sizes. The wire should be large — No. 12, preferably, or No. 14 as a minimum. Manufactured chokes such as the Mallory RF583 are more compact and give higher inductance for a given resistance because they are bank-wound, and may be substituted if obtainable. C_1 should be at least 0.5 μfd.; even more capacitance may help in bad cases of hash.

The 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 case and 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 reasonably well-shielded supply. A little care in this respect usually is more productive than experimenting with different values in the hash filters. Such experimenting should come *after* it has been found that radiation from the leads has been reduced to an absolute minimum. Shielding the leads is not particularly helpful.

The 100-$\mu\mu$fd. mica condenser, C_6, connected from the positive output lead to the "hot" side of the "A" battery, may be helpful in reducing hash in certain power supplies. A trial is necessary to see whether or not it is required. It should be mounted right on the output socket.

Testing for methods of eliminating hash should be carried out with the supply operating a receiver. Since the interference usually is picked up on the receiver antenna leads by radiation from the supply itself and the battery leads, it is advisable to keep the supply and battery as far from the receiver as the connecting cables will permit. Three or four feet should be ample. The microphone cord likewise should be kept away from the supply and leads.

The smoothing filter for battery operation can be a single-section affair, but there will be

TABLE I—BATTERY SERVICE HOURS

Estimated to 34-volt end-point per nominal 45-volt section.
Based on intermittent use of 3 to 4 hours daily.
(For batteries manufactured in U.S.A. only.)

Manufacturer's Type No.		Weight		Current Drain in Ma.											
Burgess	Eveready	Lb.	Oz.	5	10	15	20	25	30	40	50	60	75	100	150
—	386	14	—	2000	1100	690	510	400	320	200	170	130	100	50	30
	486	13	5	1700	880	550	395	300	240	165	125	100	70	45	20
21308	—	12	8	1600	1100	690	490	—	300	200	—	100	—	50	25
	586	12	2	1400	800	530	380	260	185	130	85	60	40	30	14
10308	—	11	4	1300	700	520	350	—	—	130	—	90	—	42	18
	585	8	13	900	450	290	210	130	100	60	45	25	20	11	5
2308	—	8	3	1100	500	330	180	—	100	65	—	34	—	—	—
B30	—	2	8	350	170	90	50	—	21	15	—	—	—	—	—
	762	3	3	320	140	81	54	37	27	—	—	—	—	—	—
	482	2	—	320	140	81	54	37	27	—	—	—	—	—	—
A30	—	2	—	210	80	44	24	—	14	5	—	—	—	—	—
	738	1	2	160	70	30	20	10	7	—	—	—	—	—	—
Z30N	—	1	4	155	70	30	20	15	7.5	—	—	—	—	—	—
	733	—	10	50	20	11	7	5.2	—	—	—	—	—	—	—
W30FL	—	—	11	45	19	12	7	—	3.5	—	—	—	—	—	—
	455[1]	—	8.6	70	20	11	7	5.2	—	—	—	—	—	—	—
XX30	—	—	9	70	20	12	7	—	3.5	—	—	—	—	—	—

[1] Same life figures apply to 467, 67½-volt, 10.5 oz.

Estimated to 1-volt end-point per nominal 1.5-volt unit. Based on intermittent use of 3 to 4 hours per day at room temperature. (For batteries manufactured in U.S.A. only.)

Manufacturer's Type No.		Weight		Voltage	Current Drain in Ma.											
Burgess	Eveready	Lb.	Oz.		30	50	60	120	150	175	180	200	240	250	300	350
—	A-1300	8	4	1.25	—	—	—	—	2000	1715	1500	1333	1250	1200	1000	854
—	740	6	12	1.5	—	—	—	—	1400	1200	—	1050	—	775	625	—
—	741[1]	2	14	1.5	—	—	1100	750	—	—	—	375	300	275	215	175
—	743	2	1	1.5	—	—	750	325	—	—	—	245	—	180	135	110
—	7111	2	2	1.5	—	—	700	320	—	—	200	—	120	—	90	—
—	742	1	6	1.5	—	—	500	325	—	—	155	135	100	95	85	50
8F[2]	—	2	10	1.5	—	—	1100	680	450	—	—	400	—	320	230	190
4FA[3]	—	1	4	1.5	—	—	600	350	220	—	—	160	—	110	90	60
—	A-2300	15	8	2.5	—	—	—	—	2000	1715	1500	1333	1250	1200	1000	854
—	723	1	—	3.0	—	—	240	100	—	—	70	—	40	—	30	—
20F2	—	13	12	3.0	—	—	—	—	1000	—	—	750	—	700	600	500
2F2H	—	1	6	3.0	600	—	340	130	95	—	—	60	—	42	30	—
2F2BP[4]	—	1	5	3.0	600	—	340	130	95	—	—	60	—	42	30	—
F2BP	—	—	12	3.0	340	—	130	45	30	—	—	—	—	—	—	—
G3[5]	—	1	5	4.5	370	—	150	50	35	—	—	—	—	—	—	—
—	746	1	3	4.5	—	200	—	—	—	—	—	—	—	—	—	—
—	718[6]	3	—	6.0	—	375	—	—	—	—	—	—	—	—	—	—
F4PI	—	1	6	6.0	340	—	130	45	30	—	—	—	—	—	—	—

[1] Same life figures apply to 745, wt. 3 lbs.
[2] Same life figures apply to 8FL, wt. 2 lbs. 15 oz.
[3] Same life figures apply to 4F, wt. 1 lb. 5 oz.
[4] Same life figures apply to 2F4, volts 6, wt. 2 lbs. 11 oz.
[5] Same life figures apply to G5, volts 7½, wt. 2 lbs. 2 oz.
[6] Same life figures apply to 747, wt. 3 lbs.

If batteries of another make are to be used, locate ones of similar size and weight on these tables and comparable performance may be expected.

some hum (readily distinguishable from hash because of its deeper pitch) unless the filter output capacitance is fairly large — 16 to 32 μfd.

A typical example of vibrator-supply construction is shown in the photographs of Figs. 1802 and 1803.

All components in the supply with the exception of the four-prong outlet socket are mounted on a piece of quarter-inch tempered Masonite measuring 3¾ × 9 inches. This fits into a plywood box having inside dimensions (3¾ × 9 × 5½ inches) just large enough to contain the equipment. The Masonite shelf rests on ¾-inch-square strips, 1¼ inches long, glued to the corners of the box at the bottom. The top and bottom of the box are removable. To provide shielding and thus reduce hash troubles, the box is covered with thin iron salvaged from 5-quart oil cans. Where the edges bend around the box to make a joint, the lacquer is rubbed off with steel wool so the pieces make electrical contact, and the metal is tacked to the plywood with escutcheon pins.

To make sure that the shielding will be

TABLE II—VIBRATOR SUPPLIES

Manufacturer's Type Number					Output		Rectifier	Output Filter
American Television and Radio Co.	Electronic Labs	Halli-crafters	Mallory	Radiart	Volts	Ma.		
VPM-F-7					90	10	Syn.	Yes
			VP-551[1]		125–150–175–200	100 max.	Syn.	No
				4201B[2]	250	50	Syn.	Yes
			VP-540		250	60	Syn.	Yes
				4204F[3]	100–150–250	35–40–60	Syn.	Yes
	605				150–200–250–275	35–40–50–65	Syn.	No
	604[4]		VP-552[5]		225–250–275–300	50–65–80–100	Syn.	No
				4201[6]	150–200–250–275–300	35–40–50–70–100	Syn.	No
	251[7]				300	100	Tube	Yes
			VP-555		300	200	Tube	Yes
VPM-6[8]	311[9]				250–275–300–325	50–75–100–125	Tube	Yes
		VP-2			300	170	Tube	No
		VP-4			320	70	Tube	No
			VP-557		400	150	Tube	Input cond.
				4202D	300–400	200–150	Tube	Yes
	606[10]				325–350–375–400 and 110 a.c. 60 cycle	125–150–175–200 20 watts	Tube	Input condenser

All inputs 6.3 volts d.c. unless otherwise noted.

[1] VP-553 same with tube rectifier.
[2] In weatherproof case. 4201B2 same with tube rectifier.
[3] 180-cycle vibrator, lightweight. 4204 same without filter.
[4] 601 same with tube rectifier; 602 same except 12 v. d.c. input and tube rectifier; 603 same except 32 v. d.c. input and tube rectifier.
[5] VP-554 same with tube rectifier; VP-G556 same except 12 v. d.c. input; VP-F558 same except 32 v. d.c. input.
[6] 4200D same with tube rectifier; 4200DF same with tube rectifier and output filter.
[7] 551 same with 12 v. d.c. input.
[8] Also available without filter.
[9] 511 same except 12 v. d.c. input.
[10] Input 6 v. d.c. or 110 v. a.c., 607 same except 12 v. d.c. or 110 v. a.c. input; 608 same except 32 v. d.c. or 110 v. a.c. input; 609 same except 110 v. d.c. or 110 v. a.c. input.

TABLE III—DYNAMOTORS

Manufacturer's Type No.			Input		Output		Weight Lbs.
Carter	Eicor	Pioneer	Volts	Amps.	Volts	Ma.	
210A			6	6.1	200	100	6½
MA250	102[1]	E1W272[2]	6	4.2	250	50	6½
251A		E1W339[3]	6	7.9	250	100	6½
301A	106[4]	E2W351[5]	6	9.7	300	100	6½
315A	158[6]	E2W243[5]	6	13.4	300	150	7⅞
320A		RAOW158[7]	6	18.2	300	200	9½
MA301			6	9	300	100	
351A			6	10	350	100	6½
355A	108	E2W256[5]	6	15	350	150	7⅞
352AR			6	22	350	200	9½
401A			6	13	400	100	7⅞
		E2W438	6	14.2	400	125	9¼
415A	109[8]		6	20	400	150	7⅞
420A			6	23.4	400	200	9½
425A		RA1W201[9]	6	30	400	225	9½
V450			5.5	29	400	250	
A430			6	31	400	300	13
		E3W413	6	15	500	100	11
520AS		RA1W189[10]	6	27.4	400	250	—
A650			6	40	600	250	13
AFS630			6	46.4	600	300	13

[1] Input current 4.6 amp.; wt. 4⅝ lbs.
[2] Wt. 7½ lbs.
[3] Input current 7.5 amp.; wt. 7½ lbs.
[4] Wt. 5 lbs.
[5] Wt. 9⅓ lbs.
[6] Input current 14 amp.; wt. 5¾ lbs.
[7] Wt. 16 lbs.; input current 18 amp.
[8] Input current 17 amp.
[9] Wt. 17½ lbs.; input current 25 amp.
[10] Input current 27 amp.; wt. 17½ lbs.

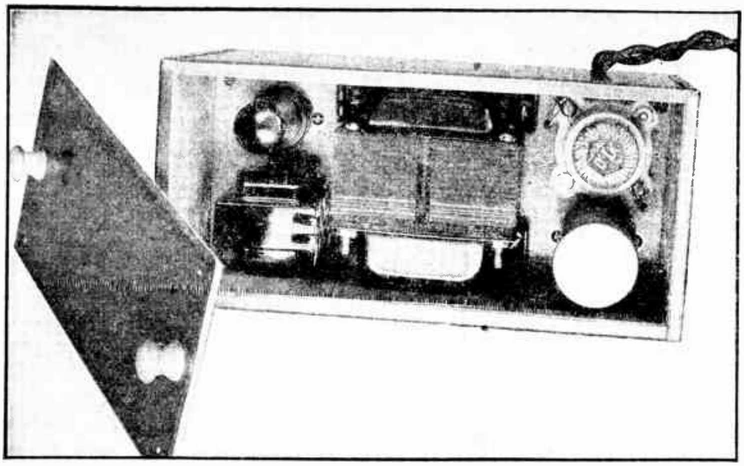


Fig. 1802 — A view inside a typical vibrator-type power supply. The rectifier tube is at the upper left with the filter choke just below. The primary fuse socket and vibrator are at the right. A synchronous-type vibrator may be substituted for the interrupter type if it is desired to eliminate the rectifier tube.

complete, the top and bottom of the box slide into place from the side, with the metal covering extending out so that it fits tightly under a lip bent over from the metal on the sides. These lips also are cleaned of lacquer to permit good electrical contact. The general construction should be quite apparent from the photographs. The bottom is provided with rubber feet, and the top has a small knob at each end so that it can be pulled out. This is essential, since the fit is good and there is no way to get either the top or bottom off, once on, without having some sort of handle to grip.

Charging Storage Batteries

If access to a.c.-operated chargers is not possible at times between actual use, some form of self-powered charging system is essential.

This need is ordinarily best met by a gasoline- or wind-driven generator. Water-power generators have been used, but their dependence on special circumstances is obvious, and they are not available in small sizes.

The wind charger consists of a small generator driven by a suitable impeller, mounted to take advantage of the free energy offered by the wind. The standard type will supply up to 16 amperes to a 6-volt battery. It will ordinarily keep fully charged a battery used to power a typical receiver and small transmitter operated from vibrator or genemotor supply in intermittent operation.

Gasoline-driven generators are also available for use in charging 6-volt or larger batteries. These ordinarily are rated at 150 or 200 watts. A ½- or ¾-h.p. single-cylinder 4-cycle engine is used, which will operate for 12 to 15 hours on a gallon of gasoline.

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. Ratings of typical gas-engine driven generator units are given in Table IV.

A variant on the generator idea is the use of

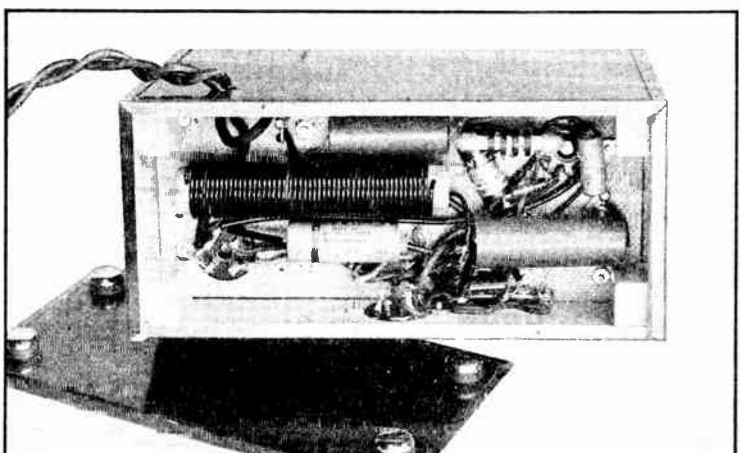


Fig. 1803 — Hash and smoothing filter components are mounted in the bottom of the low-voltage vibrator power supply. The 4-prong outlet socket is mounted on the side.

TABLE IV—GASOLINE-ENGINE DRIVEN GENERATORS, AIR-COOLED

Manufacturer				Output		Weight Lbs.	Starter
Eicor	Kato	Onan	Pioneer	Volts	Watts		
3AP6[1]			BD-6[1]	110 a.c. or 6 d.c.	300 200	100	Push-button
	JR-35[2]			110 a.c.	300	65	Push-button
	JRA-3[2]			110 a.c.	350	65	Rope crank
	19-A			110 a.c. or 6 d.c.	350 200	95	Push-button
		358[1,3]		115 a.c.	350	91	Push-button
	JR-10[2]			110 a.c.	400	—	Rope crank
		5L[3]		110 a.c.	500	165	Push-button
	23A			110 a.c. or 6 d.c.	500 200	105	Push-button
6AP1	14A		BA-6[1]	110 a.c.	600	135	Push-button
		7L[3]		115 a.c.	750	195	Push-button
10AP1		10L[3,4]	BA-10[1]	110 a.c.	1000	170	Push-button
	26A			110 a.c.	1000	265	Manual
		OTC		110 a.c.	1500	135	Manual
			BA-15	110 a.c.	1500	365	Push-button

[1] Also available in remote-control models.
[2] Intermittent-duty model.
[3] Also available in manual-started type.
[4] 115-volt output; weight 200 lbs.

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 amongst amateurs. Such generators are similar in construction and capacity to the small gas-driven units.

The home construction of generators of all the above types has been successfully attempted by amateurs at times, although the possession of a considerable knowledge of electric-motor design is essential. One especially useful possibility is the rewinding of old automobile charging generators, several hundred watts capacity being obtainable from the largest sizes. Those originally used on the old 4-cylinder Dodge cars have been successfully adapted by amateurs. Trade schools will often have their students rewind these generators for only the cost of the material, and this possibility is worth investigating.

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 line. 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.

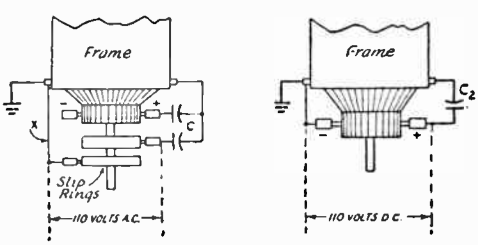


Fig. 1804 — Connections used for eliminating interference from gas-driven generator plants. C should be 1 μfd., 300 volts, paper, while C2 may be 1 μfd. with a voltage rating of twice the d.c. output voltage delivered by the generator. X indicates an added connection between the slip ring on the grounded 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. 1804, 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 are connected into the second detector.

High-Frequency Equipment

The use of high-frequency equipment for the handling of all intracommunity emergency communications is recommended not only for the purpose of limiting the interference range but also because equipment for these frequencies may be built in easily-portable form. Low-power transceivers and transmitter-receivers in the form of glove-compartment units, walkie-talkies and handie-talkies find ready application in this type of work.

Glove-compartment units and other forms of mobile installations may be operated readily from a vibrator supply or genemotor connected to the car storage battery, although a separate battery is recommended for protracted operating periods, such as in an emergency, to guard against discharging the car battery to the point where it will no longer start the car. The usefulness of a mobile unit in emergencies is apparent, since it constitutes a self-powered installation which may be placed in a strategic location with a minimum loss of time.

Handie-talkies and walkie-talkies, on the other hand have the advantage that they may be brought to points which for one reason or another may be inaccessible to a car. Handie-talkies universally operate from self-contained dry batteries, while the heavier walkie-talkie units may be designed to operate from either dry batteries or a small storage battery of the motorcycle type and a vibrator unit. In some cases, it may be desirable to build the power supply as a separate unit so that the weight which must be carried to the scene of an emergency may be distributed between two persons.

Higher-powered transmitters and more elaborate equipment of the type often used as permanent station equipment operating from a.c. are desirable as control-station equipment if a suitable source of power is available.

Portable Equipment — Low-Frequency

The weakest unit in a low-frequency portable or emergency communications installation often is the receiver.

An inadequate receiver, with poor selectivity, low sensitivity and insufficient stability, can ruin a QSO even under favorable conditions. When it is remembered that conditions in portable or emergency operation are often more severe than those at home, with poor antenna facilities, high noise levels, severe interference, etc., the fallacy of attempting to use an inferior portable receiver is apparent.

The best procedure of all is to use the home-station receiver for portable work. Headphones should be used and the output tube removed (if it isn't necessary for headphone operation), but this is no hardship. Headphones are far more satisfactory in such applications than the speaker in any event. This procedure not only ensures the availability of the high-performance receiver so vitally necessary, but the practice that has been obtained by using the receiver at home is invaluable in the specialized operating techniques of portable or emergency work. It takes as much experience to learn to run a receiver properly as it does to drive a car, and the middle of a crisis is no time to gain that experience. Even on lowered plate voltage the home superhet will be better than a makeshift set-up.

If a special portable/emergency receiver is to be built, it should be a superheterodyne. With present-day tubes and components, it is possible to build a simple superheterodyne as cheaply as a t.r.f. receiver, and there is no comparison between the two in performance. The average communications superheterodyne can be operated with storage-battery heater supply and dry-cell or vibrator-pack "B" supply. With the audio power tubes removed from the receiver, the power requirements are not too great. Some of the receivers on the amateur market have provision at the rear of the set for plugging in a d.c. supply, and those which do not can be easily modified by drilling a socket hole at the rear of the receiver and wiring it into the set. When regular a.c. operation is used, a plug in the socket completes the circuit.

The design of low-frequency transmitters for emergency, portable and rural transmitters, will depend almost entirely upon the power supply available. Considering possible defects in hastily-improvised radiation systems, etc., it seems unwise to use less than 10 watts input to a power amplifier or 15 watts to an oscillator. However, powers greater than two or three times these values are not usually necessary, so selection of the power supply will depend almost entirely upon the pocketbook and other resources. The 300-volt 100-ma. vibrator supplies and genemotors represent a nice compromise unless it is possible to step into the 200- or 300-watt gasoline-driven generator class.

Perhaps the best plan in providing for an emergency and portable transmitter is to utilize the basic exciter unit in the regular station. This not only ensures the availability of a reliable, efficient unit at all times but means a saving in parts and equipment. It represents no hardship to the permanent station to construct the exciter so it is compact, readily removable and, above all, solidly and dependably assembled. If your present exciter is not adaptable to this use, plan the new one so it will be. Provision for 6-volt tubes throughout is essential, with the heater circuit so arranged that it can be connected to a storage battery without change. A suitable plate supply using a vibrator or genemotor or similar system should be available separately, arranged for ready connection. The best method is to have a socket-and-plug connector assembly, with one plug built into the transmitter and another, wired identically, connected permanently to the emergency supply.

A Simple Modulator for Portable Work

The circuit diagram of a simple modulator for portable or mobile work is shown in Fig. 1805. In this arrangement the microphone is used directly to drive a pair of 6V6GT modulators without intermediate speech amplifiers. Such a modulator works surprisingly well to modulate Class C inputs up to 25 watts. The unit requires 75 to 100 ma. at 200 to 300 volts.

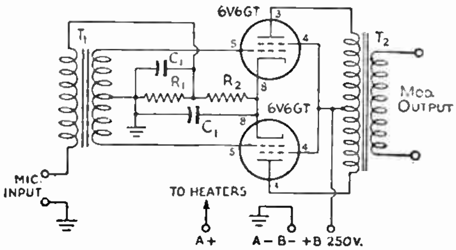


Fig. 1805 — Simple modulator for portable and general-utility work.
C_1 — 10-μfd. 25-volt electrolytic.
R_1 — 100 ohms, 1 watt.
R_2 — 150 ohms, 1 watt.
T_1 — Input transformer (Thordarson T-83A78).
T_2 — Output transformer (Thordarson T-19M13).

Voltage for the single-button carbon microphone is taken from the junction of the two cathode-biasing resistors, R_1 and R_2, thus eliminating the necessity for bulky microphone batteries. These two resistors could be replaced by a single resistor with a sliding contact. One side of the heater circuit is grounded so that only three power-supply wires are required. The complete unit may be assembled on a small chassis.

High-Frequency Antennas

In many cases, particularly at control stations, it will be necessary to use nondirective antennas because of the necessity for working field stations at random points of the compass. At field stations which normally work with only a single control station, however, it may be advantageous to use a simple form of directive array. The power gain will be worth while in bettering the signals in both directions, and in addition will minimize interference to and from other networks. The simpler forms of antennas described in Chapters Ten and Seventeen are quite suitable.

More important, perhaps, than the antenna itself is its location. Every effort should be made to get the antenna well above its surroundings and to provide, whenever possible, a clear path between the control station and the network stations with which it must communicate. Having a line of sight between antennas will ensure successful communication even though the power is very low and the antenna itself is nothing more than a simple half-wave wire. Where there are intervening obstructions, it will be helpful to use as much height as possible.

Vertical polarization is to be preferred to horizontal, since vertical polarization is better suited to mobile operation. A simple vertical antenna has practically no horizontal directivity, therefore it will work equally well in all directions except for effects attributable to its surroundings and to the terrain over which the signal must travel. The signal strength will be poor if a horizontally-polarized antenna is used to receive a vertically-polarized signal.

A half-wave antenna, two half-waves fed in phase stacked vertically, or an extended double-Zepp, all will be satisfactory, and are very simple types to construct. Design details will be found in Chapter Ten. If the station is to be operated on a fixed frequency, the antenna length should be adjusted for that frequency. If the same antenna is to work on several frequencies, the length had best be chosen midway between the two extremes.

Mobile antennas — It is probable that most networks will have one or more stations installed in cars, for dispatching to points which may be in urgent need of communication. The equipment previously described is readily adaptable to car installations; the transceiver, in particular, can be set up with little difficulty, and can get its power from the car broadcast receiver, if there is one. This would require only the installation of a suitable power socket in the car receiver, together with a switch to cut the power from the receiver when the transceiver is in use. Antennas suitable for such mobile installations are described in Chapter Seventeen.

For a solid but easily detachable mounting for a mobile antenna, the arrangement shown in Fig. 1806 is suggested. It is held in place by a panel of wood, cut to the shape of the window, on which the antenna is mounted. By running up the window the panel is held firmly in place. The antenna is of the "J" type. This type of installation places the radiator proper above the roof of the car, and has the advantage that it can be readily removed from the car when not in use or when needed elsewhere. Fig. 1808 shows a folded doublet.

The unit shown is built of ¼-inch plywood, since the usual thickness of the window glass in cars is ¼ inch. Run down the window of the car about halfway, or enough to leave at least a 6-inch opening, and make a pattern of cardboard using the top edge of the window glass for the guide. Trim the cardboard to this shape, and then push it up in the window and use the edge of the glass to mark the bottom

Fig. 1806 — A "J"-type antenna for 144-Mc. mobile operation can be mounted easily in the window of a car, allowing the radiator proper to be placed above the roof of the vehicle. The dimensions are given in the text.

edge of the pattern. From the pattern, mark the piece of plywood and cut it out with a saw. Additional small pieces to form stops in the corners are fastened to the main piece with glue and brads. A piece of plywood about $6 \times 8\frac{1}{2}$ inches should be fastened to the large piece at the point where the antenna is to be supported, using glue and brads, and the four stand-off insulators which support the antenna bolted to this piece. If the insulators are not long enough for the antenna to clear the side of the car, they can be raised by wood strips.

Two small strips should be nailed along the inside of the main piece so that they extend down below the edge a few inches and form, with the outside pieces, a yoke to keep the assembly in the proper position on the window.

The feeder can be made of flexible rubber-covered wire (obtained by splitting a length of parallel lamp cord) separated by small plastic or dry-wood spacers. The antenna ends of the wires are soldered to the heads of the large bolts in the upper stand-off insulators, and the wire is run out through holes in the wood.

The antenna and matching-section rods are regular automobile whip antennas and are supported on the stand-off insulators by small loop-shaped metal clamps. The shorting bar is made along the same lines, with bars of heavy metal on both sides of the clamp loops.

The length of the half-wave "J" antenna itself should be 38 inches for a frequency of 146 Mc. — the center of the two-meter band. Since the length of the matching section should be a quarter wavelength, or 19 inches, the total length of the right-hand element shown in Fig. 1806 should be 57 inches, while the shorter left-hand element should be 19 inches long. The spacing between elements should be 2 inches. With an open-wire transmission line consisting of two No. 18 wires spaced 2 inches, the line should be connected $5\frac{1}{2}$ inches up from the shorting bar at the bottom of the elements.

The folded-doublet antenna shown in Fig. 1808 is another simple type of antenna which may be adapted for mobile use, especially where center-feed is more convenient. It has the advantages of rather broad-band characteristic and moderately-high impedance at the feeding point. It should have an over-all length of 38 inches for 146 Mc.

A Car-Roof Antenna

Fig. 1807 shows a sketch of a fitting for a vertical v.h.f. car-roof antenna which provides a good mechanical arrangement for folding the antenna parallel to the car roof when the antenna is not in use.

The pieces *A* and *B* are made from sections of brass rod ¾ inch in diameter. One end of piece *A*, which has an over-all length of $3\frac{1}{2}$ inches, is turned down for a length of 2 inches to the diameter required to fit the inside of the bottom of the tubular antenna, which is soldered fast. At the other end of piece *A* is cut a tongue, 1 inch long and ¼ inch wide as shown in sketch.

Piece *B* has an over-all length of 6 inches. One end is turned down and threaded with a ¾-inch die, while a slot, 1 inch deep and ¼ inch wide to fit the tongue of *A*, is cut in the opposite end. The slotted end is then drilled and tapped on one side of the slot for a ¼-inch thumb screw, *C*. A vertical elongated hole is drilled and filed out in the tongue of piece *A*, so that, with the thumb screw loosened, *A* can be lifted up slightly to clear the shoulders of *B* while the antenna is being folded down. The solid seating of the two pieces, *A* and *C*, against each other when the antenna is erected in a vertical position provides little opportunity for the joint to work loose under vibration.

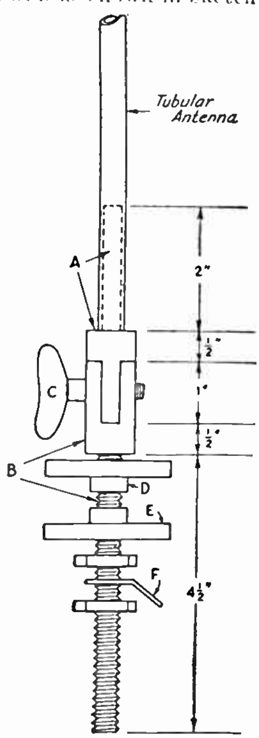


Fig. 1807 — Feed-through insulation and fittings for the folding car-roof mobile antenna. The joint hinges at *C* so that the antenna may be folded down parallel to the roof of the car.

The threaded

shank of piece *B* passes through a hole in the roof of the car. The polystyrene washers, *D* and *E*, provide the necessary insulation. Each is 2 inches in diameter and ¼ inch thick and has a collar or hub ¼ inch thick turned on one side to fit the hole in the car roof. The assembly is clamped to the roof of the car by means of the locking nuts either side of *F*. *F* is a soldering lug for making the connection to the antenna.

If the assembly is placed near the forward part of the roof, a two-meter half-wave antenna may be folded back at the hinge when not in use without the antenna overhanging the rear of the car.

Low-Frequency Emergency Antennas

Any of the simple low-frequency antennas described in Chapter Ten, or modifications of them, should be suitable for low-frequency portable and emergency work. End-fed antennas of the simple voltage-fed or Zepp types probably are the easiest to erect, although a center-fed antenna is more tolerant as to dimensions so long as the entire system including the feeders can be tuned to resonance. With such a center-fed arrangement, the feeders will stay in balance, even though the antenna portion of the system is much less than a half wavelength long.

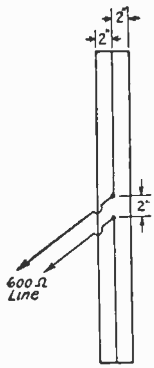


Fig. 1808 — Three-wire folded-doublet antenna for matching a 600-ohm line. The three conductors are connected together at the ends, as indicated. They may be made of wire, rod or tubing, and can be mounted on stand-off insulators on a wooden support.

For portable work at low frequencies a compact antenna which has been used successfully at 3.5 Mc. consists of about 60 feet of No. 18 enameled wire wound in a spiral around a long bamboo fishing pole. The turns are space-wound over the top 14 feet of the pole and then close-wound for about three feet. The remaining length of the pole is left free so that it may be lashed to a tree or other convenient upright, or simply stuck in the ground when no support is available. The bottom end of the winding is connected through an antenna tuner to ground.

The pi-section antenna coupler described in Chapter Ten is a good device for coupling random lengths of wire to either transmitter or receiver. An antenna of this type may be erected by tying a weight to one end of the wire and tossing it into a tree or over some other possible elevated support.

Transmission lines — At nearly all fixed locations it will be necessary to use a transmission line between the antenna and the radio equipment, since the latter will be indoors where it is easily accessible while the former will be placed on the roof of the building to secure adequate height. Low-loss coaxial or parallel (Twin-Lead) line is convenient for working into the center of a half-wave antenna, and it is readily available on the market today. The alternative is an open-wire line having an impedance of 500 to 600 ohms. It is advisable to keep the spacing between wires small at the higher frequencies; 2-inch spacing is about right, provided the line can be installed fairly rigidly so that it will not swing in a breeze and cause the transmitter frequency to change. This close separation also requires a fairly large number of spacers — at intervals of perhaps three to four feet. On lower frequencies the feeder spacing can be greater.

To make such a line nonresonant it will be necessary to install a matching stub at the antenna. The design and adjustment of such stubs also is covered in Chapter Ten. As an alternative, a multiwire doublet antenna may be used to couple directly to a line having an impedance of the order of 500 to 600 ohms without special matching provisions. Such an antenna is shown schematically in Fig. 1808. It gives a 9-to-1 impedance step-up at the line terminals, hence practically automatic matching to a 600-ohm line, assuming the normal doublet impedance of 70 ohms. In addition, it has a broad resonance characteristic and therefore is well suited to working anywhere in the band.

To avoid the necessity for impedance matching, two-wire lines may be operated as tuned lines if desired. Such operation has been successful with lines up to at least 100 feet long. Since in most cases the coupling device at the transmitter or receiver is a single-turn coil, the simplest method of tuning the line is to adjust the feeder length until the current in the line is maximum when the transmitter is operating on the chosen frequency. A small dial light or flashlight bulb, connected in series with one side of the line right at the transmitter terminals, may be used as a current indicator. The transmission line should be made about four feet longer than necessary, its length being adjusted by cutting off an inch or two at a time until maximum bulb brilliancy is obtained.

From a constructional standpoint it is desirable to use the same antenna for both transmitting and receiving. The change-over switch for this purpose should have low capacity, and preferably should have low-loss insulation. The ordinary type of wafer switch is satisfactory, particularly if it is ceramic insulated. A small porcelain-base d.p.d.t. knife switch also may be used for this purpose. If possible, the antenna switch should be combined mechanically with the power-supply change-over switches for the transmitter and receiver so that all the necessary switching from transmission to reception can be done in one simple operation.

Chapter Nineteen

Measurements and Measuring Equipment

To comply with FCC regulations it is necessary that the amateur station be equipped to make a few relatively simple measurements. For example, the regulations require that means be available for checking the transmitter frequency to make sure that it is inside the band. This means must be independent of the frequency control of the transmitter itself; it is not enough to depend on, say, the calibration of a crystal in the crystal-controlled oscillator that drives the transmitter. In addition, it is necessary to make sure that the plate power input to the final stage of the transmitter does not exceed one kilowatt. The regulations also impose certain requirements with respect to plate-supply filtering, stability and purity of the transmitted signal, and depth of modulation in the case of 'phone transmission.

In many cases all these measurements can be made to a satisfactory degree of accuracy with no more auxiliary equipment than the regular station receiver. However, a better job usually can be done by building and calibrating some relatively simple test gear. Too, the progressive amateur is interested in instruments as an aid to better performance.

Methods of making the measurements required in the amateur station will be discussed in this chapter, and design and construction of representative types of the instruments used in making these measurements will be described.

Frequency Measurement

Frequency-measuring equipment can be divided into two broad classes: oscillators of various types generating signals of known frequency that can be compared with the signal whose frequency is unknown, and adjustable resonant circuits.

Instruments in the first classification are the more accurate. Two types are commonly used by amateurs, the *secondary frequency standard* and the *heterodyne frequency meter*. The secondary frequency standard, nearly always crystal-controlled, usually generates a frequency of 100 kc. and employs a circuit that is rich in harmonic output. As a result, it supplies a series of frequencies, all multiples of 100 kc., which provides accurate calibration points throughout the communications spectrum. The more elaborate instruments of this type are provided with frequency dividers (multivibrators) to supply intermediate calibration points; a divisor commonly used is 10, thus furnishing signals at intervals of 10 kc. when the fundamental frequency is 100 kc.

The heterodyne frequency meter is a variable-frequency oscillator which is calibrated in frequency against a secondary standard or by other means. The oscillator usually is designed to cover the lowest frequency band in which measurements are to be made; measurements then can be made in higher-frequency bands by using the harmonic output of the oscillator. For example, when the oscillator is set to 3560 kc. its second harmonic is 7120 kc., its fourth harmonic is 14,240 kc., and so on. The proper frequency reading is determined by knowing the fundamental frequency of the oscillator and the number of the harmonic which falls in the desired frequency range.

Both the secondary standard and the heterodyne meter are ordinarily used in conjunction with a receiver, the signals from the instruments being picked up just as though they were from distant stations. In the case of the secondary standard, the frequency of the unknown signal can be determined by locating it between two known 100-kc. or 10-kc. multiples. With the heterodyne meter, the frequency is measured by adjusting the frequency meter until its signal is at zero-beat with the signal of unknown frequency, after which the frequency can be read from the frequency-meter calibration.

Since the secondary standard operates on a fixed frequency and can be crystal-controlled, its accuracy can be quite high. However, it simply establishes a series of known frequencies at regular intervals, and thus auxiliary methods must be used for determining frequencies between the known points. The series of fixed frequencies, when they mark the edges of amateur bands (as they do if they are multiples of 100 kc.), is quite sufficient for amateur work because the information that is required is whether or not the transmitter frequency is *inside* the band limits, rather than the exact frequency itself. On the other hand the heterodyne frequency meter, while capable of giving readings at any point in its calibrated range, is inherently less accurate than the crystal-

WWV SCHEDULES

All U. S. frequency calibration is based on the standard frequency transmissions from the National Bureau of Standards standard-frequency station, WWV. This station is on the air continuously, day and night, its radio frequencies of 5, 10 and 15 Mc. (and 2.5 Mc. from 7 P.M. to 9 A.M. EST with 440-cycle modulation only) modulated by standard audio frequencies of 440 and 4000 cycles per second, the former corresponding to A above middle C. In addition, there is a 0.005-second pulse every second, heard as a "tick," which provides an accurate time interval.

The audio frequencies are interrupted on the hour and every five minutes thereafter for one minute to give Eastern Standard Time in telegraphic code and to provide an interval for checking r.f. measurements. The station announcement is given by voice on the hour and half hour.

The accuracy of all frequencies is better than a part in 10,000,000. The 1-minute, 4-minute, and 5-minute intervals marked by the beginning and ending of the announcement periods are accurate to a part in 10,000,000. The beginnings of the periods when the audio frequencies are interrupted mark accurately the hour and the successive 5-minute periods.

controlled standard because of the lower stability of the variable-frequency oscillator.

In the absence of more elaborate frequency-measuring equipment, a calibrated receiver may be used to indicate the approximate frequency of the transmitter. 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 (S-meter), the receiver beat oscillator being turned off.

When checking the transmitter frequency the receiving antenna should be disconnected, so that the signal will not overload or "block" the receiver. 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.

Heterodyne frequency meter with built-in 100-kc. crystal calibrator — The basis of the heterodyne frequency meter is a completely-shielded oscillator with a precise frequency calibration. The oscillator must be so designed and constructed that it can be accurately calibrated and will retain its calibration over long periods of time.

The oscillator used in the frequency meter must be very stable. Mechanical considerations are most important in its construction. No matter how good the instrument may be electrically, its accuracy cannot be depended upon if the mechanical construction is flimsy. Inherent frequency stability can be improved by avoiding the use of phenolic compounds and thermoplastics (bakelite, polystyrene, etc.) in the oscillator circuit, employing only high-grade ceramics instead. Plug-in coils ordinarily are not acceptable; instead, a solidly-built and firmly-mounted tuned circuit should be permanently installed. The oscillator panel and chassis should be as rigid as possible.

A stable oscillator circuit suitable for use in a heterodyne frequency meter is the electron-coupled circuit. It is possible to take output from the plate with but negligible effect on the frequency of the oscillator, and strong harmonics are generated in the plate circuit.

The heterodyne frequency meter shown in Figs. 1901 to 1904, inclusive, combines a number of features that make it suitable for accurate frequency measurement in the amateur bands from 3.5 to 144 Mc. As shown in the circuit diagram, Fig. 1903, it consists of a 6SK7 electron-coupled oscillator followed by a 6AC7 amplifier that is used to intensify the higher-frequency harmonics. A second 6SK7 oscillator, using a crystal of the type that operates at either 100 or 1000 kc., provides check-points and a means for calibration of the frequency meter. A 6SL7 is incorporated to amplify the crystal harmonics and to provide a detector circuit in which the outputs of the crystal and e.c. oscillators can be mixed for calibration purposes. The detector also enables direct checking of the transmitter frequency.

The fundamental tuning range of the heterodyne oscillator is from 3500 to 4000 kc. By means of S_1 this range can be changed to 3500–3720 kc., approximately, so that the

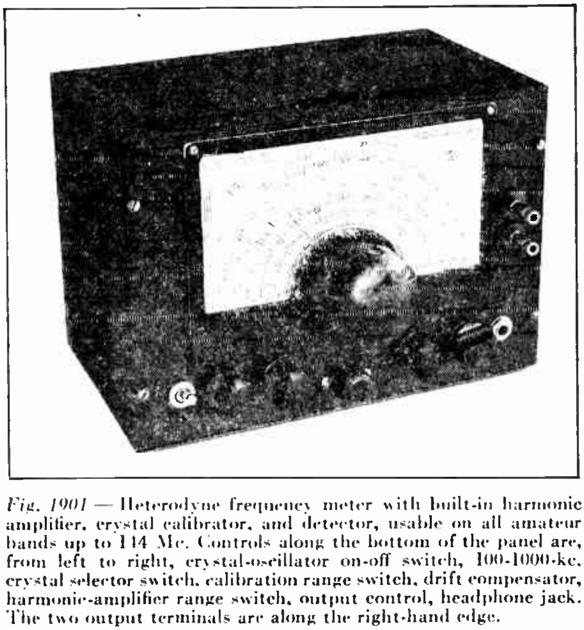


Fig. 1901 — Heterodyne frequency meter with built-in harmonic amplifier, crystal calibrator, and detector, usable on all amateur bands up to 144 Mc. Controls along the bottom of the panel are, from left to right, crystal-oscillator on-off switch, 100-1000-kc. crystal selector switch, calibration range switch, drift compensator, harmonic-amplifier range switch, output control, headphone jack. The two output terminals are along the right-hand edge.

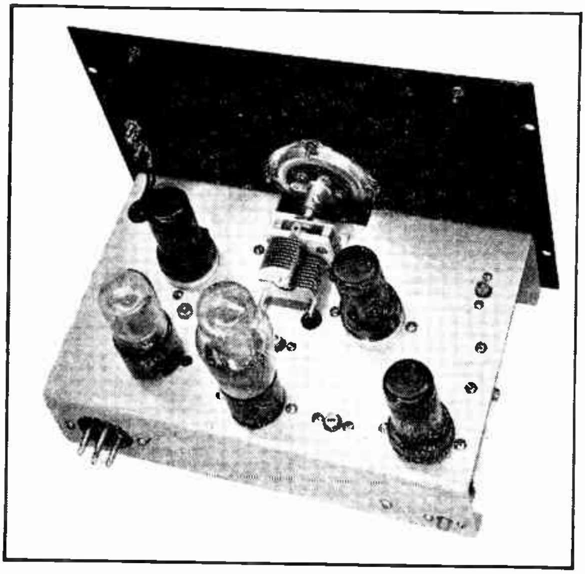


Fig. 1902 — Inside view of the heterodyne frequency meter. The main tuning condenser is in the center with the e.c. oscillator tube to its right. The crystal-oscillator tube is at the upper left, and the twin-triode amplifier-detector is in line with it at the rear edge (foreground) of the chassis. The 6AC7 is in the lower-right corner.

eighth harmonic just covers the 28–29.7-Mc. band. This avoids excessively critical tuning at the higher frequencies. The main tuning condenser, C_2, is connected across all of L_1 for the larger range and is connected to a tap on L_1 for the smaller to increase the bandspread. Simultaneously, an adjustable padding condenser, C_1, is switched in so that the oscillator frequency will be exactly 3500 kc. with C_2 set at maximum capacitance regardless of the switch position. C_4 is a fixed padding condenser to make the circuit fairly high-C, and C_5 is the band-setting condenser. C_3 is a small padder adjustable from the panel; its function is to permit resetting the oscillator frequency to the calibration check-points provided by the crystal oscillator and thus take care of drift from temperature variations and other causes.

The 6AC7 plate circuit is broadly tuned by means of switched coils resonating, with the circuit capacitances, at 144, 50 and 28 Mc., and thus increases the harmonic strength on those bands. A radio-frequency choke is connected to the fourth switch position; this gives ample signal strength at 14 Mc. and lower frequencies. Potentiometer R_5 makes it possible to reduce the strength of the signal from the meter to the value desired for measurement purposes.

In the crystal oscillator circuit, S_2 changes the frequency from 100 to 1000 kc. or vice versa. In the 100-kc. position C_{14} is connected across the crystal to provide means for adjusting the frequency to exactly 100 kc.

As shown in Figs. 1902 and 1904, the frequency meter is built on a chassis folded from a piece of sheet aluminum, the dimensions being 9 inches wide by 5½ inches deep by 2 inches high. Half-inch lips are bent along the bottom edges of the walls to make the chassis more rigid. The cabinet into which the meter fits is 10 by 7 by 6 inches. The main tuning condenser, C_2, is mounted on an aluminum bracket above the chassis and the coil, L_1, is similarly mounted below it. The band-setting condenser, C_5, is mounted on the chassis behind the coil, with its shaft protruding through the chassis for screwdriver adjustment. Trimmer C_3 is mounted on the panel and is adjusted by a knob underneath the main tuning dial. The coil is shielded from the amplifier section by the small aluminum baffle shown in Fig. 1904. The bandspread padder, C_1, is mounted to the left of the oscillator range switch and, like C_5, is screwdriver-adjusted from the top of the chassis. Wiring in the oscillator tuned circuit, including the switch, should be short, direct, and as rigid as possible.

The 100-kc. oscillator trimmer, C_{14}, does not require frequent adjustment and is therefore mounted on the rear edge of the chassis, close to the crystal unit. C_{16}, the plate tuning condenser for 1000 kc., is adjusted from the top of the chassis and is mounted to the right of the crystal-oscillator socket in Fig. 1904.

In putting the instrument into operation, the crystal oscillator should be checked first. Connect a length of wire to the crystal output terminal (from C_{18}) and listen on a receiver over the range from 3.5 to 5 Mc. With S_2 in the 1000-kc. position, signals should appear at 4000 and 5000 kc., and with S_2 in the 100-kc. position signals should be heard every 100 kc. Tune in WWV on 5000 kc., wait for the modulation to go off, and then adjust C_{14} for zero-beat. This sets the oscillator to precisely 100 kc. In the 1000-kc. position there may be a difference of a few kilocycles between the frequency of WWV and the 5-Mc. harmonic, but this is not serious since the 1000-kc. oscillator is used only as an aid in identification of the 100-kc. harmonics.

To set the range of the e.c. oscillator, put S_2 in the 1000-kc. position, plug a pair of 'phones into J_1, set S_2 on the maximum range position (C_2 across all of L_1), and set C_2 near minimum capacitance. Adjust C_5 until the 4000-kc. harmonic is heard. Then switch S_2 to 100 kc. and tune C_2 toward maximum, counting off five additional 100-kc. signals. C_5 may then be readjusted to bring the 3500-kc. marker close to the end of the tuning-dial scale. The 100-kc. points may then be marked off on the scale or the readings recorded. The second tuning range is adjusted by setting C_2 at 3500 kc. on the *first* range, then setting S_1 so that C_2 is connected to the tap, and adjusting C_1 (without touching C_2) so that the 3500-kc. marker is brought to the same point on the dial. The second range may be calibrated by the 100-kc. points in the same way as the first.

Calibration points may be obtained between the 100-kc. markers on both ranges by using a receiver as an auxiliary. For example, if the receiver is adjusted to pick up the fifth harmonic of the e.c. oscillator (17.5 to 20 Mc.) and the harmonic is beat against 100-kc. points from the crystal oscillator in that range, 100-kc. intervals on the fifth harmonic will give 20-kc. intervals on the fundamental. With a straight-line capacitance condenser at C_2, the relationship between dial divisions and frequency is almost linear, and marking off the dial at the proper intervals between actual calibration points will result in a calibration of sufficient accuracy.

The various amateur bands are covered by the following harmonics: 3.5-4 Mc., fundamental; 7-7.3 Mc., 2nd harmonic; 14-14.4 Mc., 4th; 27.185-27.245 Mc., 7th; 28-29.7 Mc., 8th; 50-54 Mc., 14th; 144-148 Mc., 40th. At lower frequencies a short length of wire connected to the output terminal will give ample signal strength under average conditions, but in the v.h.f. range closer coupling — such as running the wire in close proximity to the receiving antenna lead, or actually connecting it to the antenna post through a small fixed condenser — may be necessary to get a good signal.

With an instrument of this type the edges of amateur bands may be quite accurately determined, if care is used in setting the 100-kc. oscillator to WWV and equal care is used in setting the e.c. oscillator scale to the 100-kc. crystal points. C_3 may be used for the latter purpose each time the meter is used, and particularly during the first 30 minutes or so of operation when the temperature of the equipment is rising. The accuracy at intermediate points will depend upon the accuracy of the original calibration; it should be possible to read within 0.05 per cent under normal conditions by using the "drift corrector," C_3.

Absorption frequency meters — The simplest possible frequency-measuring device is a resonant circuit, tunable over the desired frequency range and having its tuning dial calibrated in terms of frequency. Such a frequency meter operates by extracting a small amount of energy from the oscillating circuit to be

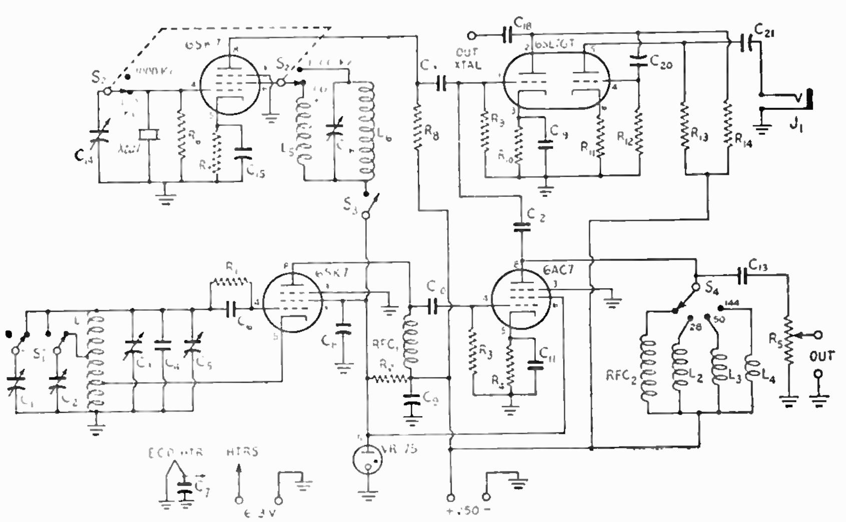


Fig. 1903 — Circuit diagram of the heterodyne frequency meter.

C_1, C_5 — 75-$\mu\mu$fd. variable.
C_2, C_{16} — 100-$\mu\mu$fd. variable.
C_3, C_{14} — 25-$\mu\mu$fd. variable.
C_4 — 220-$\mu\mu$fd. mica.
C_6, C_{10}, C_{13} — 100-$\mu\mu$fd. mica.
C_7, C_8, C_9, C_{15}, C_{20}, C_{21} — 0.01-μfd. paper.
C_{11}, C_{19} — 470-$\mu\mu$fd. mica.
C_{12} — 10-$\mu\mu$fd. mica.
C_{17} — 0.001-μfd. mica.
C_{18} — 47-$\mu\mu$fd. mica.
R_1, R_3, R_9, R_{12} — 0.47 megohm, ½ watt.
R_2 — 10,000 ohms, 1 watt.
R_4 — 330 ohms, 1 watt.
R_5 — 25,000-ohm potentiometer.
R_6 — 4.7 megohms, ½ watt.
R_7 — 470 ohms, 1 watt.
R_8 — 0.22 megohm, 1 watt.
R_{10} — 10,000 ohms, 1 watt.
R_{11} — 1500 ohms, 1 watt.
R_{13}, R_{14} — 0.1 megohm, ½ watt.
L_1 — 18 turns No. 18 on 1-inch form, length 1½ inches. Cathode tap 5 turns from ground end; band-spread tap 11 turns from ground.
L_2 — 24 turns No. 18 enam. close-wound on ¼-inch form.
L_3 — 11 turns No. 18 enam. close-wound on ¼-inch form.
L_4 — 2 turns No. 16 spaced ½ inch, diameter ¼ inch.
L_5 — 8-mh. coil (r.f. choke).
L_6 — 1 pie of 4-pie 2.5-mh. r.f. choke.
J_1 — Open-circuit jack.
RFC_1, RFC_2 — 2.5-mh. r.f. choke.
S_1 — 2-position 2-pole ceramic wafer switch.
S_2 — 2-position 2-pole switch (bakelite insulation satisfactory).
S_3 — S.p.s.t. toggle.
S_4 — 4-position 1-pole ceramic wafer switch.
XTAL — 100-1000-kc. crystal unit (Bliley SMC-100).

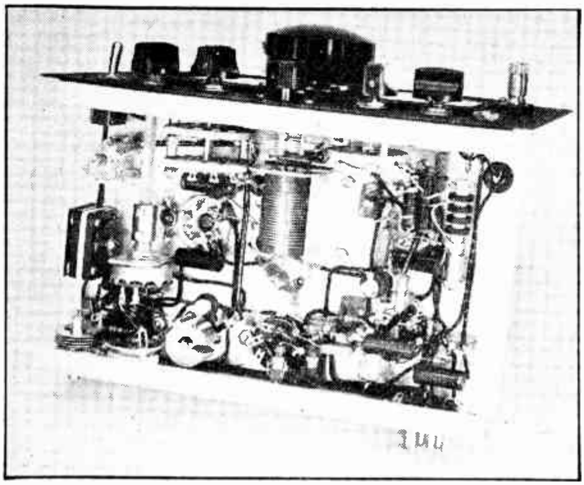


Fig. 1904 — Underneath the chassis of the heterodyne frequency meter. The parts layout is discussed in the text.

measured, the frequency then being determined by tuning the frequency-meter circuit to resonance and reading the frequency from the calibrated scale. This method is not capable of as high accuracy as the heterodyne methods for two reasons: First, the resonance indication is relatively "broad" as compared to the zero-beat of a heterodyne; second, the necessarily close coupling between the frequency meter and the circuit being measured causes some detuning in both circuits, with the result that the calibration of the frequency-meter circuit depends to some degree on the coupling to the circuit being measured.

It is necessary to have some means for indicating resonance with an absorption frequency meter. When such a meter is used for checking a transmitter, the plate current of the tube connected to the circuit being checked can provide the resonance indication. When the frequency meter is tuned through resonance the plate current will rise, and if the frequency meter is loosely coupled to the tank circuit the plate current will simply give a slight upward flicker as the meter is tuned through resonance. The greatest accuracy is secured when the loosest possible coupling is used.

A receiver oscillator may be checked by tuning in a steady signal and heterodyning it to give a beat note as in ordinary c.w. reception. When the frequency meter is coupled to the oscillator coil and tuned through resonance the beat note will change. Again, the coupling should be made loose enough so that a just-perceptible change in beat note is observed when the meter is tuned through resonance.

Although the absorption-type frequency meter should not be depended upon for accurate measurement, it is a highly-useful instrument to have in the station even when better frequency-measuring equipment is available. Since it generates no harmonics itself, it will respond only to the frequency to which it is tuned. It is therefore indispensable for distinguishing between fundamental and various harmonics, and for detecting harmonics and parasitic oscillations. When provided with a sensitive resonance indicator it is also useful for detecting r.f. in undesired places such as power wiring, for making rough measurements of field strength in adjustment of antennas, and can likewise be used as a modulation monitor.

An approximate calibration — usually sufficient — 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.

A sensitive absorption frequency meter — Figs. 1905 to 1907, inclusive, show an absorption frequency meter or "wavemeter" with a crystal-detector/milliammeter resonance indicator which provides a relatively high degree of sensitivity. As shown in the circuit diagram, Fig. 1906, a pick-up coil coupled to the resonant circuit is connected in series with a crystal

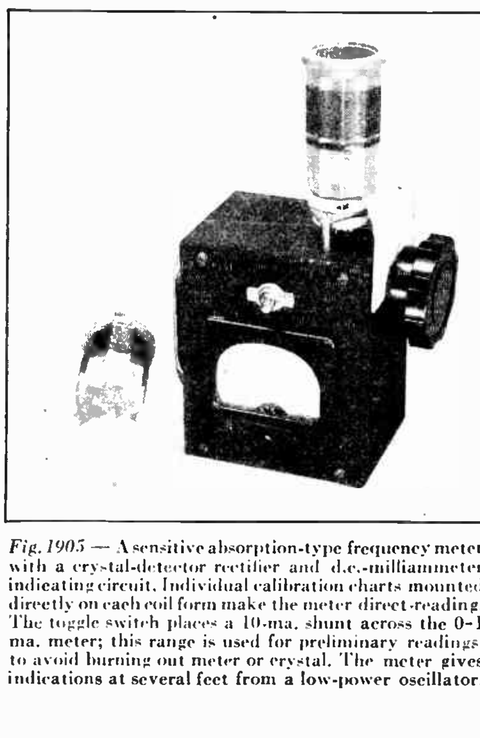


Fig. 1905 — A sensitive absorption-type frequency meter with a crystal-detector rectifier and d.c.-milliammeter indicating circuit. Individual calibration charts mounted directly on each coil form make the meter direct-reading. The toggle switch places a 10-ma. shunt across the 0-1 ma. meter; this range is used for preliminary readings, to avoid burning out meter or crystal. The meter gives indications at several feet from a low-power oscillator.

detector and 0–1 milliammeter. Plug-in coils are provided so that the unit covers the frequency spectrum from about 1 megacycle to 70 Mc. A switch, S, and shunt, R_1, are included so that the meter scale readings can be increased by a factor of 10, to reduce danger of overloading the milliammeter when making preliminary measurements. Any type of fixed crystal detector may be used, but the v.h.f. types are recommended when obtainable.

The unit is constructed in a 3- by 4- by 5-inch metal box, the milliammeter being mounted on one of the side panels. The coil socket is on top near one edge, with the tuning

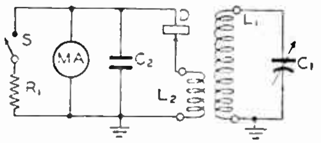


Fig. 1906—Indicating frequency-meter circuit diagram.

C_1 — 140-μμfd. variable (Hammarlund HFA-140-A).
C_2 — 0.001-μfd. mica.
R_1 — 3-ohm shunt: see general data on meter shunts.
L_1, L_2 — Plug-in coils wound on 1½-inch diameter forms:
D — Fixed crystal detector.
MA — 0.1 d.c. milliammeter (Triplett Model 321).
S — S.p.s.t. toggle switch.

Frequency Range	*Wire Size*	L_1	*Length*	L_2 [1, 2]
1.1–3.5 Mc.	No. 28 e.	81¾	1⅞″	17 turns
2.5–8.0 Mc.	No. 24 t.	37¾	1⅝″	11 "
4.5–11 Mc.	No. 20 t.	17¾	1½″	6 "
7.5–25 Mc.	No. 16 t.	8¾	1¼″	4 "
22–70 Mc.	No. 16 e.	2¾	1″	2 "

[1] Close-wound, No. 30 d.s.c., ¼ inch from primary.
[2] Because the impedance of individual crystal detectors varies considerably, experiment with the number of turns on L_2 is necessary for maximum current indication. If meter reads backward, reverse crystal connections.

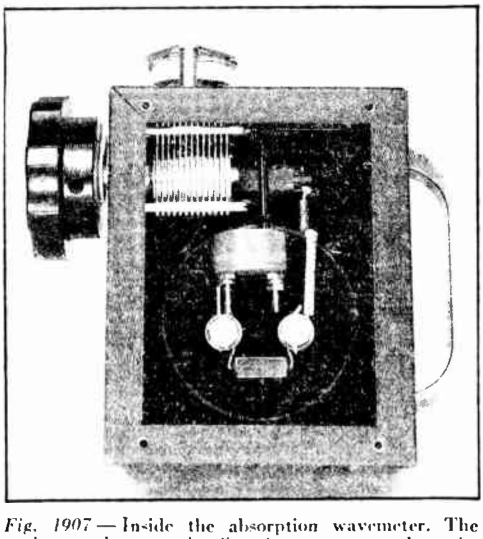


Fig. 1907 — Inside the absorption wavemeter. The tuning condenser and coil socket are mounted on the frame of the 3 by 4 by 5 box; remaining parts are fastened to one of the removable sides.

condenser just below it inside the case. This arrangement keeps the tuned-circuit leads short. A handle is mounted on the side of the box opposite the tuning control for convenience in handling. A metal plate, on which an appropriate calibration scale is pasted, is fastened to each plug-in coil so that the proper calibration automatically comes under the knob pointer when the coil is plugged in. The unit may be calibrated as described in the preceding section.

A two- or three-foot rod antenna and headphone jack may be added to the unit, using the connections shown in Fig. 1909. These additions permit the use of the instrument for field-strength measurements and for monitoring 'phone transmissions. The rod antenna is not required for ordinary frequency measurement, and its use may be undesirable when the frequencies of individual simultaneously-operating circuits are to be checked — as in the case of a multistage transmitter with frequency multipliers — because the antenna increases the sensitivity to such an extent that it may be difficult to identify the output of a particular circuit.

In addition to the uses mentioned in the preceding section, a meter of this type may be used for final adjustment of neutralization in triode r.f. amplifiers when loosely coupled to the plate tank coil.

V.H.F. wavemeter-field strength indicator-monitor — For operation at very-high frequencies a different type of construction must be adopted for wavemeters of the type described in the preceding section. An instrument suitable for the range 100 to 250 Mc. is shown in Figs. 1908 to 1910, inclusive. Provision is made in this unit for attaching an antenna so relative field-strength measurements can be made (for checking v.h.f. antenna patterns, for example) and the circuit includes a headphone jack so 'phone transmissions can be monitored.

The tuning condenser is a split-stator affair of 25 μμfd. per section. It is mounted to give short leads to the coil, and the use of a split stator condenser results in a low minimum capacity. The indicating device includes a pick-up loop loosely coupled to the tuned circuit, a 1N34 crystal and a 0–1 milliammeter. The by-pass condenser, C_2, furnishes a short r.f. return to the pick-up loop and avoids any resonances in this circuit within the frequency range of the wavemeter. For field-strength indication, an antenna is connected to one side of the pick-up loop and the wavemeter circuit, L_1C_1, is detuned, resulting in a nonselective indicator.

The wavemeter is built in a 3- by 4- by 5-inch metal cabinet, with the tuning condenser, C_1, mounted under the top. The condenser shaft comes out through a clearance hole in the side. An aluminum plate, 2⅝ by 3⅞ inches, is bolted on the side to back up the calibration scale. A polystyrene strip is used to mount the

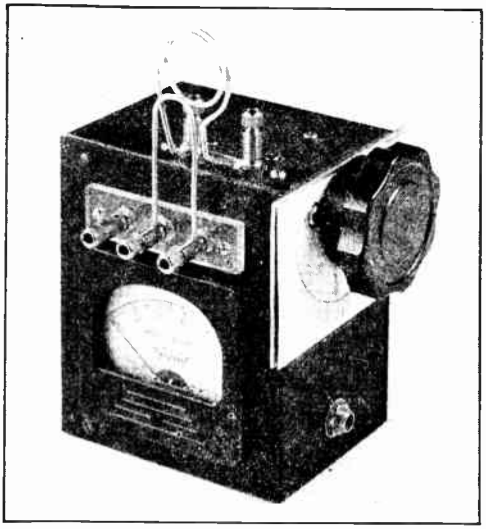


Fig. 1908 — A combination wavemeter, field-strength indicator and 'phone quality monitor for the 100–250-Mc. range. The two-turn coil is part of the wavemeter portion, and the hairpin loop provides pick-up for the 1N34 crystal detector. For field-strength work, a short antenna is connected to the binding post at the left of the hairpin loop.

two National FWA binding posts that hold the coil, L_1. The 'phone jack, J_1, is mounted on the side of the case below the tuning knob.

The wavemeter may be calibrated by using Lecher wires (see next section) in conjunction with a v.h.f. oscillator. (The oscillator may be a 144- or 220-Mc. transmitter.) Attach a two-

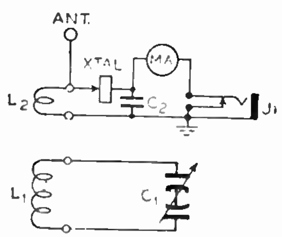


Fig. 1909 — Wiring diagram of the wavemeter and field-strength indicator.

C_1 — 25-μμfd.-per-section split-stator variable. (Cardwell ER-25-AD).
C_2 — 100-μμfd. midget mica.
L_1 — 90–180 Mc.: 2 turns No. 12 wire, 1⅛-inch diam., spaced wire diameter.
125–250 Mc.: hairpin loop of No. 12, 1¼-inch long, ¾-inch spacing.
L_2 — Hairpin loop of No. 12, 2¼ inches long, ¾-inch spacing.
J_1 — Closed-circuit telephone jack.
MA — 0–1 milliammeter.
XTAL — Type 1N34.

foot length of stiff wire to the antenna post of the wavemeter. With an oscillator capable of delivering 5 watts or so, a meter reading should be obtained several feet from the oscillator. The Lecher wires can then be very loosely coupled to the oscillator, and as the proper shorting points on the Lecher wires are found, a dip will be observed in the wavemeter current. If now the tuning knob of the wavemeter is rotated, a sharp dip in wavemeter current will be found, and this point should be marked in pencil on the scale and the frequency, as calculated from the Lecher wires, should be noted for future calibration. As a double check on the calibration of the wavemeter, remove the antenna and tune the wavemeter for *maximum* meter reading. The two points should be identical. If they are not, the pick-up loop is coupled too closely to the tuned circuit of the wavemeter.

Lecher wires — At very-high and ultra-high 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 distance between two consecutive current loops is equal to one-half wavelength. Thus the wavelength can be read directly in meters (inches × 39.37 if a yardstick is used), 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 be about one to one-and-one-half inches. The positions of the current loops are found by means of a "shorting bar," which is simply a metal strip or knife edge which can be slid along the line to vary its effective length. The

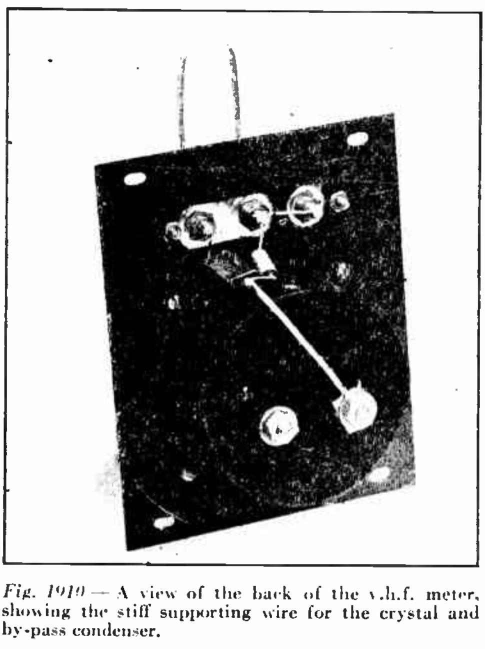


Fig. 1910 — A view of the back of the v.h.f. meter, showing the stiff supporting wire for the crystal and by-pass condenser.

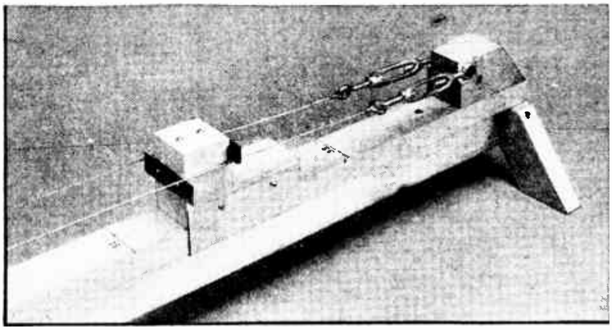


Fig. 1911 — One end of a typical Lecher-wire system. The feet at each end keep the assembly from tipping over when in use. The wires terminate in airplane-type strain insulators at one end, and at the other in small turnbuckles for maintaining tension. The wire is No. 16 bare solid copper antenna wire (hard-drawn). The turnbuckles are held in place by a 3/16 × 2-inch bolt through the anchor block. This end of the line is thus short-circuited; it does not matter whether it is open or shorted, since the other end is the one connected to the pick-up loop.

system can be used more conveniently and with greater accuracy if it is built up in permanent fashion and provided with a shorting bar maintained at right angles to the wires (Fig. 1911). The support may consist of two pieces of "1-by-2" pine fastened together with wood screws to form a "T" girder, this arrangement being used to minimize bending of the wood when the wires are tightened.

A slider holds the shorting bar and acts as a guide to keep the wire spacing constant. A piece of wood held in the hand can be used; it is an easy matter to regulate the pressure so that free movement is secured. A spring device may be arranged for the same purpose.

For convenience in measuring lengths directly in the metric system used for wavelength, the supporting beam may be marked off in decimeter (10-centimeter) units. A 10-centimeter transparent scale (obtainable at 5 & 10 cent stores) may be cemented to the slider, extending out from the front, so that readings can be taken to the nearest millimeter. The difference between any two readings gives the half wavelength directly.

Making measurements — Resonance indications can be obtained in several different ways. Let us suppose the frequency of a transmitter is to be measured. A convenient and fairly sensitive indicator can be made by soldering the ends of a one-turn loop of wire, of about the same diameter as the transmitter tank coil, to a low-current flashlight bulb, then coupling the loop to the tank coil to give a moderately-bright glow. A similar coupling loop should be connected to the ends of the Lecher wires and brought near the tank coil, as shown in Fig. 1912. 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. Marking the second spot, the distance between the two points can be measured and will be equal to half the wavelength.

If the measurement is made in inches, the frequency will be

$$F_{Mc.} = \frac{5906}{\text{length (inches)}}$$

If the length is measured in meters,

$$F_{Mc.} = \frac{150}{\text{length (meters)}}$$

In checking a superregenerative receiver, the Lecher wires may be similarly coupled to the receiver coil. In this case the resonance indication may be obtained by setting the receiver just to the point where the hiss is obtained, then as the bar is slid along the wires a spot will be found where the receiver goes out of oscillation. The distance between two such spots is equal to a half wavelength.

In either case, the most accurate readings result only when the loosest possible coupling is used between the line and the tank coil. After taking a preliminary reading to find the regions along the line in which resonance occurs, loosen the coupling until the indications are just discernible and repeat the measurement. Unless this is done the tuning of the line will affect the frequency of the oscillator and inaccurate indications will be obtained. As the coupling is loosened the resonance points will become sharper, which is a further aid to accurate determination of the wavelength.

The shorting bar must be kept at right angles to the two wires. A sharp edge on the bar is desirable, since it not only helps make good contact but also definitely locates the *point* of contact.

The accuracy with which frequency can be measured by such a system depends principally upon the technique of measurement. The necessity for using very loose coupling to the transmitter or receiver has already been mentioned. In addition, careful measurement of the exact distance between two current loops also is essential. Even if all other sources of error are eliminated, measurements within 0.1 per cent require an accuracy within 1 part in 1000, or 1 millimeter in one meter, in measuring the distance along the wires. This means that an accurate standard of length is necessary — a

Fig. 1912 — 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.

good steel tape, for instance — and that care must be used in determining the length exactly.

Signal Monitoring

Every amateur station should make provision for checking the quality of the transmitter output. This requires that some means be available in the station for reproducing the conditions existing at a distant receiving station; that is, for reducing the strength of the signal from the transmitter to such a point that its characteristics can be examined without danger of false indications from overloading the receiving equipment.

The simplest method of checking the quality of c.w. transmissions is to use the regular station receiver. If the receiver is a superheterodyne the process may simply be that of reducing the r.f. gain to minimum and tuning to the transmitter frequency. If distant signals are stable and have "pure d.c." tone in normal reception, then the local transmitter should too, when the receiver gain is reduced to the point where the receiver does not overload. If the signal is too strong with the r.f. gain "off," shorting the antenna input terminals may reduce it to suitable proportions, or the mixer circuit in the receiver may be temporarily detuned to arrive at the same result.

An alternative method is to set the receiver on the next lower-frequency band than the one in use, then tune the receiver so that the second harmonic of its oscillator beats with the transmitter signal to produce the intermediate frequency. Higher-order harmonics also may be used for this purpose. With this harmonic method there is ordinarily no danger that the receiver will overload, because the r.f. and mixer tuned circuits are so far from resonance with the transmitter frequency. The setting of the tuning dial bears no direct relation to the transmitter frequency under these conditions, since the oscillator harmonic must maintain a constant difference with the transmitter to produce the i.f. beat.

A 'phone signal may be monitored in the same way, provided a headset is used for reception. Use of a loudspeaker is not usually practicable because the sound output feeds back to the microphone and causes howling. A crystal detector and headset may also be used for the same purpose, as described in preceding sections. In monitoring a 'phone signal the best plan is to have another person speak into the microphone rather than to listen to one's own voice. It is difficult to judge quality when speaking and listening at the same time.

Measurement of Current, Voltage and Power

The amateur regulations require that when the power input to the final stage is above 900 watts, means must be provided for measuring the power input. This may be done by measuring the d.c. voltage applied to the final stage plates and the d.c. current flowing to them. The instruments required are a milliammeter and voltmeter.

Although in lower-power transmitters power-input measurements are not required, it is nevertheless true that a milliammeter is an almost indispensable instrument in the amateur station. It is invaluable in the adjustment of transmitting amplifier stages; tuning a transmitter without measuring grid and plate currents is like working in the dark. A d.c. voltmeter, although not essential, is useful in conjunction with the milliammeter in determining whether tube ratings are being exceeded or not and thus is helpful in prolonging tube life.

Besides d.c. measurements, it is also well to measure the filament voltages applied to transmitting tubes. Tube performance is dependent upon proper cathode emission, which in turn depends upon the voltage applied to the filament or heater. Also, the life of some transmitting tubes, particularly the thoriated-tungsten filament types, is critically dependent upon maintaining the filament voltage within rather close limits. Since most transmitting tube filaments are operated on a.c., an a.c. voltmeter is a worthwhile addition to amateur transmitting equipment.

Adjustment of a transmitter for maximum power output to the antenna or transmission line is facilitated by the use of instruments which measure radio-frequency current. Such instruments, although not actually essential, round out the measuring equipment used in transmitter adjustment.

D.c. instruments — D.c. ammeters and voltmeters are basically identical instruments, the difference being in the method of connection. An ammeter is connected in series with the circuit and measures the current flow. A voltmeter is a milliammeter which measures the current through a high resistance connected across the source to be measured; its calibration is in terms of the voltage drop in the resistance or *multiplier*.

If a single instrument must be used for measuring widely-different values of current or voltage, it is advisable to purchase one

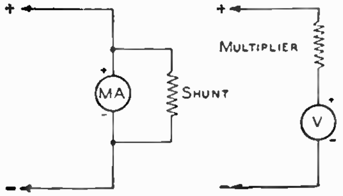


Fig. 1913 — **How voltmeter multipliers and milliammeter shunts are connected to extend the range of a d.c. meter.**

which will read, at about 75 per cent of full scale, the *smallest* value of current or voltage to be measured. Small currents cannot be read with any degree of precision on a high-scale instrument; on the other hand, the range of a low-scale instrument can be extended as desired to take care of larger values. The ranges of both voltmeters and ammeters can be extended by the use of external resistors, connected in series with the instrument in the case of a voltmeter or in shunt in the case of an ammeter. Fig. 1913 shows at the left the manner in which a shunt is connected to extend the range of an ammeter and at the right the connection of a voltmeter multiplier.

To calculate the value of a shunt or multiplier it is necessary to know the resistance of the meter. If it is desired to extend the range of a voltmeter, the value of resistance which must be added in series is given by the formula:

$$R = R_m\ (n - 1)$$

where R is the multiplier resistance, R_m the resistance of the voltmeter, and n the scale multiplication factor. For example, if the range of a 10-volt meter is to be extended to 1000 volts, n is equal to 1000/10 or 100.

If a milliammeter is to be used as a voltmeter, the value of series resistance can be found by Ohm's law:

$$R = \frac{1000\ E}{I}$$

where E is the desired full-scale voltage and I the full-scale reading of the instrument in milliamperes.

To increase the current range of a milliammeter, the resistance of the shunt is

$$R = \frac{R_m}{n - 1}$$

where the symbols have the same meanings as above.

Homemade milliammeter shunts can be constructed from any of the various special kinds of resistance wire, or from ordinary copper magnet wire if no resistance wire is available. The Copper Wire Table in Chapter Twenty gives the resistance per 1000 feet for various sizes of copper wire. After computing the resistance required, determine the smallest wire size which 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.

Precision wire-wound resistors used as voltmeter multipliers cannot readily be made by the amateur because of the much higher resistance required (as high as several megohms). As an economical substitute, standard fixed resistors may be used. Such resistors are supplied in tolerances of 5, 10 or 20 per cent ± the marked values. By obtaining matched pairs from the dealer's stock, one of which is, for example, 4 per cent low while the other is 4 per cent high, and using the pairs in parallel or series to obtain the required value of resistance, good accuracy can be obtained at small cost. High-voltage multipliers are preferably made up of several resistors in series; this not only raises the breakdown voltage but tends to average out errors in the individual resistors due to manufacturing tolerances.

When d.c. voltage and current are known, the power in a d.c. circuit can be stated by simple application of Ohm's law: $P = EI$. Thus the voltmeter and ammeter are also the instruments used in measuring d.c. power.

Multirange voltmeters and ohmmeters — A combination voltmeter-millammeter having various ranges is extremely useful for experimental purposes and for trouble shooting in receivers and transmitters. As a voltmeter such an instrument should have high resistance so that very little current will be drawn in making voltage measurements. A voltmeter taking considerable current will give inaccurate readings when connected across a high-resistance source — as is often the case in various parts of a receiver circuit. For such purposes the instrument should have a resistance of at least 1000 ohms per volt; a 0–1 milliammeter or 0–500 microammeter (0–0.5 ma.) is the basis of most multirange meters of this type. Microammeters having a range of 0–50 μa., giving a sensitivity of 20,000 ohms per volt, also are used.

The various current ranges on a multirange instrument can be obtained by using a number of shunts individually switched in parallel with the meter. Care should be used to minimize contact resistance in the switch.

It is often necessary to check the value of a resistor or to find the value of an unknown resistance, particularly in receiver servicing. An "ohmmeter" is used for this purpose. The ohmmeter is simply a low-current d.c. volt-

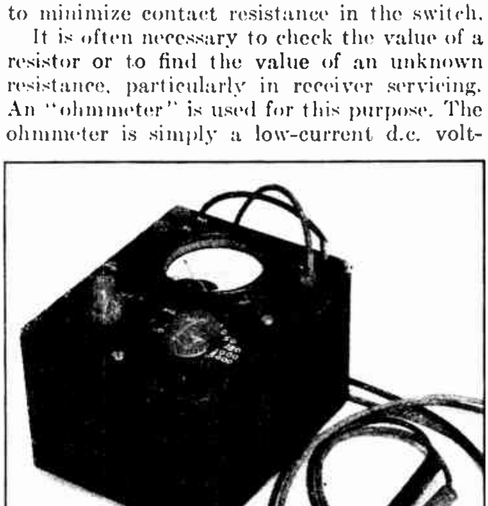
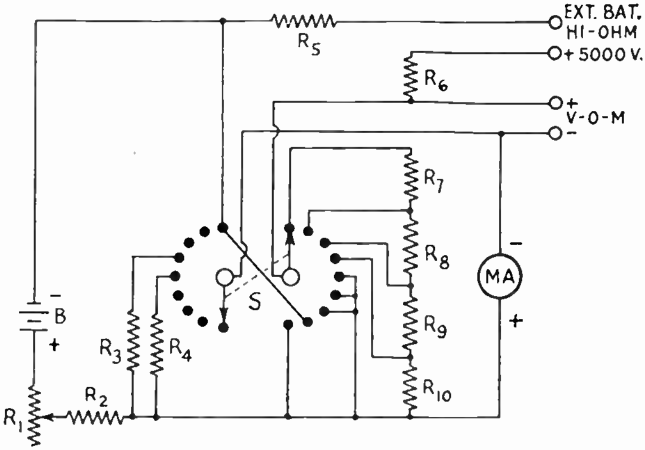


Fig. 1914 — **An inexpensive multirange volt-ohm-milliammeter housed in a standard 3 x 4 x 5 metal cabinet. Ranges are marked with number dies, the impressions being filled with white ink. High-voltage test leads are available for use on the 5000-volt range.**

Fig. 1915 — Circuit of the low-cost V-O-M.
R_1 — 2000-ohm wire-wound variable.
R_2 — 3000 ohms, ½ watt.
R_3 — 100-ma. shunt, 0.33 ohm (see text).
R_4 — 10-ma. shunt, 3.6 ohms (see text).
R_5 — 40,000 ohms, ½ watt.
R_6 — 4 megohms, 4 watts (four 1-megohm 1-watt resistors in series).
R_7 — 0.75 megohm, 1 watt (0.5 megohm and 0.25 megohm, ½ watt, in series).
R_8 — 0.2 megohm, ½ watt.
R_9 — 40,000 ohms, ½ watt.
R_{10} — 10,000 ohms, ½ watt.
B — 4.5 volts (Burgess 5360).
MA — 0–1 d.c. milliammeter.
S — 9-point 2-pole switch (Mallory-Yaxley 3109).



meter provided with a source of voltage (usually dry cells), the meter and battery being connected in series with the unknown resistance. If a full-scale deflection is obtained with the connections to the external resistance shorted, insertion of the resistance under measurement will cause the reading to decrease. The meter scale can be calibrated in ohms. 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 read on the meter,
e is the series voltage applied, and
R_m is the internal resistance of the meter.

A combination multirange volt-ohm-milliammeter, reduced to simple and inexpensive terms, is shown in Figs. 1914 to 1916, inclusive. Using a 0–1 milliammeter, the voltmeter has five ranges at 1000 ohms per volt: 0–10, 50, 250, 1000 and 5000 volts. Current ranges of 0–1, 10 and 100 ma. are provided. There are two resistance measurement ranges (three with external battery), a series range of 0–250,000 ohms, and a shunt range of 0–500 ohms. The "high-ohms" scale can be multiplied by 10 if the positive terminal of a 45-volt battery is connected to the terminal indicated in Fig. 1915, the unknown resistance being connected between the negative battery terminal and the negative terminal of the ohmmeter.

For economy, ordinary carbon resistors are used as voltmeter multipliers. These can be obtained with an accuracy within 5 per cent. The 5000-volt multiplier is four 1-watt resistors encased in heavy varnished cambric tubing to protect against flash-overs. The tubing extends over the positive "5M" terminal, which is further insulated by a wrapping of friction tape.

The 10-ma. and 100-ma. shunts are made of ordinary copper magnet wire wound on short lengths of ¼-inch diameter bakelite rod.

Measuring L and C — The ability to measure the inductance of coils, the capacitance of condensers, or the resonant frequency of a tuned circuit frequently saves time that might otherwise be spent in cut-and-try. A convenient instrument for this purpose is the grid-dip oscillator, which is simply a low-power oscillator equipped with a low-range milliammeter that measures the rectified grid current. When a resonant circuit tuned to the same frequency as the oscillator is coupled to the latter, the energy extracted by the coupled circuit reduces the amount available for feedback, with the result that the oscillator grid current decreases. Consequently there is a "dip" in grid current as either the oscillator or the circuit under measurement is tuned through resonance. The oscillator should be

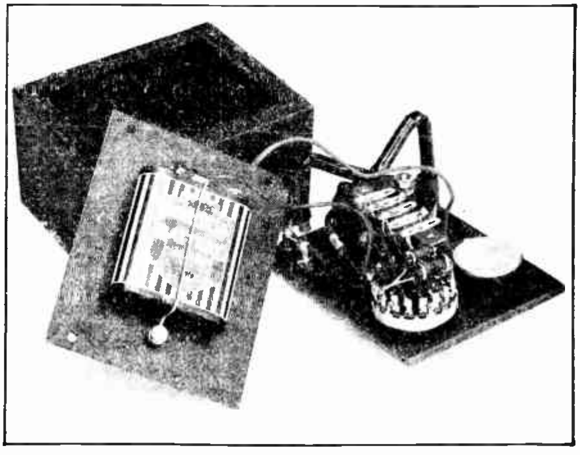


Fig. 1916 — Interior of low-cost volt-ohm-milliammeter. All parts except the internal ohmmeter battery are mounted on the 4 x 5-inch bakelite panel. The battery is attached to the bottom plate. The voltmeter multiplier is first assembled on an insulated tie-strip, then wired into the circuit. The M-shaped object in the rear is the 5000-volt multiplier — four 1-watt resistors covered with varnished cambric tubing.

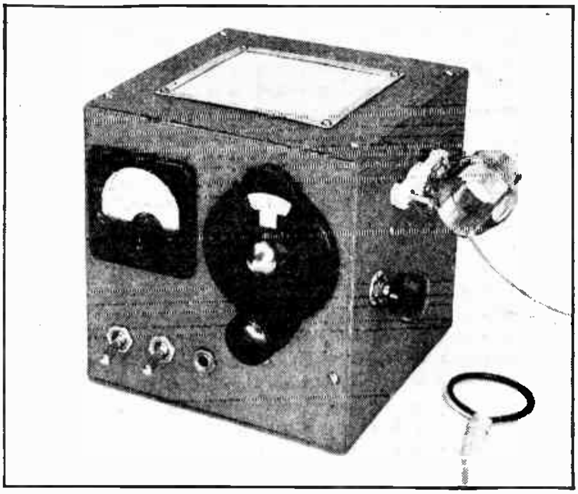


Fig. 1917 — The grid-dip meter is built in a 6 by 6 by 6-inch metal box. The tuning dial, milliammeter, "A" and "B" switches, and 'phone jack are on the front. The knob on the side controls the grid resistance. Standard plug-in coils are used.

arranged so that its frequency is continuously variable over a wide range, to make it most useful in measuring the resonant frequency of circuits whose constants are unknown or known only approximately.

A grid-dip oscillator is shown in Figs. 1917 to 1920, inclusive. As shown in the circuit diagram, Fig. 1918, it consists of a simple oscillator circuit using a dry-cell tube, battery operation being adopted to make the instrument conveniently portable. The frequency range is continuous from 3 to 60 megacycles, using standard midget air-wound plug-in coils. Grid current is measured by a 0–1 milliammeter, and is adjustable to any convenient value within this range by R_2. Separate switches are provided for the plate and filament supplies; by closing S_1 and leaving S_2 open the tube acts as a diode rectifier and the instrument thus can be used as an absorption wavemeter. The 'phone jack, J_1, also makes it possible to use it as a monitor. For convenience in measuring circuits that may be built into transmitters or receivers, the pick-up loop shown in Fig. 1917 provides the coupling. The loop is connected to the link on the oscillator coils through a few feet of 150-ohm Twin-Lead. The instrument may be calibrated by checking its frequency at a number of dial settings on a calibrated receiver.

For measuring inductance, the coil to be measured is connected to a condenser of known capacitance as shown at A in Fig. 1918. A mica condenser may be used as a standard; a 100-μμfd. 5-per-cent tolerance unit will serve for most purposes. With the unknown coil connected to the standard condenser, the pick-up loop is coupled to the coil and the oscillator frequency adjusted for the grid-current dip, using the loosest coupling that gives a detectable indication. The inductance is then given by the formula

$$L_{\mu h} = \frac{25{,}300}{C_{\mu\mu fd}\, f^2_{Mc.}}$$

A calibrated variable condenser is required for measuring capacitance. The circuit used is shown at B in Fig. 1918. The frequency of the circuit, using any convenient coil, is first measured with the unknown capacitance disconnected and the calibrated condenser set near maximum. The unknown is then connected and the calibrated condenser readjusted to resonance. The unknown capacitance is then equal to the difference between the capacitances at the two settings of the calibrated condenser. Obviously only capacitances smaller than the maximum capacitance of the calibrated condenser can be measured by this method. Since high accuracy in capacitance measurement is not ordinarily required, a satisfactory standard is any condenser of the straight-line capacitance type, for which a sufficiently good calibration curve can be constructed by noting the dial divisions at which the plates just start to mesh and are completely meshed, and assuming that the capaci-

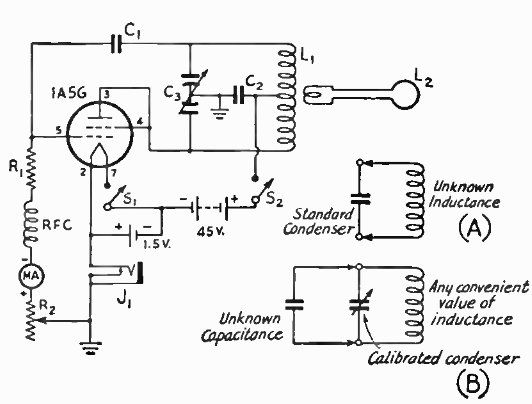


Fig. 1918 — Circuit of the grid-dip meter.

C_1, C_2 — 0.001-μfd. mica.
C_3 — 100-μμfd.-per-section variable (Hammarlund HFD-100).
R_1 — 4700 ohms, ½ watt.
R_2 — 25,000-ohm potentiometer.
L_1 — Center-tapped coils with center link. National AR-16 series or any equivalent coils may be used.
L_2 — Pick-up loop; one turn No. 14, diameter 1½ inches.
J_1 — Closed-circuit jack.
MA — 0-1 milliammeter.
RFC — 2.5-mh. r.f. choke.
S_1, S_2 — S.p.s.t. toggle switch.

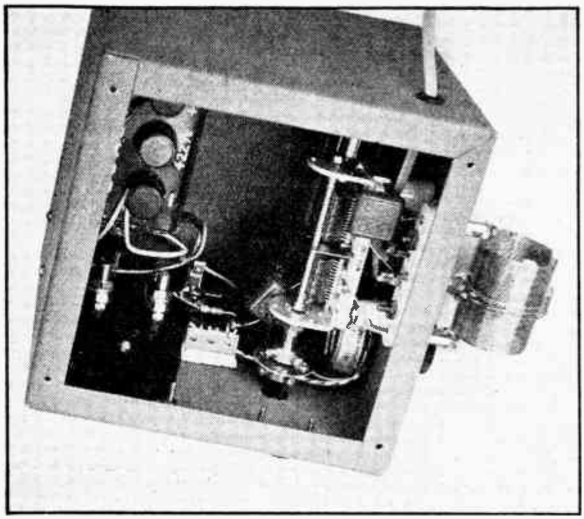


Fig. 1919 — Top view of the grid-dip meter. The tuning condenser is mounted on small stand-off insulators primarily to space it sufficiently from the side to make room for the dial on the front.

tance change is linear within those limits. The minimum and maximum capacitance (corresponding closely enough to these condenser settings) can be obtained from the manufacturer's data on the particular condenser used.

The Oscilloscope

The cathode-ray oscilloscope is an instrument of great versatility, and in conjunction with the instruments herein described, should be a valuable addition to the practical amateur station. The oscilloscope is useful on d.c., and audio and radio frequencies, and is particularly suited to a.f. and r.f. measurements because, compared to other types of measuring equipment, it introduces relatively little error at such frequencies.

Probably the chief use of the oscilloscope in amateur work is in measuring the percentage modulation in 'phone transmitters and in serving as a continuous monitor of modulation percentage. An oscilloscope for this purpose may be quite simple and inexpensive, consisting only of a small cathode-ray tube and an appropriate power supply. However, by providing amplifiers for the deflection plates and furnishing a linear sweep circuit, the possibilities of the instrument are greatly extended. It then becomes possible, for example, to examine audio-frequency waveforms and to check and locate the cause of distortion in a.f. amplifiers.

Constructional considerations — In building an oscilloscope, care should be taken to see that the tube is shielded from stray electric and magnetic fields which might deflect the beam, and means should be provided to protect the operator from accidental shock, since the voltages employed with the larger tubes are quite high. In general, the preferable form of construction is to enclose the instrument completely in a metal cabinet. It is good practice to provide an interlock switch which automatically disconnects the high-voltage supply when the cabinet is opened for servicing or other reasons.

In laying out the unit, the cathode-ray tube must be placed so that the alternating magnetic field from the power transformer has no effect on the electron beam. The transformer should be mounted directly behind the base of the tube, with the axes of the transformer windings and of the tube on a common line.

It is important that provision be included either for switching off the electron beam or reducing the spot intensity when no signal voltage is being applied. A thin, bright line or a spot of high intensity will "burn" the tube screen.

If trouble is experienced in obtaining a clean pattern from a high-power transmitter because of r.f. voltage introduced by the 115-volt line, by-pass condensers (0.01 or 0.1 μfd.) should be connected in series across the primary of the power transformer, the common connection between the two being grounded to the case.

A simple oscilloscope — The circuit of a simple cathode-ray oscilloscope is shown in Fig. 1921. Either a 1-inch 913 or a 2-inch 902 tube can be used. The cathode-ray tube may

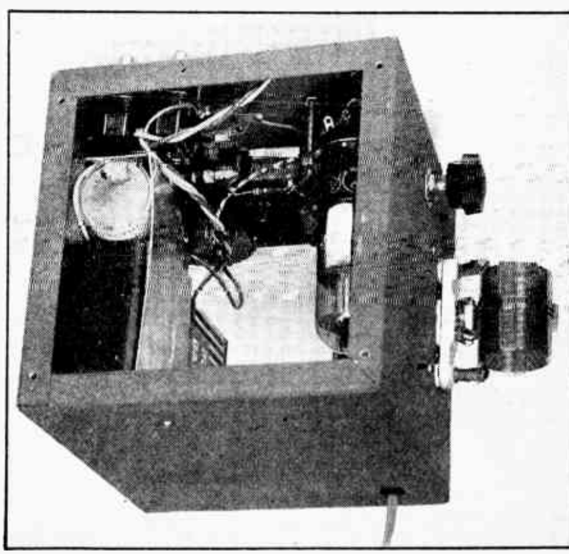


Fig. 1920 — A view from the bottom of the grid-dip meter. The oscillator tube is mounted underneath and parallel to the tuning condenser. Batteries are held in place by a metal strip fastened to the cabinet.

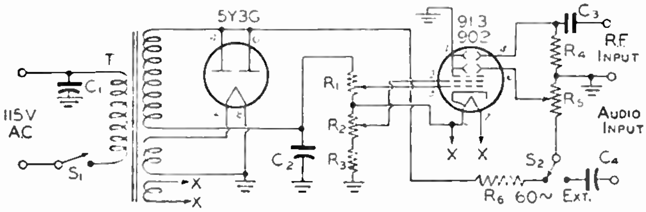


Fig. 1921 — An oscilloscope circuit for modulation monitoring.

C_1 — 0.01-μfd. 400-volt paper.
C_2 — 0.5-μfd. 800-volt paper or oil-filled.
C_3 — 0.005-μfd. mica.
C_4 — 0.1-μfd. 600-volt paper.
R_1 — 50,000-ohm variable.
R_2, R_5 — 0.5-megohm variable.
R_3 — 1 megohm, 1 watt.
R_4, R_6 — 0.5 megohm, 1 watt.
S_1 — S.p.s.t. toggle switch.
S_2 — S.p.d.t. toggle switch.
T — Replacement-type transformer: 350 volts, 40 ma.; 5 volts, 3 amperes; 6.3 volts, 2 amperes.

be mounted, together with the associated rectifier tube and other components, in a cabinet made of a standard 3 × 5 × 10-inch steel chassis with bottom plate.

This circuit is useful primarily for modulation checking in radiotelephone transmitters. Horizontal sweep voltage may be obtained either from an audio-frequency source, such as the modulator stage of the transmitter, or from the 60-cycle a.c. line, as selected by S_2. Using the modulator output for the sweep, the pattern on the screen will be in the form of a trapezoid, as described in Chapter Five.

R_5 controls the amplitude of the applied horizontal sweep. R_1 is the intensity control and R_2 the focusing control. If needed, a 2.5-mh. 125-ma. r.f. choke may be connected in series with the lead to the rotor of R_5 to correct leaning of patterns caused by r.f. coupling.

A complete oscilloscope — The usefulness of the oscilloscope is enhanced by providing a linear sweep circuit or time base, together with amplifiers for the horizontal and vertical deflection-plate signals so that sufficient voltage will be available at the deflection plates to give a pattern of suitable size. An inexpensive oscilloscope so equipped is shown in Figs. 1922 to 1925, inclusive. It uses the 1-inch Type 913 tube, but the 2-inch Type 902 readily can be substituted in the circuit.

As shown in Fig. 1923, the high-voltage d.c. is furnished by two 6H6s connected as half-wave voltage doublers. One supplies 300 volts positive for the amplifiers and sweep generator, and the other furnishes 300 volts negative for the cathode-ray tube voltage-divider network. The current drain is 2 ma. from the positive and 0.75 ma. from the negative supply.

The horizontal sweep generator is a 1/25-watt neon bulb (General Electric NE-51) used in a saw-tooth oscillator circuit. The frequency is determined by R_{24} plus R_{25} and the shunt capacity selected by S_3, and is variable between 12 and 700 cycles. A synchronizing voltage can be coupled in through C_{12} and its amplitude adjusted by R_{26}. The "Sync-Sweep" switch, S_2, allows five different conditions of sweep and synchronization, as follows: (1) external synchronization, (2) line synchronization, (3) internal synchronization, (4) line (sine-wave) sweep, and (5) external sweep.

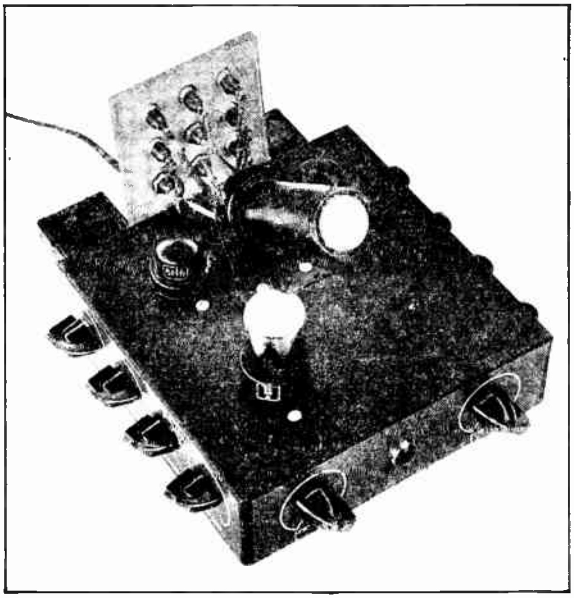


Fig. 1922 — A simple oscilloscope using a 1-inch tube. The controls on the front, from left to right, are "Sync Amplitude," pilot light and "Fine Frequency." Note the small neon tube, used for generating the sweep voltages, to the right of the 6SL7. A hood mounts over the 913 and the terminal panel at the rear of the chassis. The controls along the side, from back to front, are "Focus," "Vertical Centering," "Sync-Sweep" and "Vertical Gain."

The positive sawtooth from the generator becomes a negative sawtooth after amplification through the horizontal amplifier (one section of a 6SL7), and to make the trace sweep from left to right in the conventional fashion the cathode-ray tube must be turned so that the No. 1 pin is at the bottom, with pins No. 3 and No. 7 horizontal. Used in this manner a waveform will appear in the correct polarity when passed through the vertical amplifier but it will be inverted when applied directly to the vertical plates.

The unit is built on a 7- by 7- by 2-inch chassis. The ten controls and the pilot light are mounted along the front and sides, and the two heater transformers are mounted on the back. The external connections are brought to

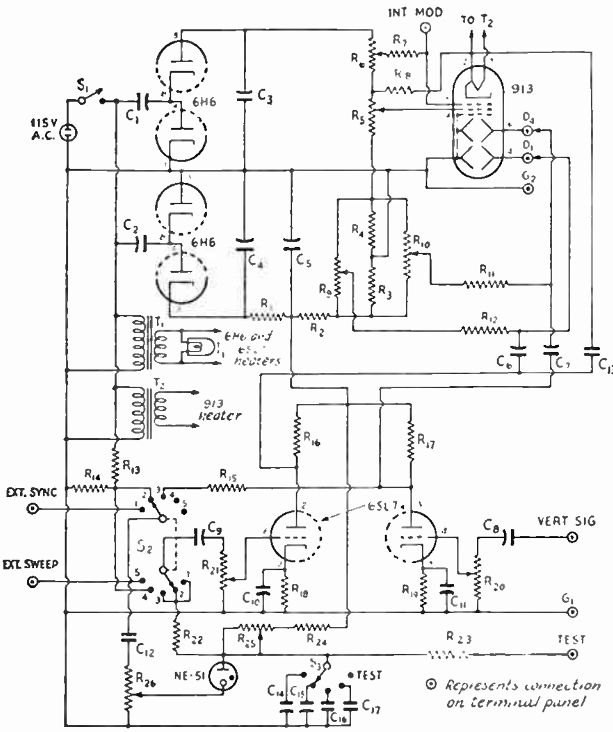


Fig. 1923 — Wiring diagram of the 1-inch oscilloscope. Terminals G_1 and G_2 should be connected to chassis.

C_1, C_2, C_3, C_4, C_5 — 8-μfd. 250-volt electrolytic.
C_6, C_7, C_8, C_9 — 0.1-μfd. 600-volt paper.
C_{10}, C_{11} — 25-μfd. 25-volt electrolytic.
C_{12} — 0.001-μfd. mica.
C_{13} — 100-μμfd. mica.
C_{14} — 0.05-μfd. 400-volt paper.
C_{15} — 0.02-μfd. 400-volt paper.
C_{16} — 0.006-μfd. mica.
C_{17} — 0.002-μfd. mica.
R_1 — 10,000 ohms.
R_2, R_{23} — 0.2 megohm.
R_3, R_4 — 0.1 megohm.
R_5 — 0.25-megohm variable, "Focus" control.
R_6 — 50,000 ohms, variable, "Intensity" control.
R_7, R_8 — 0.5 megohm.
R_9, R_{10}, R_{20} — 1.0-megohm variable. "Horizontal Centering," "Vertical Centering" and "Vertical Gain" controls.
R_{11}, R_{12}, R_{13} — 2.0 megohms.
R_{14} — 50,000 ohms.
R_{15} — 1.0 megohm.
R_{16}, R_{17} — 0.25 megohm.
R_{18}, R_{19} — 5000 ohms.
R_{21} — 3-megohm variable. "Horizontal Gain" control.
R_{22}, R_{24} — 3.0 megohms.
R_{25} — 10.0 megohm variable. "Fine Frequency" control.
R_{26} — 0.1-megohm variable. "Sync Amplitude" control.
All fixed resistors are ½-watt carbon.
I_1 — 6.3-volt pilot lamp.
S_1 — S.p.s.t. snap switch mounted on R_6.
S_2 — Two-pole 5-position rotary. "Sync-Sweep."
S_3 — Single-pole 5-position rotary. "Coarse Frequency."
T_1, T_2 — 6.3-volt 1.0-ampere heater transformer.

nine tip jacks on a polystyrene panel which is also mounted on the back of the chassis. Mounting the jacks for connections at the back of the chassis keeps the leads clear of the controls.

The arrangement of the tubes on the chassis can be seen in the photographs. The leads in the sweep generator, amplifier grid circuits and all heaters should be shielded to minimize a.c. pick-up. Too much pick-up in the sweep circuit will cause it to synchronize with the line frequency and produce unstable sweeps at other frequencies. The outputs of the amplifiers are brought out in flexible leads terminated in pin tips which can be plugged into the proper jacks on the terminal panel, thus making it a simple matter to remove them when working directly into the 'scope deflection plates.

Since one side of the a.c. line is common to the d.c. voltages and chassis, it is necessary to know when the chassis is connected to the grounded side of the line. The "Test" terminal is a means for checking this. With S_1 turned to the "Off" position and S_3 set to "Test," connect the "Test" terminal to an actual ground or the common of the unit to be tested with the 'scope. If the neon tube glows, the a.c. plug should be reversed.

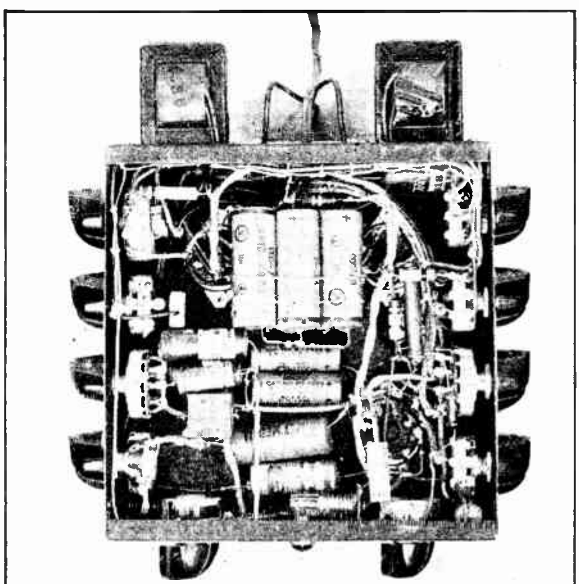


Fig. 1924 — View showing the arrangement of parts underneath the oscilloscope chassis. The controls along the left-hand side, from top to bottom, are "Intensity," "Horizontal Centering," "Coarse Frequency" and "Horizontal Gain." (McCormick, Jan., 1946, *QST*.)

The direct sensitivity of the vertical plates is 125 volts/inch and 175 volts/inch for the horizontal. Working through the amplifiers at maximum gain, the vertical sensitivity is 0.9

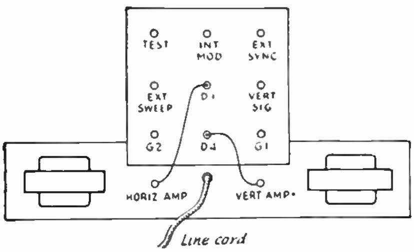


Fig. 1925 — A sketch of the back of the 'scope, showing the arrangement of terminals.

volts/inch and 1.1 volts/inch for the horizontal. The a.c. power consumption of the unit is approximately 20 watts.

ℂ Signal Generators

Test oscillators — A simple test oscillator for receiver checking and similar uses is shown in Fig. 1926. It uses the electron-coupled oscillator circuit with provision for suppressor-grid a.f. modulation. The output attenuator is a potentiometer so connected as to present a constant input resistance to the receiver.

For suppressor-grid modulation, apply approximately 10 volts of audio (for 50-per-cent modulation), where shown in the diagram.

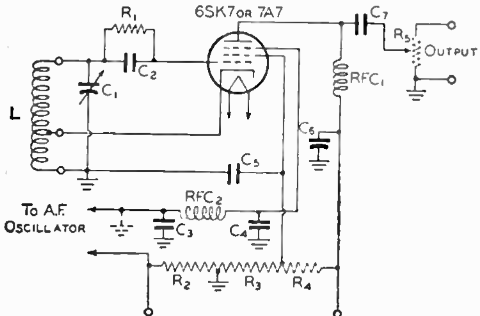


Fig. 1926 — Electron-coupled i.f. test-oscillator circuit diagram.

C_1 — 100-$\mu\mu$fd. variable with 200-$\mu\mu$fd. fixed silver-mica zero drift in parallel.
C_2 — 100-$\mu\mu$fd. midget mica.
C_3, C_4 — 250-$\mu\mu$fd. midget mica.
C_5 — 0.005-μfd. mica.
C_6 — 0.1-μfd. 400-volt paper.
C_7 — 500-$\mu\mu$fd. midget mica.
R_1 — 50,000 ohms, ½ watt.
R_2 — 2000 ohms, ½ watt.
R_3 — 20,000 ohms, 1 watt.
R_4 — 20,000 ohms, 2 watt.
R_5 — 500-ohm carbon potentiometer.
L — 440–510 kc.: 140 turns No. 30 enameled, close-wound on 1½-inch diameter plug-in form. Cathode tap 35 turns from ground end.
1400–1550 kc.: 42 turns No. 20 d.s.c. tapped 10 turns from ground.
4500–5500 kc.: 11 turns No. 18 enameled, turns spaced diameter of wire, tapped 3 turns from ground.
RFC_1 — 2.5-mh. r.f. choke.
RFC_2 — 25-mh. r.f. choke.

The suppressor grid is biased 10 volts negative for modulated use; if an unmodulated signal is desired, the upper terminal may be grounded as indicated. This will increase the output from the oscillator. Conversely, if the output potentiometer does not attenuate the signal sufficiently, additional d.c. negative bias may be applied between the modulation terminals.

In aligning a receiver it is important that the test signal be prevented from entering circuits where it can cause false indications. This will occur if the signal can enter the receiver by any other means than through the output leads from the test oscillator. The test oscillator must be thoroughly shielded, and the output lead likewise should be a shielded cable with the center wire the "hot" lead. Make all ground returns to a heavy copper strap connected to the cabinet at the output ground terminal. The plug-in coil should be separately shielded.

The i.f. ranges of the test oscillator can be calibrated by beating against signals of known frequency in the b.c. band. Frequencies between 465 kc. and 275 kc. can be spotted by using the second harmonic of the oscillator, the remainder of the range to 175 kc. being checked by using the third harmonic.

The a.f. modulating source for the test oscillator can be any audio oscillator capable of delivering 10 to 20 volts at the standard receiver-checking frequency of 400 cycles.

A useful audio-oscillator circuit is shown in Fig. 1927. It employs a two-terminal or "transitron" circuit using a pentagrid tube. A frequency of approximately 400 cycles is generated with the tuned-circuit values shown. The frequency may be changed by substituting a different value for C_1; several values of capacitance may be arranged to be selected by a switch so that an assortment of frequencies is available.

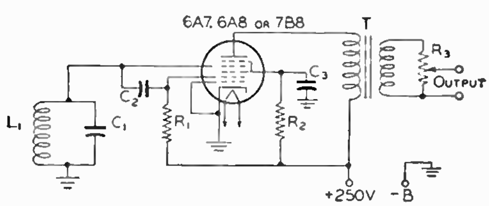


Fig. 1927 — Simple negative-resistance audio oscillator.

C_1 — 0.15-μfd. 400-volt paper.
C_2 — 0.1-μfd. 400-volt paper.
C_3 — 0.25-μfd. 200-volt paper.
R_1, R_2 — 50,000 ohms, 1 watt.
R_3 — 50,000-ohm volume control.
L_1 — 1.2-henry choke (Thordarson T-14C61 with iron core removed).
T — Output transformer (interstage audio, 1:3 ratio).

ℂ Antenna Measurements

Antenna measurements are made for the purpose (a) of securing maximum transfer of power to the antenna from the transmitter, and (b) of adjusting directional antennas to conform with design conditions. Related to measurements of the antenna system proper is the measurement of transmission-line performance.

Checking the transmission line for standing waves can be done by measuring the current in the wires, using a device of the type pictured in Fig 1928. The hooks (which should be sharp enough to cut through the insulation, if any, on the wires) are placed on one of the wires, the spacing between them being adjusted to give a suitable reading on the meter. At any one position along the line the currents in the two wires should be identical. Readings taken at intervals of a quarter wavelength will indicate whether or not standing waves are present.

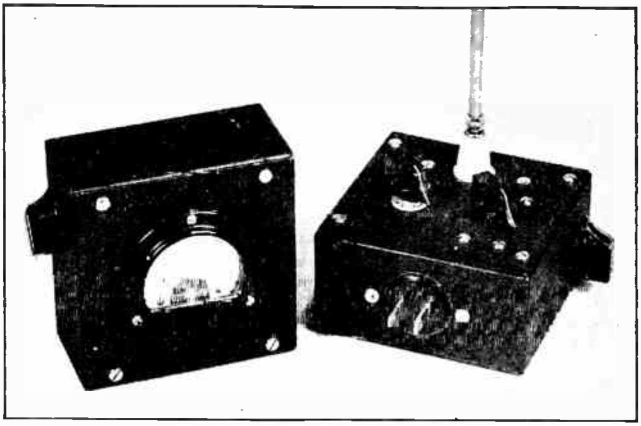


Fig. 1929 — **Remote-indicating field-strength meter, consisting of an r.f. pick-up and rectifier unit, and a meter unit. The knob on the left side of the meter unit is the switch for the shunt. On the pick-up unit the two controls are the bandswitch (left) and tuning. The knob at the right is for the resistor-shorting switch.**

Field-intensity meters — In adjusting antenna systems for maximum radiation and in determining radiation patterns, use is made of field-intensity meters. Fundamentally the field-intensity meter

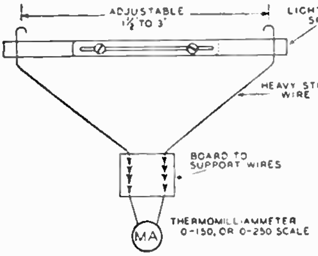


Fig. 1928 — **Line-current measuring device for checking standing waves on transmission lines.**

consists of a small pick-up antenna and an indicating device such as a rectifier and microammeter or a vacuum-tube voltmeter provided with a tuned input circuit. It is used to indicate the relative intensity of the radiation field under actual *radiating* conditions. It is particularly useful on the very-high frequencies and in adjusting directional antennas. Field-intensity checks should be made at points several wavelengths distant from the antenna and at heights corresponding with the desired angle of radiation.

The absorption frequency meters shown in Figs. 1905 and 1908 may be used as field-strength meters if provided with pick-up antennas. However, it is convenient to have the indicating device separate from the actual pick-up. This arrangement allows the pick-up unit to be set up out in the field to pick up radiation from the antenna under test, while the meter unit is near where adjustments are to be made. Antenna adjustment thus becomes a one-man job. The unit shown in Figs. 1929–1931, inclusive, is in two sections, one containing the usual tuned circuit, crystal rectifier, and antenna connection, and the other housing a microammeter for registering the rectified current from the crystal. The two units are fitted with matching plug and socket, permitting them to be used together, or they may be interconnected by means of a cable which can be any length up to several hundred feet. Three coils are used, so that measurements may be made on 28, 50, and 144 Mc. with the snap of a switch. A resistor is inserted in series with the crystal and meter, to lessen the loading effect on the tuned circuit and to make the response of the crystal more linear with variations in radiated power. As the resistor reduces the sensitivity somewhat, a switch is provided to short it out in case measurements are to be made with extremely low power or at large distances from the transmitting antenna. A

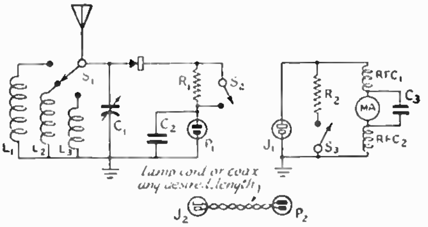


Fig. 1930 — **Wiring diagram of the remote-indicating field-strength meter.**

C_1 — 25-μμfd. midget variable.
C_2, C_3 — 0.001-μfd. mica.
R_1 — 1000 ohms, ½ watt.
R_2 — 250 ohms, ½ watt.
L_1 — 28-Mc. coil — 7 turns No. 22 enamel, ¼ inch long, on ¾-inch dia. form (National PRF-1).
L_2 — 50-Mc. coil — 6 turns No. 22 enamel, ¼ inch long, on 9/16-inch dia. form (National PRE-1).
L_3 — 144-Mc. coil — 3 turns No. 18 enamel, ¼ inch long, ⅜-inch dia., self-supporting.
J_1, J_2 — Universal receptacle, two-pole retainer-ring type (Amphenol 61-F).
MA — 0-100 microammeter (0-500 microammeter or 0-1 milliammeter may be used, with reduced sensitivity).
P_1, P_2 — Polarized plug, two-pole retainer-ring type (Amphenol 61-MP).
S_1 — 3-position wafer-type switch.
S_2, S_3 — S.p.s.t. snap switch.
RFC_1, RFC_2 — 2.5 mh. choke (National R-100).

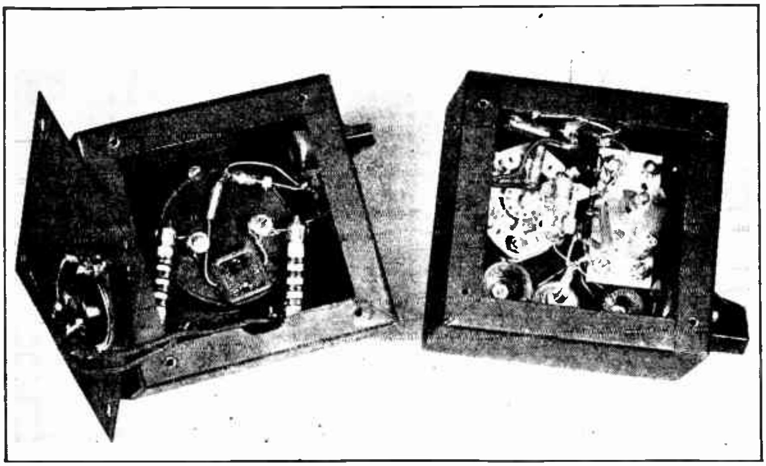


Fig. 1931 — Inside view of the two units of the remote-indicating field-strength meter.

100-microampere meter is used to give high sensitivity, and a shunt is available to multiply the range of the meter by three. This shunt is also provided with a switch so that low or high readings can be taken without making a trip to the pick-up unit. The crystal is the 1N21 type. Germanium crystals (1N34) also may be used with good results.

The two units are housed in 2 by 4 by 4-inch steel boxes with front and back removable. In the pick-up unit all parts except the resistor shorting switch and connecting plug are mounted on the top panel, permitting easy wiring of the assembly. The interconnecting plug and socket are the polarized type, with one prong on the plug slightly larger than the other. The plug will fit a standard a.c. outlet, so the interconnecting cable (ordinary rubber-covered lamp cord) doubles as a long a.c. extension cord when not in use for its intended purpose.

The antenna connection is a steatite feed-through bushing fitted with a "banana plug" socket. A convenient pick-up antenna is made by drilling and tapping a ¼-inch rod for 6/32 thread to take the threaded end of a banana plug. The length of the antenna will vary the sensitivity of the unit. If measurements are to be made with high power levels, a rod a few inches in length will suffice, but for ordinary work a length of 24 inches or so will be about right.

Chapter Twenty

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Vacuum-Tube Characteristics and Miscellaneous Data

Inductance and Capacity

Inductance **(L)** — The formula for computing the inductance of air-core coils is:

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

where a is the mean diameter of the coil in inches, b is the length of the winding in inches, c is the radial depth of the winding in inches, and n is the number of turns. The quantity c may be neglected if the coil is a single-layer solenoid.

For 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 (page 416), 35 turns of No. 30 d.s.c. will occupy 0.5 inch. Therefore, $a = 1.5$, $b = 0.5$, $n = 35$, and

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

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

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

Straight round wires:

To calculate the high-frequency inductance of a straight round wire:

$$L = 0.00508\ l\ (2.303 \log_{10} \frac{4l}{d} - 1)$$

l = length in inches
d = diameter in inches
L = inductance in microhenrys

Condenser capacity **(C)** — The formula for determining the capacity of a condenser is:

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

where A is the area of one side of one plate in square inches, n is the total number of plates, d is the separation between plates in inches, and K is the dielectric constant (= 1 for air; see the table on page 415 for values for other materials).

The dielectric constant is the ratio of the capacity of a condenser with a given dielectric to its capacity with air dielectric.

ABBREVIATIONS FOR ELECTRICAL AND RADIO TERMS

Alternating current	a.c.	Medium frequency	m.f.
Ampere (amperes)	a.	Megacycles (per second)	Mc.
Amplitude modulation	a.m.	Megohm	MΩ
Antenna	ant.	Meter	m.
Audio frequency	a.f.	Microfarad	μfd.
Centimeter	cm.	Microhenry	μh.
Continuous waves	c.w.	Micromicrofarad	μμfd.
Cycles per second	c.p.s.	Microvolt	μv.
Decibel	db.	Microvolt per meter	μv/m.
Direct current	d.c.	Microwatt	μw.
Electromotive force	e.m.f.	Milliampere	ma.
Frequency	f.	Millivolt	mv.
Frequency modulation	f.m.	Milliwatt	mw.
Ground	gnd.	Modulated continuous waves	m.c.w.
Henry	h.	Ohm	Ω
High frequency	h.f.	Power	P.
Intermediate frequency	i.f.	Power factor	p.f.
Interrupted continuous waves	i.c.w.	Radio frequency	r.f.
Kilocycles (per second)	kc.	Ultrahigh frequency	u.h.f.
Kilovolt	kv.	Very-high frequency	v.h.f.
Kilowatt	kw.	Volt (volts)	v.
Magnetomotive force	m.m.f.	Watt (watts)	w.

RMA Radio Color Codes

Standard color codes have been adopted by the Radio Manufacturers Association for the ready identification of values and connections for standard components.

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.

Mica condensers:

If one row of three colored markers appears on the condenser, the voltage rating is 500 volts and the capacity is expressed to two significant figures, in micromicrofarads, as follows: First dot on left, first significant figure. Second dot, second significant figure. Third dot, decimal multiplier.

Example: A condenser has one row of colored markers, as follows: *brown, black* and *brown*. Its capacity is 100 μμfd.

When two rows of three colored markers appear on the condenser the top row represents the significant figures, reading from left to right; the bottom row indicates the decimal multiplier, tolerance and voltage rating, reading from right to left. Capacity is in μμfd.

Example: A condenser has two rows of colored markers, as follows: Top row: left, *brown;* center, *black;* right, *no color*. Bottom row: right, *brown;* center, *green;* left, *blue*. Its ratings are 100 μμfd.; ± 5%, 600 volts.

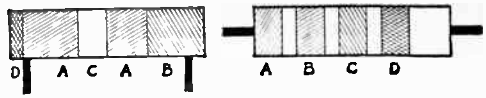
Tubular condensers:

Two groups of colored bands are used on tubular condensers. Viewed with the wide bands on the right, the wide bands indicate significant figures (from left to right); narrow bands indicate the decimal multiplier, tolerance and voltage rating, from right to left, respectively.

Resistors:

Values of resistance and tolerances are indicated by colored dots, bands or stripes on the resistor.

Two types of resistors are commonly used, one having radial and the other axial leads. The following illustration shows the two types of resistors and the system of identification.



Radial leads	*Axial leads*	*Color*
Body A	*Band A*	*Indicates first significant figure.*
End B	*Band B*	*Indicates second significant figure.*
Band C (or dot)	*Band C*	*Indicates decimal multiplier.*
Band D	*Band D*	*Indicates tolerance in per cent.*

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.

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.

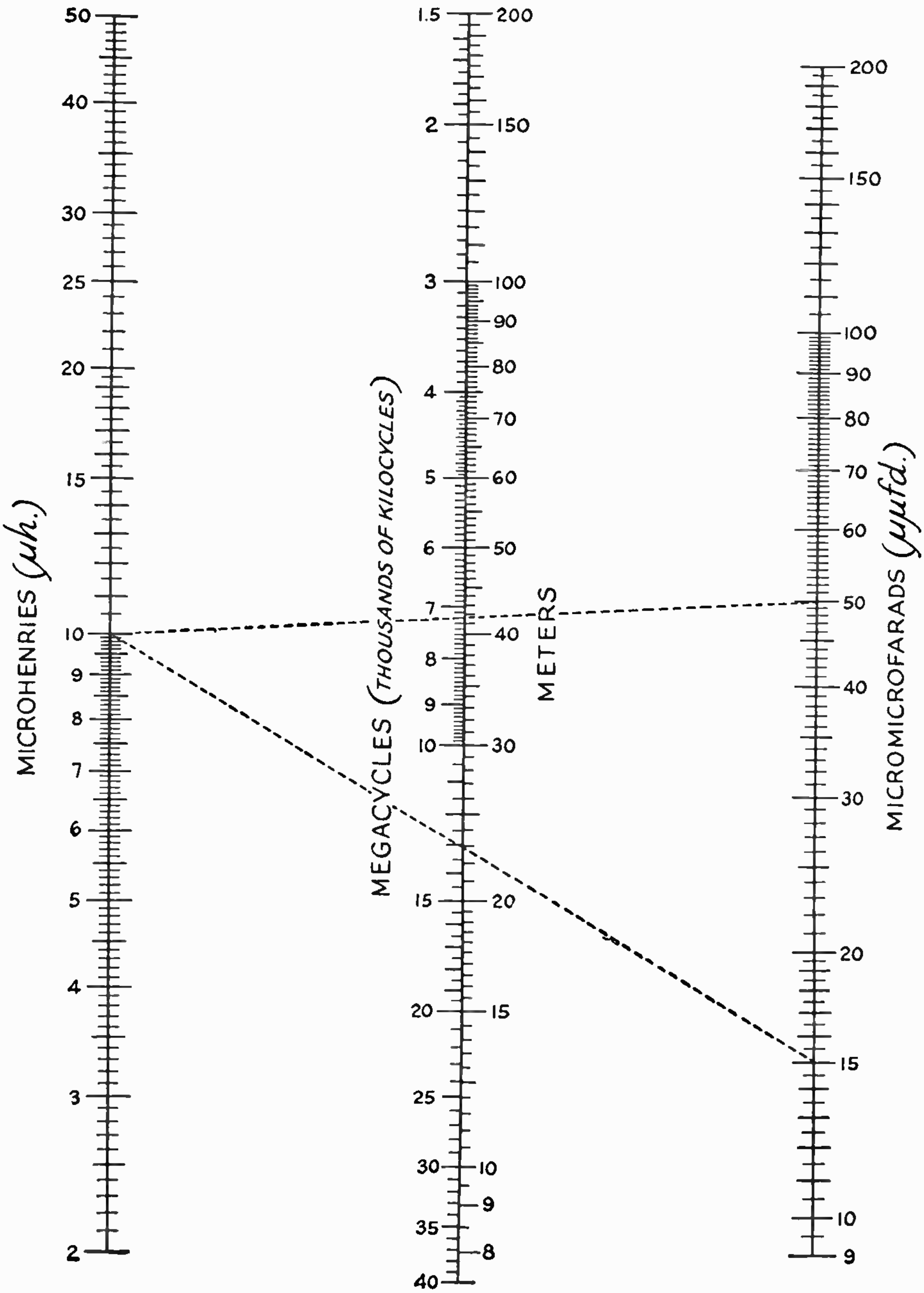
Field coils:

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*

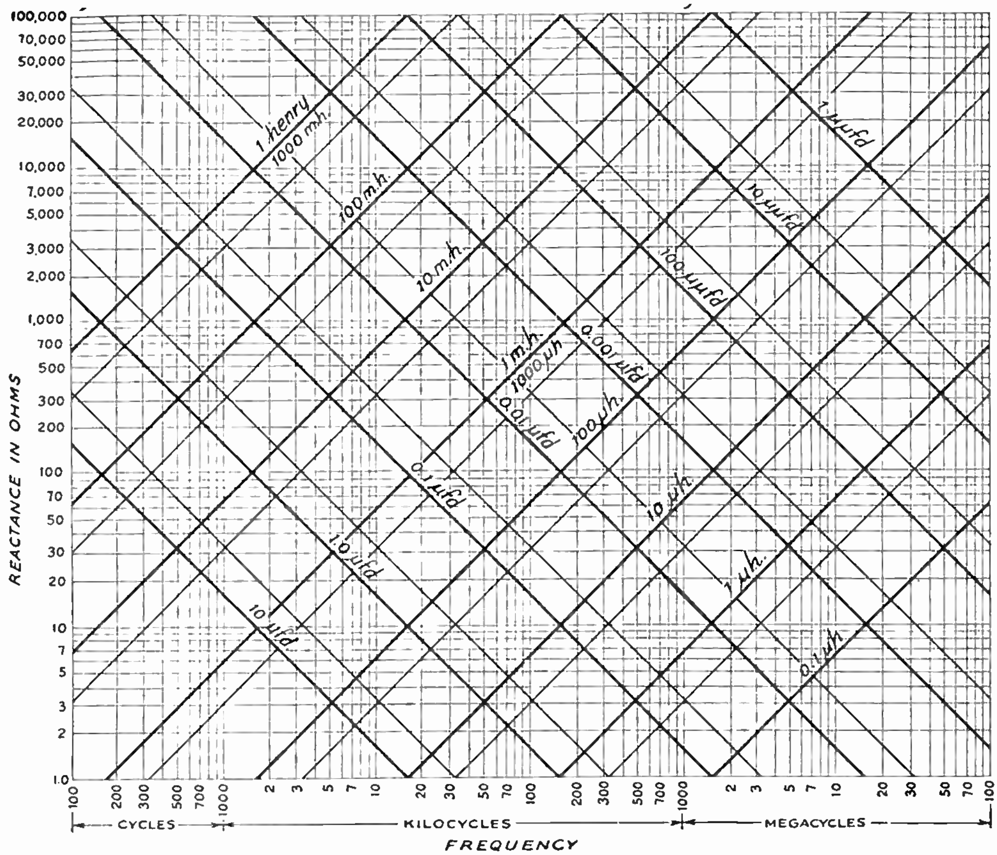
INDUCTANCE, CAPACITY AND FREQUENCY CHART — 1.5–40 MC.



This chart may be used to find the values of inductance and capacity required to resonate at any given frequency in the medium- or high-frequency ranges; or, conversely, to find the frequency to which any given coil-condenser combination will tune. In the example shown by the dashed lines, a condenser has a minimum capacity of 15 μμfd. and a maximum capacity of 50 μμfd. If it is to be used with a coil of 10-μh. inductance, what frequency range will be covered? The straight-edge is connected between 10 on the left-hand scale and 15 on the right, giving 13 Mc. as the high-frequency limit. Keeping the straight-edge at 10 on the left-hand scale, the other end is swung to 50 on the right-hand scale, giving a low-frequency limit of 7.1 Mc. The tuning range would, therefore, be from 7.1 Mc. to 13 Mc., or 7100 kc. to 13,000 kc. The center scale also serves to convert frequency to wavelength.

The range of the chart can be extended by multiplying each of the scales by 0.1 or 10. In the example above, if the capacities are 150 and 500 μμfd. and the inductance 100 μh., the range becomes approximately 231 to 422 meters or 0.7 to 1.3 Mc. Alternatively, 1.5 to 5 μμfd. and 1 μh. will give a range of approximately 71 to 130 Mc.

INDUCTIVE AND CAPACITIVE REACTANCE VS. FREQUENCY CHART



By use of the chart above, the approximate reactance of any capacity 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 capacity 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 capacity 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. Coëffi.* [2] *of Resistance*
Aluminum (2S; pure)	59	0.0049
Aluminum (alloys):		
Soft-annealed	45–50	
Heat-treated	30–45	
Brass	28	0.002–0.007
Cadmium	19	
Chromium	55	
Climax	1.83	
Cobalt	16.3	
Constantin	3.24	0.00002
Copper (hard drawn)	89.5	0.004
Copper (annealed)	100	
Everdur	6	
German Silver (18%)	5.3	0.00019
Gold	65	
Iron (pure)	17.7	0.006
Iron (cast)	2–12	
Iron (wrought)	11.4	
Lead	7	0.0041
Manganin	3.7	0.00002
Mercury	1.66	0.00089
Molybdenum	33.2	0.0033
Monel	4	0.0019
Nichrome	1.45	0.00017
Nickel	12–16	0.005
Phosphor Bronze	36	0.004
Platinum	15	
Silver	106	0.004
Steel	3–15	
Tin	13	0.0042
Tungsten	28.9	0.0045
Zinc	28.2	0.0035

[1] At 20° C., based on copper as 100. [2] Per °C. at 20° C.

Approximate relations:

An increase of 1 in A. W. G. or B. & S. wire size increases resistance 25%.
An increase of 2 increases resistance 60%.
An increase of 3 increases resistance 100%.
An increase of 10 increases resistance 10 times.

Table of Dielectric Characteristics

Dielectric material [1]	*Dielectric constant (K)*	*Power factor*					*Dielectric strength (puncture voltage)* [2]	*Volume resistivity* [3] *(ρ)*
		60 cycles	*1 kc.*	*1 Mc.*	*10 Mc.*	*100 Mc.*		
Air (normal pressure)	1.0						19.8–22.8	
AlSiMag A196	5.7–6.3	2.9		0.21	0.15		240	10^{14}
Aniline formaldehyde	3–5	1–6					400	
Asphalts	2.7–3.1		2.3				25–30	
Bakelite — *See* Phenol								
Beeswax	2.9–3.2							
Casein plastics [4]	6.1–6.4			5.2–6			165	
Castor oil	4.3–4.7			7			380	
Celluloid	4–16			5–10				
Cellulose acetate [5]	6–8	3–6	4–6	4–6	5.5		300–1000	4.5×10^{10}
Cellulose nitrate [6]	4–7			2.8–5			300–780	$2\text{–}30 \times 10^{10}$
Ceresin wax	2.5–2.6			0.12–0.21				
Cresol formaldehyde	6	10					400	
Dilectene	3.57					0.33		
Ethyl cellulose	2–2.7	0.7	1.2	1.5			1500	10^{15}
Fiber	5–7.5			4.5–5			150–180	5×10^{9}
Formica MF-66	4.6–4.9		1.5	1.1			450	
Glass:								
Cobalt	7.3			0.7				
Common window	7.6–8			1.4			200–250	
Crown	6.2–7		1	1 [3]			500	
Electrical	4–5			0.5			2000	8×10^{14}
Flint	7–10		0.45	0.4				
Nonex	4.2			0.25		0.28		
Photographic	7.5			0.8–1				
Plate	6.8–7.6			0.6–0.8				
Pyrex	4.2–4.9		0.5	0.7		0.54	335	10^{11}
Gutta percha	2.5–4.9						200–500	$5 \times 10^{14}\text{–}10^{15}$
Lucite [7]	2.5–3	7	5	1.5–3	1.9		480–500	
Melamine formaldehyde	8	16					300	
Mica	2.5–8	0.2	0.3	0.2–6	0.02			2×10^{17}
Mica (clear India)	6.4–7.5	2	2	2	2		600–1500	
Mycalex	7.4			0.18			250	10^{13}
Mycalex (British)	6			0.3			350	
Mykroy	6.5–7			0.1–0.2			630	
Nylon	3.6			2.2				
Paper	2.0–2.6						1250	
Paraffin wax (solid)	1.9–2.6			0.1–0.3			300	$10^{15}\text{–}10^{19}$
Pemque	7.21			0.2				
Phenol: [8]								
Pure	5			1			400–475	1.5×10^{12}
Asbestos base	7.5			15			90–150	
Black molded	5–5.5			3.5			400–500	
Fabric base	5–6.5			3.5–11			150–500	
Mica-filled	5–6			0.8–1			475–600	
Paper base	3.8–5.5			2.5–4			650–750	$10^{10}\text{–}10^{18}$
Yellow	5.3–5.4			0.36–0.7			500	
Polyethylene	2.3–2.4	0.02	0.02	0.02–0.05			1000	10^{17}
Polyindene	3	0.04						
Polyisobutylene	2.4–2.5	0.04 5	0.05				500	10^{16}
Polystyrene [9]	2.4–2.9(2.6)	0.02	0.018	0.02	0.02	0.02	500–2500	10^{20}
Porcelain (dry process)	6.2–7.5			0.7–15			40–100	5×10^{8}
Porcelain (wet process)	6.5–7			0.6			150	
Pressboard (untreated)	2.9–4.5						125–300	
Pressboard (oiled)	5						750	
Quartz (fused)	3.5–(3.8)	0.01	0.01	0.015–0.03	0.01	0.05	200	$10^{14}\text{–}10^{18}$
Rubber (hard) [10]	2–3.5(3)			0.5–1			450	$10^{12}\text{–}10^{15}$
Shellac	2.5–4			0.09			900	10^{16}
Steatite: [11]								
"Commercial" grade	4.9–6.5	0.02	0.2	0.2	0.4	0.5		
"Low-loss" grade	4.4	0.02	0.2	0.2	0.18	0.13	150–315	$10^{14}\text{–}10^{15}$
Titanium dioxide [12]	90–170		0.1	0.1				
Urea formaldehyde [13]	5–7	3–5	2–3	2–4	4		300–550	$10^{12}\text{–}10^{13}$
Varnished cloth [14]	2–2.5			2–3			440–550	
Vinyl resins	4			1.4–1.7			400–500	10^{14}
Vitrolex	6.4			0.3				
Wood (dry oak)	2.5–6.8(3)		3.8	4.2				
Wood (paraffined maple)	4.1						115	

[1] Most data taken at 25° C.
[2] Puncture voltage, in volts per mil. Most data applies to relatively thin sections and cannot be multiplied directly to give breakdown for thicker sections without added safety factor.
[3] In ohm-cm.
[4] Includes such products as Aladdinite, Ameroid, Galalith, Erinoid, Lactoid, etc.
[5] Includes Fibestas, Lumerith, Nixonite, Plastacele, Tenite, etc.
[6] Includes Amerith, Nitron, Nixonoid, Pyralin, etc.
[7] Methylmethacrylate resin.
[8] Phenolaldehyde products include Acrolite, Bakelite, Catalin, Celeron, Dielecto, Durez, Durite, Formica, Gemstone, Heresite, Indur, Makalot, Marblette, Micarta, Opalon, Prystal, Resinox, Synthane, Textolite, etc. Yellow bakelite is so-called "low-loss" bakelite.
[9] Includes Amphenol 912A, Distrene, Intelin IN 45, Loalin, Lustron, Quartz Q, Rezoglas, Rhodolene M, Ronilla L, Styraflex, Styron, Trolitul, Victron, etc.
[10] Also known as Ebonite.
[11] Soapstone — Alberene, Alsimag, Isolantite, Lava, etc.
[12] Rutile. Used in low temperature-coefficient fixed condensers.
[13] Includes Aldur, Beetle, Plaskon, Pollopas, Prystal, etc.
[14] Includes Empire cloth.

COPPER-WIRE TABLE

Gauge No. B. & S.	Diam. in Mils[1]	Circular Mil Area	Turns per Linear Inch[2]				Turns per Square Inch[2]			Feet per Lb.		Ohms per 1000 ft. 25° C.	Current Carrying Capacity at 1500 C.M. per Amp.[3]	Diam. in mm.	Nearest British S.W.G. No.
			Enamel	S.C.C.	D.S.C. or S.C.C.	D.C.C.	S.C.C.	Enamel S.C.C.	D.C.C.	Bare	D.C.C.				
1	289.3	83690	—	—	—	—	—	—	—	3.947	—	.1264	55.7	7.348	1
2	257.6	66370	—	—	—	—	—	—	—	4.977	—	.1593	44.1	6.544	3
3	229.4	52640	—	—	—	—	—	—	—	6.276	—	.2009	35.0	5.827	4
4	204.3	41740	—	—	—	—	—	—	—	7.914	—	.2533	27.7	5.189	5
5	181.9	33100	—	—	—	—	—	—	—	9.980	—	.3195	22.0	4.621	7
6	162.0	26250	—	—	—	—	—	—	—	12.58	—	.4028	17.5	4.115	8
7	144.3	20820	—	—	—	—	—	—	—	15.87	—	.5080	13.8	3.665	9
8	128.5	16510	7.6	—	7.4	7.1	—	—	—	20.01	19.6	.6405	11.0	3.264	10
9	114.4	13090	8.6	—	8.2	7.8	—	—	—	25.23	24.6	.8077	8.7	2.906	11
10	101.9	10380	9.6	—	9.3	8.9	87.5	84.8	80.0	31.82	30.9	1.018	6.9	2.588	12
11	90.74	8234	10.7	—	10.3	9.8	110	105	97.5	40.12	38.8	1.284	5.5	2.305	13
12	80.81	6530	12.0	—	11.5	10.9	136	131	121	50.59	48.9	1.619	4.4	2.053	14
13	71.96	5178	13.5	—	12.8	12.0	170	162	150	63.80	61.5	2.042	3.5	1.828	15
14	64.08	4107	15.0	—	14.2	13.8	211	198	183	80.44	77.3	2.575	2.7	1.628	16
15	57.07	3257	16.8	—	15.8	14.7	262	250	223	101.4	97.3	3.247	2.2	1.450	17
16	50.82	2583	18.9	18.9	17.9	16.4	321	306	271	127.9	119	4.094	1.7	1.291	18
17	45.26	2048	21.2	21.2	19.9	18.1	397	372	329	161.3	150	5.163	1.3	1.150	18
18	40.30	1624	23.6	23.6	22.0	19.8	493	454	399	203.4	188	6.510	1.1	1.024	19
19	35.89	1288	26.4	26.4	24.4	21.8	592	553	479	256.5	237	8.210	.86	.9116	20
20	31.96	1022	29.4	29.4	27.0	23.8	775	725	625	323.4	298	10.35	.68	.8118	21
21	28.46	810.1	33.1	32.7	29.8	26.0	940	895	754	407.8	370	13.05	.54	.7230	22
22	25.35	642.4	37.0	36.5	34.1	30.0	1150	1070	910	514.2	461	16.46	.43	.6438	23
23	22.57	509.5	41.3	40.6	37.6	31.6	1400	1300	1080	648.4	584	20.76	.34	.5733	24
24	20.10	404.0	46.3	35.3	41.5	35.6	1700	1570	1260	817.7	745	26.17	.27	.5106	25
25	17.90	320.4	51.7	50.4	45.6	38.6	2060	1910	1510	1031	903	33.00	.21	.4547	26
26	15.94	254.1	58.0	55.6	50.2	41.8	2500	2300	1750	1300	1118	41.62	.17	.4049	27
27	14.20	201.5	64.9	61.5	55.0	45.0	3030	2780	2020	1639	1422	52.48	.13	.3606	29
28	12.64	159.8	72.7	68.6	60.2	48.5	3670	3350	2310	2067	1759	66.17	.11	.3211	30
29	11.26	126.7	81.6	74.8	65.4	51.8	4300	3900	2700	2607	2207	83.44	.084	.2859	31
30	10.03	100.5	90.5	83.3	71.5	55.5	5040	4660	3020	3287	2534	105.2	.067	.2546	33
31	8.928	79.70	101	92.0	77.5	59.2	5920	5280	—	4145	2768	132.7	.053	.2268	34
32	7.950	63.21	113	101	83.6	62.6	7060	6250	—	5227	3137	167.3	.042	.2019	36
33	7.080	50.13	127	110	90.3	66.3	8120	7360	—	6591	4697	211.0	.033	.1798	37
34	6.305	39.75	143	120	97.0	70.0	9600	8310	—	8310	6168	266.0	.026	.1601	38
35	5.615	31.52	158	132	104	73.5	10900	8700	—	10480	6737	335.0	.021	.1426	38–39
36	5.000	25.00	175	143	111	77.0	12200	10700	—	13210	7877	423.0	.017	.1270	39–40
37	4.453	19.83	198	154	118	80.3	—	—	—	16660	9309	533.4	.013	.1131	41
38	3.965	15.72	224	166	126	83.6	—	—	—	21010	10666	672.6	.010	.1007	42
39	3.531	12.47	248	181	133	86.6	—	—	—	26500	11907	848.1	.008	.0897	43
40	3.145	9.88	282	194	140	89.7	—	—	—	33410	14222	1069	.006	.0799	44

[1] A mil is 1/1000 (one thousandth) of an inch.
[2] The figures given are approximate only, since the thickness of the insulation varies with different manufacturers.
[3] The current-carrying capacity at 1000 C.M. per ampere is equal to the circular-mil area (Column 3) divided by 1000.

VACUUM-TUBE CLASSIFIED DATA TABLES AND INDEX

In the tables on pages 426–459 will be found essential characteristics and typical operating data for U. S.-made standard receiving, transmitting and special-purpose vacuum tubes. Base diagrams are shown on pages 419–425. For convenience in locating types the index below lists them in numerical alphabetical order with the page on which data is to be found. The base-diagram number for each tube follows the page number. When a tube appears in more than one table it is listed twice.

Type	Page	Base
42	431	6B
43	437	6B
45	432	4D
45Z3	444	5AM
45Z5GT	445	6AD
46	432	5C
47	432	5B
48	437	6A
49	433	5C
50	438	4D
50A5	437	6AA
50B5	440	7BZ
50C6G	437	7AC
50L6GT	437	7AC
50T	450	2D
50Y6GT	444	7Q
50Z6G	445	7Q
50Z7G	445	8AN
51	432	5E
52	431	Fig. 33
53	432	7B
53A	448	T-4B
55	432	6G
56	432	5A
56AS	431	5A
57	432	6F
57AS	431	6F
58	432	6F
58AS	431	6F
59	432	7A
70A7GT	437	8AB
70A7GT	445	8AB
70L7GT	437	8AA
70L7GT	445	8AA
71-A	438	4D
72	445	4P
73	445	4Y
75	431	6G
75TH	450	2D
75TL	450	2D
76	431	5A
77	431	6F
78	431	6F
79	431	6H
80	445	4C
81	445	4B
82	445	4C
83	445	4C
83-V	445	4AD
84/6Z4	445	5D
85	431	6G
85AS	431	6G
89	438	4D
99	438	4D
100TH	451	2D
100TL	451	2D
112-A	438	4D
117L7GT	437	8AO
117L7GT	445	8AO
117M7GT	437	8AO
117M7GT	445	8AO
117N7GT	437	8AV
117N7GT	445	8AV
117P7GT	437	8AV
117P7GT	445	8AV
117Z3	445	4BR
117Z4GT	445	5AA
117Z6GT	445	7Q
150T	453	T-3AC
152TH	453	4BC
152TL	453	4BC
182-B	438	4D
183	438	4D
203-A	451	4E
203-H	451	4E
203-H	451	T-3AB
204-A	454	T-1A
205D	446	4D
211	451	4E
212-E	454	T-2A
217A	445	T-3A
217C	445	T-3A
227A	451	T-4B
241B	454	T-2AA
242A	450	4E
242B	451	4E
242C	452	4E
249B	445	Fig. 53
250TH	454	T-3AC
250TL	454	T-3AC
254A	456	T-4C
254B	457	T-4C
261A	452	4E
270A	455	T-1A
276A	452	4E
282A	458	T-4C
284B	452	T-3AB
284D	450	4E
295A	452	4E
300T	454	T-3AC
303A	451	4E
304A	454	T-1A
304B	454	T-1A
304TH	455	T-4BF
304TL	455	T-4BF
305A	458	T-4CE
306A	456	T-5CB
307A	456	T-5C
308B	454	T-2A
310	447	4D
311	451	4E
312A	458	T-6C
312E	454	T-2AA
316A	448	—
327A	451	T-4AD
327B	450	T-4AB
342B	451	4E
356A	449	T-4BD
361A	452	4E
376A	452	4E
482B	438	4D
483	438	4D
485	438	5A
527	454	T-4B
705A	445	T-3AA
717A	429	8BK
756	448	4D
800	448	2D
801	447	4D
801A	447	4D
802	456	T-7C
803	459	T-5C
804	458	T-5C
805	453	T-3AB
806	454	T-3AC
807	457	T-5BB
808	449	2D
809	448	3G
810	453	T-3AC
811	449	3G
812	449	3G
812H	450	3G
813	458	Fig. 28
814	458	T-5D
815	457	T-8FA
816	445	4P
822	454	T-3AB
822S	454	T-3AC
826	450	T-9A
828	458	T-5C
829	457	7BP
829A	457	7BP
829B	457	7BP
830	448	4D
830B	450	3G
831	455	T-4AA
832	456	7BP
832A	456	7BP
833A	455	T-4AB
834	449	2D
835	451	4E
836	445	4P
837	456	T-7C
838	452	4E
840	433	5J
841	440	4D
841A	449	3G
841SW	449	3G
843	446	5A
844	456	T-5BB
849	455	T-1A
850	458	T-3B
852	452	2D
860	458	T-4CB
861	459	T-4
864	438	4D
865	456	T-4C
866	445	4P
866A	445	4P
866B	445	4P
866 Jr.	445	4B
871	445	4P
872	445	T-3A
872A	445	T-3A
874	441	4S
876	441	—
878	445	4P
879	445	4P
884	441	6Q
885	441	5A
886	441	—
902	443	Fig. 1
903	443	6AL
904	443	Fig. 3
905	443	Fig. 6
906P1	442	7AN
907	443	Fig. 6
908	443	7AN
909	443	Fig. 6
910	443	7AN
911	443	7AN
912	443	Fig. 8
913	443	Fig. 1
914	443	Fig. 12
930B	450	3G
938	452	4E
950	433	5R
951	432	4M
954	438	5BB
955	438	5BC
955	446	5BC
956	438	5BB
957	438	5BD
958	446	5BD
958A	438	5BD
958A	438	5BO
959	438	5BE
967	441	3G
975A	445	T-3A
991	441	—
1003	445	4R
1201	438	8BM
1203	438	4AH
1204	438	Fig. 5
1206	430	8BV
1221	431	6F
1223	429	7R
1229	433	4K
1231	430	8V
1232	429	8V
1265	441	—
1266	441	4AJ
1267	441	4V
1275	445	4C
1276	438	4D
1284	437	Fig. 4
1291	434	7BE
1293	434	Fig. 2
1294	434	4AH
1299	434	6BB
1602	447	4D
1603	431	6F
1608	447	4D
1609	438	5B
1610	455	T-5CA
1611	427	7S
1612	427	7T
1613	455	7S
1614	457	7AC
1616	445	4P
1619	456	7AC
1620	427	7R
1621	427	7S
1622	427	7AC
1623	448	3G
1624	457	T-5DC
1625	457	Fig. 29
1626	446	6Q
1627	453	T-3AC
1628	448	T-4BB
1629	437	8RA
1631	437	7AC
1632	437	7AC
1633	437	8BD
1634	437	8S
1635	429	8B
1641	445	T-4AG
1642	430	7BH
1644	437	Fig. 7
1654	445	Fig. 41
1800	443	6AL
1801	443	Fig. 13
1802P1	442	11A
1803P4	443	6AL
1804P4	443	6AL
1805P1	442	11A
1806P1	442	11A
1809P1	443	8BR
1811P1	443	6AZ
1851	427	7R
1852	426	8N
1853	426	8N
2001	443	Fig. 2
2002	443	Fig. 1
2005	443	Fig. 1
2050	441	8BA
2051	441	8BA
2523N/128AS	441	5A
7000	429	7R
7193	446	4AM
7700	431	6F
8000	453	T-3AC
8001	458	T-7CB
8003	452	T-3AB
8005	450	3G
8008	445	Fig. 11
8010-R	448	—
8012	448	T-4BB
8013-A	445	4P
8016	445	T-4BC
8020	445	4P
8025	447	4AQ
9001	440	7PM
9002	440	7TM
9002	446	7TM
9003	440	7PM
9004	438	4BJ
9005	438	Fig. 16
9006	440	6BH
AT-340	459	Fig. 27
BA	444	4J
BH	444	4J
BR	444	4H
CE220	444	4P
CK501	434	—
CK502	434	—
CK503	434	—
CK504	434	—
CK505	434	—
CK506	434	—
CK507	434	—
CK509	435	—
CK510	435	—
CK515	435	—
CK1005	445	T-9F
CK1006	445	4C
CK1007	445	T-9G
CK1009	445	—
DR123C	452	Fig. 26
EF50	439	Fig. 14
F123A	452	Fig. 26
F127A	453	Fig. 26
G84	444	2B
GL2C44	439	Fig. 17
GL2C44	446	Fig. 17
GL146	452	T-4BG
GL152	452	T-4BG
GL159	454	T-4BG
GL169	454	T-4BG
GL446A	439	Fig. 19
GL446B	439	Fig. 19
GL464A	439	Fig. 17
GL559	439	Fig. 18
GL592	454	Fig. 52
GL8012A	448	T-4BB
HD203A	453	T-3AB
HF60	449	2D
HF75	450	2D
HF100	450	2D
HF120	451	—
HF125	451	—
HF130	452	—
HF140	451	—
HF150	452	—
HF175	452	—
HF200	453	T-3AC
HF250	453	T-3AC
HF300	454	T-3AC
HK24	447	3G
HK54	449	2D
HK154	449	2D
HK158	449	2D
HK252L	453	T-4BF
HK253	445	T-3A
HK254	451	T-3AC
HK257	458	T-7CB
HK257B	458	T-7CB
HK304L	454	T-4BF
HK354	453	T-3AC
HK354C	453	T-3AC
HK354C	453	T-3AC
HK354D	453	T-3AC
HK354E	453	T-3AC
HK354F	453	T-3AC
HK454H	454	T-3AC
HK454L	454	T-3AC
HK654	455	T-3AC
HV12	454	T-3AB
HV18	453	T-3AC
HV27	454	T-3AB
HY6J5GTX	446	6Q
HY6L6GTX	456	7AC
HY6V6GTX	456	7AC
HY24	446	4A
HY25	447	3G
HY30Z	447	T-4BE
HY31Z	447	T-4D
HY40	448	3G
HY40Z	448	3G
HY51A	450	3G
HY51B	450	3G
HY51Z	450	T-4BE
HY57	448	3G
HY60	456	T-5BB
HY61	457	T-5BB
HY63	455	T-8DB
HY65	456	T-8DB
HY67	458	T-5DB
HY69	457	T-5D
HY75	446	T-8AC
HY113	435	5K
HY114B	446	T-8AC
HY115	435	5K
HY123	435	5K
HY125	435	5K
HY145	435	5K
HY155	435	5K
HY615	446	T-8AG
HY801A	447	4P
HY866Jr.	445	5P
HY1231Z	447	T-4
HY1269	457	T-5DB
HYE1148	446	T-8AG
KY21	441	—
KY866	441	Fig. 8
M54	439	—
M64	439	—
M74	439	—
RK10	447	4D
RK12	447	3G
RK15	432	4D
RK16	432	5A
RK17	432	5F
RK18	448	3G
RK19	445	T-3A
RK20	457	T-5C
RK20A	457	T-5C
RK21	445	4P
RK22	445	T-4AG
RK23	455	T-7C
RK24	433	4D
RK24	446	4D
RK25	455	T-7C
RK25B	455	T-7C
RK28	458	T-5C
RK28A	459	T-5C
RK30	448	2D
RK31	448	3G
RK32	448	2D
RK33	446	T-7DA
RK34	446	T-7DC
RK35	449	2D
RK36	451	2D
RK37	449	2D
RK38	451	2D
RK39	457	T-5BB
RK41	457	T-5BB
RK42	435	4D
RK43	435	6C
RK44	456	T-7C
RK46	457	T-5C
RK47	457	T-5D
RK48	458	T-5D
RK48A	458	T-5D
RK49	457	T-6B
RK51	449	3G
RK52	449	3G
RK56	455	T-5BB
RK57	452	T-3AB
RK58	451	T-3AB
RK59	446	T-4D
RK60	445	T-4AG
RK62	441	4D
RK63A	453	T-3AC
RK64	455	T-5BB
RK65	459	T-3BC
RK66	457	T-5C
RK75	456	T-5C
RK100	447	T-6B
RK705A	445	T-3AA
RK866	445	4P
RM208	441	—
RM209	441	—
T20	447	3G
T21	457	T-6B
T40	448	3G
T55	449	3G
T60	449	2D
T100	450	2D
T125	452	T-3AC
T200	453	T-3AC
T814	454	T-3AB
T822	454	T-3AB
TUF20	447	T-8AC
TW75	450	2D
TW150	453	T-3AC
TZ20	447	3G
TZ40	448	3G
UH35	450	3G
UH50	449	2D
UH51	449	2D
V70	450	T-3AB
V70-A	450	T-3AB
V70B	450	3G
V70C	450	3G
V70D	451	3G
VR75	441	4AJ
VR90	441	4AJ
VR105	441	4AJ
VR150	441	4AJ
VT127A	451	T-4B
WE304A	449	2D
XXB	439	Fig. 9
XXD	437	8AC
XXL	430	5AC
XXFM	439	Fig. 10
Z225	445	4P
ZB120	450	4E

VACUUM-TUBE BASE DIAGRAMS

The diagrams on the following pages show standard socket connections corresponding to the base designations given in the column headed "Socket Connections" in the classified tube data tables. Bottom views are shown throughout. Terminal designations are as follows:

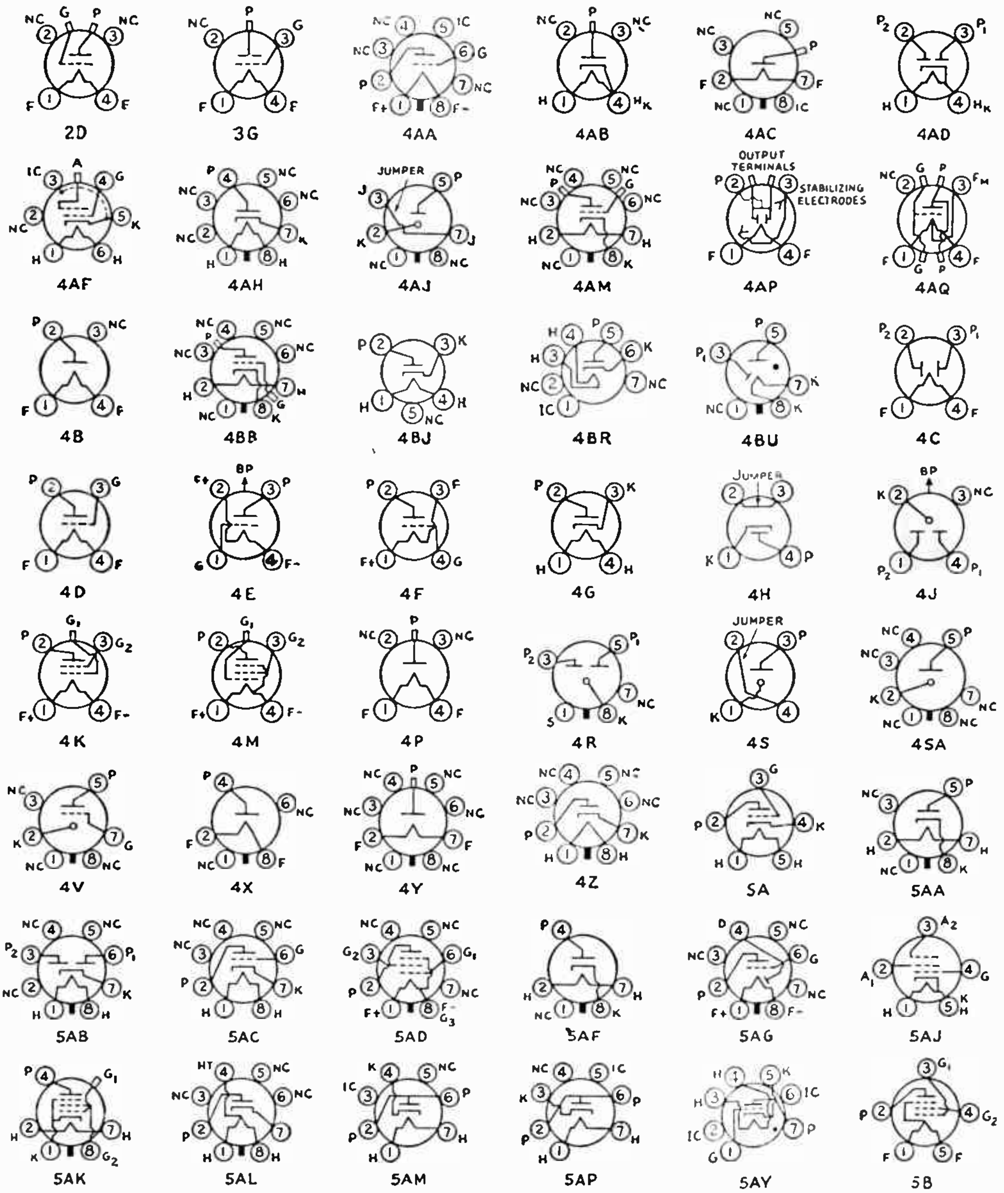
A	= Anode	F	= Filament	IS	= Internal Shield	P_I	= Starter-Anode	S	= Shell
BP	= Bayonet Pin	G	= Grid	K	= Cathode	P_{BF}	= Beam-Forming Plates	TA	= Target
BS	= Base sleeve	H	= Heater	NC	= No Connection	RC	= Ray-Control Electrode	●	= Gas-Type Tube
D	= Deflecting Plate	IC	= Internal Connection	P	= Plate (Anode)			U	= Unit
								SH	= Internal Shield

Alphabetical subscripts D, P, T and HX indicate, respectively, diode unit, pentode unit, triode unit or hexode unit in multi-unit types. Subscript M, T or CT indicates filament or heater tap.

Wherever the No. 1 pin of a metal-type tube in Table I is shown connected to the shell, the No. 1 pin in the glass (G or GT) equivalent is connected to an internal shield.

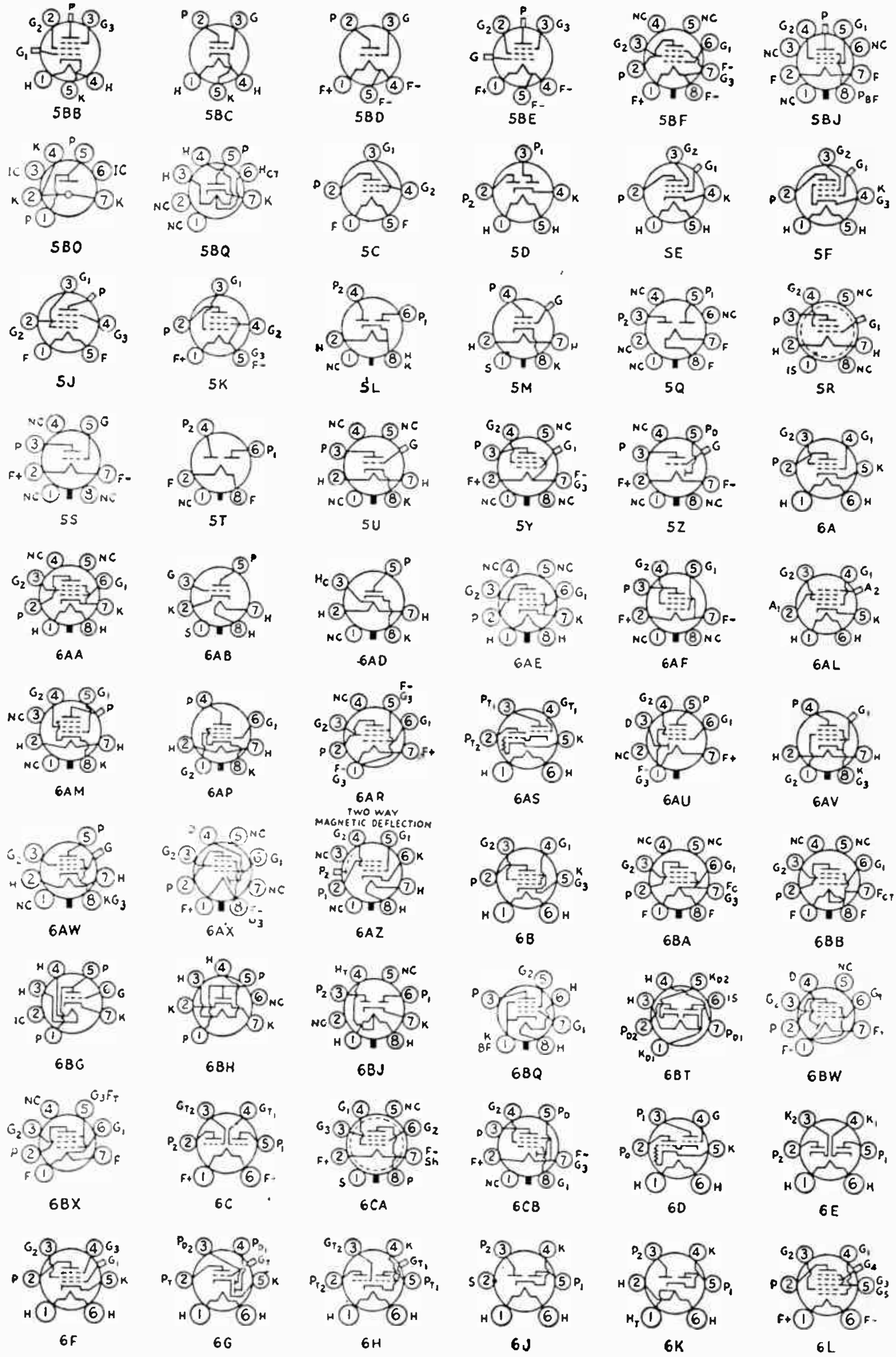
RMA TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are shown above.



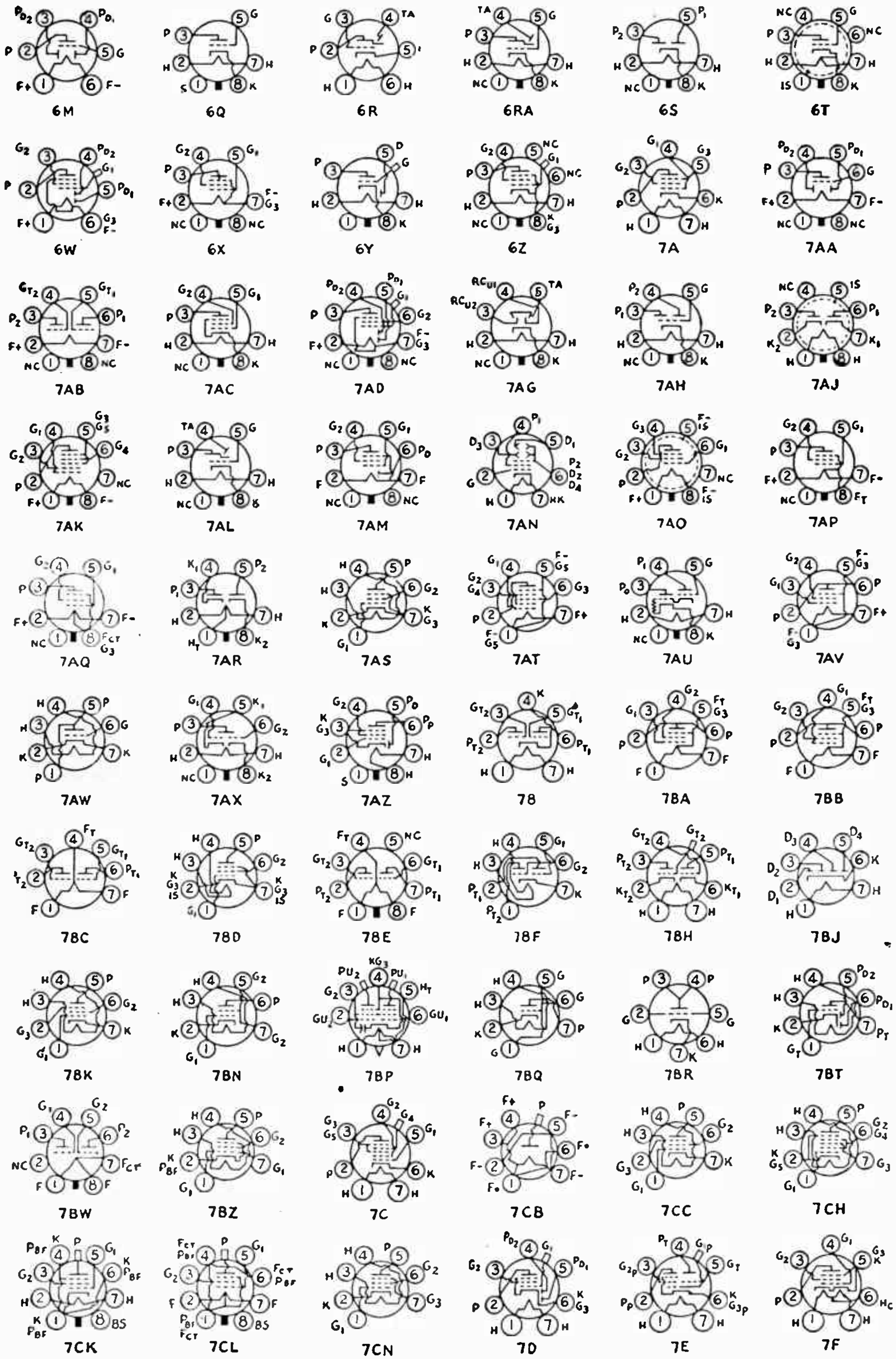
RMA TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 419.



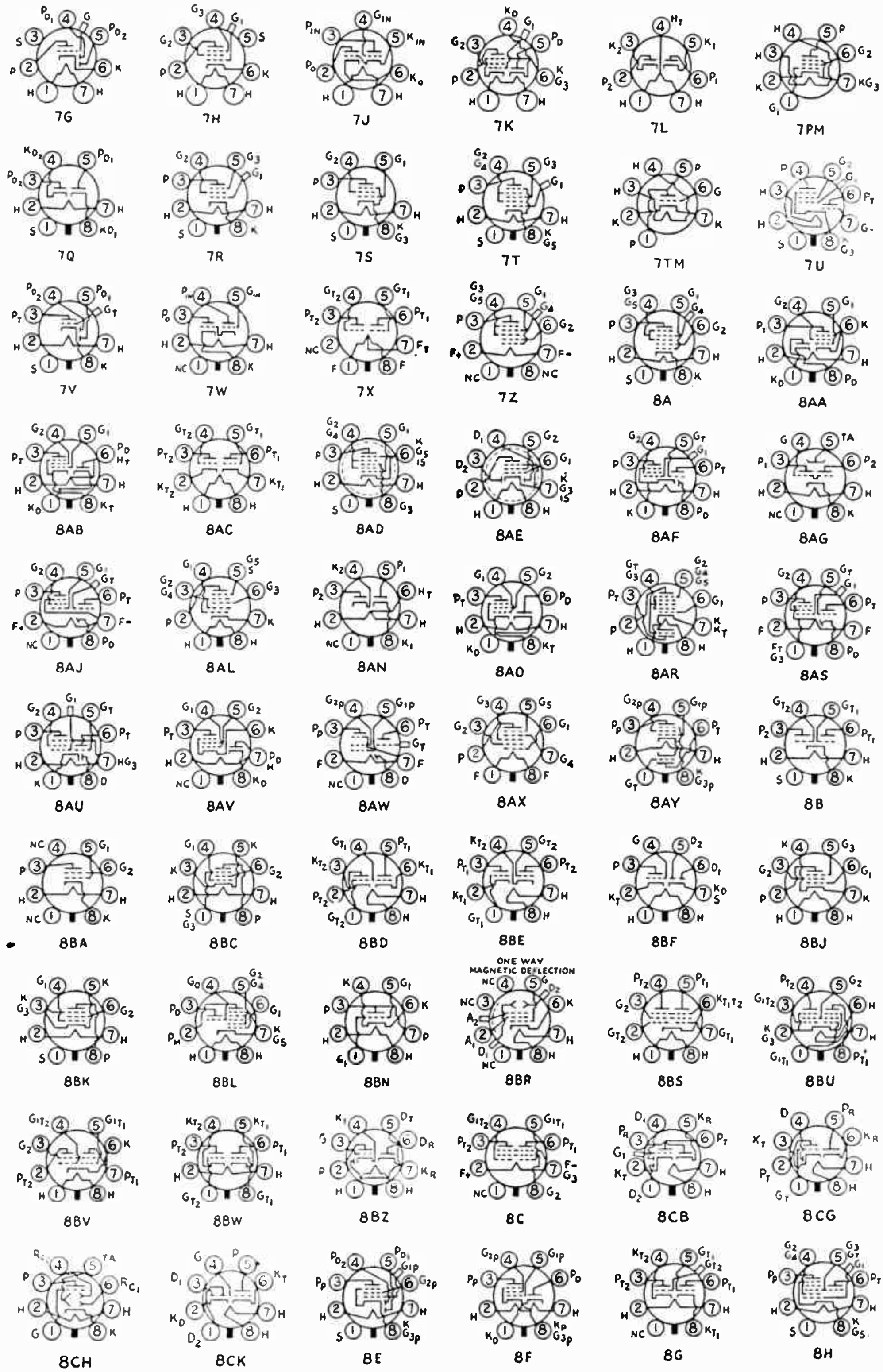
RMA TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 419.



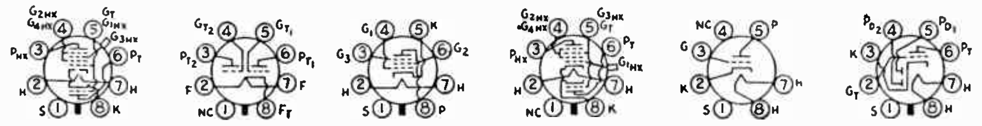
RMA TUBE BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 419.

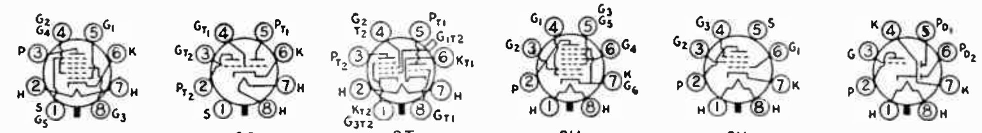


RMA TUBE BASE DIAGRAMS

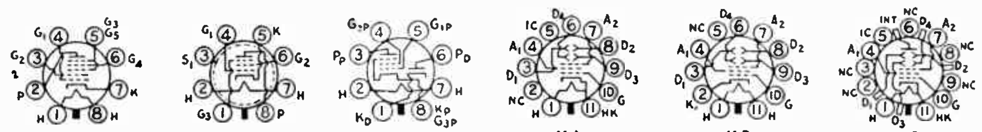
Bottom views are shown. Terminal designations on sockets are given on page 419.



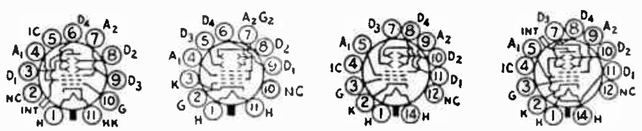
8K 8L 8N 8O 8P 8Q



8R 8S 8T 8U 8V 8W



8X 8Y 8Z 11A 11B 11E



11F 11M 14A 14B

SUPPLEMENTARY BASE DIAGRAMS

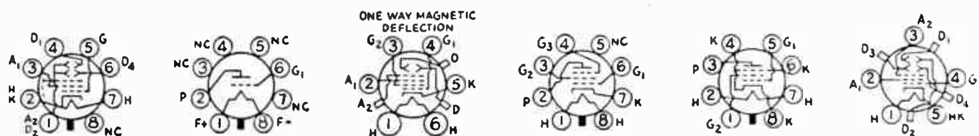


FIG. 1 FIG. 2 FIG. 3 FIG. 4 FIG. 5 FIG. 6

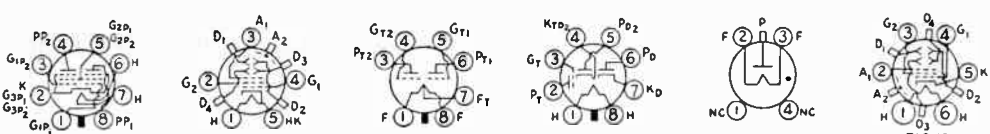


FIG. 7 FIG. 8 FIG. 9 FIG. 10 FIG. 11 FIG. 12

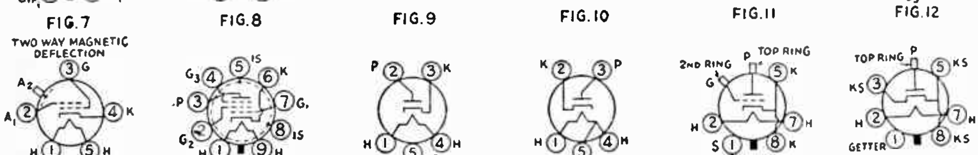


FIG. 13 FIG. 14 FIG. 15 FIG. 16 FIG. 17 FIG. 18

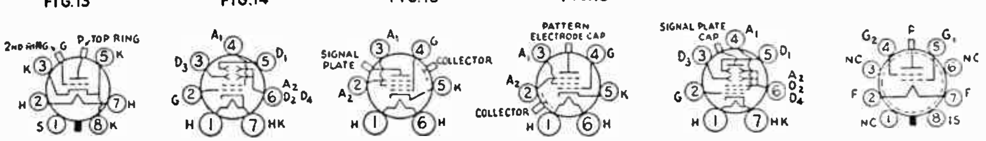


FIG. 19 FIG. 20 FIG. 21 FIG. 22 FIG. 23 FIG. 24

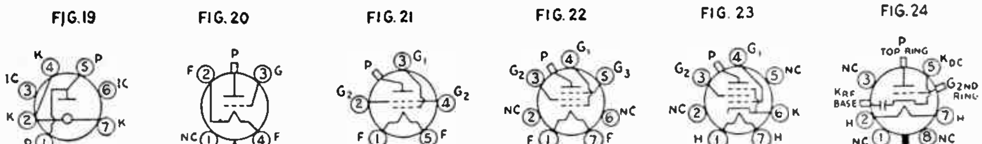


FIG 25 FIG. 26 FIG. 27 FIG. 28 FIG 29 FIG. 30

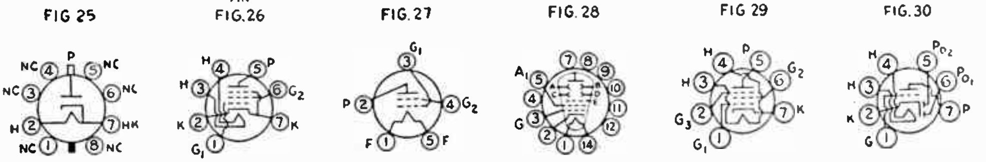
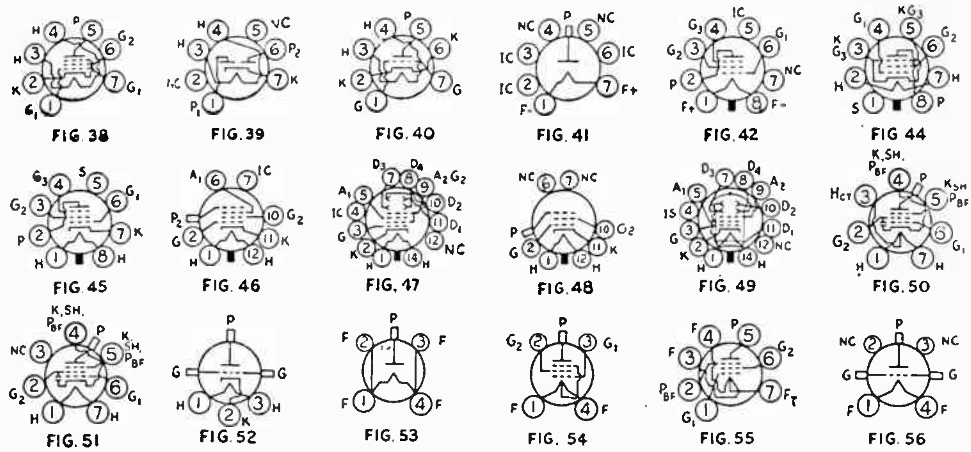


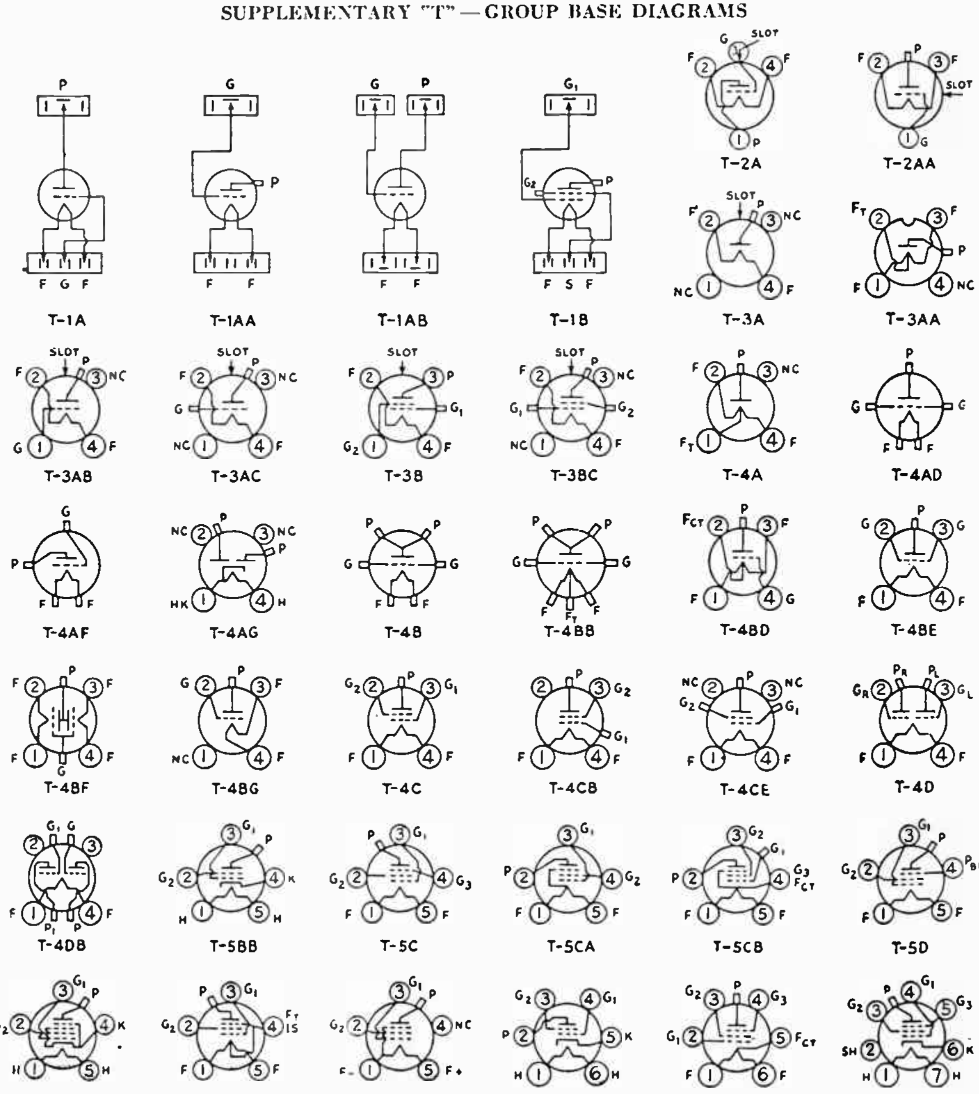
FIG 31 FIG 32 FIG 33 FIG 34 FIG 36 FIG 37

SUPPLEMENTARY BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 419.

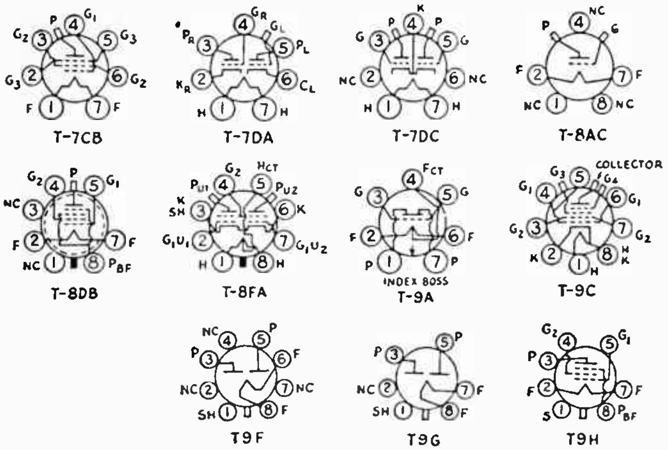


SUPPLEMENTARY "T"—GROUP BASE DIAGRAMS



SUPPLEMENTARY "T"—GROUP BASE DIAGRAMS

Bottom views are shown. Terminal designations on sockets are given on page 419.



BASE TYPE—DESIGNATIONS

The type of base used on each tube listed in the tables is indicated in the base column by a letter. The meaning of each letter is as follows:

A = Acorn
B = Glass button miniature
J = Jumbo
L = Lock-in
M = Medium
N = None or special type
O = Octal
S = Small
W = Wafer

TUBE RATINGS

The data in the classified tube tables are of two kinds, maximum ratings, and typical operating conditions.

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 addition to the maximum ratings for each type, performance data are given in the form of typical operating conditions.

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.

The maximum ratings given for each transmitting type apply only when the tube is operated at frequencies up to the specified maximum frequency for full rating as listed in the column so headed. As the frequency is raised above the specified value, the radio-frequency current, dielectric losses, and heating effects increase rapidly. Most types can be operated above their specified maximum frequency provided the plate voltage and plate input are reduced.

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 given in the tables are ICAS ratings when applicable.

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 Amps.	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.-Mixer	250	−3.0	100	3.2	3.3	Anode-grid (No. 2) 250 volts max. thru 20,000-ohms					6A8
6AB7 1853	Television Amp. Pentode	8N	6.3	0.45	8	5	0.015	Class-A Amp.	300	−3.0	200	3.2	12.5	700000	5000	3500	—	—	6AB7 1853
6AC7 1852	Television Amp. Pentode	8N	6.3	0.45	11	5	0.015	Class-A Amp.	300	−2.0	150	2.5	10	750000	9000	6750	—	—	6AC7 1852
6AG7	Video Beam Power Amp.	8Y	6.3	0.65	13	7.5	0.06	Class-A 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 Detector, Amplifier	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	—	—	—	Class-A Pent.	250 315	−16.5 −22.0	250 315	6.5 8.0	34 42	80000 75000	2500 2650	200 200	7000 7000	3.0 5.0	6F6
								Triode Amp.[1]	250	−20.0	—	—	31	2600	2700	7.0	4000	0.85	
								P.P. Pentodes P.P. Triodes[1]	375 350	−26.0 −38.0	250 —	2.5 —	17 22.5	Power output for 2 tubes at stated load, plate-to-plate			10000 6000	19.0 18.0	
6H6	Twin Diode	7Q	6.3	0.3	—	—	—	Rectifier	Max. a.c. voltage per plate = 100 r.m.s. Max. output current 4.0 ma. d.c.										6H6
6J5	Detector Amplifier Triode	6Q	6.3	0.3	3.4	3.6	3.4	Class-A Amp.	250	−8.0	—	—	9	7700	2600	20	—	—	6J5
6J7	Triple-Grid Detector, Amp.	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	Triple-Grid Variable-μ Amp.	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 Converter	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	—	—	—	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	270	125*	270	11/17	134/145	—	—	—	5000	18.5	
								P.P. Class A_1	250 270	−16.0 −17.5	250 270	10/16 11/17	120/140 134/155	24500 23500	5500 5700	—	5000 5000	14.5 17.5	
								P.P. Class AB_1	360	250*	270	5/17	88/100	Power output for 2 tubes. Load plate-to-plate			9000	24.5	
								P.P. Class AB_1	360	−22.5	270	5/15	88/132				6600	26.5	
								P.P. Class AB_2	360 360	−18.0 −22.5	225 270	3.5/11 5/16	78/142 88/205				6000 3800	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 = −15.0				
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	Triple-Grid Variable-μ	7R	6.3	0.15	6.5	10.5	0.005	Class-A Amp.	250	−3.0	100	2.0	8.5	1000000	1750	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
6SC7	Twin Triode Amplifier	8S	6.3	0.3	—	—	—	Class-A Amp.	250	−2.0	—	—	2.0	53000	1325	70	—	—	6SC7
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	Triple-Grid Semi-Variable-μ	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

TABLE I—METAL RECEIVING TUBES—*Continued*

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amps.	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
6SH7	Triple-Grid Amplifier	8BK	6.3	0.3	8.5	7	0.003	H.F. Amp.	250	– 1.0	150	4.1	10.8	900000	4900	—	—	—	6SH7
6SJ7[4]	Triple-Grid Amplifier	8N	6.3	0.3	6	7	0.005	Class-A Amp.	250	– 3.0	100	0.8	3	1500000	1650	2500	—	—	6SJ7
6SK7	Triple-Grid Variable-μ	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	3.6	3.2	1.80	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	Triple-Grid Variable-μ	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	—	—	—	Class-A Amp.	250	–12.5	250	4.5/7.0	45/47	52000	4100	218	5000	4.5	6V6
								Class-AB Amp.	250	–15.0	250	5/13	70/79	60000	3750	—	10000	10.0	
									285	–19.0	285	4/13.5	70/92	65000	3600	—	8000	14.0	
1611	Pentode Power Amplifier	7S	6.3	0.7	—	—	—	Relay Tube	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	Triple-Grid Det.-Amp.	7R	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6J7										1620
1621	Power Amplifier Pentode	7S	6.3	0.7	—	—	—	P.P. Pentodes	300	–30.0	300	6.5/13	38/69	—	—	—	4000	5.0	1621
								P.P. Triodes[1]	330	500*	—	—	55/59	—	—	—	5000	2.0	
1622	Beam Power Amplifier	7AC	6.3	0.9	—	—	—	Class-A 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

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* Cathode resistor-ohms. [1] Screen tied to plate. [2] For 6SA7GT, use Base Diagram 8AD. Grid bias—2 volts if separate oscillator excitation is used. [4] Also type "6SJ7Y".

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 Amps.	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
2C22	Triode Amplifier	4AM	6.3	0.3	2.2	0.7	3.60	Class-A Amp.	300	–10.5	—	—	11	6600	3000	20	—	—	2C22
6A5G	Triode Power Amplifier	6T	6.3	1.0	—	—	—	Class-A Amp.	250	–45.0	—	—	60	800		4.2	2500	3.75	6A5G
								P.P. Class AB	325	–68.0	—	—	80	—	5250	—	3000	15.0	
								P.P. Class AB	325	850*			80	—		—	5000	10.0	
6AB6G	Direct-Coupled Amplifier	7AU	6.3	0.5	—	—	—	Class-A Amp.	250	0	Input		5.0	40000	1800	72	8000	3.5	6AB6G
									250	0	Output		34						
6AC5G	High-μ Power Amplifier Triode	6Q	6.3	0.4	—	—	—	P.P. Class B	250	0	—	—	5.0	36700	3400	125	10000	8.0	6AC5G
								Dyn.-Coupled	250	—	—	—	32				7000	3.7	
6AC6G	Direct-Coupled Amplifier	7AU	6.3	1.1	—	—	—	Class-A Amp.	180	0	Input		7.0	—	3000	54	4000	3.8	6AC6G
									180	0	Output		45						
6AD5G	High-μ Triode	6Q	6.3	0.3	—	—	—	Class-A Amp.	250	– 2.0	—	—	0.9	—	1500	100	—	—	6AD5G
6AD6G	Electron-Ray Tube	7AG	6.3	0.15	—	—	—	Indicator	100		0 for 90°; –23 for 135°; 45 for 0°. Target current 1.5 ma.								6AD6G
6AD7G	Triode-Pentode	8AY	6.3	0.85	—	—	—	Triode Amp.	250	–25.0	—	—	4.0	19000	325	6.0	—	—	6AD7G
								Pentode Amp.	250	–16.5	250	6.5	34	80000	2500	—	7000	3.2	
6AE5G	Triode Amplifier	6Q	6.3	0.3	—	—	—	Class-A Amp.	95	–15.0	—	—	7.0	3500	1200	4.2	—	—	6AE5G
6AE6G	Twin-Plate Triode with Single Grid	7AH	6.3	0.15	Remote cut-off			Class-A Amp.	250	– 1.5			6.5	25000	1000	25	—	—	6AE6G
					Sharp cut-off			Class-A Amp.	250	– 1.5			4.5	35000	950	33			

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amps.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
6AE7GT	Twin-Input Triode	7AX	6.3	0.5	—	—	—	Driver Amplifier	250	−13.5	—	—	5.0	9300	1500	14	—	—	6AE7GT
6AF5G	Triode Amplifier	6Q	6.3	0.3	—	—	—	Class-A Amplifier	180	−18.0	—	—	7.0	—	1500	7.4	—	—	6AF5G
6AF7G	Twin Electron Ray	8AG	6.3	0.3	—	—	—	Indicator Tube											6AF7G
6AG6G	Power Amplifier Pentode	7S	6.3	1.25	—	—	—	Class-A Amplifier	250	− 6.0	250	6.0	32	—	10000	—	8500	3.75	6AG6G
6AH5G	Beam Power Amplifier	6AP	6.3	0.9	—	—	—	Class-A Amplifier	350	−18	250	—	—	33000	5200	—	4200	10.8	6AH5G
6AH7GT	Twin Triode	8BE	6.3	0.3	—	—	—	Converter & Amp.	250	− 9.0	—	—	12[1]	6600	2400	16	—	—	6AH7GT
6AL6G	Beam Power Amplifier	6AM	6.3	0.9	—	—	—	Class-A Amplifier	250	−14.0	250	5.0	72	22500	6000	—	2500	6.5	6AL6G
6AL7GT	Electron-Ray Tube	8CH	6.3	0.15	—	—	—	Indicator	Outer edge of any of the three illuminated areas displaced 1/16 in. min. outward with +5 volts to its electrode. Similar inward disp. with −5 volts. No pattern with −6 volts grid.										6AL7GT
6AQ7GT	Duplex Diode Triode	8CK	6.3	0.3	2.3	1.5	2.8	Class-A Amplifier	250	− 2.0	—	—	2.3	44000	1600	70	—	—	6AQ7GT
6AR6	Beam Power Amp.	6BQ	6.3	1.2	11	7	0.8	Class-B Amplifier	300	−36	300	4.0	58	22000	4300	95	—	—	6AR6
6AR7GT	Diode Triode Rectifier	8CG	6.3	0.3	1.4	1	2	Class-A Amplifier	250	− 2	—	—	1.3	66500	1050	70	—	—	6AR7GT
6AS7G	Low-Mu Twin Triode	8BD	6.3	2.5	—	—	—	D.C. Amplifier	135	250*	—	—	125	280	7500	2.1	—	—	6AS7G
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
6C8G	Twin Triode	8G	6.3	0.3	—	—	—	Amp. 1 Section	250	− 4.5	—	—	3.1	26000	1450	38	—	—	6C8G
6D8G	Pentagrid Converter	8A	6.3	0.15	—	—	—	Converter	250	− 3.0	100	Cathode current 13.0 Ma.			Anode grid (No. 2) Volts = 250[3]				6D8G
6E8G	Triode-Hexode Converter	8O	6.3	0.3	—	—	—	Osc.-Mixer	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	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	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
6K6G	Pentode Power Amplifier	7S	6.3	0.4	—	—	—	Class-A Amplifier	Characteristics same as Type 41—Table IV										6K6G
6L5G	Triode Amplifier	6Q	6.3	0.15	—	—	—	Class-A Amplifier	250	− 9.0	—	—	8.0	—	1900	17	—	—	6L5G
6M6G	Power Amplifier Pentode	7S	6.3	1.2	—	—	—	Class-A Amplifier	250	− 6.0	250	4.0	36	—	9500	—	7000	4.4	6M6G
6M7G	Triple-Grid 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	Direct-Coupled Amplifier	7AU	6.3	0.8	—	—	—	Power Amplifier	Characteristics same as Type 6B5—Table IV							—	—	—	6N6G
6P5GT	Triode Amplifier	6Q	6.3	0.3	3.4	5.5	2.6	Class-A Amplifier	250	−13.5	—	—	5.0	9500	1450	13.8	—	—	6P5GT
6P7G	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	—	—	—	Osc.-Mixer	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	—	—	—	Class-A Amplifier	250	− 3.0	100	1.7	7.0	—	1450	1160	—	—	6R6G
6S6GT	Triple-Grid Variable-μ	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	—	100	—	—	6S8GT
6SD7GT	Triple-Grid Semi-Variable-μ	8M	6.3	0.3	9	7.5	.0035	R.F. Amplifier	250	− 2.0	100	1.9	6.0	1000000	3600	—	—	—	6SD7GT
6SE7GT	Triple-Grid Amplifier	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.	Fig. 44	6.3	0.3	—	—	—	Class-A Amplifier	100	− 1.0	100	2.1	5.3	350000	4000	—	—	—	6SH7L
									250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	
6SL7GT	Twin Triode	8BD	6.3	0.3	—	—	—	Amplifier	250	− 2.0	—	—	2.3[1]	44000	1600	70	—	—	6SL7GT
6SN7GT	Twin Triode	8BD	6.3	0.6	—	—	—	Amplifier	250	− 8.0	—	—	9.0[1]	7700	2600	20	—	—	6SN7GT
6SU7GTY	Twin Triode	8BD	6.3	0.3	—	—	—	Class-A Amplifier	250	− 2.0	—	—	2.3	44000	1600	70	—	—	6SU7GTY

TABLE II—6.3-VOLT GLASS TUBES WITH OCTAL BASES—*Continued*

Type	Name	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amps.	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
6T6GM	Triple-Grid Amplifier	6Z	6.3	0.45	—	—	—	R.F. 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	Triple Grid Variable-μ	7R	6.3	0.3	5	9	.007	R.F. Amplifier	Characteristics same as Type 6D6—Table III										6U7G
6V7G	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	Triple-Grid 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	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	Pentode Amplifier	8BK	6.3	0.175	—	—	—	Class-A Amplifier	120	−2.0	120	2.5	7.5	390000	4000	—	—	—	717A
1223	Pentode Amplifier	7R	6.3	0.3	—	—	—	Class-A Amplifier	Characteristics same as 6C6—Table IV										1223
1635	Twin Triode Amplifier	8B	6.3	0.6	—	—	—	Class-B Amplifier	400	0	—	—	10/63	—	—	—	14000	17	1635
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.

TABLE III—7-VOLT LOCK-IN-BASE TUBES

For other lock-in-base types see Tables VIII, IX, X and XIII

Type	Name	Socket Connections	Heater Volts	Heater Amps.	Capacitance μμfd. In	Capacitance μμfd. Out	Capacitance μμfd. Plate-Grid	Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
7A4	Triode Amplifier	5AC	7.0	0.32	3.4	3	4	Class-A Amplifier	250	−8.0	—	—	9.0	7700	2600	20	—	—	7A4
7A5	Beam Power Amplifier	6AA	7.0	0.75	—	—	—	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	R.F. Amplifier	250	−3.0	100	2.0	8.6	800000	2000	1600	—	—	7A7
7A8	Multigrid Converter	8U	7.0	0.16	—	—	—	Osc.-Mixer	250	−3.0	100	3.1	3.0	50000	Anode-grid 250 volts max.[1]				7A8
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	Fig. 45	7.0	0.16	—	—	—	Class-A$_1$ Amp.	250	250*	250	2.0	6.0	750000	4200	—	—	—	7AG7
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	—	—	—	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	—	—	—	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	R.F. Amplifier	250	−3.0	100	2.0	8.5	700000	1700	1200	—	—	7B7
7B8	Pentagrid Converter	8X	7.0	0.32	—	—	—	Osc.-Mixer	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	—	—	—	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	R.F. Amplifier	250	−3.0	100	0.5	2.0	2 meg.	1300	—	—	—	7C7
7D7	Triode-Hexode Converter	8AR	7.0	0.48	—	—	—	Osc.-Mixer	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	250	−2.0	—	—	2.3	44000	1600	70	—	—	7F7
7F8	Twin Triode	8BW	6.3	0.30	2.8	1.8	1.2	R.F. Amplifier	250	−2.5	—	—	10.0	10400	5000	—	—	—	7F8
									180	−1.0			12.0	8500	7000				
7G7/1232	Triple-Grid Amplifier	8V	7.0	0.48	9	7	.007	Class-A Amplifier	250	−2.0	100	2.0	6.0	800000	4500	—	—	—	7G7/1232

TABLE III—7-VOLT LOCK-IN-BASE TUBES—*Continued*

Type	Name	Socket Connections	Heater		Capacitance $\mu\mu$fd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
			Volts	Amps.	In	Out	Plate-Grid												
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	Triple-Grid Semi-Variable-μ	8V	7.0	0.32	8	7	.007	R.F. Amplifier	250	− 2.5	150	2.5	9.0	1000000	3500	—	—	—	7H7
7J7	Triode-Hexode Converter	8AR	7.0	0.32	—	—	—	Osc.-Mixer	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	Triple-Grid Amplifier	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	—	—	—	Class-A Amplifier	250	− 8.0	—	—	9.0	7700	2600	20	—	—	7N7
7Q7	Pentagrid Converter	8AL	7.0	0.32	—	—	—	Osc.-Mixer	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	—	—	—	Osc.-Mixer	250	− 2.0	100	2.2	1.7	2000000	Triode Plate 250 v. Max.[1]				7S7
7T7	Triple-Grid Amplifier	8V	7.0	0.32	8	7	.005	Class-A Amplifier	250	− 1.0	150	4.1	10.8	900000	4900	—	—	—	7T7
7V7	Triple-Grid Amplifier	8V	7.0	0.48	—	—	—	Class-A Amplifier	300	160*	150	3.9	9.6	300000	5800	—	—	—	7V7
7W7	Triple-Grid Variable-μ	8BJ	7.0	0.48	—	—	—	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
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.

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES

Type	Name	Base	Socket Connections	Fil. or Heater		Capacitance $\mu\mu$fd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transconductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
				Volts	Amps.	In	Out	Plate-Grid												
2C21/1642	Twin-Triode Amplifier	M.	7BH	6.3	0.6	—	—	—	Class-A Amp.	250	−16.5	—	—	8.3	7600	1375	10.4	—	—	2C21/1642
6A3	Triode Power Amplifier	M.	4D	6.3	1.0	—	—	—	Class-A Amp.	250	−45	—	—	60	800	5250	4.2	2500	3.5	6A3
									Push-Pull Amp.	300 300	−62 780*	Fixed Bias Self Bias		40 40	Power output for 2 tubes load plate-to-plate			3000 5000	15 10	
6A4	Pentode Power Amplifier	M.	5B	6.3	0.3	—	—	—	Class-A Amp.	180	−12.0	180	3.9	22	45500	2200	100	8000	1.4	6A4
6A6	Twin Triode Amplifier	M.	7B	6.3	0.8	—	—	—	Class-B Amp.	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	—	—	—	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. Push-Pull Amp.	300 400	0 −13.0	— —	6[1] 4.5[1]	45 40	241000 —	2400 —	58 —	7000 10000	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	Triple-Grid Amplifier	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	Triple-Grid Variable-μ	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	Triple-Grid Amplifier	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	Triple-Grid Variable-μ	S.	7H	6.3	0.3				R.F. Amplifier			Characteristics same as 6U7G—Table II								6E7

TABLE IV—6.3-VOLT GLASS RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amps.	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
6F7	Triode Pentode	S.	7E	6.3	0.3	—	—	—	Triode Unit Amp.	100	− 3.0	—	—	3.5	16000	500	8	—	—	6F7
									Pentode Unit Amplifier	250	− 3.0	100	1.5	6.5	850000	1100	900	—	—	
6U5/6G5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	250	Cut-off Grid Bias = −22 v.			0.24	Target Current 4 ma.				—	6U5/6G5
										100	Cut-off Grid Bias = −8 v.			0.19	Target Current 1 ma.					
6H5	Electron-Ray Tube	S.	6R	6.3	0.3	—	—	—	Indicator Tube	Same characteristics as Type 6G5—Circular Pattern										6H5
6SB7Y	Pentagrid Converter	O.	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[6]	12000[7]	12.6/12.5	6.8/6.5	—	—	—	—	—	
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	Variable-μ R.F. Amplifier	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	2-Grid Triode	M.	Fig. 33	6.3	0.3	—	—	—	Class-A Preomp.[1]	110	0	—	—	43.0	1750	3000	5.2	2000	1.5	52
									Class-B, 2 tubes[5]	180	0	—	—	3.0				10000	5.0	
56AS	Triode Amplifier	S.	5A	6.3	0.4	—	—	—	Class-A Amp.	Characteristics same as 56										56AS
57AS	Pentode	S.	6F	6.3	0.4	—	—	—	R.F. Amplifier	Characteristics same as 57										57AS
58AS	Triple-Grid Variable-μ	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	Triple-Grid Detector	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	Triple-Grid Variable-μ	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	—	—	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	Triple-Grid Power Amp.	S.	6F	6.3	0.4	—	—	—	Triode Amp.[2]	250	−31.0	—	—	32.0	2600	1800	4.7	5500	0.9	89
									Pentode Amp.[8]	250	−25.0	250	5.5	32.0	70000	1800	125	6750	3.4	
1221	Pentode R.F. Amplifier	S.	6F	6.3	0.3	—	—	—	Class-A Amp.	Special non-microphonic. Characteristics same as 6C6										1221
1603[3]	Triple-Grid Amplifier	M.	6F	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6C6										1603
7700[3]	Triple-Grid Amplifier	S.	6F	6.3	0.3	—	—	—	Class-A Amp.	Characteristics same as 6C6										7700

* Cathode bias resistor-ohms.
[1] Current to input plate (P_1).
[2] Grids Nos. 2 and 3 connected to plate.
[3] Low noise, non-microphonic, 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.

TABLE V—2.5-VOLT RECEIVING TUBES

Type	Name	Base	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amps.	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
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	—	—	—	Osc.-Mixer	Characteristics same as Type 6A7, Table IV										2A7

TABLE V—2.5-VOLT RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amps.	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
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	Variable-μ Amplifier	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	—	—	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	Triple-Grid Amplifier	S.	6F	2.5	1.0	—	—	—	R.F. Amplifier	250	− 3.0	100	0.5	2.0	1500000	1225	1500	—	—	57
58	Triple-Grid Variable-μ	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	Triple-Grid Power Amplifier	M.	7A	2.5	2.0	—	—	—	Class-A Triode[4]	250	−28.0	—	—	26.0	2300	2600	6.0	5000	1.25	59
									Class-A Pentode[5]	250	−18.0	250	9.0	35.0	40000	2500	100	6000	3.0	
RK15	Triode Power Amplifier	M.	4D[1]	2.5	1.75	—	—	—	Characteristics same as Type 46 with Class-B connections											RK15
RK16	Triode Power Amplifier	M.	5A	2.5	2.0	—	—	—	Characteristics same as Type 59 with Class-A triode connections											RK16
RK17	Pentode Power Amplifier	M.	5F	2.5	2.0	—	—	—	Characteristics same as Type 2A5											RK17

[1] Grid connection to cap; no connection to No. 3 pin. [2] Grid No. 2 tied to plate. [3] Grids Nos. 1 and 2 tied together. [4] Grids Nos. 2 and 3 connected to plate. [5] Grid No. 2, screen; grid No. 3, suppressor.

TABLE VI—2.0-VOLT BATTERY RECEIVING TUBES

Type	Name	Base	Socket Connec-tions	Fil. or Heater Volts	Fil. or Heater Amps.	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
1A4P	Variable-μ Pentode	S.	4M	2.0	0.06	5	11	.007	R.F. Amplifier	180	− 3.0	67.5	0.8	2.3	1000000	750	750	—	—	1A4P
1A4T	Variable-μ Tetrode	S.	4K	2.0	0.06	5	11	.007	R.F. Amplifier	180	− 3.0	67.5	0.7	2.3	960000	750	720	—	—	1A4T
1A6	Pentagrid Converter	S.	6L	2.0	0.06	—	—	—	Converter	180	− 3.0	67.5	2.4	1.3	500000	Anode grid (No. 2) 180 max. volts				1A6
1B4P/951	Pentode R.F. Amplifier	S.	4M	2.0	0.06	5	11	.007	R.F. Amplifier	180 90	− 3.0 − 3.0	67.5 67.5	0.6 0.7	1.7 1.6	1500000 1000000	650 600	1000 550	—	—	1B4P/951
1B5/25S	Duplex-Diode Triode	S.	6M	2.0	0.06	1.6	1.9	3.6	Triode Class-A	135	− 3.0	—	—	0.8	35000	575	20	—	—	1B5/25S
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	0.6	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			

TABLE VI—2.0-VOLT BATTERY RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Filament Volts	Filament Amps.	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
15	R.F. 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	Tetrode R.F. Amplifier	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	R.F. 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	100	13500	0.45	950
RK24	Triode Amplifier	M.	4D	2.0	0.12	—	—	—	Class-A Amp.	180	−13.5	—	—	8.0	5000	1600	8.0	12000	0.25	RK24
1229	Tetrode R.F. Amplifier	M.	4K	2.0	0.06	—	—	—	Class-A Amp.	Special type 32 for low grid current applications										1229

[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	Fil. or Heater Volts	Fil. or Heater Amps.	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	Pentagrid Converter	7Z	2.0	0.06	—	—	—	Converter	Characteristics same as Type 1C6—Table VI										1C7G
1D5GP	Variable-μ R.F. Pentode	5Y	2.0	0.06	5	11	.007	R.F. Amplifier	Characteristics same as Type 1A4P—Table VI										1D5GP
1D5GT	Variable-μ R.F. 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	—	—	—	Converter	Characteristics same as Type 1A6—Table VI										1D7G
1E5GP	R.F. Amplifier Pentode	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
1F7GV[2]	Duplex-Diode Pentode	7AD	2.0	0.06	3.8	9.5	0.01	Detector-Amplifier	Characteristics same as Type 1F6—Table VI										1F7GV
1G5G	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	135	−13.5	135	2.5	8.7	160000	1550	250	9000	0.55	1G5G
1H4G	Triode Amplifier	5S	2.0	0.06	—	—	—	Detector-Amplifier	Characteristics same as Type 30—Table VI										1H4G
1H6G	Duplex-Diode Triode	7AA	2.0	0.06	1.6	1.9	3.6	Detector-Amplifier	Characteristics same as Type 1B5—Table VI										1H6G
1J5G	Pentode Power Amplifier	6X	2.0	0.12	—	—	—	Class-A Amplifier	135	−16.5	135	2.0	7.0	—	950	100	13500	0.45	1J5G
1J6G	Twin Triode	7AB	2.0	0.24	—	—	—	Class-B Amplifier	Characteristics same as Type 19—Table VI										1J6G
4A6G	Twin Triode	8L	2.0	0.12	—	—	—	Class-A, 1 section	90	−1.5	—	—	1.1	26600	750	20			4A6G
			4.0	0.06	—	—	—	Class-B, 2 sections	90	−1.5	—	—	1.1[3]				8000	1.0	

[1] Total current for both sections; no signal. [2] Also type G or GH. [3] Max. signal plate current = 10.8 Ma.

TABLE VIII—1.5-VOLT FILAMENT DRY-CELL TUBES

See also Table X for Special 1.4-volt Tubes

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amps.	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
1A5G	Pentode Power Amplifier	O.	6X	1.4	0.05	—	—	—	Class-A Amp.	90	−4.5	90	0.8	4.0	300000	850	240	25000	115	1A5G
1A7G	Pentagrid Converter	O.	7Z	1.4	0.05	—	—	—	Osc.-Mixer	90	0	45	0.6	0.55	600000	Anode-grid volts 90				1A7G

TABLE VIII—1.5-VOLT FILAMENT DRY-CELL TUBES—*Continued*

Type	Name	Base	Socket Connec-tions	Filament Volts	Filament Amps.	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 M-watts	Type
1AB5	Pentode R.F. Amplifier	O.	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
1B7G	Pentagrid Converter	O.	7Z	1.4	0.1	—	—	—	Converter	90	0	45	1.3	1.5	350000	Grid No. 1 resistor 200,000 ohms				1B7G
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
1C5G	Pentode Power Amplifier	O.	6X	1.4	0.1	—	—	—	Class-A₁ Amp.	90	−7.5	90	1.6	7.5	115000	1550	165	8000	240	1C5G
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
1G4G	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	—	—	1G4G
1G6G	Twin Triode	O.	7AB	1.4	0.1	—	—	—	Class-A Amp.	90	0	—	—	1.0	45000	675	30	—	—	1G6G
									Class-B Amp.	90	0	—	—	1/7	34 volts input per grid			12000	675	
1H5G	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	—	—	1H5G
1LA4	Pentode Power Amplifier	L.	5AD	1.4	0.05	—	—	—	Class-A Amp.	90	Characteristics same as 1A5G									1LA4
1LA6	Pentagrid Converter	L.	7AK	1.4	0.05	—	—	—	Converter	90	0	45	0.6	0.55		Anode Grid Volts 90				1LA6
1LB4	Pentode Power Amplifier	L.	5AD	1.4	0.05	—	—	—	Class-A Amp.	90	−9	90	1.0	5.0	200000	925	—	12000	200	1LB4
1LB6	Heptode Converter	L.	8AX	1.4	0.05	—	—	—	Converter	90	0	67.5	2.2	0.4	Grid No. 4—67.5 v., No. 5—0 v.					1LB6
1LC5	Triple-Grid Variable-μ	L.	7AO	1.4	0.05	3.2	7	.007	R.F. Amplifier	90	0	45	0.2	1.15	1500000	775	—	—	—	1LC5
1LC6	Pentagrid Converter	L.	7AK	1.4	0.05	—	—	—	Converter	90	0	35[4]	0.7	0.75		Anode Grid Volts 45				1LC6
1LD5	Diode Pentode	L.	6AX	1.4	0.05	3.2	6	0.18	Class-A Amp.	90	0	45	0.1	0.6	950000	600	—	—	—	1LD5
1LE3	Triode Amplifier	L.	4AA	1.4	0.05	1.7	3	1.70	Class-A Amp.	90 90	0 −3	—	—	4.5 1.3	11200 19000	1300 760	14.5	—	—	1LE3
1LG5	Pentode R.F. Amp.	L.	Fig. 42	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	Triple-Grid Amplifier	L.	7AO	1.4	0.05	3.4	8	.007	Class-A Amp.	90	0	90	0.3	1.2	1500000	750	—	—	—	1LN5
1N5G	Pentode R.F. Amplifier	O.	5Y	1.4	0.05	3	10	.007	Class-A Amp.	90	0	90	0.3	1.2	1500000	750	1160	—	—	1N5G
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
1P5G	Triple-Grid Pentode	O.	5Y	1.4	0.05	3	10	.007	R.F. Amplifier	90	0	90	0.7	2.3	800000	800	640	—	—	1P5G
1Q5G	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	1Q5G
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	R.F. Pentode	O.	6CA	1.4	0.05	5.2	8.6	0.01	R.F. Amplifier	90	0	67.5	0.68	2.45	800000	970	—	—	—	1SA6GT
1SB6GT	Diode Pentode	O.	6CB	1.4	0.05	3.2	3	0.25	Class-A Amp.	90	0	67.5	0.38	1.45	700000	665	—	—	—	1SB6GT
									R.C. Amplifier	90	0	90	Screen resistor 5 meg., grid 10 meg.					1 meg.	110[6]	
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	1.4	0.22	—	—	—	Class-A Amp.	90	0	—	—	5.2	11350	1850	21	—	—	3B7/1291
1293	U.h.f. Triode	L.	Fig. 2	1.4	0.11	—	—	—	Class-A Amp.	90	0	—	—	4.7	10750	1300	14	—	—	1293
3D6/1299	U.h.f. Tetrode	L.	6BB	1.4	0.22	7.5	6.5	0.30	Class-A Amp.	135	−6	90	0.7	5.7	—	2200	—	13000	0.5	3D6/1299
CK501	Pentode Voltage Amplifier	—[1]	——[7]	1.25	0.033	—	—	—	Class-A Amp.	30 45	0 −1.25	30 45	0.06 0.055	0.3 0.28	1000000 1500000	325 300	—	—	—	CK501
CK502	Pentode Output Amplifier	—[1]	——[7]	1.25	0.033	—	—	—	Class-A Amp.	30	0	30	0.13	0.55	500000	400	—	60000	3	CK502
CK503	Pentode Output Amplifier	—[1]	——[7]	1.25	0.033	—	—	—	Class-A Amp.	30	0	30	0.33	1.5	150000	600	—	20000	6[2]	CK503
CK504	Pentode Output Amplifier	—[1]	——[7]	1.25	0.033	—	—	—	Class-A Amp.	30	−1.25	30	0.09	0.4	500000	350	—	60000	3[2]	CK504
CK505	Pentode Voltage Amplifier	—[1]	——[7]	0.625[5]	0.03	—	—	—	Class-A Amp.	30 45	0 −1.25	30 45	0.07 0.08	0.17 0.2	1100000 2000000	140 150	—	—	—	CK505
CK506	Pentode Output Amplifier	—[1]	——[7]	1.25	0.05	—	—	—	Class-A₁ Amp.	45	−4.5	45	0.4	1.25	120000	500	—	30000	25	CK506
CK507	Pentode Output Amplifier	—[1]	——[7]	1.25	0.05	—	—	—	Class-A₁ Amp.	45	−2.5	45	0.21	0.6	360000	500	—	50000	12	CK507

TABLE VIII—1.5-VOLT FILAMENT DRY-CELL TUBES—*Continued*

Type	Name	Base	Socket Connec-tions	Filament Volts	Filament Amps.	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 M-watts	Type
CK509	Triode Voltage Amplifier	—[1]	——[7]	0.625[5]	0.03	—	—	—	Class-A Amp.	45	0	—	—	0.15	150000	160	16	1000000	—	CK509
CK510	Dual Space-Charge Tetrode	—[1]	——[7]	0.625[5]	0.05	—	—	—	Class-A Amp.	45	0	0.2	200 μa	60 μa	500000	65	32.5	—	—	CK510
CK515BX	Triode Voltage Amplifier	—[1]	——[7]	0.625[5]	0.03	—	—	—	Class-A Amp.	45	0	—	—	0.15	—	160	24	1000000	—	CK515BX
HY113 HY123	Triode Amplifier	—[1]	5K[3]	1.4	0.07	—	—	—	Class-A Amp.	45	−4.5	—	—	0.4	25000	250	6.3	40000	6.5	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	11.5 90	HY125 HY155
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

[1] Special miniature peanut base.
[2] With 5-megohm grid resistor and 0.02-μfd. grid coupling condenser.
[3] No screen connection.
[4] Through series resistor. Screen voltage must be at least 10 volts lower than oscillator anode.
[5] Two tubes connected in series for 1.4-volt operation.
[6] Voltage gain.
[7] Tinned wire leads extend from bottom of tube. Connections are labeled on tube.

TABLE IX—HIGH-VOLTAGE HEATER TUBES

Type	Name	Base	Socket Connec-tions	Heater Volts	Heater Amps.	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
12A5	Pentode Power Amplifier	M.	7F	12.6 6.3	0.3 0.6	—	—	—	Class-A_1 Amp.	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	Pentagrid Converter	O.	8A	12.6	0.15	—	—	—	Converter	Characteristics same as 6A8—Table										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	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 6F5—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	Pentode Voltage Amplifier	O.	7R	12.6	0.15	—	—	—	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-C Amp.	Characteristics same as 6Q7—Table I										12Q7GT
12SA7	Pentagrid Converter	O.	8R	12.6	0.15	—	—	—	Converter	Characteristics same as 6SA7—Table I										12SA7
12SC7	Twin Triode	O.	8S	12.6	0.15	—	—	—	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	Triple-Grid Variable-μ	O.	8BK	12.6	0.15	8.5	7.0	.003	Class-A Amp.	Characteristics same as 6SG7—Table I										12SG7

TABLE IX—HIGH-VOLTAGE HEATER TUBES—*Continued*

Type	Name	Base	Socket Connections	Heater Volts	Heater Amps.	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
12SH7	H-F Amplifier Pentode	O.	8BK	12.6	0.15	8.5	7.0	.003	H-F Amplifier	Characteristics same as 6SH7—Table I										12SH7
12SJ7	Pentode Voltage Amplifier	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
12SY7	Heptode Converter	O.	8R	12.6	0.15	Osc.-Grid leak 20000 ohms			Converter	250	−2	100	8.5	3.5	1000000	450	—	—	—	12SY7
14A4	Triode Amplifier	L.	5AC	14	0.16	3.4	3.0	4.00	Class-A Amp.	Characteristics same as 7A4—Table III										14A4
14A5	Beam Power Amplifier	L.	6AA	14	0.16	—	—	—	Class-A Amp.	250	−12.5	250	3.5/5.5	30/32	70000	3000	—	7500	2.8	14A5
14A7/12B7	Triple-Grid Variable-μ	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	Triple-Grid Amplifier	L.	8V	14	0.16	6.0	6.5	.007	Class-A Amp.	250	−3.0	100	0.7	2.2	1000000	1575	—	—	—	14C7
14E6	Duplex-Diode Triode	L.	8W	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7E6—Table III										14E6
14E7	Duplex-Diode Pentode	L.	8AE	14	0.16	4.6	5.3	.005	Class-A Amp.	Characteristics same as 7E7—Table III										14E7
14F7	Twin Triode	L.	8AC	14	0.16	—	—	—	Class-A Amp.	Characteristics same as 7F7—Table III										14F7
14H7	Triple-Grid Semi-Variable-μ	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.	8AR	14	0.16	Ipt=5 Ma.			Converter	Characteristics same as 7J7—Table III										14J7
14N7	Twin Triode	L.	8AC	14	0.32	—	—	—	Class-A Amp.	Characteristics same as 7N7—Table III										14N7
14Q7	Heptode Pentagrid Converter	L.	8AL	14	0.16	—	—	—	Converter	Characteristics same as 7Q7—Table III										14Q7
14R7	Duplex-Diode Pentode	L.	8AE	14	0.16	5.6	5.3	.004	Class-A Amp.	Characteristics same as 7R7—Table III										14R7
14S7	Triode Heptode	L.	8BL	14	0.16	Ipt=5 Ma.			Converter	250	−2.0	100	3	1.8	1250000	525	—	—	—	14S7
14V7	H.f. Pentode	L.	8V	14	0.24	—	—	—	Class-A Amp.	300	−2.0	150	3.9	9.6	300000	5800	—	—	—	14V7
14W7	Pentode	L.	8BJ	14	0.24	Rk=160 ohms			Class-A Amp.	300	−2.2	150	3.9	10	300000	5800	—	—	—	14W7
18	Pentode	M.	6B	14	0.30	—	—	—	Class-A Amp.	Characteristics same as 6F6G										18
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	250 150	−3.0 −3.0	100 Triode	2.8	1.3 3.5	— —	275 1900	— 32	— —	— —	21A7
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
25A7G	Rectifier-Amplifier	O.	8F	25	0.3	—	—	—	Class-A Amp.	100	−15.0	100	4	20.5	50000	1800	90	4500	0.77	25A7G
25AC5G	Triode Power Amplifier	O.	6Q	25	0.3	—	—	—	Class-A Amp.	110 165	+15.0 Used in dynamic-coupled circuit with 6AF5G driver	—	—	45	—	3800	58	2000 3500	2.0 3.3	25AC5G
25B5	Direct-Coupled Triodes	S.	6D	25	0.3	—	—	—	Class-A Amp.	110	0	110	7	45	11400	2200	25	2000	2.0	25B5
25B6G	Pentode Power Amplifier	O.	7S	25	0.3	—	—	—	Class-A Amp.	95	−15.0	95	4	45	—	4000	—	2000	1.75	25B6G
25B8GT	Triode Pentode	O.	8T	25	0.15	—	—	—	Class-A Amp.	Characteristics same as 12B8GT										25B8GT
25C6G	Beam Power Amplifier	O.	7AC	25	0.3	—	—	—	Class-A Amp.	135	−13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	25C6G
25D8GT	Diode Triode Pentode	O.	8AF	25	0.15	—	—	—	Triode Amp. Pentode Amp.	100 100	−1.0 −3.0	— 100	— 2.7	0.5 8.5	91000 200000	1100 1900	100 —	— —	— —	25D8GT
25L6	Beam Power Amplifier	O.	7AC	25	0.3	16	13.5	0.30	Class-A Amp.	110	−8.0	110	3.5/10.5	45/48	10000	8000	80	2000	2.2	25L6
25N6G	Direct-Coupled Triodes	O.	7W	25	0.3	—	—	—	Class-A Amp.	110	0	110	7	45	11400	2200	25	2000	2.0	25N6G

TABLE IX—HIGH-VOLTAGE HEATER TUBES—*Continued*

Type	Name	Base	Socket Connec-tions	Heater		Capacitance $\mu\mu$fd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
				Volts	Amps.	In	Out	Plate-Grid												
26A7GT	Twin Beam-Power Audio Amplifier	O.	8BU	26.5	0.6	Each Unit Push-Pull[3]			Class-A Amp.	26.5	− 4.5	26.5	2/5.5	20/20.5	2500	5500	—	1500	0.2	26A7GT
									Class-AB Amp.	26.5	− 7.0	26.5	2/8.5	19/30	—	—	—	2500[4]	0.5	
32L7GT	Diode-Beam Tetrode	O.	8Z	32.5	0.3	—	—	—	Class-A Amp.	110	− 7.5	110	3	40	15000	6000	—	2500	1.5	32L7GT
35A5	Beam Power Amplifier	L.	6AA	35	0.15	—	—	—	Class-A$_1$ Amp.	110	− 7.5	110	3/7	40/41	14000	5800	—	2500	1.5	35A5
35L6G	Beam Power Amplifier	O.	7AC	35	0.15	13	9.5	0.80	Class-A$_1$ Amp.	110	− 7.5	110	3/7	40/41	13800	5800	—	2500	1.5	35L6G
43	Pentode Power Amplifier	M.	6B	25	0.3	8.5	12.5	0.20	Class-A Amp.	95	−15.0	95	4.0	20.0	45000	2000	90	4500	0.90	43
48	Tetrode Power Amplifier	M.	6A	30	0.4	—	—	—	Class-A Amp.	96	−19.0	96	9.0	52.0	—	3800	—	1500	2.0	48
50A5	Beam Power Amplifier	L.	6AA	50	0.15	—	—	—	Class-A$_1$ Amp.	110	− 7.5	110	4/11	49/50	10000	8200	—	2000	2.2	50A5
50C6GT	Beam Power Amplifier	O.	7AC	50	0.15	—	—	—	Class-A$_1$ Amp.	135	−13.5	135	3.5/11.5	58/60	9300	7000	—	2000	3.6	50C6GT
50L6GT	Beam Power Amplifier	O.	7AC	50	0.15	—	—	—	Class-A Amp.	110	− 7.5	110	4/11	49/50	—	8200	82	2000	2.2	50L6GT
70A7GT	Diode-Beam Tetrode	O.	8AB[1]	70	0.15	—	—	—	Class-A Amp.	110	− 7.5	110	3.0	40	—	5800	80	2500	1.5	70A7GT
70L7GT	Diode-Beam Tetrode	O.	8AA	70	0.15	—	—	—	Class-A$_1$ Amp.	110	− 7.5	110	3/6	40/43	15000	7500	—	2000	1.8	70L7GT
117L7GT/117M7GT	Rectifier-Amplifier	O.	8AO	117	0.09	—	—	—	Class-A Amp.	105	− 5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117L7GT/117M7GT
117N7GT	Rectifier-Amplifier	O.	8AV	117	0.09	—	—	—	Class-A Amp.	100	− 6.0	100	5.0	51	16000	7000	—	3000	1.2	117N7GT
117P7GT	Rectifier-Amplifier	O.	8AV	117	0.09	—	—	—	Class-A Amp.	105	− 5.2	105	4/5.5	43	17000	5300	—	4000	0.85	117P7GT
1284	U.h.f. Pentode	O.	Fig. 4	12.6	0.15				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	Twin Triode	L.	8AC	12.6	0.15	—	—	—	Class-A Amp.	250	−10	—	—	9.0	—	2100	16	—	—	XXD
28D7	Double Beam Power Amplifier	L.	8BS	28.0	0.4	—	—	—	Class-A$_2$ Amp.	28	390*	28[2]	0.7[2]	9.0[2]	—	—	—	4000[2]	0.08[2]	28D7
											180*	28[3]	1.2[3]	18.5[3]	—	—	—	6000[4]	0.175[3]	

* Cathode resistor—ohms.

[1] 6.3-volt pilot lamp must be connected between pins 6 and 7.
[2] Per section (except heater)—resistance coupled.
[3] P.P. operation—values for both sections, resistance coupled.
[4] Plate to plate.
[5] Each unit.

TABLE X—SPECIAL RECEIVING TUBES

Type	Name	Base	Socket Connec-tions	Fil. or Heater		Capacitance $\mu\mu$fd.			Use	Plate Supply Volts	Grid Bias	Screen Volts	Screen Current Ma.	Plate Current Ma.	Plate Resistance Ohms	Transcon-ductance Micromhos	Amp. Factor	Load Resistance Ohms	Power Output Watts	Type
				Volts	Amps.	In	Out	Plate-Grid												
00-A	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	Triode Detector Amplifier	M.	4D	5.0	0.25	—	—	—	Class-A Amp.	135	− 9.0	—	—	3.0	10000	800	8.0	—	—	01-A
2E32	Sub-miniature Pentode	[1]	—	1.25	0.05				Class-A Amp.	22.5	0	22.5	0.3	0.4	350000	500	—	—	—	2E32
2E36	Sub-miniature Pentode	[1]	—	1.25	0.03				Class-A$_1$ Amp.	22.5	0	22.5	0.07	0.27	220000	385	—	150000	0.0012	2E36
										45	−1.25	45	0.11	0.45	250000	500	—	100000	0.006	
2E42	Sub-miniature Diode Pent.	[1]	—	1.25	0.03				Detector Amp.	22.5	0	22.5	0.12	0.35	250000	375	—	1 meg.	—	2E42
2G22	Sub-miniature Converter	[1]	—	1.25	0.05				Converter	22.5	0	22.5	0.3	0.2	500000	60	—	—	—	2G22
3A8GT	Diode Triode Pentode	O.	8AS	1.4 2.8	0.1 0.05	—	—	—	Class-A Triode	90	0	—	—	0.15	240000	275	65	—	—	3A8GT
									Class-A Pentode	90	0	90	0.3	1.2	600000	750	—	—	—	
3B5GT	Beam Power Amplifiers	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

TABLE X—SPECIAL RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amps.	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
3C5GT	Power Output Pentode	O.	7AQ	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	− 9.0	90	1.4	6.0	—	1550 1450	—	8000 10000	0.24 0.26	3C5GT
3C6	Twin Triode	L.	7BW	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	0	—	—	4.5	11200	1300	14.5	—	—	3C6
3LE4	Power Amplifier Pentode	L.	6BA	2.8	0.05	—	—	—	Class-A Amp.	90	− 9.0	90	1.8	9.0	110000	1600	—	6000	0.30	3LE4
3LF4	Power Amplifier Tetrode	L.	6BB	1.4 2.8	0.1 0.05	—	—	—	Class-A Amp.	90	− 4.5	90	1.3 1.0	9.5 8.0	75000 80000	2200 2000	—	8000 7000	0.27 0.23	3LF4
3Q5GT	Beam Power Amplifier	O.	7AP	1.4 2.8	0.1 0.05	Parallel Filaments Series Filaments			Class-A Amp.	90	− 4.5	90	1.3 1.0	9.5 7.5	—	2100 1800	—	8000	0.27 0.25	3Q5GT
4A6G	Twin Triode Amplifier	O.	8L	4	0.06	Triodes Parallel			Class-A Amp.	90	− 1.5	—	—	2.2	13300	1500	20	—	—	4A6G
				2	0.12	Both Sections			Class-B Amp.	90	0	—	—	4.6	—	—	—	8000	1.0	
6F4	Acorn Triode	A.	7BR	6.3	0.225	2.0	0.6	1.90	Class-A Amp.	80	150*	—	—	13.0	2900	5800	17	—	—	6F4
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	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	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	Tetrode R.F. Amplifier	M.	4K	3.3	0.132	3.5	10	0.02	Class-A Amp.	135	− 1.5	67.5	1.3	3.7	325000	500	160	—	—	22
26	Triode Amplifier	M.	4D	1.5	1.05	2.8	2.5	8.10	Class-A Amp.	180	−14.5	—	—	6.2	7300	1150	8.3	—	—	26
40	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[10]	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	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	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	Triode	S.	5A	3.0	1.3	—	—	—	Class-A Amp.	180	− 9.0	—	—	6.0	9300	1350	12.5	—	—	485
864	Triode Amplifier	S.	4D	1.1	0.25	—	—	—	Class-A Amp.	90	− 4.5	—	—	2.9	13500	610	8.2	—	—	864
954	Pentode Detector, Amplifier	A.	5BB	6.3	0.15	3.4	3.0	0.007	Class-A Amp.	250	− 3.0	100	0.7	2.0	1.5 meg.	1400	2000	—	—	954
									Bias Detector	250	− 6.0	100	—	Plate current to be adjusted to 0.1 ma. with no signal					—	
955	Triode Detector, Amplifier, Oscillator	A.	5BC	6.3	0.15	1.0	0.6	1.40	Class-A Amp.	250	− 7.0	—	—	6.3	11400	2200	25	—	—	955
										90	− 2.5	—	—	2.5	14700	1700	25	—	—	
956	Triple-Grid Variable-μ R.F. Amplifier	A.	5BB	6.3	0.15	3.4	3.0	.007	Class-A Amp.	250	− 3.0	100	2.7	6.7	700000	1800	1440	—	—	956
									Mixer	250	−10.0	100	—	—	Oscillator peak volts—7 min.					
957	Triode Detector, Amplifier, Oscillator	A.	5BD	1.25	0.05	0.3	0.7	1.20	Class-A Amp.	135	− 5.0	—	—	2.0	20800	650	13.5	—	—	957
958 958-A	Triode A.F. Amplifier, Oscillator	A.	5BD	1.25	0.1	0.6	0.8	2.60	Class-A Amp.	135	− 7.5	—	—	3.0	10000	1200	12	—	—	958 958-A
959	Pentode Detector, Amplifier	A.	5BE	1.25	0.05	1.8	2.5	.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	U.h.f. Pentode	L.	Fig. 5	6.3	0.15	3.5	4.0	0.06	Class-A Amp.	250	− 2	100	0.6	1.75	800000	1200	—	—	—	7AB7/ 1204
1276	Triode Power Amplifier	M.	4D	4.5	1.14	—	—	—	Class-A Amp.	Characteristics similar to 6A3										1276
1609	Pentode Amplifier	S.	5B	1.1	0.25	—	—	—	Class-A Amp.	135	− 1.5	67.5	0.65	2.5	400000	725	300	—	—	1609
9004	U.h.f. Diode	A.	4BJ	6.3	0.15	—	—	—	Detector	Max. a.c. voltage—117. Max. d.c. output current—5 ma.										9004
9005	U.h.f. Diode	A.	5BG	3.6	0.165	—	—	—	Detector	Max. a.c. voltage—117. Max. d.c. output current—1 ma.										9005

TABLE X—SPECIAL RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amps.	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
EF-50	High Frequency Pentode Amplifier	L.	Fig. 14	6.3	0.3	—	—	—	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
M54	Tetrode Power Amplifier	[1]	—	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]	—	0.625	0.02	—	—	—	Class-A Amp.	30	0	—	—	0.03	200000	110	25	—	—	M64
M74	Tetrode Voltage Amplifier	[1]	—	0.625	0.02	—	—	—	Class-A Amp.	30	0	7.0	0.01	0.02	500000	125	70	—	—	M74
XXB	Twin Triode Frequency Converter	L.	Fig. 9	2.8/ 1.4 3.2[6]/ 1.6	0.05/ 0.10 — —	—	—	—	Converter[5]	90[2]	0	—	—	4.5[7] 4.5[9]	11200[7] 11200[9]	1300[7] 1300[9]	14.5[2]	—	—	XXB
											−3	—	—	1.4[7] 1.4[9]	1900[7] 1900[9]	760[7] 760[9]	14.5[2]	—	—	
XXFM	Twin-Diode Triode	L.	Fig. 10	6.3	0.3	—	—	—	Special Detector Amplifier	250[8]	−1	—	—	1.9	6700	1500	100	—	—	XXFM
										100[8]	0	—	—	1.2	85000	1000	85	—	—	
										100[3]	—	—	—	4[4]	—	—	—	—	—	

* Cathode resistor ohms.
[1] No base; tinned wire leads.
[2] Both Sections.
[3] Diode plates (A.C. max. volts per plate).
[4] Max. D.C. output.
[5] Section No. 2 recommended for h.f.o.
[6] Dry battery operation.
[7] Section No. 1.
[8] Amplifier plate.
[9] Section No. 2.
[10] Same as X99. Type V99 is same, but socket connections are 4E.

TABLE XI—MINIATURE RECEIVING TUBES

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amps.	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
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.										1A3
1L4	R.F. Pentode Amplifier	B.	6AR	1.4	0.05	3.6	7.5	.008	Class-A Amp.	90	0	90	2.0	4.5	350000	1025	—	—	—	1L4
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			1R5
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	1S4
1S5	Diode Pentode	B.	6AU	1.4	0.05	—	—	—	Class-A Amp.	67.5	0	67.5	0.4	1.6	600000	625	—	—	—	1S5
									R-Coupled Amp.	90	0	90	Screen resistor 3 meg., grid 10 meg.					1 meg.	0.050	
1T4	Triple-Grid Variable-μ	B.	6AR	1.4	0.05	3.6	7.5	0.01	Class-A Amp.	90	0	45	0.65	2.0	800000	750	—	—	—	1T4
1U4	Pentode R.F. Amplifier	B.	6AR	1.4	0.05	3.6	7.5	.008	Class-A Amp.	90	0	90	0.45	1.6	1500000	900	—	—	—	1U4
1U5	Diode Pentode	B.	6BW	1.4	0.05	—	—	—	Class-A Amp.	67.5	0	67.5	0.4	1.6	600000	625	—	—	—	1U5
3A4	Power Amplifier Pentode	B.	7BB	1.4 2.8	0.2 0.1	4.8	4.2	0.20	Class-A Amp.	135 150	−7.5 −8.4	90 90	2.6 2.2	14.8 13.3	90000 100000	1900	—	8000	0.6 0.7	3A4
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	—	—	3A5
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	3Q4
				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	3S4
				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	3V4
				2.8	0.05	Series Filaments			Class-A Amp.	90	−4.5	90	1.7	7.7	120000	2000	—	10000	0.24	

TABLE XI—MINIATURE RECEIVING TUBES—*Continued*

Type	Name	Base	Socket Connections	Fil. or Heater Volts	Fil. or Heater Amps.	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
6AG5	Pentode R.F. Amplifier	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	— —	— —	—	6AG5
6AJ5	U.h.f. Pentode	B.	7PM	6.3	0.175	—	—	—	R.F. Amplifier	28	200*	28	1.2	3.0	90000	2750	250	—	—	6AJ5
									Class-AB Amp.	180	−7.5	75	—	—	—	—	—	28000	1.0	
6AK5	H.F. Pentode	B.	7BD	6.3	0.175	—	—	—	R.F. Amplifier	180	200*	120	2.4	7.7	690000	5100	3500	—	—	6AK5
										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	6AK6
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]										6AL5
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 μμf load; peak current per plate = 10 ma.; peak inverse voltage = 210.										6AN6
6AQ5	Beam Power Tetrode	B.	Fig. 38	6.3	0.45	—	—	—	Class-A Amp.	180 250	−8.5 −12.5	180 250	3.0 4.5	29 45	58000 52000	3700 4100	— —	5500 5000	2.0 4.5	6AQ5
6AQ6	Duodiode Hi-mu Triode	B.	7BT	6.3	0.15	1.7	1.5	1.80	Class-A Triode	250 100	−3.0 −1.0	— —	— —	1.0 0.8	58000 61000	1200 1150	70 70	— —	— —	6AQ6
6AS6	Sharp Cut-off Pentode	B.	7CN	6.3	0.175	4.0	3.0	0.02	Class-A Amp.	120	−2	120	3.5	5.5	—	3500	—	—	—	6AS6
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	—	—	6AT6
6AU6	Pentode R.F. Amp.	B.	7BK	6.3	0.3	5.5	5.0	.0035	Class-A Amp.	250	−1	150	4.3	10.8	2000000	5200	—	—	—	6AU6
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	—	—	—	6BA6
6BD6	Pentode R.F. Amplifier	B.	Fig. 36	6.3	0.3	—	—	—	Class-A Amp.	100	−1	100	5	13	120000	2350	—	—	—	6BD6
										250	−3	100	3.5	9	700000	2000	—	—	—	
6BE6	Pentagrid Converter	B.	7CH	6.3	0.3	Osc. Grid 20000 Ω			Converter	250	−1.5	100	7.1	3.0	1000000	475	—	—	—	6BE6
6C4	Triode Amplifier	B.	6BG	6.3	0.15	1.8	1.3	1.60	Class-A Amp.	250	−8.5	—	—	10.5	7700	2200	17	—	—	6C4
6J4	U.h.f. Grounded-Grid R.F. Amplifier	B.	7BQ	6.3	0.4	—	—	—	Grounded-Grid Class-A Amp.	150 100	200* 100*	— —	— —	15.0 10.0	4500 5000	12000 11000	55 55	— —	— —	6J4
6J6	Twin Triode	B.	7BF	6.3	0.45	—	—	—	Class-A Amp. Mixer, Oscillator	100	50*	—	—	8.5	7100	5300	38	—	—	6J6
6N4	U.h.f. Triode Amplifier	B.	Fig. 40	6.3	0.2	3.0	1.6	1.10	Class-A R.F. Amplifier	180	−3.5	—	—	12.0	—	6000	32	—	—	6N4
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	—	—	12AT6
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	—	—	—	12BA6
12BD6	Pentode Amplifier	B.	Fig. 36	12.6	0.15	4.3	5.0	.004	Class-A Amp.	250	−3	100	3.5	9.0	700000	2000	—	—	—	12BD6
12BE6	Pentagrid Converter	B.	7CH	12.6	0.15	Osc. Grid 20000 Ω			Converter	250	−1.5	100	7.1	3.0	1000000	475	—	—	—	12BE6
12BF6	Duodiode Triode	B.	Fig. 37	12.6	0.15	1.8	1.1	2.00	Class-A Amp.	250	−9	—	—	9.5	8500	1900	16	—	—	12BF6
50B5	Beam Power Amplifier	B.	7BZ	50.0	0.15	13	6.5	0.50	Class-A Amp.	110	−7.5	110	4.0	49.0	14000	7500	—	3000	1.9	50B5
9001	Triple-Grid Detector, Amplifier	B.	7PM	6.3	0.15	3.6	3.0	0.01	Class-A Amp.	250	−3.0	100	0.7	2.0	Over 1 meg.	1400	—	—	—	9001
									Mixer	250	−5.0	100	Osc. peak voltage 4 volts			550	—	—	—	
9002	Triode Detector, Amplifier, Oscillator	B.	7TM	6.3	0.15	1.2	1.1	1.40	Class-A Amp.	250	−7.0	—	—	6.3	11400	2200	25	—	—	9002
										90	−2.5	—	—	2.5	14700	1700	25	—	—	
9003	Triple-Grid Variable-μ R.F. Amplifier	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	—	—	—	9003
									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.										9006

* Cathode resistor—ohms.

[1] Per Plate.

TABLE XII—CONTROL AND REGULATOR TUBES

Type	Name	Base	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Use	Peak Anode Voltage	Max. Anode Ma.	Minimum Supply Voltage	Operating Voltage	Operating Ma.	Grid Resistor	Tube Voltage Drop	Type
0A2	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	—	—	185	150	5–30	—	—	0A2
0B2	Voltage Regulator	7-pin B.	5BO	Cold	—	—	Voltage Regulator	—	—	133	108	5–30	—	—	0B2
0A4G	Gas Triode Starter-Anode Type	6-pin O.	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
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
2B4	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	2B4
6Q5G		5-pin M.	5A	Htr.	2.5	1.4									6Q5G
2C4	Gas Triode	7-pin B.	—	Fil.	2.5	—	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[5]	15	3C23
										—	100	1500	−2.5[5]	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[2]	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	—	—	—	—	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
KY21	Gas Triode Grid Type	4-pin M.	—	Fil.	2.5	10.0	Grid-Controlled Rectifier	—	—	—	3000	500	—	—	KY21
RK62	Gas Triode Grid Type	4-pin S.	4D	Fil.	1.4	0.05	Relay Tube	45	1.5	—	30–45	0.1–1.5	—	15	RK62
RM208	Permatron	4-pin M.	—	Fil.	2.5	5.0	Controlled Rectifier[1]	7500[2]	1000	—	—	—	—	15	RM208
RM209	Permatron	4-pin M.	—	Fil.	5.0	10.0	Controlled Rectifier[1]	7500[2]	5000	—	—	—	—	15	RM209
OA3/VR75	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	105	75	5–40	—	—	OA3/VR75
OB3/VR90	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	125	90	5–40	—	—	OB3/VR90
OC3/VR105	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	135	105	5–40	—	—	OC3/VR105
OD3/VR150	Voltage Regulator	6-pin O.	4AJ	Cold	—	—	Voltage Regulator	—	—	185	150	5–40	—	—	OD3/VR150
KY866	Mercury Vapor Triode	4-pin M.	Fig. 8	Fil.	2.5	5.0	Grid-Controlled Rectifier	10000	1000	0–150	—	—	—	—	KY866

[1] For use as grid-controlled rectifier or with external magnetic control. RM-208 has characteristics of 866, RM-209 of 872.
[2] When under control peak inverse rating is reduced to 2500.
[3] At 1000 anode volts.
[4] Grid tied to plate.
[5] Peak inverse voltage.
[6] Grid.
[7] Megohms.
[3] Grid voltage.

TABLE XIII—CATHODE-RAY TUBES AND KINESCOPES

Type	Name	Socket Connections	Heater		Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage [1]	Screen Input Power [2]	Deflection Sensitivity [6]		Anode No. 3 Voltage	Pattern Color	Type
			Volts	Amps.										D1 D2	D3 D4			
2AP1	Electrostatic Cathode-Ray	11B	6.3	0.6	Oscillograph Television	2"	1000	250	−60	—	—	660	—	0.11	0.13	—	Green	2AP1
							500	125	−30	—	—			0.22	0.26			
3AP1/ 906-P1-4-5-11	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3"	1500	430	−50	—	—	550	10	0.22	0.23	—	Green Blue White	3AP1/ 906-P1-4-5-11
							1000	285	−33	—	—			0.33	0.35			
							600	170	−20	—	—			0.55	0.58	—		
3BP1-4-11	Electrostatic Cathode-Ray	14A	6.3	0.6	Oscillograph	3"	2000	575	−60	—	—	550	—	0.13	0.17	—	Green	3BP1-4-11
							1500	430	−45	—	—			0.17	0.23			
3DP1	Electrostatic Cathode-Ray	Fig. 49	6.3	0.6	Oscillograph	3"	2000	575	−60	—	—	550	—	200 [3]	148 [3]	—	Green	3DP1
							1500	430	−40	—	—			150 [3]	111 [3]	—		
3EP1/ 1806-P1	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph Television	3"	2000	575	−60	—	—	550	—	0.115	0.154	—	Green	3EP1/ 1806-P1
							1500	430	−45	—	—			0.153	0.205			
3GP1-4-5-11	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph	3"	1500	350	−50	—	—	550	—	0.21	0.24	—	White Green Blue	3GP1-4-5-11
							1000	234	−33	—	—		—	0.32	0.36			
3JP1-2-4-11	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph	3"	2000	575	−60	—	—	550	—	0.13	0.17	4000	Green Blue White	3JP1-2-4-11
							1500	430	−45	—	—		—	0.17	0.23	3000		
3KP1	Electrostatic Cathode-Ray	11M	6.3	0.6	Oscillograph	3"	1000	300	−45	1000	—	500	—	68 [3]	136 [3]	—	Green	3KP1
							2000	600	−90	2000	—			52 [3]	104 [3]			
5AP1/ 1805-P1 5AP4/ 1805-P4	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph Television	5"	2000	575	−35	—	—	500	10	0.17	0.21	—	Green White	5AP1/ 1805-P1 5AP4/ 1805-P4
							1500	430	−27					0.23	0.28	—		
5BP1/ 1802-P1-2-4-5-11	Electrostatic Picture Tube	11A	6.3	0.6	Oscillograph	5"	2000	450	−40	—	—	500	10	0.3	0.33	—	Green White Blue	5BP1/ 1802-P1-2-4-5-11
							1500	337	−30	—	—			0.4	0.45			
5CP1-2-4-5-11	Electrostatic Cathode-Ray	14B	6.3	0.6	Oscillograph Television	5"	2000	575	−60	—	—	550	—	0.28	0.32	4000	White Green Blue	5CP1-2-4-5-11
							1500	430	−45	—	—		—	0.37	0.43	3000		
							2000	575	−60	—	—		—	0.36	0.41	2000		
5FP1-2-4-11	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Oscillograph Television	5'	7000	250	−45	—	—	—	—	—	—	—	Green White Blue	5FP1-2-4-11
							4000	250	−45	—	—		—	—	—			
5HP1 5HP4	Electrostatic Cathode-Ray	11A	6.3	0.6	Oscillograph	5"	2000	425	−40	—	—	500	—	0.3	0.33	—	Green White	5HP1 5HP4
							1500	310	−30					0.4	0.44	—		
5JP1-2-4-5-11	Electrostatic Cathode-Ray	11E	6.3	0.6	Oscillograph	5"	2000	520	−75	—	—	500	—	0.25	0.28	4000	White Green Blue	5JP1-2-4-5-11
							1500	390	−56	—	—		—	0.33	0.37	3000		
5LP1-2-4-5-11	Electrostatic Cathode-Ray	11F	6.3	0.6	Oscillograph Television	5"	2000	500	−60	—	—	500	—	0.25	0.28	4000	White Green Blue	5LP1-2-4-5-11
							1500	375	−45	—	—		—	0.33	0.37	3000		
							1000	250	−30	—	—		—	0.49	0.56	2000		
5MP1-4-5-11	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	5"	1500	375	−50	—	—	660	—	0.39	0.42	—	White Green Blue	5MP1-4-5-11
							1000	250	−33	—	—		—	0.58	0.64			
5RP1-2-4-11	Electrostatic Cathode-Ray	Fig. 34	6.3	0.6	Oscillograph	5"	3000	—	−90	—	—	1200	—	0.12	0.12	15000	Green White Blue	5RP1-2-4-11
							2000	575	−60	—	—		—	0.18	0.18	10000		
5TP4	Projection Kinescope	Fig. 46	6.3	0.6	Television	5"	27000	4900	−70	200	—	—	—	—	—	—	White	5TP4
7AP4	Electromagnetic Picture Tube	5AJ	2.5	2.1	Television	7"	3500	1000	−67.5	—	—	—	2.5	—	—	—	White	7AP4
7BP1-2-4-11	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Oscillograph Television	7"	7000	250	−45	—	—	—	—	—	—	—	White Green Blue	7BP1-2-4-11
							4000	250	−45	—	—							

TABLE XIII—CATHODE-RAY TUBES AND KINESCOPES—*Continued*

Type	Name	Socket Connections	Heater Volts	Heater Amps.	Use	Size	Anode No. 2 Voltage	Anode No. 1 Voltage	Cut-Off Grid Voltage	Grid No. 2 Voltage	Signal-Swing Voltage	Max. Input Voltage [1]	Screen Input Power [2]	Deflection Sensitivity [6] $D_1 D_2$	Deflection Sensitivity [6] $D_3 D_4$	Anode No. 3 Voltage	Pattern Color	Type
7CP1/ 1811-P1	Electromagnetic Cathode-Ray	6AZ	6.3	0.6	Oscillograph	7″	7000 4000	1470 840	−45 −45	250 250	— —	—	—	— —	— —	—	Green	7CP1/ 1811-P1
7DP4	Kinescope	Fig. 46	6.3	0.6	Television	7″	6000	1430	−45	250	—	—	—	—	—	—	White	7DP4
7GP4	Electrostatic Kinescope	Fig. 47	6.3	0.6	Television	7″	3000	1200	−84	3000	—	—	—	123 [3]	102 [3]	—	White	7GP4
9AP4/ 1804-P4	Electromagnetic Picture Tube	6AL	2.5	2.1	Television	9″	7000 6000	1425 1225	−40 −38	250	25	—	10	—	—	—	White	9AP4/ 1804-P4
9CP4	Electromagnetic Picture Tube	4AF	2.5	2.1	Television	9″	7000	—	−110	—	25	—	10	—	—	—	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	Fig. 48	6.3	0.6	Television	10″	—	9000	−45	250	—	—	—	—	—	—	—	10BP4
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	Electromagnetic Picture Tube	4AF	2.5	2.1	Television	12″	7000	—	−110	—	25	—	10	—	—	—	White	12CP4
12DP4	Electromagnetic Cathode-Ray	5AN	6.3	0.6	Television	12″	7000 4000	250 250	−45 −45	— —	— —	— —	— —	— —	— —	— —	White	12DP4
902	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	2″	600	150	−60	—	—	350	5	0.19	0.22	—	Green	902
903 [5]	Electromagnetic Cathode-Ray	6AL	2.5	2.1	Oscillograph	9″	7000	1360	−120	250	—	—	10	—	—	—	Green	903
904	Electrostatic-Magnetic Cathode-Ray	Fig. 3	2.5	2.1	Oscillograph	5″	4600	970	−75	250	—	4000	10	0.09	—	—	Green	904
905	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″	2000	450	−35	—	—	1000	10	0.19	0.23	—	Green	905
907	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″	Characteristics same as Type 905							—	—	—	Blue	907
908	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″	Characteristics same as Type 3AP1/906P1							—	—	—	Blue	908
909 [6]	Electrostatic Cathode-Ray	Fig. 6	2.5	2.1	Oscillograph	5″	Characteristics same as Type 905							—	—	—	Blue	909
910 [6]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″	Characteristics same as Type 3AP1/906P1							—	—	—	Blue	910
911 [5]	Electrostatic Cathode-Ray	7AN	2.5	2.1	Oscillograph	3″	Characteristics same as Type 3AP1/906P1							—	—	—	Green	911
912	Electrostatic Cathode-Ray	Fig. 8	2.5	2.1	Oscillograph	5″	10000	2000	−66	250	—	7000	10	0.041	0.051	—	Green	912
913	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	1″	500	100	−65	—	—	250	5	0.07	0.10	—	Green	913
914	Electrostatic Cathode-Ray	Fig. 12	2.5	2.1	Oscillograph	9″	7000	1450	−50	250	—	3000	10	0.073	0.093	—	Green	914
1800 [5]	Electromagnetic Kinescope	6AL	2.5	2.1	Television	9″	6000	1250	−75	250	25	—	10	—	—	—	Yellow	1800
1801 [5]	Electromagnetic Kinescope	Fig. 13	2.5	2.1	Television	5″	3000	450	−35	—	20	—	10	—	—	—	Yellow	1801
2001	Electrostatic Cathode-Ray	Fig. 2	6.3	0.6	Oscillograph	1″	Characteristics essentially same as 913											2001
2002	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscillograph	2″	600	120	—	—	—	—	—	0.16	0.17	—	Green	2002
2005	Electrostatic Cathode-Ray	Fig. 1 [4]	2.5	2.1	Television	5″	2000	1000	−35	200	—	—	10	0.5	0.56	—	—	2005
24-XH	Electrostatic Cathode-Ray	Fig. 1	6.3	0.6	Oscilloscope	2″	600	120	−60	—	—	—	10	0.14	0.16	—	Blue	24-XH

[1] Between Anode No. 2 and any deflecting plate. [2] In mw./sq. cm., max. [3] D.C. Volts/in. [4] Cathode connected to pin 7. [5] Discontinued. [6] In mm./volt d.c.

TABLE XIV—RECTIFIERS—RECEIVING AND TRANSMITTING

See also Table XI—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater		Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
					Volts	Amps.					
BA	Full-Wave Rectifier	4-pin M.	4J	Cold	—	—	350	350	Tube drop 80 v.		G
BH	Full-Wave Rectifier	4-pin M.	4J	Cold	—	—	350	125	Tube drop 90 v.		G
BR	Half-Wave Rectifier	4-pin M.	4H	Cold	—	—	300	50	Tube drop 60 v.		G
CE-220	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	3.0	—	20	20000	100	HV
OY4	Half-Wave Rectifier	5-pin O.	4BU	Cold	Connect Pins 7 and 8		95	75	300	500	G
OZ4	Full-Wave Rectifier	5-pin O	4R	Cold	—	—	350	30–75	1250	200	G
1	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	1000	400	MV
1-V	Half-Wave Rectifier	4-pin S.	4G	Htr.	6.3	0.3	350	50	—	—	HV
1B48	Half-Wave Rectifier	7-pin B.	—	Cold	—	—	800	6	2700	50	G
1Z2	Half-Wave Rectifier	7-pin B.	7CB	Fil.	1.5	0.3	7800	2	20000	10	HV
2V3G	Half-Wave Rectifier	6-pin O.	4AC	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	Half-Wave Rectifier	4-pin M.	4AB	Fil.	2.5	1.75	4500	7.5	—	—	HV
2Y2	Half-Wave Rectifier	4-pin M.	4AB	Fil.	2.5	1.75	4400	5.0	—	—	HV
2Z2/G84	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	1.5	350	50	—	—	HV
3B24	Half-Wave Rectifier	4-pin M.	T-4A	Fil.	5.0 2.5 [9]	3.0 3.0	— —	60 30	20000 20000	300 150	HV
3B25	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	500	4500	2000	G
3B26	Half-Wave Rectifier	8-pin O.	Fig. 31	Htr.	2.5	4.75	—	20	15000	8000	HV
DR-3B27	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	5.0	3000	250	8500	1000	HV
5R4GY	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	900 [4] 950 [7]	150 [4] 175 [7]	2800	650	HV
5T4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	3.0	450	250	1250	800	HV
5U4G	Full-Wave Rectifier	8-pin O.	5T	Fil.	5.0	3.0	Same as Type 5Z3				HV
5V4G	Full-Wave Rectifier	8-pin O.	5L	Htr.	5.0	2.0	Same as Type 83V				HV
5W4	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	1.5	350	110	1000	—	HV
5X3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	2.0	1275	30	—	—	HV
5X4G	Full-Wave Rectifier	8-pin O.	5Q	Fil.	5.0	3.0	Same as 5Z3				HV
5Y3G	Full-Wave Rectifier	5-pin O.	5T	Fil.	5.0	2.0	Same as Type 80				HV
5Y4G	Full-Wave Rectifier	8-pin O.	5Q	Fil.	5.0	2.0	Same as Type 80				HV
5Z3	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	3.0	500	250	1400	—	HV
5Z4	Full-Wave Rectifier	5-pin O.	5L	Htr.	5.0	2.0	400	125	1100	—	HV
6W5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.9	350	100	1250	350	HV
6X4	Full-Wave Rectifier	7-pin B.	Fig. 39	Htr.	6.3	0.6	325	70	1250	210	HV
6X5	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.5	350	75	—	—	HV
6Y5	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	Full-Wave Rectifier	6-pin S.	6K	Htr.	6.3	0.6	230	60	—	—	HV
6ZY5G	Full-Wave Rectifier	6-pin O.	6S	Htr.	6.3	0.3	350	35	1000	150	HV
7Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	6.3	0.5	350	60	—	—	HV
7Z4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	6.3	0.9	450 [1] 325 [1]	100	1250	300	HV
12A7	Rectifier-Pentode	7-pin S.	7K	Htr.	12.6	0.3	125	30	—	—	HV
12Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	12.6	0.3	250	60	—	—	HV
12Z5	Voltage Doubler	7-pin M.	7L	Htr.	12.6	0.3	225	60	—	—	HV
14Y4	Full-Wave Rectifier	8-pin L.	5AB	Htr.	12.6	0.3	450 [1] 325 [1]	70	1250	210	HV
14Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	12.6	0.3	250	60	—	—	HV
25A7G	Rectifier-Pentode	8-pin O.	8F	Htr.	25	0.3	125	75	—	—	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	Voltage Doubler	6-pin S.	6E	Htr.	25	0.3	250	85	—	—	HV
25Z3	Half-Wave Rectifier	4-pin S.	4G	Htr.	25	0.3	250	50	—	—	HV
25Z4	Half-Wave Rectifier	6-pin O.	5AA	Htr.	25	0.3	125	125	—	—	HV
25Z5	Rectifier-Doubler	6-pin S.	6E	Htr.	25	0.3	125	100	—	500	HV
25Z6	Rectifier-Doubler	7-pin O.	7Q	Htr.	25	0.3	125	100	—	500	HV
28Z5	Full-Wave Rectifier	8-pin L.	5AB	Htr.	28	0.24	450 [7] 325 [4]	100	—	300	HV
32L7GT	Rectifier-Tetrode	8-pin O.	8Z	Htr.	32.5	0.3	125	60	—	—	HV
35W4	Half-Wave Rectifier	7-pin B.	5BQ	Htr.	35 [2]	0.15	125	100 [5]	330	600	HV
35Y4	Half-Wave Rectifier	8-pin O.	5AL	Htr.	35 [2]	0.15	235	60 100 [8]	700	600	HV
35Z3	Half-Wave Rectifier	8-pin L.	4Z	Htr.	35	0.15	250 [5]	100	700	600	HV
35Z4GT	Half-Wave Rectifier	6-pin O.	5AA	Htr.	35	0.15	250	100	700	600	HV
35Z5G	Half-Wave Rectifier	6-pin O.	6AD	Htr.	35 [2]	0.15	125	60 100 [9]	—	—	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

See also Table XI—Control and Regulator Tubes

Type No.	Name	Base	Socket Connections	Cathode	Fil. or Heater Volts	Fil. or Heater Amps.	Max. A.C. Voltage Per Plate	D.C. Output Current Ma.	Max. Inverse Peak Voltage	Peak Plate Current Ma.	Type
45Z5GT	Half-Wave Rectifier	6-pin O.	6AD	Htr.	45[2]	0.15	125	60 100[8]	—	—	HV
50Y6GT	Full-Wave Rectifier	7-pin O.	7Q	Htr.	50	0.15	125	85	—	—	HV
50Z6G	Voltage Doubler	7-pin O.	7Q	Htr.	50	0.3	125	150	—	—	HV
50Z7G	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	Half-Wave Rectifier	4-pin J.	T-3A	Fil.	10	3.25	—	—	3500	600	HV
217-C	Half-Wave Rectifier	4-pin J.	T-3A	Fil.	10	3.25	—	—	7500	600	HV
Z225	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	250	10000	1000	MV
249-B	Half-Wave Rectifier	4-pin M.	Fig. 53	Fil.	2.5	7.5	3180	375	10000	1500	MV
HK253	Half-Wave Rectifier	4-pin J.	T-3A	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	1750	125	5000	500	MV
836	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	5.0	—	—	5000	1000	HV
866A/866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	3500	250	10000	1000	MV
866B	Half-Wave Rectifier	4-pin M.	4P	Fil.	5.0	5.0	—	—	8500	1000	MV
866 Jr.	Half-Wave Rectifier	4-pin M.	4B	Fil.	2.5	2.5	1250	250[3]	—	—	MV
HY866 Jr.	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	2.5	1750	250[3]	5000	—	MV
RK866	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	3500	250	10000	1000	MV
871[10]	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	2.0	1750	250	5000	500	MV
878	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	7100	5	20000	—	HV
879	Half-Wave Rectifier	4-pin S.	4P	Fil.	2.5	1.75	2650	7.5	7500	100	HV
872A/872	Half-Wave Rectifier	4-pin J.	T-3A	Fil.	5.0	7.5	—	1250	10000	5000	MV
975A	Half-Wave Rectifier	4-pin J.	T-3A	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.	T-9F	Fil.	6.3	0.1	—	70	450	—	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
1275	Full-Wave Rectifier	4-pin M.	4C	Fil.	5.0	1.75	Same as 5Z3				HV
1616	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	130	6000	800	HV
1641/ RK60	Full-Wave Rectifier	4-pin M.	T-4AG	Fil.	5.0	3.0	—	50 250	4500 2500	— —	HV
1654	Half-Wave Rectifier	7-pin B.	Fig. 41	Fil.	1.4	0.05	2500	1	7000	6	HV
8008	Half-Wave Rectifier	4-pin[6]	Fig. 11	Fil.	5.0	7.5	—	1250	10000	5000	MV
8013A	Half-Wave Rectifier	4-pin M.	4P	Fil.	2.5	5.0	—	20	40000	150	HV
8016	Half-Wave Rectifier	6-pin O.	4AC	Fil.	1.25	0.2	—	2.0	10000	7.5	HV
8020	Half-Wave Rectifier	4-pin M.	4P	Fil.	5.0 5.8	5.5 6.5	10000 12500	100 100	40000 40000	750 750	HV
RK19	Full-Wave Rectifier	4-pin M.	T-3A	Htr.	7.5	2.5	1250	200[1]	3500	600	HV
RK21	Half-Wave Rectifier	4-pin M.	4P	Htr.	2.5	4.0	1250	200[1]	3500	600	HV
RK22	Full-Wave Rectifier	4-pin M.	T-4AG	Htr.	2.5	8.0	1250	200[4]	3500	600	HV

[1] With input choke of at least 20 henrys.
[2] Tapped for pilot lamps.
[3] Per pair with choke input.
[4] Condenser input.
[5] With 100 ohms min. resistance in series with plate; without series resistor, maximum r.m.s. plate rating is 117 volts.
[6] Same as 872A/872 except for heavy-duty push-type base. Filament connected to pins 2 and 3, plate to top cap.
[7] Choke input.
[8] Without panel lamp.
[9] Using only one-half of filament.
[10] Discontinued.

TABLE XV—TRIODE TRANSMITTING TUBES

Type	Max. Plate Dissi- pation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
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	958-A
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	RK24
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)	150	− 10	30	16	0.35	3.5	6J6
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	9002
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	955
HY114B	1.8	1.4	0.155	180	12	3.0	13	1.0	1.3	1.0	300	O.	T-8AC	Class-C Amp.-Oscillator	180	− 30	12	2.0	0.2	1.4[3]	HY114B
														Class-C Amp. Plate-Mod.	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	150	− 35	30	5.0	0.2	2.2	3A5
6F4	2.0	6.3	0.225	150	20	8.0	17	2.0	1.9	0.6	500	A.	7BR	Class-C Amp.-Oscillator	150	− 15 550* 2000[4]	20	7.5	0.2	1.8	6F4
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	HY24
														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	250	− 60	20	6.0	0.54	3.5	RK33
6N4	3.0	6.3	0.2	180	12	—	32	3.1	2.35	0.55	500	B.	Fig. 40	Class-C Amp.-Oscillator	180	—	—	—	—	—	6N4
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	HY6J5GTX
														Class-C Amp. (Plate Mod.)	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)	—	—	—	—	—	—	2C22/7193
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]	HY615 HY-E1148
														Class-C Amp. Plate-Mod.	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-446A GL-446B
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	—	—	—	—	—	GL-2C44 GL-464A
6C4	5.0	6.3	0.15	300	25	8.0	17	1.8	1.6	1.3	150	B.	6BG	Class-C Amp.-Oscillator	300	− 27	25	7.0	0.35	5.5	6C4
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	1626
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	250	− 60	40	12	1.0	7	2C21/ RK33
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	2C40
2C43	12	6.3	0.9	500	40	—	48	2.9	1.7	0.05	1250	O.	Fig. 19	Class-C Amp.-Oscillator	470[7]	—	38[7]	—	—	9[7]	2C43
2C26A	10	6.3	1.10	3500[5]	—	—	16.3	2.6	2.8	1.1	250	O.	4BB	Pulse Oscillator	400	− 15	16	—	—	—	2C26A
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	300	− 36	80	20	1.8	16	2C34/ RK34
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	205D
														Class-C Amp. (Plate-Mod.)	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	2C25
														Class-C Amp. Plate-Mod.	350	−100	50	12	2.2	12	
10Y	15	7.5	1.25	450	65	15	8	4.1	7.0	3.0	—	M.	4D	Class-C Amp.-Oscillator	450	−100	65	15	3.2	19	10Y
														Class-C Amp. Plate-Mod.	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	843
														Class-C Amp. (Plate-Mod.)	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	RK59
HY75	15	6.3	2.5	450	80	20	10	1.6	3.8	0.6	60	O.	T-8AC	Class-C Amp.-Oscillator	450	− 50	80	12	—	21[3]	HY75
														Class-C Amp. Plate-Mod.	450	− 60	80	12	—	16[3]	

TABLE XV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
1602	15	7.5	1.25	450	60	15	8.0	4.0	7.0	3.0	6	M.	4D	Class-C Amp. (Telegraphy)	450	−115	55	15	3.3	13	1602
														Class-C Amp. (Telephony)	350	−135	45	15	3.5	8.0	
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	841
														Class-C Amp. (Telephony)	350	− 47	50	15	2.0	11	
10 RK10 [1]	15	7.5	1.25	450	65	15	8.0	3.0	8.0	4.0	—	M.	4D	Class-C Amp. (Telegraphy)	450	−100	65	15	3.2	19	10 RK10
											60			Class-C Amp. (Telephony)	350	−100	50	12	2.2	12	
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	RK100
														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.	T-8AC	Class-C Amp.-Oscillator	750	−150	75	20	1.5/2.5	40	TUF-20
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	1608
														Class-C Amp. (Telephony)	350	− 80	85	20	3.0	18	
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	310
														Class-C Amp. (Telephony)	500	−190	55	15	4.5	18	
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	801-A/801
														Class-C Amp. (Telephony)	500	−190	55	15	4.5	18	
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	HY801-A
														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	T20
														Class-C Amp. Plate-Mod.	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	TZ20
														Class-C Amp Plate-Mod.	750	−100	70	23	4.8	38	
15E	20	5.5	4.2	10000 [5]	—	—	25	1.4	1.15	0.3	600	N.	T-4AF	Oscillator at 400 Mc.	10000	4500 [3]	3	1	—	10000 [6]	15E
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	3-25A3 25T
															1500	− 95	67	13	2.2	75	
															1000	− 70	72	9	1.3	47	
3-25D3 3C24 24G	25	6.3	3.0	2000	75	25	23	2.0	1.6	0.2	60	S.	2D	Class-C Amp.-Oscillator	2000	−170	63	17	4.5	100	3-25D3 3C24 24G
								1.7	1.5	0.3					1500	−110	67	15	3.1	75	
															1000	− 80	72	15	2.6	47	
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						3C28
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						3C34
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	RK11
														Class-C Amp. Plate-Mod.	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	RK12
														Class-C Amp. Plate-Mod.	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	HK24
														Class-C Amp. Plate-Mod.	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	HY25
														Class-C Amp. Plate-Mod.	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	8025
	20				65	20								Class-C Amp. (Plate Mod.)	800	−105	40	10.5	1.4	22	
	30				80	20								Class-C Amp. (Telegraphy)	1000	− 90	50	14	1.6	35	
HY30Z	30	6.3	2.25	850	90	25	87	6.0	4.9	1.0	60	M.	T-4BE	Class-C Amp.-Oscillator	850	− 75	90	25	2.5	58	HY30Z
														Class-C Amp. Plate-Mod.	700	− 75	90	25	3.5	47	
HY31Z [2] HY1231Z [2]	30	6.3	3.5	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	HY31Z HY1231Z
		12.6	1.7											Class-C Amp. (Telephony)	400	−100	150	30	3.5	45	

TABLE XV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
316A	30	2.0	3.65	450	80	12	6.5	1.2	1.6	0.8	500	N.	—	Class-C Amp. (Telegraphy)	450	—	80	12	—	7.5	316A
														Class-C Amp. Plate-Mod.	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	809
														Class-C Amp. Plate-Mod.	750	−60	100	32	4.3	55	
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	1623
														Class-C Amp. Plate-Mod.	750	−125	100	20	4.0	55	
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						53A
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	RK30
														Class-C Amp. Plate-Mod.	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	800
														Class-C Amp. Plate-Mod.	1000	−200	70	15	4.0	50	
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	1628
														Class-C Amp. Plate-Mod.	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	2.8	0.35	500	N.	T-4BB	Class-C Amp.-Oscillator	1000	−90	50	14	1.6	35	8012 GL-8012-A
								2.7	2.5	0.4				Class-C Amp. Plate-Mod.	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	RK18
														Class-C Amp. Plate-Mod.	1000	−160	80	13	3.1	64	
RK31	40	7.5	3.0	1250	100	35	170	7.0	1.0	2.0	30	M.	3G	Class-C Amp. (Telegraphy)	1250	−80	100	30	3.0	90	RK31
														Class-C Amp. Plate-Mod.	1000	−80	100	28	3.5	70	
HY40	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	HY40
														Class-C Amp. Plate-Mod.	850	−90	125	25	5.0	82	
														Grid-Modulated Amp.	1000	—	125	—	—	20	
HY40Z	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	HY40Z
														Class-C Amp. Plate-Mod.	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	T40
														Class-C Amp. Plate-Mod.	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	TZ40
														Class-C Amp. Plate-Mod.	1250	−100	125	30	7.5	116	
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	HY57
														Class-C Amp. Plate-Mod.	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	—	—	756
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	830
														Grid-Modulated Amp.	1000	−200	50	2.0	3.0	15	
3-50A4 35T 3-50D4 35TG	50	5.0	4.0	2000	150	50	39	4.1	1.8	0.3	100	M.	3G	Class-C Amp. (Telegraphy)	2000	−135	125	45	13	200	3-50A4 35T 3-50D4 35TG
								2.5	1.8	0.4	100	M.	2D	Class-C Amp. Plate Mod.	1500	−120	100	30	5.0	120	
														Grid Modulated Amp.	2000	−400	60	3.0	3.0	50	
8010-R	50	6.3	2.4	1350	150	20	30	2.3	1.5	0.07	350	N.	—	Class-C Amplifier	—	—	—	—	—	—	8010-R
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	RK32
														Class-C Amp. Plate-Mod.	1000	−310	100	21	8.7	70	

TABLE XV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
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	RK35
														Class-C Amp. Plate-Mod.	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	RK37
														Class-C Amp. Plate-Mod.	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	UH50
														Class-C Amp. Plate-Mod.	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	UH51
														Class-C Amp. Plate-Mod.	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	HK54
														Class-C Amp. Plate-Mod.	2500	−250	100	20	8.0	210	
														Grid-Modulated Amp.	2000	−150	39	1.5	3.0	28	
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	HK154
														Class-C Amp. Plate-Mod.	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	HK158
														Class-C Amp. Plate-Mod.	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	WE304A 304B
														Class-C Amp. Plate-Mod.	1000	−180	100	—	—	65	
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	356A
														Class-C Amp. Plate-Mod.	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	808
														Class-C Amp. Plate-Mod.	1250	−225	100	32	10.5	105	
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	834
														Class-C Amp. Plate-Mod.	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	841A
841SW	50	10	2.0	1000	150	30	14.6	—	9.0	—	—	M.	3G	Class-C Amplifier	—	—	—	—	—	—	841SW
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	T55
														Class-C Amp. Plate-Mod.	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	811
														Class-C Amp. Plate-Mod.	1250	−125	125	50	11	120	
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	812
														Class-C Amp. Plate-Mod.	1250	−125	125	25	6.0	120	
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	RK51
														Class-C Amp. Plate-Mod.	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	RK52
														Class-C Amp. Plate-Mod.	1250	−120	115	47	8.5	102	
T-60 HF60	60	10	2.5	1600	150	50	20	5.5	5.2	2.5	60 30	M.	2D	Class-C Amp.-Oscillator	1500	−150	150	50	9.0	100	T-60 HF60

TABLE XV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
826	60	7.5	4.0	1000	125	40	31	3.7	2.9	1.4	250	N.	T-9A	Class-C Amp.-Oscillator	1000	−70	125	35	5.8	86	826
														Class-C Amp. Plate-Mod.	800	−98	94	35	6.2	53	
														Class-B Amp. (Telephony)	1000	−50	65	8.5	3.7	22	
														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	830B 930B
														Class-C Amp. Plate-Mod.	800	−150	95	20	5.0	50	
														Class-B Amp. (Telephony)	1000	−35	85	6.0	6.0	26	
HY51A HY51B	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	HY51A HY51B
														Class-C Amp. Plate-Mod.	1000	−67.5	130	15	7.5	104	
														Grid-Modulated Amp.	1000	—	100	—	—	33	
HY51Z	65	7.5	3.5	1000	175	35	85	7.9	7.2	0.9	60	M.	T-4BE	Class-C Amp. (Telegraphy)	1000	−22.5	175	35	10	131	HY51Z
														Class-C Amp. Plate-Mod.	1000	−30	150	35	10	104	
														Grid-Modulated Amp.	1000	—	100	—	—	33	
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	UH35
														Class-C Amp. Plate-Mod.	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.	T-3AB 3G	Class-C Amp. (Telegraphy)	1500	−215	130	6.0	3.0	140	V70 V70B
														Class-C Amp. Plate-Mod.	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.	T-3AB 3G	Class-C Amp. (Telegraphy)	1000	−110	140	30	7.0	90	V70A V70C
														Class-C Amp. Plate-Mod.	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	50T
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	3-75A3 75TH
3-75A2 75TL						35	12	2.6	2.4	0.4				Class-C Amp. (Telegraphy)	2000	−300	150	21	8	225	3-75A2 75TL
HF75	75	10	3.25	2000	120	—	12.5	—	2.0	—	75	M.	2D	Class-C Oscillator-Amp.	2000	—	120	—	—	150	HF75
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	TW75
														Class-C Amp. Plate-Mod.	2000	−260	125	32	13.2	198	
T-100 HF100	75	10	2.0	1500	150	30	23	3.5	4.5	1.4	30	M.	2D	Class-C Amp. (Telegraphy)	1500	−200	150	18	6.0	170	T-100 HF100
														Class-C Amp. Plate-Mod.	1250	−250	110	21	8.0	105	
														Grid-Modulated Amp.	1500	−280	72	1.5	6.0	42	
111H	75	10	2.25	1500	160	—	23	—	4.6	—	25	M.	2D	Class-C Osc.-Amp.	1500	—	160	—	—	175	111H
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	ZB120
														Class-C Amp. Plate-Mod.	1000	−150	120	21	5.0	95	
														Grid-Modulated Amp.	1250	—	95	8.0	1.5	45	
327B	75	10.5	10.6	15000	—	—	30	3.4	2.45	0.3	—	N.	T-4AD	—	—	—	—	—	—	—	327B
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	242A
														Class-C Amp. Plate-Mod.	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	284D
														Class-C Amp. Plate-Mod.	1000	−450	150	50	—	100	
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	812-H
															1250	−125	125	25	5.0	116	
														Class-C Amp. (Plate-Mod.)	1500	−125	165	21	6.0	180	
															1250	−125	125	25	6.0	120	
8005	85	10	3.25	1500	200	45	20	6.4	5.0	1.0	60	M.	3G	Class-C Amp.-Oscillator	1500	−130	200	32	7.5	220	8005
														Class-C Amp. Plate-Mod.	1250	−195	190	28	9.0	170	
														Class-B Amp. (Telephony)	1500	−80	83	1.0	5.0	45	

TABLE XV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
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	V-70-D
															1500	− 90	165	19	3.9	195	
														Class-C Amp. (Plate-Mod.)	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	RK36
														Class-C Amp. (Telephony)	2000	−360	150	30	15	200	
														Grid-Modulated Amp.	2000	−270	72	1.0	3.5	42	
														Class-B Amp. (Telephony)	2000	−180	75	3.0	10	50	
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	RK38
														Class-C Amp. (Telephony)	2000	−200	160	30	10	225	
														Grid-Modulated Amp.	2000	−150	80	2.0	5.5	60	
														Class-B Amp. (Telephony)	2000	−100	75	2.0	7.0	55	
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	3-100A4 100TH
														Class-C Amp. Plate-Mod.	3000	−210	167	45	18	400	
														Class-B Amp. (Telephony)	3000	− 70	50	2.0	5.0	50	
														Grid-Modulated Amp.	3000	−400	70	3.0	7.0	100	
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	3-100A2 100TL
														Class-C Amp. Plate-Mod.	3000	−600	167	35	18	400	
														Class-B Amp. (Telephony)	3000	−280	50	1.0	5.0	50	
														Grid-Modulated Amp.	3000	−560	60	2.0	7.0	90	
VT127A	100	5.0	10.4	3000	—	30	15.5	2.7	2.3	0.35	150	N.	T-4B	Class-C Amp.-Oscillator	Characteristics similar to 100TL						VT127A
227A	100	10.5	10.7	15000[5]	—	—	31	3.0	2.2	0.30	—	N.	T-4B	Oscillator at 200 Mc.	15000	1200[4]	10	3	—	50000[6]	227A
327A	100	10.5	10.7	15000[5]	—	—	31	3.4	2.3	0.35	—	N.	T-4AD	Oscillator at 200 Mc.	15000	1200[4]	10	3	—	50000[6]	327A
HK254	100	5.0	7.5	4000	200	40	25	3.3	3.4	1.1	50	J.	T-3AC	Class-C Amp. (Telegraphy)	4000	−380	120	35	20	475	HK254
														Class-C Amp. Plate-Mod.	3000	−290	135	40	23	320	
														Class-B Amp. (Telephony)	3000	−125	51	2.0	3.0	54	
														Grid-Modulated Amp.	3000	—	51	3.0	4.0	58	
RK58	100	10	3.25	1250	175	70	—	8.5	6.5	10.5	—	J.	T-3AB	Class-C Amp. (Telegraphy)	1250	− 90	150	30	6.0	130	RK58
														Class-C Amp. Plate-Mod.	1000	−135	150	50	16	100	
														Class-B Amp. (Telephony)	1250	—	106	15	6.0	42.5	
HF120	100	10	3.25	1250	175	—	12	—	10.5	—	20	J.	—	Class-C Amp.-Oscillator	1250	—	175	—	—	150	HF120
HF125	100	10	3.25	1500	175	—	25	—	11.5	—	30	J.	—	Class-C Amp.-Oscillator	1500	—	175	—	—	200	HF125
HF140	100	10	3.25	1250	175	—	12	—	12.5	—	15	J.	—	Class-C Amp.-Oscillator	1250	—	175	—	—	150	HF140
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	203A 303A
														Class-C Amp. (Telephony)	1000	−135	150	50	14	100	
														Class-B Amp. (Telephony)	1250	− 45	105	3.0	3.0	42.5	
203H	100	10	3.25	1500	175	60	25	6.5	11.5	1.5	15	J.	T-3AB	Class-C Amp. (Telegraphy)	1500	−200	170	12	3.8	200	203H
														Class-C Amp. (Telephony)	1250	−160	167	19	5.0	160	
														Class-B Amp. (Telephony)	1500	− 48	100	3.0	2.0	52	
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	211 311 835
														Class-C Amp. (Telephony)	1000	−260	150	35	14	100	
														Class-B Amp. (Telephony)	1250	−100	106	1.0	7.5	42.5	
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	242B 342B
														Class-C Amp. Plate-Mod.	1000	−160	150	50	—	100	
														Class-B Amp. (Telephony)	1250	− 80	120	—	—	50	

TABLE XV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode		Max. Plate Voltage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factor	Interelectrode Capacitances (μμfd.)			Max. Freq. Mc. Full Ratings	Base	Socket Connections	Typical Operation	Plate Voltage	Grid Voltage	Plate Current Ma.	D.C. Grid Current Ma.	Approx. Grid Driving Power Watts	Approx. Carrier Output Power Watts	Type
		Volts	Amps.					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	242C
														Class-C Amp. Plate-Mod.	1000	−160	150	50	—	100	
														Class-B Amp. (Telephony)	1250	−90	120	—	—	50	
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	261A 361A
														Class-C Amp. Plate-Mod.	1000	−160	150	50	—	100	
														Class-B Amp. (Telephony)	1250	−100	125	—	—	50	
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	276A 376A
														Class-C Amp. Plate-Mod.	1000	−160	125	50	—	85	
														Class-B Amp. (Telephony)	1250	−100	125	—	—	50	
284B	100	10	3.25	1250	150	100	5.0	4.2	7.4	5.3	—	J.	T-3AB	Class-C Amp. (Telegraphy)	1250	−500	150	—	—	125	284B
														Class-C Amp. Plate-Mod.	1000	−430	150	50	—	100	
														Class-B Amp. (Telephony)	1250	−270	120	—	—	50	
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	295A
														Class-C Amp. Plate-Mod.	1000	−125	150	50	—	100	
														Class-B Amp. (Telephony)	1250	−75	105	—	—	42.5	
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	838 938
														Class-C Amp. (Telephony)	1000	−135	150	60	16	100	
														Class-B Amp. (Telephony)	1250	0	106	15	6.0	42.5	
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	852
														Class-C Amp. (Telephony)	2000	−500	67	30	23	75	
														Class-B Amp. (Telephony)	3000	−250	43	0	7.0	40	
8003	100	10	3.25	1500	250	50	12	5.8	11.7	3.4	30	J.	T-3AB	Class-C Amp.-Oscillator	1350	−180	245	35	11	250	8003
														Class-C Amp. Plate-Mod.	1100	−260	200	40	15	167	
														Class-B Amp. (Telephony)	1350	−110	110	1.5	8	50	
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	3X100A11 2C39
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	3C22
4C36	125	5	7.5	4000	—	—	29	3.2	3.0	0.4	60	J.	Fig. 56	Class-C Amp.-Oscillator	—	—	—	—	18	480	4C36
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	F-123-A DR-123C
														Class-C Amp. Plate-Mod.	1500	−290	160	25	10	200	
														Class-B Amp. (Telephony)	1500	−100	120	1	6	65.5	
RK57/805	125	10	3.25	1500	210	70	—	6.5	8.0	5.0	30	J.	T-3AB	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	215	RK57/805
														Class-C Amp. (Telephony)	1250	−160	160	60	16	140	
														Class-B Amp. (Telephony)	1500	−10	115	15	7.5	57.5	
T125	125	10	4.5	2500	250	60	25	6.3	6.0	1.3	60	J.	T-3AC	Class-C Amp. (Telegraphy)	2500	−200	240	31	11	475	T125
														Class-C Amp. Plate-Mod.	2000	−215	200	28	10	320	
HF130	125	10	3.25	1250	210	—	12.5	—	9.0	—	20	J.	—	Class-C Amp.-Oscillator	1250	−210	—	—	—	170	HF130
HF150	125	10	3.25	1500	210	—	12.5	—	7.2	—	30	J.	—	Class-C Amp.-Oscillator	1500	—	210	—	—	200	HF150
HF175	125	10	4.0	2000	250	—	18	—	6.3	—	25	J.	—	Class-C Amp.-Oscillator	2000	—	250	—	—	300	HF175
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	GL146
														Class-C Amp. Plate-Mod.	1000	−200	160	40	—	100	
														Class-B Amp. (Telephony)	1250	0	132	—	—	55	
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	GL152
														Class-C Amp. Plate-Mod.	1000	−200	160	30	—	100	
														Class-B Amp. (Telephony)	1250	−40	132	—	—	55	

TABLE XV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
805	125	10	3.25	1500	210	70	40/60	8.5	6.5	10.5	30	J.	T-3AB	Class-C Amp. (Telegraphy)	1500	−105	200	40	8.5	215	805
														Class-C Amp. Plate-Mod.	1250	−160	160	60	16	140	
														Class-B Amp. (Telephony)	1500	−10	115	15	7.5	57.5	
3X150A3 3C37	150	6.3	2.5	1000	—	—	23	4.2	3.5	0.6	500	N.	—	—	—	—	—	—	—	—	3X150A3 3C37
150T	150	5.0	10	3000	200	50	13	3.0	3.5	0.5	—	J.	T-3AC	Class-C Amp. (Telegraphy)	3000	−600	200	35	—	450	150T
3-150A3 152TH	150	5/10	12.5/6.25	3000	450	85	20	5.7	4.5	0.8	40	J.	4BC	Class-C Amp. (Telegraphy)	3000	−300	250	70	27	600	3-150A3 152TH
3-150A2 152TL						75	12	4.5	4.4	0.7				Class-C Amp. (Telegraphy)	3000	−400	250	40	20	600	3-150A2 152TL
TW150	150	10	4.1	3000	200	60	35	3.9	2.0	0.8	—	J.	T-3AC	Class-C Amp.-Oscillator	3000	−170	200	45	17	470	TW150
														Class-C Amp. Plate-Mod.	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.	T-4BF	Class-C Amp.-Oscillator	3000	−400	250	30	15	610	HK252-L
														Class-C Amp. Plate-Mod.	2500	−350	250	35	16	500	
HF200 HV18	150	10–11	3.4	2500	200	50	18	5.2	5.8	1.2	20	J.	T-3AC	Class-C Amp. (Telegraphy)	2500	−300	200	18	8.0	380	HF200 HV18
														Class-C Amp. Plate-Mod.	2000	−350	160	20	9.0	250	
														Class-B Amp. (Telephony)	2500	−140	90	—	4.0	80	
HD203A	150	10	4.0	2000	250	60	25	—	12	—	15	J.	T-3AB	Class-C Amplifier	—	—	—	—	—	375	HD203A
HF250	150	10.5	4.0	2500	200	—	18	—	5.8	—	20	J.	T-3AC	Class-C Amp.-Oscillator	2500	—	200	—	—	375	HF250
HK354 HK354C	150	5.0	10	4000	300	50	14	4.5	3.8	1.1	30	J.	T-3AC	Class-C Amp. (Telegraphy)	4000	−690	245	50	48	830	HK354 HK354C
														Class-C Amp. Plate-Mod.	3000	−550	210	50	35	525	
														Class-B Amp. (Telephony)	3000	−205	78	2.0	10	82	
														Grid-Modulated Amp.	3000	−400	78	3.0	12	85	
HK354D	150	5.0	10	4000	300	55	22	4.5	3.8	1.1	30	J.	T-3AC	Class-C Amp. (Telegraphy)	3500	−490	240	50	38	690	HK354D
														Class-C Amp. Plate-Mod.	3500	−425	210	55	36	525	
HK354E	150	5.0	10	4000	300	60	35	4.5	3.8	1.1	30	J.	T-3AC	Class C Amp. (Telegraphy)	3500	−448	240	60	45	690	HK354E
														Class-C Amp. Plate-Mod.	3000	−437	210	60	45	525	
HK354F	150	5.0	10	4000	300	75	50	4.5	3.8	1.1	30	J.	T-3AC	Class-C Amp. (Telegraphy)	3500	−368	250	75	50	720	HK354F
														Class-C Amp. Plate-Mod.	3000	−312	210	75	45	525	
810 1627	150	10 5.0	4.5 9.0	2250	275	70	36	8.7	4.8	12	30	J.	T-3AC	Class-C Amp. (Telegraphy)	2250	−160	275	40	12	475	810 1627
														Class-C Amp. Plate-Mod.	1800	−200	250	50	17	335	
														Class-B Amp. (Telephony)	2250	−70	100	2.0	4.0	75	
														Grid-Modulated Amp.	2250	−140	100	2.0	4.0	75	
8000	150	10	4.5	2250	275	40	16.5	5.0	6.4	3.3	30	J.	T-3AC	Class-C Amp.-Oscillator	2250	−210	275	25	9.0	475	8000
														Class-C Amp. Plate-Mod.	1800	−320	250	20	8.8	335	
														Class-B Amp. (Telephony)	2250	−145	100	0	5.4	75	
														Grid-Modulated Amp.	2250	−265	100	0	2.5	75	
RK63 RK63A	200	5.0 6.3	10 14	3000	250	60	37	2.7	3.3	1.1	—	J.	T-3AC	Class-C Amp. (Telegraphy)	3000	−200	233	45	17	525	RK63 RK63A
														Class-C Amp. Plate-Mod.	2500	−200	205	50	19	405	
														Class-B Amp. (Telephony)	3000	−150	100	1.0	12	100	
														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.	T-3AC	Class-C Amp. (Telegraphy)	2500	−280	350	54	25	685	T200
														Class-C Amp. Plate-Mod.	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	F-127-A
														Class-C Amp. Plate-Mod.	2500	−300	200	58	25.2	420	

TABLE XV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
822 822S	200	10	4.0	2500	300	60	30	8.5	13.5	2.1	20 30	J.	T-3AB T-3AC	Class-C Amp. (Telegraphy)	2500	−190	300	51	17	600	822 822S
														Class-C Amp. Plate-Mod.	2000	−75	250	43	13.7	405	
4C32	200	10	4.5	3000	300	60	30	5.5	5.8	1.1	60	J.	T-3AC	Class-C Amp.-Oscillator	2000	−165	275	20	10	400	4C32
														Class-C Amp. (Plate-Mod.)	2000	−200	250	20	15	375	
GL-592	200	10	5.0	3500	250	50	24	3.6	3.3	0.41	110	N.	Fig. 52	Class-C Amp.-Oscillator	2600	−240	250	45	18	425	GL-592
														Class-C Amp. (Plate-Mod.)	2000	−500	250	50	—	—	
4C34 HF300	200	11–12	4.0	3000	275	60	23	6.0	6.5	1.4	60	J.	T-3AC	Class-C Amp. (Telegraphy)	3000	−400	250	28	16	600	4C34 HF300
														Class-C Amp. Plate-Mod.	2000	−300	250	36	17	385	
											20			Class-B Amp. (Telephony)	2500	−100	120	0.5	6.0	105	
T814 HV12	200	10	4.0	2500	200	60	12	8.5	12.8	1.7	30	J.	T-3AB	Class-C Amp. (Telegraphy)	2500	−240	300	30	10	575	T814 HV12
														Class-C Amp. Plate-Mod.	2000	−370	300	40	20	485	
T822 HV27	200	10	4.0	2500	300	60	27	8.5	13.5	2.1	30	J.	T-3AB	Class-C Amp. (Telegraphy)	2500	−175	300	50	15	585	T822 HV27
														Class-C Amp. Plate-Mod.	2000	−195	250	45	15	400	
														Class-B Amp. (Telephony)	2500	−95	125	5.0	8.0	110	
806	225	5.0	10	3300	300	50	12.6	6.1	4.2	1.1	30	J.	T-3AC	Class-C Amp. (Telegraphy)	3300	−600	300	40	34	780	806
														Class-C Amp. Plate-Mod.	3000	−670	195	27	24	460	
														Class-B Amp. (Telephony)	3300	−280	102	—	10.3	115	
3-250A4 250TH	250	5.0	10.5	4000	350	100	37	5.0	2.9	0.7	40	J.	T-3AC	Class-C Amp. (Telegraphy)	2000	−120	350	100	34	750	3-250A4 250TH
														Class-C Amp. Plate-Mod.	3000	−210	330	75	42	750	
														Class-B Amp. (Telephony)	3000	−80	125	4.0	15	125	
														Grid-Modulated Amp.	3000	−160	125	4.5	20	125	
3-250A2 250TL	250	5.0	10.5	4000	350	50	14	3.7	3.1	0.7	40	J.	T-3AC	Class-C Amp. (Telegraphy)	3000	−350	335	45	29	750	3-250A2 250TL
														Class-C Amp. Plate-Mod.	3000	−350	335	45	29	750	
														Class-B Amp. (Telephony)	3000	−225	125	2.0	15	125	
														Grid-Modulated Amp.	3000	−450	125	2.0	15	125	
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	GL159
														Class-C Amp. Plate-Mod.	1500	−240	400	23	9.0	450	
														Class-B Amp. (Telephony)	2000	−90	190	—	2.5	130	
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	GL169
														Class-C Amp. Plate-Mod.	1500	−100	400	45	10	450	
														Class-B Amp. (Telephony)	2000	−10	190	—	3.5	130	
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	204A 304A
														Class-C Amp. Plate-Mod.	2000	−250	250	35	20	350	
														Class-B Amp. (Telephony)	2500	−70	160	—	15	100	
308B	250	14	4.0	2250	325	75	8.0	13.6	17.4	9.3	1.5	N.	T-2A	Class-C Amplifier	3500 2000	−600 −300	300 500	60 —	— —	800 800	308B
HK454H	250	5.0	11	5000	375	85	30	4.6	3.4	1.4	100	J.	T-3AC	Class-C Amp. (Telegraphy)	1750	−400	300	—	—	350	HK454H
HK454-L	250	5.0	11	5000	375	60	12	4.6	3.4	1.4	100	J.	T-3AC	Class-C Amp. Plate-Mod.	1250	−320	300	75	—	250	HK454-L
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-B Amp. (Telephony)	1750	−230	215	—	—	125	212E 241B 312E
														Class-C Amp. (Telegraphy)	3500	−275	270	60	28	760	
														Class-C Amp. Plate-Mod.	3500	−450	270	45	30	760	
300T[1]	300	8.0	11.5	3500	350	75	16	4.0	4.0	0.6	—	J.	T-3AC	Class-C Amp. (Telegraphy)	2000	−225	300	—	—	400	300T
HK304-L	300	5/10	26/13	3000	1000	150	10	12	9.0	0.8	—	N.	T-4BF	Class-C Amp. Plate-Mod.	1500	−200	300	75	—	300	HK304-L
														Class-B Amp. (Telephony)	2000	−120	300	—	—	200	
527	300	5.5	135.0	20000[2]	—	—	38	19.0	12.0	1.4	200	N.	T-4B	Oscillator at 200 Mc.	Approximately 250 watts output						527

TABLE XV—TRIODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-patian Watts	Cathade Valts	Cathade Amps.	Max. Plate Valtage	Max. Plate Current Ma.	Max. D.C. Grid Current Ma.	Amp. Factar	Interelectrade Capacitances (μμfd.) Grid ta Fil.	Grid ta Plate	Plate to Fil.	Max. Freq. Mc. Full Ratings	Base	Sacket Connec-tians	Typical Operatian	Plate Valtage	Grid Valtage	Plate Current Ma.	D.C. Grid Current Ma.	Apprax. Grid Driving Pawer Watts	Apprax. Carrier Output Pawer Watts	Type
HK654	300	7.5	15	4000	600	100	22	6.2	5.5	1.5	20	J.	T-3AC	Class-C Amp. (Telegraphy)	2000	−380	500	75	57	720	HK654
														Class-C Amp. Plate-Mad.	2000	−365	450	110	70	655	
														Class-8 Amp. (Telephany)	3500	−137	150	13	13	210	
														Grid-Madulated 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.	T-4BF	Class-C Amplifier	1500	−125	667	115	25	700	3-300A3 304TH
3-300A2 304TL						150	12	8.5	9.1	0.6	40	N.	T-4BF	Class-C Amplifier	1500	−250	665	90	33	700	3-300A2 304TL
833A	300	10	10	3000	500	100	35	12.3	6.3	8.5	30	N.	T-1AB	Class-C Amp. (Telegraphy)	2000	−200	475	65	25	740	833A
														Class-C Amp. (Telephany)	2500	−300	335	75	30	635	
														Class-B Amp. (Telephony)	3000	− 70	150	2.0	10	150	
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	270A
														Class-C Amp. Plate-Mad.	2250	−300	300	80	—	450	
														Class-B Amp. (Telephany)	3000	−180	175	—	—	175	
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	849
														Class-C Amp. (Telephany)	2000	−300	300	30	14	425	
														Class-B Amp. (Telephany)	2500	−125	216	1.0	12	180	
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	831
														Class-C Amp. (Telephany)	3000	−500	200	60	50	360	
														Class-B Amp. (Telephony)	3500	−220	146	—	—	160	

* Cathade resistar in ahms.

[1] Discantinued.
[2] Twin triade. Values, except interelement capacities, are far bath sections in push-pull.
[3] Output at 112 Mc.
[4] Grid-leak resistar in ahms.
[5] Max. peak valts, plate pulsed.
[6] Pulse power output.
[7] Values are for two tubes.

TABLE XVI—TETRODE AND PENTODE TRANSMITTING TUBES

Type	Max. Plate Dissi-patian Watts	Cathade Valts	Cathade Amps.	Max. Plate Valt-age	Max. Screen Valt-age	Max. Screen Dissi-pation Watts	Interelectrade Capacitances (μμfd.) Grid ta Fil.	Grid ta Plate	Plate ta Fil.	Max. Freq. Mc. Full Ratings	Base	Sacket Can-nec-tians	Typical Operation	Plate Valt-age	Screen Valt-age	Sup-pressar Valt-age	Grid Valt-age	Plate Current Ma.	Screen Current Ma.	Grid Current Ma.	Screen Resistar Ohms	Apprax. Grid Driving Pawer Watts	Apprax. Carrier Output Pawer Watts	Type
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.-Oscillatar	150	135	0	− 26	18.3	6.5	0.13	2300	—	1.2	3A4
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.-Osc.	200	100	—	−22.5	20	4.0	2.0	—	0.1	3.0	HY63
													Class-C Amp. Plate-Mad.	180	100	—	− 35	15	3.0	2.0	—	0.2	2.0	
RK64[1]	6.0	6.3	0.5	400	100	3.0	10	0.4	9.0	60	M.	T-5BB	Class-C Amp. (Telegraphy)	400	100	30	− 30	35	10	3.0	—	0.18	10	RK64
													Class-C Amp. Plate-Mad.	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.-Osciliatar	400	150	—	− 50	22.5	7.0	1.5	—	0.1	5.0	1610
RK56	8.0	6.3	0.55	300	300	4.5	10	0.2	9.0	60	M.	T-5BB	Class-C Amp. (Telegraphy)	400	300	—	− 40	62	12	1.6	—	0.1	12.5	RK56
													Class-C Amp. Plate-Mad.	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.	T-7C	Class-C Amp. (Telegraphy)	500	200	45	− 90	55	38	4.0	—	0.5	22	RK23 RK25 RK25B
													Class-C Amp. (Telephany)	400	150	0	− 90	43	30	6.0	8300	0.8	13.5	
													Suppressar-Madulated 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	1613
													Class-C Amp. Plate-Mad.	275	200	—	− 35	42	10	2.8	10000	0.16	6.0	

TABLE XVI—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
2E30	10	6.0	0.7	250	250	2.5	10	0.5	4.5	160	B.	Fig. 55	Class-C Amp.-Oscillator	250	250	—	−60	55	9.0	0.8	—	0.07	7.5	2E30
6F6 6F6G	11	6.3	0.7	375	285	3.75	6.5	0.2	13	—	O.	7AC	Class-C Amp.-Oscillator	350	200	—	−35	50	10	3.5	—	0.22	9.0	6F6 6F6G
							8.0	0.5	6.5				Class-C Amp. Plate-Mod.	275	200	—	−35	42	10	2.8		0.16	6.0	
837 RK44[1]	12	12.6	0.7	500	300	8	16	0.2	10	20	M.	T-7C	Class-C Amp. (Telegraphy)	500	200	40	−70	80	15	4.0	20000	0.4	28	837 RK44
													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	
2E24	9.0	6.3[4]	0.65	500	200	2.3	8.5	0.11	6.5	125	O.	7CL	Class-C Amp. Plate-Mod.	400	180	—	−45	50	8.0	2.5	27500	0.15	13.5	2E24
														500	180	—	−45	54	8.0	2.5	40000	0.16	18.0	
	13.5			600	200	2.5							Class-C Amp. (Telegraphy)	400	200	—	−45	75	10.0	3.0	20000	0.19	20	
														600	195	—	−50	66	10	3.0	40500	0.21	27	
2E26	13.5	6.3	0.8	600	200	2.5	13	0.2	7.0	125	O.	7CK	Class-C Amp. (Telegraphy)	400	190	—	−30	75	11	3.0	19000	0.12	20	2E26
														600	185	—	−45	66	10	3.0	41500	0.17	27	
	9.0			500	200	2.3							Class-C Amp. (Plate-Mod.)	400	160	—	−50	50	7.5	2.5	32000		13.5	
														500	180	—	−50	54	9.0	2.5	35500		18	
802	13	6.3	0.9	600	250	6.0	12	0.15	8.5	30	M.	T-7C	Class-C Amp. (Telegraphy)	600	250	40	−120	55	16	2.4	22000	0.30	23	802
													Class-C Amp. Plate-Mod.	500	245	40	−40	40	15	1.5	16300	0.10	12	
													Suppressor-Modulated Amp.	600	250	−45	−100	30	24	5.0	14500	0.6	6.3	
HY6V6-GTX	13	6.3	0.5	350	225	2.5	9.5	0.7	9.5	60	O.	7AC	Class-C Amp.-Oscillator	300	200	—	−45	60	7.5	2.5	—	0.3	12	HY6V6-GTX
													Class-C Amp. Plate-Mod.	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.	T-5BB	Class-C Amp. (Telegraphy)	425	200	—	−62.5	60	8.5	3.0	—	0.3	18	HY60
													Class-C Amp. Plate-Mod.	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	HY65
													Class-C Amp. Plate-Mod.	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	2E25
													Class-C Amp. (Plate-Mod.)	400	200	—	−45	60	12	3.0	—	0.4	16	
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	306A
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	307A RK-75
													Suppressor-Modulated Amp.	500	200	−50	−35	40	20	1.5	14000	—	6.0	
832[3]	15	6.3	1.6	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	832
		12.6	0.8										Class-C Amp. (Telephony)	425	200	—	−60	52	16	2.4	14000	0.15	16	
832A[3]	15	6.3	1.6	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	832A
		12.6	0.8										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.	T-5BB	Class-C Amp. (Telegraphy)	500	175	—	−125	25	—	5.0	—	—	9.0	844
													Class-C Amp. (Telephony)	500	150	—	−100	20	—	—	—	—	4.0	
B65	15	7.5	2.0	750	175	3.0	8.5	0.1	8.0	15	M.	T-4C	Class-C Amp. (Telegraphy)	750	125	—	−80	40	—	5.5	—	1.0	16	865
													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.	7AC	Class-C Amp. (Telegraphy)	400	300	—	−55	75	10.5	5.0	9500	0.36	19.5	1619
													Class-C Amp. Plate-Mod.	325	285	—	−50	62	7.5	2.8	5000	0.18	13	
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	254A
6L6 6L6G	21	6.3	0.9	375	300	3.5	10	0.4	12		O.	7AC	Class-C Amp.-Oscillator	375	200	—	−35	88	9.0	3.5	—	0.18	17	6L6 6L6G
							11.5	0.9	9.5				Class-C Amp. Plate-Mod.	325	—	—	−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	6L6GX
													Class-C Amp. Plate-Mod.	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	HY6L6-GTX
													Class-C Amp. Plate-Mod.	400	225	—	−45	90	9.0	3.0	16000	0.8	20	

TABLE XVI—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
T21	21	6.3	0.9	400	300	3.5	13	0.7	12	30	M.	T-6B	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	25	T21
													Class-C Amp. Plate-Mod.	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.	T-6B	Class-C Amp. (Telegraphy)	400	250	—	−50	95	8.0	3.0	—	0.2	25	RK49
													Class-C Amp. (Telephony)	300	200	—	−45	60	15	5.0	6700	0.34	12	
1614	21	6.3	0.9	375	300	3.5	10	0.4	12.5	80	O.	7AC	Class-C Amp. (Telegraphy)	375	250	—	−40	80	10	2.0	12500	0.1	21	1614
													Class-C Amp. Plate-Mod.	325	—	—	−40	70	8.0	2.0	10000	0.1	15	
RK41[1] RK39	25	2.5 6.3	2.4 0.9	600	300	3.5	13	0.2	10	60	M.	T-5BB	Class-C Amp. (Telegraphy)	600	300	—	−90	93	10	3.0	—	0.38	36	RK41 RK39
													Class-C Amp. (Telephony)	475	250	—	−50	85	9.0	2.5	25000	0.2	26	
HY61/ 807	25	6.3	0.9	600	300	3.5	11	0.2	7.0	60	M.	T-5BB	Class-C Amp. (Telegraphy)	600	250	—	−50	85	9.0	4.0	39000	0.4	40	HY61/ 807
													Class-C Amp. (Telephony)	475	250	—	−50	100	9.0	3.5	25000	0.2	27	
815[3]	25	12.6 6.3	0.8 1.6	500	200	4.0	13.3	0.2	8.5	125	O.	T-8FA	Class-C Amp.-Oscillator	500	200	—	−45	150	17	2.5	—	0.13	56	815
													Class-C Amp. Plate-Mod.	400	175	—	−45	150	15	3.0	—	0.16	45	
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	254B
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	1624
													Class-C Amp. Plate-Mod.	500	275	—	−50	75	9.0	3.3	25000	0.25	24	
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	RK66
													Class-C Amp. Plate-Mod.	500	—	—	−50	75	8.0	3.2	25000	0.23	25	
807 1625	30	6.3 12.6	0.9 0.45	750	300	3.5	11	0.2	7.0	60	M.	T-5BB Fig. 29	Class-C Amp. (Telegraphy)	750	250	—	−50	100	8.0	3.0	—	0.22	50	807 1625
													Class-C Amp. Plate-Mod.	600	275	—	−90	100	6.5	4.0	—	0.4	42.5	
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	2E22
													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	3D23 TB-35
													Class-C Amp. (Plate-Mod.)	1000	300	—	−200	85	14	10	—	2.0	60	
RK20[1] RK20A RK46[1]	40	7.5 7.5 12.6	3.0 3.25 2.5	1250	300	15	14	0.01	12	—	M.	T-5C	Class-C Amp. (Telegraphy)	1250	300	45	−100	92	36	11.5	—	1.6	84	RK20 RK20A RK46
													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	HY69
													Class-C Amp. Plate-Mod.	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	
829[1,3]	40	6.3 12.6	2.25 1.12	500	225	40	14.5	0.1	7.0	200	N.	7BP	Class-C Amp. (Telegraphy)	500	200	—	−45	240	32	12	9300	0.7	83	829
													Class-C Amp. Plate-Mod.	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	829A
													Class-C Amp. Plate-Mod.	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	225	6	14.5	0.1	7.0	200	N.	7BP	Class-C Amp. (Grid Mod.)	750	200	—	−55	80	5.0	0	—	0.7	24	829B 3E29
	28			600	225	7							Class-C Amp. (Plate-Mod.)	600	200	—	−70	150	30	12.0	13300	0.9	70	
	40			750	225	7							Class-C Amp. (Telegraphy)	750	200	—	−55	160	30	12.0	18300	0.8	87	
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	HY1269
													Class-C Amp. Plate-Mod.	600	250	—	−70	100	12.5	5	35000	0.5	42	
													Grid-Modulated Amp.	750	300	—	—	80	—	—	—	—	20	
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	RK47
													Class-C Amp. Plate-Mod.	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	

TABLE XVI—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissipation Watts	Cathode Volts	Cathode Amps.	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	Approx. Carrier Output Power Watts	Type
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	312A
													Class-C Amp. Plate-Mod.	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	804
													Class-C Amp. Plate-Mod.	1250	250	50	−90	75	20	6.0	50000	0.75	65	
													Grid-Modulated Amp.	1500	300	45	−130	50	13.5	3.7	—	1.3	28	
													Suppressor-Modulated Amp	1500	300	−50	−115	50	32	7.0	—	0.95	28	
4D22 4D32	50	25.2 12.6 6.3	0.8 1.6 3.75	750	350	14	28	0.27	13	60	N.	Fig. 50	Class-C Amp. (Telegraphy)	750	300	—	−100	240	26	12	—	1.5	135	4D22 4D32
														600	300	—	−100	215	30	10	—	1.25	100	
												Fig. 51	Class-C Amp. (Plate Mod.)	600	—	—	−100	220	28	10	10000	1.25	100	
														550	—	—	−100	175	17	6	15000	0.6	70	
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	305A
													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	HY67
													Class-C Amp. Plate-Mod.	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	814
													Class-C Amp. Plate-Mod.	1250	300	—	−150	145	20	10	48000	3.2	130	
													Grid-Modulated Amp.	1500	250	—	−120	60	3.0	2.5	—	4.2	35	
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	282A
													Class-C Amp. Plate-Mod.	750	150	—	−180	100	—	50	—	—	50	
4E27/ 8001	75	5.0	7.5	4000	750	30	12	0.06	6.5	75	J.	T-7CB	Class-C Amp. (Telegraphy)	2000	750	—	−200	150	18	0.7	300000	0.2	230	4E27/ 8001
													Class-C Amp. Plate-Mod.	2000	600	60	−200	100	8	0.6	240000	0.1	200	
													Suppressor-Modulated Amp	2000	500	−300	−130	55	45	3.0	—	0.4	35	
HK257 HK257B	75	5.0	7.5	4000	500	25	13.8	0.04	6.7	75 120	J.	T-7CB	Class-C Amp. (Telegraphy)	2000	500	60	−200	150	11	6.0	—	1.4	230	HK257 HK257B
													Class-C Amp. Plate-Mod.	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.	T-5C	Class-C Amp. (Telegraphy)	1500	400	75	−100	180	28	12	40000	2.2	200	828
													Class-C Amp. Plate-Mod.	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	
RK28	100	10	5.0	2000	400	35	15	0.02	15	—	J.	T-5C	Class-C Amp. (Telegraphy)	2000	400	45	−100	150	55	13	21000	2.0	210	RK28
													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	RK48 RK48A
													Class-C Amp. (Telephony)	1500	400	—	−100	148	50	6.5	22000	1.0	165	
													Grid-Modulated Amplifier	1500	400	—	−145	77	10	1.5	—	1.6	40	
813	100	10	5.0	2000	400	22	16.3	0.2	14	30	J.	Fig. 28	Class-C Amp. (Telegraphy)	2000	400	—	−90	180	15	3.0	107000	0.5	260	813
													Class-C Amp. (Telephony)	1600	400	—	−130	150	20	6.0	60000	1.2	175	
													Grid-Modulated Amplifier	2000	400	—	−120	75	3.0	—	—	—	50	
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	850
													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	860
													Class-C Amp. Plate-Mod.	2000	220	—	−200	85	25	38	100000	17	105	

TABLE XVI—TETRODE AND PENTODE TRANSMITTING TUBES—*Continued*

Type	Max. Plate Dissi-pation Watts	Cathode Volts	Cathode Amps.	Max. Plate Volt-age	Max. Screen Volt-age	Max. Screen Dissi-pation Watts	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 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	Approx. Carrier Output Power Watts	Type
4-125A	125	5.0	6.2	3000	400	20	10.3	0.03	3.0	120	N.	Fig. 27	Class-C Amp. (Telegraphy)	3000	350	—	−150	167	30	9	—	2.5	375	4-125A
													Class-C Amp. Plate-Mod.	2500	350	—	−330	150	30	13	—	6	300	
RK28A	125	10	5.0	2000	400	35	15	0.02	15	—	J.	T-5C	Class-C Amp. (Telegraphy)	2000	400	45	−100	170	60	10	—	1.6	250	RK28A
													Class-C Amp. Plate-Mod.	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.	T-5C	Class-C Amp. (Telegraphy)	2000	500	40	− 90	160	45	12	—	2.0	210	803
													Class-C Amp. (Telephony)	1600	500	100	− 80	150	20	4.0	20000	4.0	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	
AT-340	150	5	7.0	4000	400	—	9.04	0.19	4.16	120	J.	Fig. 27	Class-C Amp.-Oscillator	3000	400	—	−500	165	75	—	—	2.4	—	AT-340
RK65	215	5.0	14	3000	500	35	10.5	0.24	4.75	60	J.	T-38C	Class-C Amp. (Telegraphy)	3000	400	—	−100	240	70	24	—	6.0	510	RK65
													Class-C (Plate & Screen Mod.)	2500	—	—	−150	200	70	22	30000	6.3	380	
4-250A	250	5.0	14.5	4000	600	50	12.7	0.06	4.5	85	N.	Fig. 27	Class-C Amp. (Telegraphy)	4000	500	—	−250	250	22	13	—	4.1	750	4-250A
														2500	500	—	−100	325	70	22	—	3.7	562	
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	861
													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.

[4] Terminals 3 and 6 must be connected together.

[5] Filament limited to intermittent operation.

Chapter Twenty-One

Radio Operating

The object of most radio communication is the transmission of intelligence from one point to another, accurately and in as short a time as possible. For efficiency in communication, each class of radio service has set up operating methods and procedure best suited to its needs. Operators should not only be expert in transmitting and receiving code or voice signals, but thoroughly familiar with the uniform practices of their service.

ℭ Memorizing the Code

One of the amateur operator-license requirements covers ability to send and receive Continental (International Morse) code at the rate of 13 words per minute.

The serious student of code — sending, receiving, operating practices, copying on the typewriter, etc. — would be best advised to purchase a copy of the ARRL booklet, *Learning the Radiotelegraph Code* (price, 25 cents, postpaid).

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

Period: didahdidahdidah. Comma: dahdahdididahdah. Question mark: dididahdahdidi. Error: didididididididi. Double dash: dahdididi-dah. Wait: didahdididi. End of message: didahdidahdi. Invitation to transmit: dahdidah. End of work: didididahdidah. Fraction bar: dahdididahdi.

Fig. 2101 — The Continental (International Morse) code.

The first step is to *memorize* the code. The complete Continental alphabet is shown in the table of Fig. 2101. All of the characters should be learned, starting with letters and going on to numerals and punctuation marks. Take a few at a time. Review at intervals all the letters learned up to that time.

Think of the letters in terms of *sound* rather than their appearance as actual dot-and-dash combinations. Think of A as the sound "di*dah*" — not as a "dot-dash." Make the sound "di" staccato, allowing stress to fall equally on every "*dah*." There should never be a space or hesitation between "dits" and "dahs" of the same letter.

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 every character to the point where you can recognize each of them without hesitation. Concentrate on any difficult letters until they become as familiar as the rest.

ℭ Acquiring Speed by Buzzer Practice

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

Either the buzzer set shown in Figs. 2102 and 2103 or the audio oscillator described will give satisfactory results as a practice set.

The battery-operated audio oscillator in Figs. 2104 and 2105 is easy to construct and is effective. If nothing is heard in the headphones when the key is depressed, reverse the leads going to *either* transformer winding (do not reverse both windings).

With a practice set ready, send single letters at first. When each character can be read quickly follow this by slow sending of complete words and sentences. Have the material sent at just a little faster rate than you can copy easily; this speeds up your mind. *Write* down each letter you recognize. Do *not* try to write

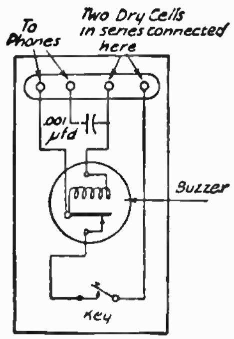


Fig. 2102 — The headphones are connected across the coils of the buzzer, with a condenser in series. If the value shown gives an excessively loud signal, it may be reduced to 470 μμfd. or 220 μμfd.

down the dots and dashes; write down letters. Don't stop to compare the sounds of different letters, or think too long about a letter or word that has been missed. Go right on to the next one, or each "miss" will cause you to lose several characters. If you exercise a little patience you will soon be getting every character. When you can receive 13 words a minute (65 letters a minute), have the sender transmit code groups rather than English text. This will prevent you from recognizing a word "on the way" and filling it in before you've really listened to the letters themselves.

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

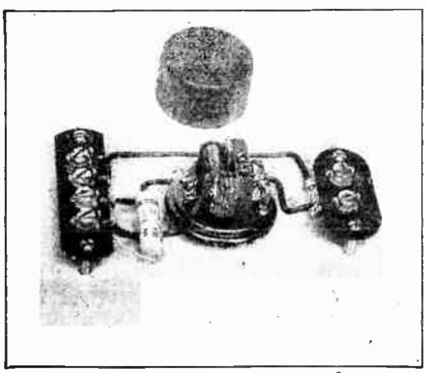


Fig. 2103 — The cover of the buzzer unit has been removed in this view of the buzzer code-practice set.

Learning by Listening

W1AW conducts practice transmissions nightly Monday through Friday at speeds of 15, 20, 25, 30 and 35 w.p.m. Such practice tapes start at ten P.M. EST (EDST in summer). In addition, the Official Bulletins, also sent from W1AW, give added practice at 15 and 25 w.p.m. See the Operating News section announcements of W1AW Operating Schedule and Code Proficiency Practice notes, in the latest copy of *QST*. Practise until you can send in what you have copied over the air on W1AW's monthly "qualifying run" to get a 15-word-per-minute Code Proficiency Certificate or a sticker for advanced speeds. As soon as you can, listen on a real communications receiver (with beat oscillator) and have the fun of learning by listening.

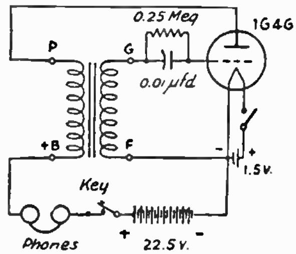


Fig. 2104 — Wiring diagram of a simple vacuum-tube audio-frequency oscillator for use as a code-practice set.

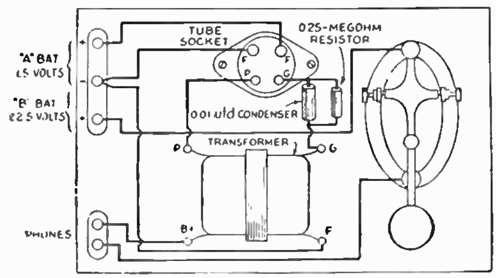


Fig. 2105 — Layout of the audio-oscillator code-practice set. All parts may be mounted on a wooden baseboard, approximately 5 × 7 inches in size.

Using a Key

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

Do not hold the key tightly. Let the hand rest lightly on the key. The thumb should be against the left side of the knob. The first and second fingers should be bent a little. They should hold the middle and right sides of the knob, respectively. The fingers are partly on top and partly over the side of the knob. The other two fingers should be free of the key. Fig. 2106 shows the correct way to hold a key.

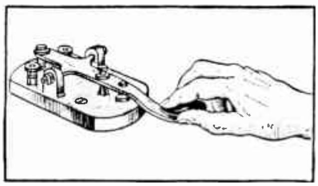


Fig. 2106 — This sketch illustrates the correct position of the hand and fingers for good sending with a telegraph key.

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

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

General Procedure

Calling — The call signal of the calling station must be inserted at frequent intervals for identification purposes. Repeating the call signal of the called station five times and signing not more than twice (this repeated not more than five times) has proved excellent for telegraph or voice practice (the receiver being kept tuned to the frequency of the called station). The use of a break-in system (c.w.) or push-to-talk (voice) is highly recommended to save time and reduce unnecessary interference to a minimum. Example:

W6EY W6EY W6EY W6EY W6EY DE W1AW W1AW.

Stations desiring communication with any station may use the signal of inquiry, CQ, in place of the call signal of the station called. The general inquiry call (CQ) should be sent not more than five times without interspersing one's station identification, and the length of repeated calls is carefully limited in intelligent amateur operating. Too many insertions of one's own call in a CQ will decrease its effectiveness. CQ is not to be used when testing or when the sender is not expecting or looking for an answer. After a CQ the dial should be covered thoroughly for two or three minutes looking for replies. For voice work "Calling any amateur station" is considered superior to CQ, one of the attributes of voice operation being the ability to "say it with words."

FCC regulations require all amateur operators to send the call of the station called or worked and their own call at the beginning and end of each transmission, and in any event at least once each ten minutes during long transmissions. Where break-in is used and exchanges of sequences of 3 minutes or less are taking place, the calls are required (additional to beginning and end) only each ten minutes. "This is" or "from" must be used by voice stations in place of "DE." Portables and mobiles must give their geographical designation after their calls.

Answering a call — The above example, when replying to a call, may be cut down to three (or less) calls, DE, and one or two repetitions of your own call, with further reduction to a one-times-one call when conditions permit during communication. Example:

WØEFC DE W1AW GE OM K (good evening, old man, go ahead.)

Ending signals — After a CQ, a transmission should end with K (invitation to transmit):

CQ CQ (etc.) DE W7BG W7BG K.

After a call to a specific station (contact not yet established) use $\overline{AR}$:

VE3CAR VE3CAR (etc.) DE W1BDI $\overline{AR}$.

At the end of each transmission during QSO use K:

. . . W5BMI DE W6RBQ K.

At the conclusion of a QSO use $\overline{VA}$ or $\overline{SK}$:

. . . Tnx data ur rig 73 $\overline{VA}$ W1AW DE W4IP.

If closing station, add CL.

Voice calls — An initial voice call may be made as follows: "Calling any amateur station, this is W 6 BAKER KING YOUNG in Whittier, California. Go ahead."

W1LVQ calls W6BKY: "W6BKY, this is W 1 LEWIS VICTOR QUEEN in Hartford, Connecticut. Go ahead."

W6BKY answers W1LVQ: "W1LVQ from W6BKY" (proceeds with contact).

During the contact as above, transmissions may be ended: "W1LVQ from W6BKY, over."

In concluding a contact: "W1LVQ, this is W 6 BAKER KING YOUNG in Whittier, California, signing off."

If W6BKY is closing his station, he concludes: ". . . signing off and closing station."

Tuning procedure after CQs — The use of special abbreviations after a CQ call to indicate from what part of the band tuning will start is a valuable aid to the receiving operator in determining frequency to use and how long to call. ARRL recommends the following abbreviations for this purpose:

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 should include: By c.w. — CQ DE W6RBQ MH K. By voice — Simply use the words for which the abbreviation MH stands.

Directional CQs — If interested in a particular direction or locality for a contact or message relay, so indicate in your call. A CQ call must be long enough to attract one or more operators, but not long enough to cause listeners to tire and tune away from your signal. Examples: CQ W5, CQ DALLAS, CQ WEST.

ℂ Voice and Telegraph Operation

Radiotelegraph code is used for reliable accurate communication of intelligence, even at great distances. The good operator is noted for his neatness and accuracy of copy. It is desirable to copy exactly what is sent. If there is any doubt about a letter or word one should query the transmitting operator. Never send R (for OK) until *all that has been sent* is successfully received (copied down or understood).

Procedure in telegraph and in radiotelephone operation is similar. However, in voice work the operator makes little use of the special abbreviations available for code work, of course, since he may directly speak out their full meaning. Radiotelephony is used by other services mainly for discussion or command-control purposes. Telegraph operation is generally preferred for message work and extreme DX under difficult conditions.

Repeats — When a few word-groups in conversation or message handling have been missed, a selection of one or more of the following abbreviations are used to ask for a repeat on the parts in doubt:

Abbreviation	*Meaning*
?AA..........	Repeat all after....................
?AB..........	Repeat all before....................
?AL..........	Repeat all that has been sent........
?BN...AND...	Repeat all between. . .and..........
?WA..........	Repeat the word after...............
?WB	Repeat the word before..............

The good operator will ask only for what fills are needed, separating different requests for repetition by using the break sign or double dash (— · · · —) between these parts. There is seldom any excuse for repeating a whole message just to get a few lost words.

Another interrogation method is sometimes used, the question signal (· · — — · ·) being sent between the last word received correctly and the first word (or first few words) received after the interruption.

Unusual words should be avoided, in the interest of accuracy, when drafting messages. When they unavoidably turn up difficult words may be repeated, or *repeated and spelled*. The operator says "I will repeat" or "I say again" when thus retransmitting a difficult word or expression.

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 sta-

FOR RADIOTELEPHONE

As a service to all amateurs, the ARRL Word List printed herewith has been devised. A phonetic alphabet or special word list is recommended for use as needed in identifying station calls or difficult words.

The list helps to avoid facetious word combinations. This gives it greatest acceptability to all amateurs.

Use of this standard list is recommended by ARRL. Haphazard selection of words often results in confusion. A degree of uniformity in use of phonetic words reflects favorably on your individual operating, and on the whole amateur service.

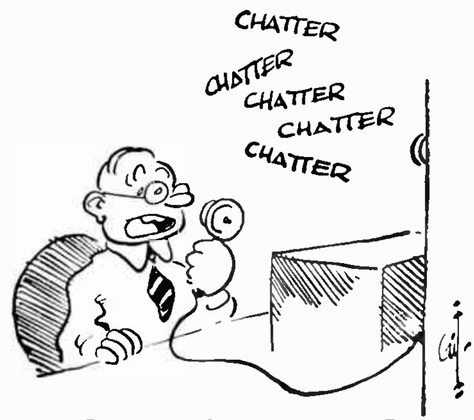
A — ADAM	N — NANCY
B — BAKER	O — OTTO
C — CHARLIE	P — PETER
D — DAVID	Q — QUEEN
E — EDWARD	R — ROBERT
F — FRANK	S — SUSAN
G — GEORGE	T — THOMAS
H — HENRY	U — UNION
I — IDA	V — VICTOR
J — JOHN	W — WILLIAM
K — KING	X — X-RAY
L — LEWIS	Y — YOUNG
M — MARY	Z — ZEBRA

Example: W1EH . . . W 1 EDWARD HENRY.

It is recommended that use of Q-code and special abbreviations be minimized in voice work insofar as possible, and the full expression (with conciseness) be substituted.

tions should use a *standard* list as needed to identify call signals or unfamiliar expressions.

Using a microphone — Many of the principles for getting operating results are similar to those set down for key operation. However, the ability to phrase clearly and concisely counts. Good push-to-talk technique differs considerably from broadcasting. Where possible, controls or on-off switches should be arranged to permit fast back-and-forth exchanges. This will help to reduce the length of transmissions, enable us to note quickly when interference comes on a frequency, and will keep brother amateurs from calling us a "monologuist" — an individual who likes to monopolize a channel and hear himself talk!



USE PUSH-TO-TALK AND AVOID BEING CALLED A MONOLOGUIST

Here is a short tabulation of the points of good result-getting technique:

1) **Listen much . . . with care.** Avoid distractions in your operating room. Tune the band well after each call.

2) **Time your calls: monitor your own frequency.** Call only when a station is free.

3) **Make short calls, with breaks to listen.** Speak clearly, at a steady, modest rate. Three short calls are better than one long one.

4) **Use push-to-talk technique . . . speak near the microphone.** Watch the modulation indicator. Keep local background noise at a minimum.

5) **Make notes. Avoid missing points for comment.** Jot down topics to avoid repeats.

6) **Talk in connected thoughts and phrases.** Notes will help avoid mixing up subjects. Push-to-talk technique will keep brother amateurs from calling you a monologuist.

7) **Speak naturally.** QSOs need not be cut and dried. Make them interesting. Avoid exhibitionism. Use proper operating form to promote efficiency in communication and add respect and prestige for your station.

Voice equivalents to code procedure — "Go ahead" or "Over" (K) indicates receipt or further transmission is expected.

Wait, stand by (AS-QRX).

Okay (R) indicates receipt for a correctly-transcribed message, or that transmission was received "solid" with no missing portions.

Make transmissions through twice (QSZ). Repeat each word twice.

All After . . . (AA). Repeat all after . . . word).

All Before . . . (AB). Repeat all before . . . (word).

Repeat BetweeN . . . and . . . (BN). Repeat between . . . and . . . (words).

Message handling — Each service — commercial, military, amateur — prescribes a message form, but all are generally similar. A message is broadly divided into four parts: (1) the preamble; (2) the address; (3) the text; (4) the signature. The preamble of all amateur radiograms includes:

a) Number (of this message).
b) Station of origin.
c) Check (number of words in text).
d) Place of origin.
e) Time filed.
f) Date.

Therefore, it might look like this:

NR 34 W9AND 13 CHICAGO ILL 450 PM MAY 12
WILLIAM MONTGOMERY
2159 ROXY ST NW
WASHINGTON DC $\overline{\text{BT}}$
LOCAL EMERGENCY COORDINATORS HAVE 123 MEN AVAILABLE FOR ACTIVE DUTY CASE OF EMERGENCY $\overline{\text{BT}}$
BLAKE

This is obviously the 34th message (of that day or that month, as the policy of the station prescribes) from station W9AND. The check is 13. The signal $\overline{\text{BT}}$ (double dash) is used to separate the text from address and signature.

Several radiograms may be transmitted in series (QSG . . .) with the consent of the station which is to receive them. As a general rule long radiograms should be transmitted in sections of approximately fifty words, each ending with · · — — · · (?), meaning, "Have you received the message correctly thus far?"

If the first part of a message is received but substantially all of the latter portions lost, the request for the missing parts is simply RPT TXT AND SIG, meaning, "Repeat text and signature." PBL and ADR may be used similarly for the preamble and address. RPT ALL or RPT MSG should not be sent unless nearly all of the message is lost.

The service message — When one station has a message to transmit to another concerning the handling of a previous message, the message is titled "service" and is indicated by "SVC" in the preamble when sent. Such a message may refer to nondelivery, delayed transmission, errors, or to any phase of message-handling activity. Words may be abbreviated in the text of the service message to conserve time. Do not abbreviate to the point where misunderstanding may arise.

Land-line check — The land-line or "text" count, consisting of count only of the words in the body or text of the message, is probably now most widely used. (The "cable" count covers all words in the address and signature, as well, probably accounting for its unpopularity.) When in the case of a few exceptions to the basic rule in land-line checking, certain words in the address, signature or preamble are counted, they are known as extra words and all such are so designated in the check right after the total number of words.

The check includes:

1) All words, figures and letters in the body, and,

2) the following extra words:

a) Signature except the first, when there are more than one (a title with signature does not count extra, but an address following a signature does).

b) Words "report delivery," or "rush" in the check.

c) Alternate names and/or street address, and such extras as "personal" or "attention."

Dictionary words in most languages count as one word irrespective of length of the word. In counting figures, a group of five digits or less counts as one word. Bars of division and decimal points may constitute one or more of the digits in such a group. It is recommended that, where feasible, words be substituted for figures to reduce the possibility of error in transmission.

Net Operation

Amateurs can add much experience and pleasure to their amateur lives, and substance and accomplishment to the credit of *all* amateur radio, when organized into effective interconnection of the cities and towns of a state.

The selection of suitable stations to be invited to work together is important. Operating ability is required. All individuals must be willing to contribute unselfishly to the success of the group objectives, permitting operations to be guided absolutely by the word of the NCS (Net Control Station).

"Break-in" is advantageously employed here — the receiver is kept running during transmissions, so that nearly-simultaneous two-way communication is possible.

Briefly, the procedure in net operation is as follows: The NCS calls the net together at a preannounced time and using a predetermined call. Immediately, station members of the net reply in alphabetical (or some other predetermined) order, reporting on the NCS's signal strength and stating what traffic is on hand, and for whom. The NCS acknowledges, meanwhile keeping an account of all traffic on hand, by stations. He then directs the transfer of messages from one station to another, giving preference to any urgent traffic so indicated at roll call. When all traffic has been distributed and it is apparent there is no further business the NCS will close the net, in most cases maintaining watch on the net frequency for any special traffic which might appear. In general the operation of all net stations is conducted for highest efficiency, on the same, or on closely-adjacent frequencies.

Keeping a log — FCC regulations require nearly every radio-communication station to keep a complete operating record or "log," including such data as times and dates of transmissions, stations contacted, message traffic handled, input power to the transmitter, frequency used, and signature or "sine" of the operator in charge.

Secrecy of correspondence — Provisions in the Communications Act make it a misdemeanor (with heavy penalties) to give out information of any sort to any person except the addressee of a message or his authorized agents. Remember that any addressed point-to-point communication (call-to-call) is covered by the law. Only when sent after a CQ call or *QST* (to all amateurs) can a conversation or message be used or divulged without the express consent of the originator or recipient.

Time Systems

While many telegraph and radio circuits use local standard (or daylight) time in log-keeping and message-handling, international radiocommunication stations and the military services follow the 24-hour system of time-keeping. Greenwich Civil Time (24-hour system) is based on the time in Greenwich, England, the city at the 0° meridian. Midnight in Greenwich is represented by 0000; 0600 represents 6 A.M. there; 1200 is noon; 1800 is 6 P.M.; 2400 is again midnight and the same as 0000 of the next day. The figures must be corrected to each individual time zone. Eastern Standard Time is five hours behind Greenwich, so that 0630 GCT (6:30 A.M. in Greenwich) would represent 1:30 A.M. EST, for example. As an example of reverse translation, 9:30 A.M. EST would be designated in the log as 1430 GCT. EDST is four hours behind GCT; MDST, six hours; PDST, seven.

The military services use simply a 24-hour clock, based on local time, without correcting to Greenwich or any other longitude. The principal advantage of this system is the elimination of the necessity for the use of P.M. or A.M. abbreviations. Each 15° zone of longitude around the globe is designated by a letter which is sent in messages with the numerals giving the time.

ARRL Operating Organization

The purpose of station-building is to communicate. To assist amateurs to get the most from their communication by amateur radio ARRL maintains a Communications Department with 70 territorial Sections (U.S.A., P.I., Cuba, Canada). A member-elected Section Manager administers appointments and handles correspondence and activity reports (published monthly in *QST*) from the active reporting stations in each Section.

All posts in the organization are dedicated to fulfilling certain specified objectives. A high standard of operation, telegraph or voice, is called for in each "station" appointment. In certain of the activities or station tests, results may be achieved in a week-end or two of operation that are the equivalent of "months" of average amateur work. *Organization* permits superior results through the mutual coöperation and collaboration of each member of a group. Our ARRL is a mutual-benefit association for the representation of the amateur, striving to add in every way to the effectiveness of the individual station and to increase the pleasure and profit of the member in his hobby.

The following abbreviated descriptions indicate the types of ARRL-SCM appointments that are made with the purpose of each. See page 6 in any *QST* for the address of your Section Communications Manager. Every reader of these pages is cordially invited to report his station activities to his SCM for *QST* mention. All who meet the qualifications and will assist in the objectives set down in the ARRL Constitution, and the book *Operating an Amateur Radio Station* are urged to secure appropriate forms for appointments from the SCM or ARRL Headquarters and to fully participate in their operating organization.

Leadership and station appointments—SEC (Section Emergency Coördinator). Promotes and administers Section emergency radio organization.

EC (Emergency Coördinator). Organizes amateurs of a community or other area for radio emergency service; liaison with officials of agencies served and with representatives of other communication facilities locally.

ORS (Official Relay Station). Traffic service, operates nets and trunk lines.

OPS (Official 'Phone Station). Voice-operating, assists in establishing high operating standards.

OES (Official Experimental Station). Experimental operating, collects reports on v.h.f.-u.h.f.-s.h.f. propagation data or contacts; some engage in fax, f.m., tv., etc. experiments.

OBS (Official Broadcasting Station). Transmits ARRL Bulletins to amateurs.

OO (Official Observer). Sends mail (or radios) coöperative notices to amateurs to assist in frequency observance, insure high-quality signals, and prevent FCC trouble for the individual or the fraternity.

RM (Route Manager). Organizes traffic nets and coördinates schedules.

PAM ('Phone Activities Manager). Organizes activities for OPS.

RST System of Signal Reports

The RST system is an abbreviated method of indicating the main characteristics of a received signal, the Readability, Signal Strength, and Tone. The letters RST determine the order of sending the report. In asking for this form of report, one transmits RST? or simply QRK?

READABILITY

1 — Unreadable
2 — Barely readable, occasional words distinguishable
3 — Readable with considerable difficulty
4 — Readable with practically no difficulty
5 — Perfectly readable

SIGNAL STRENGTH

1 — Faint — signals barely perceptible
2 — Very weak signals
3 — Weak signals
4 — Fair signals
5 — Fairly good signals
6 — Good signals
7 — Moderately strong signals
8 — Strong signals
9 — Extremely strong signals

TONE

1 — Extremely-rough hissing note
2 — Very rough a.c. note, no trace of musicality
3 — Rough low-pitched a.c. note, slightly musical
4 — Rather rough a.c. note, moderately musical
5 — Musically-modulated note
6 — Modulated note, slight trace of whistle
7 — Near d.c. note, smooth ripple
8 — Good d.c. note, just a trace of ripple
9 — Purest d.c. note

(If the note appears to be crystal-controlled simply add an X after the appropriate number.)

If there is evidence of a chirp, the letter C may be added to so indicate.

Examples

By Telegraph: RST 359; RST 567X; RST 198C. The letters RST need not be sent, if it is clearly understood that the RST System is being used.
By Voice: Say simply, "I am receiving you Readability . . . (1–5), Strength . . . (1–9)."

Emergency Operating

One of the most interesting and practical fields for the active amateur, adding to his enjoyment as well as his prestige and record for successful and constructive communication, is that of emergency operating work. Before World War II individual amateurs and groups had scores of recorded instances of participation, handling information of critical value by *amateur radio* in sudden emergencies resulting from hurricane, flood, earthquakes, blizzards and other natural and man-made disasters that severed wire communication and transportation.

Following World War II, the FCC reopened amateur facilities in a limited manner just one week after V-J Day, to permit the reactivation of the ARRL Emergency Corps and the restoration of the widespread amateur radio capabilities to help local communities and the nation through the wide geographical availability of amateur stations. Even if amateurs do not find radio drills and activities pertinent to emergency preparedness (on 144 Mc. and every low-frequency band) of the greatest interest and pleasure, amateurs should wish to participate in AEC organization and planning in order to continue such FCC approbation and action in their behalf! So every reader who is an amateur is urged to subscribe to the Emergency Corps and participate in every local and national activity in any manner related to emergency preparation!

A communications emergency occurs whenever normal facilities are interrupted or overloaded, and may or may not involve general public participation or require FCC recognition or declaration. A communications emergency need not involve a public relief or welfare emergency, but the latter condition usually is accompanied by a communications emergency.

Relief problems of the community at large, official messages from Red Cross, military and civic officials, have absolute priority in emergency. Radio circuits must carry the important messages first, and when personal-safety messages are permissible, in the judgment of operators in the affected area, it is even then much more profitable to carry the burden of traffic and outgoing messages of safety, rather than requests for investigating safety which cannot be acted upon except at a deferred date.

When FCC declares a condition of general communications emergency, special amateur regulations (§12.156) govern absolutely, with the following provisions effective until the Commission declares the emergency ended:

1) No transmissions in the 80-meter band may be made *except* those relating to the relief or emergency service. Casual conversation, incidental calling or testing, remarks not pertinent to the constructive handling of the emergency communications, shall be prohibited.

2) Band-edge segments of 25 kc. shall be reserved at all times for (a) emergency calling channels, (b) initial calls from the isolated, (c) first calls initiating dispatch of important priority relief matters. All stations shall, for general communication, shift to other within-band frequencies for carrying on communications.

3) Hourly observance of mandatory quiet or listening periods, the first five minutes of each hour. (No calls may be answered in this period. Only "utmost priority" traffic may continue.)

4) For promulgating the emergency-declaration, and for policing-warning-observing work, FCC may designate certain amateur stations. Announcements from these stations will be identified by their reference to §12.156 by number, and their specification of the date of the FCC's declaration, with statement of the area and nature of the emergency.

Emergency calling frequencies — Regarding QRR, which call is limited to use of isolated stations for first emergency calls, special provisions and methods are necessary to assist the stations under handicap of no commercial power, in remote sections, in getting contact and help.

It is recommended by ARRL that frequencies *at the band edges* be utilized for emergency calls when no general emergency is declared or in effect. This lends point and specification to builders of emergency equipment. This spot on all bands is well covered continuously by receivers. It gives hope to the isolated operator that he be heard. All listeners are instructed to hunt for weak signals on such frequencies, during general emergency, for taking account of the isolated and establishing new important connections.

AMERICAN RADIO RELAY LEAGUE

EMERGENCY CORPS

FOR PUBLIC SERVICE

This Certifies that John J. Doe

is a Full member of the ARRL Emergency Corps for one year from date below or endorsement on reverse side.

In the event of failure of regular communication facilities due to storms, floods, and similar disasters, this operator offers the use of his amateur radio station and services to his country and community.

He will coöperate closely in Emergency Corps activities, such as plans for rendering emergency communications service, and will participate as possible in appropriate preparedness drills and tests.

Dated Jan. 1, 1946

R. J. Clark — A.R.R.L. Emergency Coördinator F. E. Handy — Comms. Mgr. A.R.R.L.

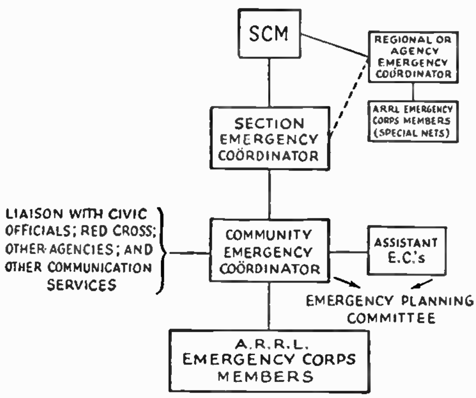
EMERGENCY CORPS MEMBERSHIP CARD
Have You Got Yours?

ARRL Emergency Corps — The ARRL Emergency Corps (AEC) is dedicated to organization of the amateur radio service for top performance in supplying emergency radio communication whenever and wherever needed. The Emergency Corps has been organized and strengthened to insure maximum effectiveness at the same time it provides operating enjoyment for its members.

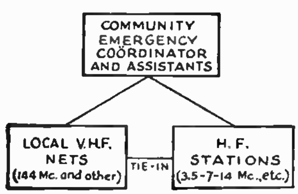
Emphasis is on radio activity and simulated emergency nets. The organization chart and radio functional diagram will help you to understand the operation of the Corps. V.h.f. is the accepted medium for local emergency communication. The 144-Mc. band is recommended for local nets where practicable. H.f.-band stations will be recruited for long-haul emergency requirements. Drills and simulated emergency work are the aim in each community. Activity in these will be required to keep in the *Full Membership* group.

Here is an official activity in which you, as an amateur, will want to participate. If you have an operative station on 144-Mc. or other amateur frequencies, aim to join the ARRL Emergency Corps. Work closely with the Emergency Coördinator (wherever appointed) and the SCM.

Why you should join — Amateur radio must carry forward its rôle of furnishing emergency communications. Disaster *can* and *does* strike where least expected! To cope with emergency problems wherever they arise, the support of amateurs throughout the nation is required. Public service in emergencies is part of the tradition of amateur radio, and substantial justification for the frequency assignments granted by our government. The ARRL Emergency Corps is an important activity.



ORGANIZATION CHART



RADIO FUNCTIONAL DIAGRAM

How to join — Application forms are available from your local EC, the local ARRL-affiliated club, your SCM or from League headquarters. One of these forms properly filled out and returned to the address indicated thereon, entitles you to receive a card certifying membership in the Emergency Corps. You will then be included in plans for on-the-air tests, drills, and other interesting activities. Join now! A postal will bring you the application form.

◖ Operating Activities

Operating in the amateur bands offers many thrills. The "unexpected" is always around the corner. Special activities are sponsored by the American Radio Relay League, adding to ham interest and fraternalism.

Within the ARRL field organization there are all-season and quarterly activities. The first Saturday night each month is set aside for all ARRL officials, officers, and directors to get together over the air from their own stations, wherever located. The 3.5-Mc. band is used and this first Saturday night is known to the gang as ARRL Officials Nite.

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 shoulder the communication load as emergencies turn up and the occasion requires. A premium is placed on the use of equipment without connection to commercial sources of power.

The Worked All States (WAS) award is made available by ARRL to all amateurs who have confirmed evidence of contacts with all states from one location — as one example of available certificate awards. A DX "Century Club" certificate likewise is given to all amateurs proving contact with 100 countries in a like manner. Code Proficiency Certificates are available for submitted copy of aural reception at 15 to 35 words per minute, provided bona fide "copy" of monthly qualifying runs checks.

Progress in proficiency of code reception is shown after the initial test and the ARRL certificate award by a separate dated-and-initialed endorsement. This is arranged for display on the certificate. Every licensee is invited to go "all out" for our awards by sending in copy transcribed by his personal efforts on one of the qualifying runs. See the latest issue of *QST* for the current schedule of W1AW Qualifying Runs. Get your certificate . . . then the progress awards!

Follow *QST* each month for current announcements of special simulated-emergency tests, concerning ARRL Trunk Line operation, A-1 Operator Club, Rag Chewers Club, Old Timers Club, Field Day, International DX and All-Section Sweepstakes competitions, and others.

The booklet *Operating an Amateur Radio Station* is sent gratis on request to League members, and covers the rules for different ARRL Awards as well as the several leadership and station appointments granted amateurs-members of the League who are conducting particular types of services in an exemplary manner, to assist brother amateurs or build the ability of amateur radio to serve the community and the nation. This 19-page book deals consecutively with Operating Practice, Emergency Communication, Operating Activities, ARRL Field Organization, Leadership Appointments, Station Appointments, Handling Messages, Network Organizing and NCS Duties, Abbreviations, and FCC Regulations, Orders and Miscellany. If you are a League member mail a card for your free copy today.

INTERNATIONAL AMATEUR PREFIXES

To make possible identification of calls heard on the air, the international telecommunications conferences assign to each nation certain alphabetical blocks, from which *all* classes of stations are assigned prefixes. The following prefixes are used by amateurs.

C	China (used unofficially)
CE	Chile
CM–CO	Cuba
CN	French Morocco
CP	Bolivia
CR	Portuguese colonies: 4, Cape Verde Ids.; 5, Port. Guinea; 6, Angola; 7, Mozambique; 8, Port. India; 9, Macao; 10, Timor.
CT	Portugal: 1, Portugal proper; 2, Azores Ids.; 3, Madeira Ids.
CX	Uruguay
D	Germany
EA	Spain and colonies: 1, 2, 3, 4, 5, 7, Spain proper; 6, Balearic Ids.; 8, Canary Ids.; 9, Span. Morocco & No. Africa.
EI	Eire
EL	Liberia
EP	Iran (Persia)
ES	Estonia
F	France and colonies: F3, F8, France proper; FA, Algeria; FB8, Madagascar; FD8, Togo; FE8, Cameroons; FF8, Fr. West Africa; FG8, Guadeloupe; FI8, Fr. Indo-China; FK8, New Caledonia; FL8, Fr. Somaliland; FM8, Martinique; FN8, Fr. India; FO8, Fr. Oceania; FP8, St. Pierre & Miquelon; FQ8, Fr. Equatorial Africa; FR8, Reunion Ids.; FT4, Tunisia; FU8, New Hebrides; FY8, Fr. Guiana & Inini.
G	Great Britain except: GI, Northern Ireland; GM, Scotland; GW, Wales.
HA	Hungary
HB	Switzerland
HC	Ecuador
HH	Haiti
HI	Dominican Republic
HJ–HK	Colombia
HP	Republic of Panama
HR	Honduras
HS	Siam
HZ	Hedjaz
I	Italy
J	Japan
K	Continental United States of America
KA	Philippine Ids.
KB–KZ	Territories and possessions of the U.S.: KB6, Baker, Howland, American Phoenix Ids.; KG6, Guam; KH6, Hawaii; KJ6, Johnston Island; KL7, Alaska; KM6, Midway Islands; KP4, Puerto Rico; KP6, Palmyra Group, Jarvis Id.; KS6, American Samoa; KV4, Virgin Islands; KW6, Wake Group; KZ5, Canal Zone (Army).
LA	Norway
LU	Argentina
LX	Luxembourg
LY	Lithuania
LZ	Bulgaria
MX	Manchuria
NY	U.S. Navy yards: NY1–2, Canal Zone; NY4, Guantanamo, Cuba.
OA	Peru
OH	Finland
OK	Czechoslovakia
ON	Belgium
OQ	Belgian Congo
OX	Greenland
OY	The Faeroes
OZ	Denmark
PA	Netherlands
PJ	Curacao
PK	Netherlands Indies: 1, 2, 3, Java; 4, Sumatra; 5, Dutch Borneo; 6, Celebes-New Guinea.
PX	Andorra
PY	Brazil
PZ	Surinam (Neth. Guiana)
SM	Sweden
SP	Poland
ST–SU	Egypt: ST, Egyptian Sudan; SU, Egypt proper
SV–SX	Greece
TA	Turkey
TF	Iceland
TG	Guatemala
TI	Costa Rica
U–UC	Union of Socialistic Soviet Republics: 1–7, European; 8, 9, 0, Asiatic. UB, Ukraine; UC, White Russian.
VE	Canada
VK	Australia: 2, 3, 5, 6, 8, Aust. proper; 4, Papua Terr.; 7, Tasmania; 9, New Guinea Terr.
VO	Newfoundland and Labrador
VP to VS	British colonies and protectorates: VP1, Brit. Honduras; 2, Leeward & Windward Ids.; 3, Brit. Guiana; 4, Trinidad and Tobago; 5, Jamaica and Cayman Ids.; 6, Barbados; 7, Bahamas; 8, Falkland Ids.; 9, Bermuda; VQ1, Zanzibar; 2, Northern Rhodesia; 3, Tanganyika; 4, Kenya; 5, Uganda; 6, Brit. Somaliland; 8, Mauritius and Chagos; 9, Seychelles; VR1, Gilbert & Ellice Ids. and Ocean Id.; 2, Fiji Ids.; 3, Fanning Id.; 4, Solomon ids; 5, Tonga (Friendly) Ids.; 6, Pitcairn Id.; VS1, Straits Settlements; 2, Federated Malay States; 3, Non-federated Malay States; 4, Brit. North Borneo; 5, Sarawak; 6, Hongkong; 7, Ceylon; 8, Bahrein Id.; 9, Maldive Ids.
VU	British India
W	Continental United States of America
XE	Mexico
XU	China
XZ	Burma
YA	Afghanistan
YI	Iraq
YL	Latvia
YM	Free City of Danzig
YN	Nicaragua
YR	Roumania
YS	El Salvador
YT–YU	Yugoslavia
YV	Venezuela
ZA	Albania
ZB to ZJ	British colonies and protectorates: ZB1, Malta; 2, Gibraltar; ZC1, Transjordania; 2, Cocos Ids.; 3, Christmas Id.; 4, Cyprus; 6, Palestine; ZD1, Sierra Leone; 2, British Cameroons, Nigeria; 3, Gambia; 4, Gold Coast (Brit. Togoland); 6, Nyasaland; 7, St. Helena; 8, Ascension Id.; 9, Tristan da Cunha; ZE1, Southern Rhodesia.
ZK–ZL–ZM	New Zealand: ZK1, Cook Ids., Zanzibar; ZK2, Niue; ZL, New Zealand proper; ZM, Brit. Samoa.
ZP	Paraguay
ZS–ZT–ZU	Union of South Africa: ZS1–2, 4–6, South Africa proper; ZS3, Southwest Africa.

Index

Notes

The Catalog Section

☆ ☆ ☆

In the following pages is a catalog-file of products of the principal manufacturers who serve the short-wave field. Appearance in these pages is by invitation—space has been sold only to those dependable firms whose established integrity and whose products have met with the approval of the American Radio Relay League.

INDEX OF MANUFACTURERS

★ CATALOG SECTION ★

The Radio Amateur's Handbook

NATIONAL RADIO PRODUCTS
•
1947

CATALOG NO. 700

NATIONAL COMPANY, INC.

MALDEN, MASSACHUSETTS, U. S. A.

NATIONAL DIALS

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 or 5 scale. Fits ¼" shaft. **Specify scale.**

N Dial — **List $**
AD Dial — **List $**

"Velvet Vernier" Dial, Type B, has a compact variable ratio 6 to 1 minimum, 20 to 1 maximum drive that is smooth and trouble free. The case is black bakelite. 1 or 5 scale. 4" diam. Fits ¼" shaft.

Specify scale.

B Dial — **List $**

The original "Velvet Vernier" mechanism is now available in a metal skirted dial 3" in diameter. The planetary drive has a ratio of 5 to 1. It is available with 2, 3, 4, 5 or 6 scale and fits ¼" shaft.

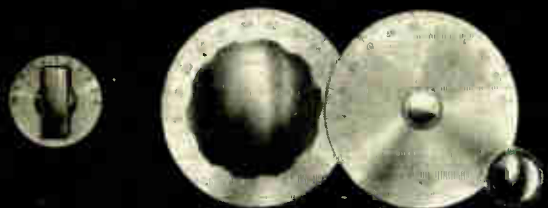
AM Dial — **List $**

The BM Dial is a smaller version of the B Dial (described in the opposite column) for use where space is limited. The drive ratio is fixed. Although small in size, the BM Dial has the same smooth action as the larger units. 1 or 5 scale. 3" diam. Fits ¼" shaft.

Specify Scale.

BM Dial — **List $**

INEXPENSIVE DIALS



TYPE R
List $
1⅝" Dia.

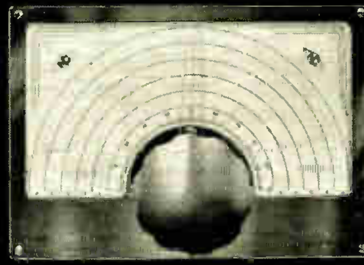
TYPE O
List $
3½" Dia.

TYPE L
List $
5" Dia.

TYPE K
List $
3½" Dia.

TYPE M
List $
5" Dia.

FOR INDIVIDUAL CALIBRATING



For experimenters who "build their own" and desire direct calibration. Fine for Freq. Monitors and VFO's.

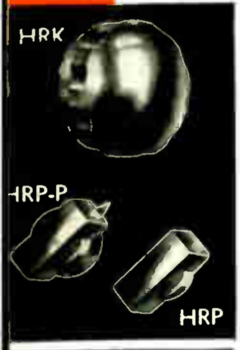
- **Dial bezel size 5" x 7¼"**
- **Five blank ranges for direct calibration**
- **Employs Velvet Vernier Drive**
- **5 to 1 ratio**

TYPE ACN — **List $**

R Dial scale 3 only but marked 10-0; O, K, L, M scale 2. All fit ¼" shafts.

KNOBS

HRK (Fits ¼" shaft) — **List $**
Black bakelite knob 2⅜" diam.



HRP-P (Fits ¼" shaft) — **List $**
Black bakelite knob 1¼" long and ½" wide. Equipped with pointer.

HRP **List $**
The Type HRP knob has no pointer, but is otherwise the same as the knob above.

ACCESSORIES

ODL — **List $**
A locking device which clamps the rim of O, K, L and M Dials. Brass, nickel plated.

ODD — **List $**
Vernier drive for O, K, L, M or other plain dials.

SB (Fits ¼" shaft) — **List $**
A nickel plated brass bushing ½" diam.

RSL (Fits ¼" shaft) — **List $**
Rotor Shaft Lock for TMA, TMC and similar condensers.

DIAL SCALES

Scale	Divisions	Rotation	Direction of Condenser Rotation for Increase of dial reading
1	0-100-0	180°	Either
2	0-100	180°	Counter Clockwise
3	100-0	180°	Clockwise
4	150-0	270°	Clockwise
5	200-0	360°	Clockwise
6	0-150	270°	Counter Clockwise

The HRT is a new gray plastic tuning knob with a chrome plated appearance circle. The HRT knob fits a ¼" dia. shaft and is 2⅛" in dia.

HRT Knob — **List $**

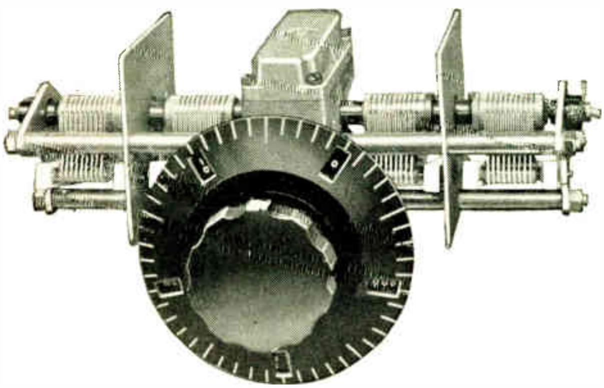
The HRS Knobs are a new gray plastic knob with a 1¾" dia. chrome plated skirt. HRS Knobs fit ¼" dia. shafts. Three types are available as follows:

HRS-1 Knob ON-OFF through 30° rotation — **List $**

HRS-2 Knob 5-0-5 through 180° rotation — **List $**

HRS-3 Knob 0-10 through 300° rotation — **List $**

NATIONAL PRECISION CONDENSERS



The Micrometer dial reads direct to one part in 500. Division lines are approximately 1/4" apart. The dial revolves ten times in covering the tuning range, and the numbers visible through the small windows change every revolution to give consecutive numbering by tens from 0 to 500. The condenser is of extremely rigid construction, with four bearings on the rotor shaft. The drive, at the mid-point of the rotor, is through an enclosed preloaded worm gear with 20 to 1 ratio. Each rotor is individually insulated from the frame, and each has its own individual rotor contact. Stator insulation is Steatite. Plate shape is straight-line frequency when the frequency range is 2:1.

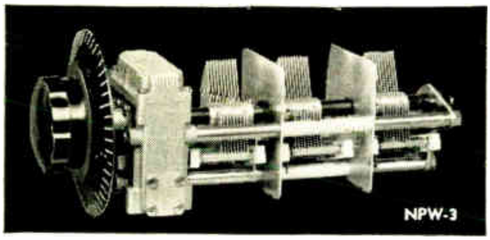
PW Condensers are available in 2, 3 or 4 sections, in either 160 or 225 mmf per section. Larger capacities cannot be supplied.

A single-section PW condenser with grounded rotor is supplied in capacities of 150, 200, 350 and 500 mmf, single spaced, and capacities up to 125 mmf, double spaced.

PW condensers are all with rotor shaft parallel to the panel.

PW-1R	Single section right	**List $**	**PW-3R**	Double section right; single left	**List $**
PW-1L	Single section left	**List $**	**PW-3L**	Double section left; single right	**List $**
PW-2R	Double section right	**List $**	**PW-4**	Double section each side	**List $**
PW-2L	Double section left	**List $**			
PW-2S	Single section each side	**List $**			

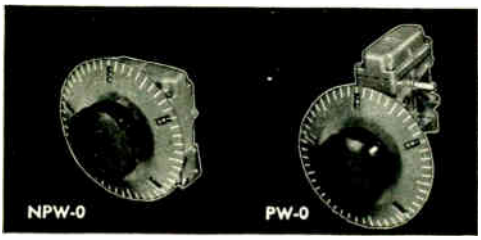
NPW MODEL with micrometer dial.



Similar to PW models, except that rotor shaft is perpendicular to panel.

NPW-3. Three sections, each 225 mmf. **List $**

GEAR DRIVE UNITS with micrometer dial



NPW-O **List $**

Uses parts similar to the NPW condenser. Drive shaft perpendicular to panel. One TX-9 coupling supplied.

PW-O **List $**

Uses parts similar to the PW condenser. Drive shaft parallel to panel. Two TX-9 couplings supplied.

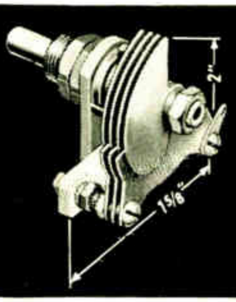
MICROMETER DIAL

PW-D **List $**

Identical with the dials used on the condensers and drives above. It revolves ten times in covering the complete range and as there is no gear reduction unit furnished, the driven shaft will revolve ten times, also. The PW-D dial fits a shaft 5/16" in diameter.

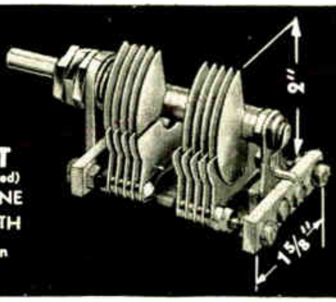
NATIONAL RECEIVING CONDENSERS

TYPE STHS
STRAIGHT-LINE WAVELENGTH
180 Rotation



Capacity	Minimum Capacity	No. of Plates	Air Gap	Length	Catalog Symbol	List
SINGLE BEARING MODELS						
15 Mmf.	3 Mmf.	3	.018"	1 3/16"	STHS- 15	$
25	3.25	4	.018"	1 3/16"	STHS- 25	
50	3.5	7	.018"	1 3/16"	STHS- 50	
DOUBLE BEARING MODELS						
35 Mmf.	6 Mmf.	8	.026"	2 1/4"	ST- 35	$
50	7	11	.026"	2 1/4"	ST- 50	
75	8	15	.026"	2 1/4"	ST- 75	
100	9	20	.026"	2 1/4"	ST-100	
140	10	27	.026"	2 1/4"	ST-140	
150	10.5	29	.026"	2 3/4"	ST-150	
200	12.0	27	.018"	2 1/4"	STH-200	
250	13.5	32	.018"	2 3/4"	STH-250	
300	15.0	39	.018"	2 3/4"	STH-300	
335	17.0	43	.018"	2 1/4"	STH-335	
SPLIT STATOR DOUBLE BEARING MODELS						
50–50	5–5	11–11	.026"	2 3/4"	STD- 50	$
100–100	5.5–5.5	14–14	.018"	2 3/4"	STHD-100	

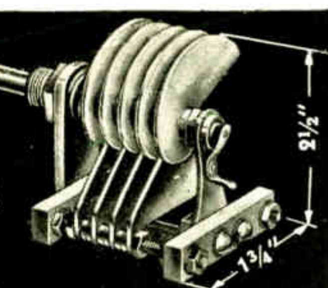
TYPE ST
(Type STD Illustrated)
STRAIGHT-LINE WAVELENGTH
180° Rotation



The ST Type condenser has Straight-Line Wavelength plates. All double-bearing models have the front bearing insulated to prevent noise. On special order a shaft extension at each end is available, for ganging. On double-bearing single shaft models, the rotor contact is through a constant impedance pigtail. Steatite insulation.

NOTE — Type SS Condensers, having straight-line-capacity plates but otherwise similar to the Type ST, are available. Capacities and Prices same as Type ST.

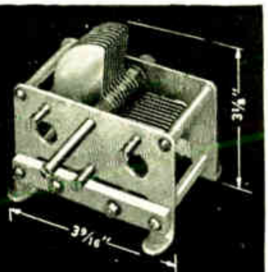
TYPE SE
(Type SEU Illustrated)
STRAIGHT-LINE FREQUENCY
270° Rotation



Capacity	Minimum Capacity	No. of Plates	Air Gap	Length	Catalog Symbol	List
15 Mmf.	7 Mmf.	6	.055"	2 1/4"	SEU- 15	$
20	7.5	7	.055"	2 1/4"	SEU- 20	
25	8	9	.055"	2 1/4"	SEU- 25	
50	9	11	.026"	2 1/4"	SE- 50	
75	10	15	.026"	2 1/4"	SE- 75	
100	11.5	20	.026"	2 1/4"	SE-100	
150	13	29	.026"	2 3/4"	SE-150	
200	12	27	.018"	2 1/4"	SEH-200	
250	14	32	.018"	2 3/4"	SEH-250	
300	16	39	.018"	2 1/4"	SEH-300	
335	17	43	.018"	2 1/4"	SEH-335	

TYPE SE — All models have two rotor bearings, the front bearing being insulated to prevent noise. A shaft extension at each end, for ganging, is available on special order. On models with single shaft extension, the rotor contact is through a constant impedance pigtail. The SEU models (illustrated) are suitable for high voltages as their plates are thick polished aluminum with rounded edges. Other SE condensers do not have polished edges on the plates. Steatite insulation.

TYPE EMC
STRAIGHT-LINE WAVELENGTH
180 Rotation



Capacity	Minimum Capacity	No. of Plates	Length	Catalog Symbol	List
150 Mmf.	9 Mmf	9	2 15/16"	EMC-150	$
250	11	15	2 15/16"	EMC-250	
350	12	20	2 15/16"	EMC-350	
500	16	29	4 3/8"	EMC-500	
1000	22	58	6 3/4"	EMC-1000	

TYPE EMC — A general purpose condenser available in large sizes and having Straight-Line wavelength plates. They are similar in construction to the TMC Transmitting condenser, and have high efficiency and rugged frames. Insulation is Steatite, and Peak Voltage Rating is 1000 volts. Same sizes available with straight line capacity plates, type DXC condenser.

NATIONAL MINIATURE CONDENSERS

PSR — See table — Type PSR condensers are small, compact, low-loss units with silver plating on conducting parts. Their soldered construction makes them particularly suitable for applications where vibration is present. Adjustment is made with a screw driver. Steatite base.

PSE — See table — Type PSE condensers are similar to Type PSR, but are provided with a 1/4" diameter shaft extension at each end.

PSL — See table — Type PSL condensers are similar to Type PSR, but are provided with a rotor shaft lock, so that the rotor can be clamped at any setting.

M-30 **List $**

Type M-30 is a small adjustable mica condenser with a maximum capacity of 30 mmf. Dimensions 13/16" x 9/16" x 1/2". Isolantite base.

W-75, 75 mmf. **List $**
W-100, 100 mmf. **List $**

Capacity	Catalog Symbol			List
25 mmf.	PSR-25	PSE-25	PSL-25	$
50	PSR-50	PSE-50	PSL-50	
75	PSR-75	PSE-75	PSL-75	
100	PSR-100	PSE-100	PSL-100	
140	PSR-140	PSE-140	PSL-140	

Capacity	Minimum Capacity	No. of Plates	Air Gap	Catalog Symbol	List
15 mmf.	1.5	6	.017"	UM-15	$
35	2.5	12	.017"	UM-35	
50	3	16	.017"	UM-50	
75	3.5	22	.017"	UM-75	
100	4.5	28	.017"	UM-100	
10	1	8	.042"	UM-10D	
25	3 4	14	.042"	UMA-25	
BALANCED STATOR MODEL					
25	2	4-4-4	.017"	UMB-25	$
50	5	8-8-8	.017"	UMB-50	

Small padding condensers having very low temperature coefficient. Mounted in an aluminum shield 1 1/4" in diameter.

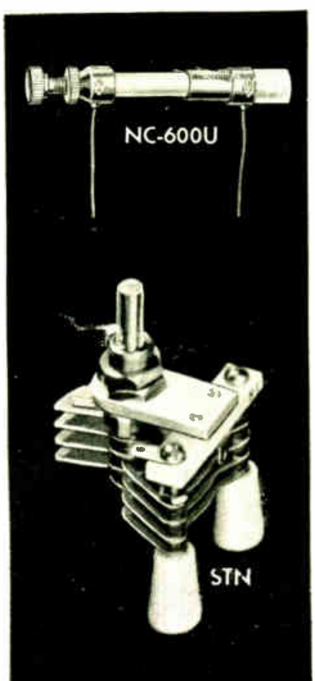
The UM CONDENSER is designed for ultra high frequency use and is small enough for convenient mounting in PB-10 and RO shield cans. They are particularly useful for tuning receivers, transmitters, and exciters. Shaft extensions at each end of the rotor permit easy ganging when used with one of our flexible couplings. The UMB-25 Condenser is a balanced stator model, two stators act on a single rotor. The UM can be mounted by the angle foot supplied or by bolts and spacers. See table for sizes.

Dimensions: Base 1" x 2 1/4", Mounting holes 5/8" x 1 23/32", Axial length 2 1/8" overall.

Plates: Straight line capacity, 180° rotation.

The UM-10D and UMA-25 condensers are double spaced versions of the UM condenser. The UMA-25 is assembled with nuts and bolts so that the capacity may be reduced if desired.

NATIONAL NEUTRALIZING CONDENSERS



NC-600U **List $**
With standoff insulator

NC-600 **List $**
Without insulator

For neutralizing low power beam tubes requiring from .5 to 4 mmf., and 1500 max. total volts such as the 6L6. The NC-600U is supplied with a GS-10 standoff insulator screwed on one end, which may be removed for pigtail mounting.

STN **List $**

The Type STN has a maximum capacity of 18 mmf. (3000 V), making it suitable for such tubes as the 10 and 45. It is supplied with two standoff insulators.

NC-800A **List $**

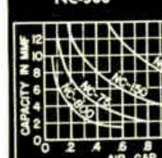
The NC-800A disk-type neutralizing condenser is suitable for the RCA-800, 35T, HK-54 and similar tubes. It is equipped with a clamp to lock its setting. The chart below gives capacity and air gap for different settings.

NC-75 **List $**
For 75T, 808, 811, 812 & similar tubes.

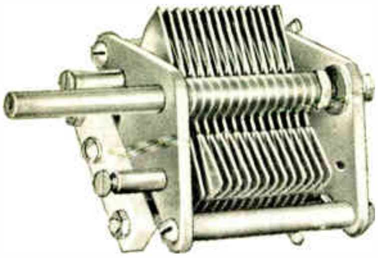
NC-150 **List $**
For HK354, RK36, 300T, 852, etc.

NC-500 **List $**
For WE-251, 450TH, 450TL, 750TL, etc.

These larger disk type neutralizing condensers are for the higher powered tubes. Disks are aluminum, insulation steatite.



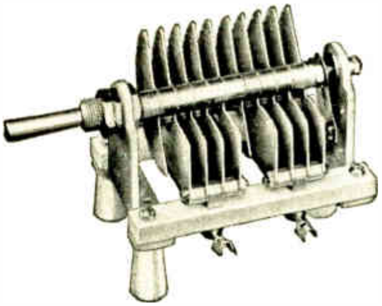
NATIONAL TRANSMITTING CONDENSERS



TYPE TMS

is a condenser designed for transmitter use in low power stages. It is compact, rigid, and dependable. Provision has been made for mounting either on the panel, on the chassis, or on two stand-off insulators. Insulation is Steatite. Voltage ratings listed are conservative.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
SINGLE STATOR MODELS							
100 Mmf.	9.5	3″	.026″	1000v.	9	TMS-100	
150	11	3″	.026″	1000v.	14	TMS-150	
250	13.5	3″	.026″	1000v.	22	TMS-250	
300	15	3″	.026″	1000v.	27	TMS-300	
35	8	3″	.065″	2000v.	7	TMSA-35	
50	11	3″	.065″	2000v.	11	TMSA-50	
DOUBLE STATOR MODELS							
50–50 Mmf.	6–6	3″	.026″	1000v.	5–5	TMS-50D	
100–100	7–7	3″	.026″	1000v.	9–9	TMS-100D	
50–50	10.5–10.5	3″	.065″	2000v.	11–11	TMSA-50D	



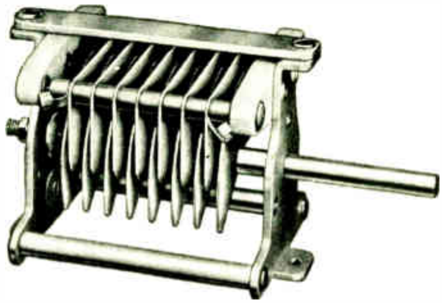
TYPE TMH

features very compact construction, excellent power factor, and aluminum plates .040″ thick with polished edges. It mounts on the panel or on removable stand-off insulators. Steatite insulators have long leakage path. Stand-offs included in listed price.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List
SINGLE STATOR MODELS							
50 Mmf.	9	3¾″	.085″	3500v.	15	TMH-50	
75	11	3¾″	.085″	3500v.	19	TMH-75	
100	12.5	5⅛″	.085″	3500v.	25	TMH-100	
150	18	6½″	.085″	3500v.	37	TMH-150	
35	11	5⅛″	.180″	6500v.	17	TMH-35A	
DOUBLE STATOR MODELS							
35–35 Mmf.	6–6	3¾″	.085″	3500v.	9–9	TMH-35D	
50–50	8–8	5⅛″	.085″	3500v.	13–13	TMH-50D	
75–75	11–11	6½″	.085″	3500v.	19–19	TMH-75D	

NATIONAL TRANSMITTING CONDENSERS

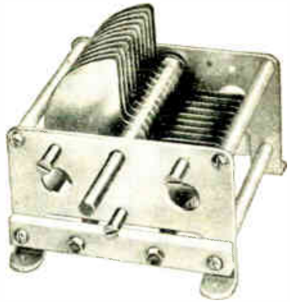
TYPE TMK



is a new condenser for exciters and low power transmitters. Special provision has been made for mounting AR-16 coils in a swivel plug-in mount on either the top or rear of the condenser, (see page 10). For panel or stand-off mounting; Steatite insulation.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
			SINGLE STATOR MODELS				
35 Mmf.	7.5	2 7/32"	.047"	1500v.	7	TMK-35	
50	8	2 3/8"	.047"	1500v.	9	TMK-50	
75	9	2 11/16"	.047"	1500v.	13	TMK-75	
100	10	3"	.047"	1500v.	17	TMK-100	
150	10.5	3 5/8"	.047"	1500v.	25	TMK-150	
200	11	4 1/4"	.047"	1500v.	33	TMK-200	
250	11.5	4 7/8"	.047"	1500v.	41	TMK-250	
			DOUBLE STATOR MODELS				
35–35 Mmf.	7.5–7.5	3"	.047"	1500v.	7–7	TMK-35D	
50–50	8–8	3 5/8"	.047"	1500v.	9–9	TMK-50D	
100–100	10–10	4 1/4"	.047"	1500v.	17–17	TMK-100D	
Swivel Mounting Hardware for AR 16 Coils						SMH	

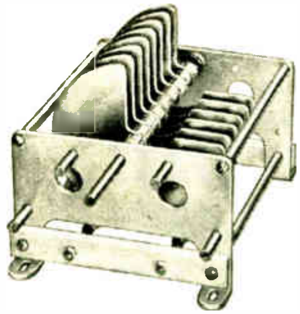
TYPE TMC



is designed for use in the power stages of transmitters where peak voltages do not exceed 3000. The frame is extremely rigid and arranged for mounting on panel, chassis or standoff insulators. The plates are aluminum with buffed edges. Insulation is Steatite. The stator in the split stator models is supported at both ends.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
			SINGLE STATOR MODELS				
50 Mmf.	10	3"	.077"	3000v.	7	TMC-50	
100	13	3 1/2"	.077"	3000v.	13	TMC-100	
150	17	4 5/8"	.077"	3000v.	21	TMC-150	
250	23	6"	.077"	3000v.	32	TMC-250	
300	25	6 3/4"	.077"	3000v.	39	TMC-300	
			DOUBLE STATOR MODELS				
50–50 Mmf.	9–9	4 5/8"	.077"	3000v.	7–7	TMC-50D	
100–100	11–11	6 3/4"	.077"	3000v.	13–13	TMC-100D	
200–200	18.5–18.5	9 1/4"	.077"	3000v.	25-25	TMC-200D	

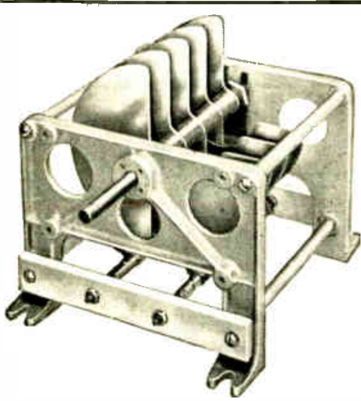
NATIONAL TRANSMITTING CONDENSERS



TYPE TMA

is a larger model of the popular TMC. The frame is extremely rigid and arranged for mounting on panel, chassis or stand-off insulators. The plates are of heavy aluminum with rounded and buffed edges. Insulation is Steatite located outside of the concentrated field.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
SINGLE STATOR MODELS							
300 Mmf.	19.5	4 9/16"	.077"	3000v.	23	TMA-300	
50	15	4 9/16"	.171"	6000v.	7	TMA-50A	
100	19.5	6 7/8"	.171"	6000v.	15	TMA-100A	
150	22.5	6 7/8"	.171"	6000v.	21	TMA-150A	
230	33	9 5/16"	.171"	6000v.	33	TMA-230A	
100	30	9 1/4"	.265"	9000v.	23	TMA-100B	
150	40.5	12 1/2"	.265"	9000v.	33	TMA-150B	
50	21	7 1/8"	.359"	12000v.	13	TMA-50C	
100	37.5	12 7/8"	.359"	12000v.	25	TMA-100C	
DOUBLE STATOR MODELS							
200-200 Mmf.	15-15	6 7/8"	.077"	3000v.	16-16	TMA-200D	
180-180	10-10	12 3/4"	.140"	4000v.	24-24	TMA-180D	
50-50	12.5-12.5	6 7/8"	.155"	6000v.	8-8	TMA-50DA	
100-100	17-17	9 5/16"	.155"	6000v.	14-14	TMA-100DA	
60-60	19.5-19.5	12 1/2"	.249"	9000v.	15-15	TMA-60DB	
40-40	18-18	12 7/8"	.343"	12000v.	11-11	TMA-40DC	



TYPE TML

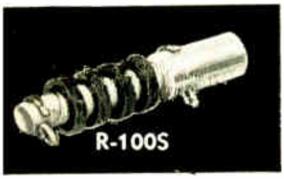
condenser is a 1 KW job throughout. Steatite insulators, specially treated against moisture absorption, prevent flash-overs. A large self-cleaning rotor contact provides high current capacity. Thick capacitor plates, with accurately rounded and polished edges, provide high voltage ratings. Sturdy cast aluminum end frames and dural tie bars permit an unusually rigid structure. Precision end bearings insure smooth turning and permanent alignment of the rotor. End frames are arranged for panel, chassis or stand-off mountings.

Capacity	Minimum Capacity	Length	Air Gap	Peak Voltage	No. of Plates	Catalog Symbol	List Price
SINGLE STATOR MODELS							
75 Mmf.	25	18 1/16"	.719"	20,000v.	17	TML-75E	
150	60	18 1/16"	.469"	15,000v.	27	TML-150D	
100	45	13 5/8"	.469"	15,000v.	19	TML-100D	
50	22	8 5/16"	.459"	15,000v.	9	TML-50D	
245	54	18 1/16"	.344"	10,000v.	35	TML-245B+	
150	45	13 5/8"	.344"	10,000v.	21	TML-150B+	
100	32	10 15/16"	.344"	10,000v.	15	TML-100B+	
75	23.5	8 5/16"	.344"	10,000v.	11	TML-75B+	
500	55	18 1/16"	.219"	7,500v.	49	TML-500A+	
350	45	13 5/8"	.219"	7,500v.	33	TML-350A+	
250	35	10 15/16"	.219"	7,500v.	25	TML-250A+	
DOUBLE STATOR MODELS							
30–30 Mmf.	12–12	18 1/16"	.719"	20,000v.	7–7	TML-30DE	
60–60	26–26	18 1/16"	.469"	15,000v.	11–11	TML-60DD	
100–100	27–27	18 1/16"	.344"	10,000v.	15–15	TML-100DB+	
60–60	20–20	13 5/8"	.344"	10,000v.	9–9	TML-60DB+	
200–200	30–30	18 1/16"	.219"	7,500v.	21–21	TML-200DA+	
100–100	17–17	10 15/16"	.219"	7,500v.	11–11	TML-100DA+	

NATIONAL RF CHOKES

R-100

R-100U



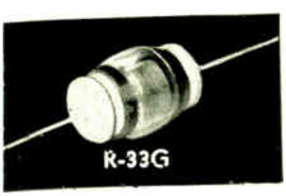
R-100S

R-300

R-300U

R-300S

R-33



R-33G

R-100 **List $**
R-100U **List $**
R-100S **List $**

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 standoff insulator; the R-100S has cotter-pin lug terminals and a stand-off insulator. These chokes are available in 2.5, 5 and 10 millihenry sizes and are rated at 125 milliamperes.

R-300 **List $**
R-300U **List $**
R-300S **List $**

RF chokes R-300, R-300U and R-300S are similar in size to R-100 series but have higher current capacity. The R-300U s provided with a removable stand-off insulator at one end. The R-300S has non-removable stand-off insulator and cotter-pin lug terminals. 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 and R-300S are identical electrically.

R-33 **List $**

The R-33 series chokes are 2-section RF chokes and available in 1, 10, 50 and 100 microhenry sizes. They are rated at 33 milliamperes. The chokes are wound on a 5/8" long form and range in diameter up to 5/16" maximum diameter.

The R-33G choke is a 2-section 750 microhenry RF choke hermetically sealed in glass with a current rating of 33 milliamperes. The choke body is 1" long by 5/8" diameter.

R-152 **List $**

For the 80 and 160 meter bands. Inductance 4 m.h., DC resistance 10 ohms, DC current 600 ma. Coils honeycomb wound on Isolantite core.

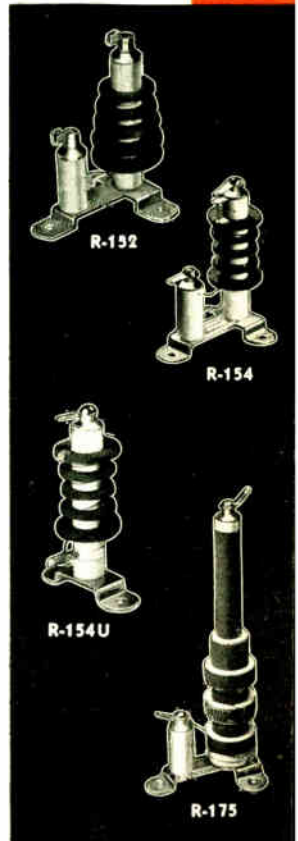
R-154 **List $**

R-154U **List $**

For the 20, 40 and 80 meter bands. Inductance 1 m.h., DC resistance 6 ohms, DC current 600 ma. Coils honeycomb wound on Isolantite core. The R-154U does not have the third mounting foot and the small insulator, but is, otherwise the same as R-154. See illustration.

R-175 **List $**

The R-175 Choke is suitable for parallel-feed as well as series-feed in transmitters with plate supply up to 3000 volts modulated or 4000 volts unmodulated. Unlike conventional chokes, the reactance of the R-175 is high throughout the 10 and 20 meter bands as well as the 40, 80 and 160 meter bands. Inductance 225 μh, distributed capacity 0.6 mmf., DC resistance 6 ohms, DC current 800 ma., voltage breakdown to base 12,500 volts.



R-50 RF Choke **List $**

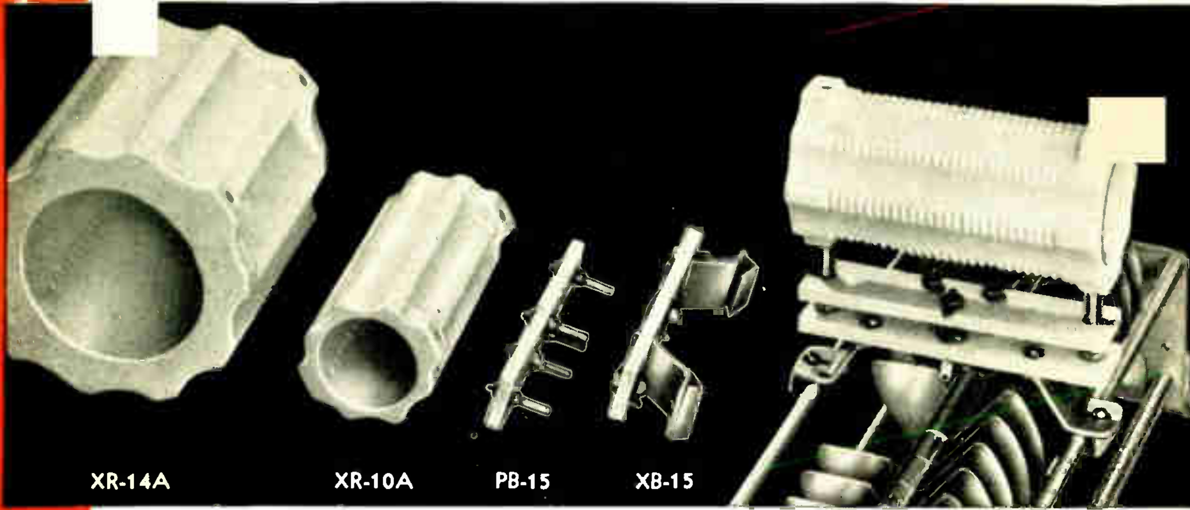
The R-50 series chokes are 4-section RF chokes and available in 0.5, 1, 2.5, and 10 millihenry sizes. They are rated at 50 milliamperes. The chokes are wound on a 1" long form and have a maximum diameter of 15/32". The 10 millihenry choke is wound on an iron core.

R-50

R-60 RF Choke **List $**

The R-60 choke is a high current RF choke (500 milliamperes) available in 2 and 4 microhenry sizes. The choke is 1 1/8" long by 5/16" diameter.

R-60



TRANSMITTER COIL FORMS

The Transmitter Coil Forms and Mounting are designed as a group, and mount conveniently on the bars of a TMA condenser. The larger coil form, Type XR-14A, has a winding diameter of 5″, a winding length of 3¾″ (30 turns total) and is intended for the 80 meter band. The smaller form, Type XR-10A, has a winding length of 3¾″ and a winding diameter of 2½″ (26 turns total). It is intended for the 20 and 40 meter bands.

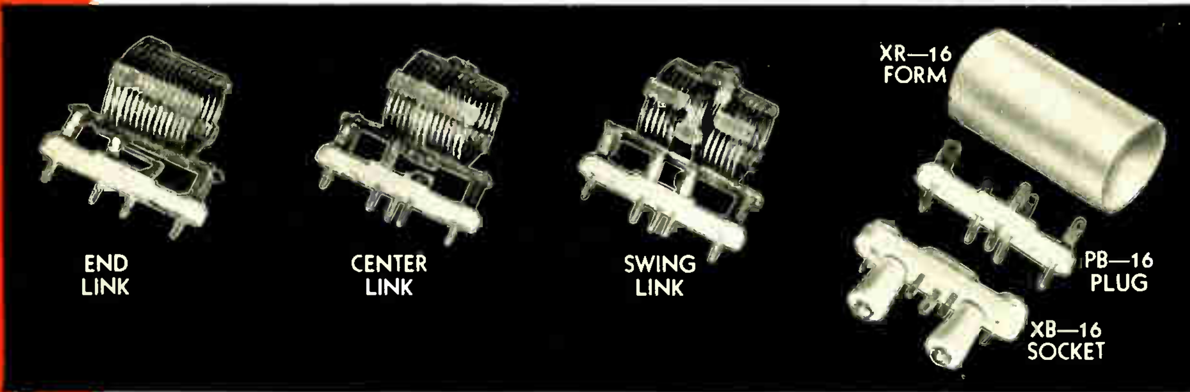
Either coil form fits the PB-15 plug. For higher frequencies, the plug may be used with a self-supporting coil of copper tubing. The XB-15 Socket may be mounted on breadboards or chassis, as well as on the TMA Condenser.

SINGLE UNITS

XR-10A, Coil Form only List $
XR-14A, Coil Form only List $
PB-15, Plug only List $
XB-15, Socket only List $

ASSEMBLIES

UR-10A, Assembly (Including small Coil Form, Plug and Socket) List $
UR-14A, Assembly (including large Coil Form, Plug and Socket) List $



EXCITER COILS AND FORMS — TYPE AR-16 (Air Spaced)

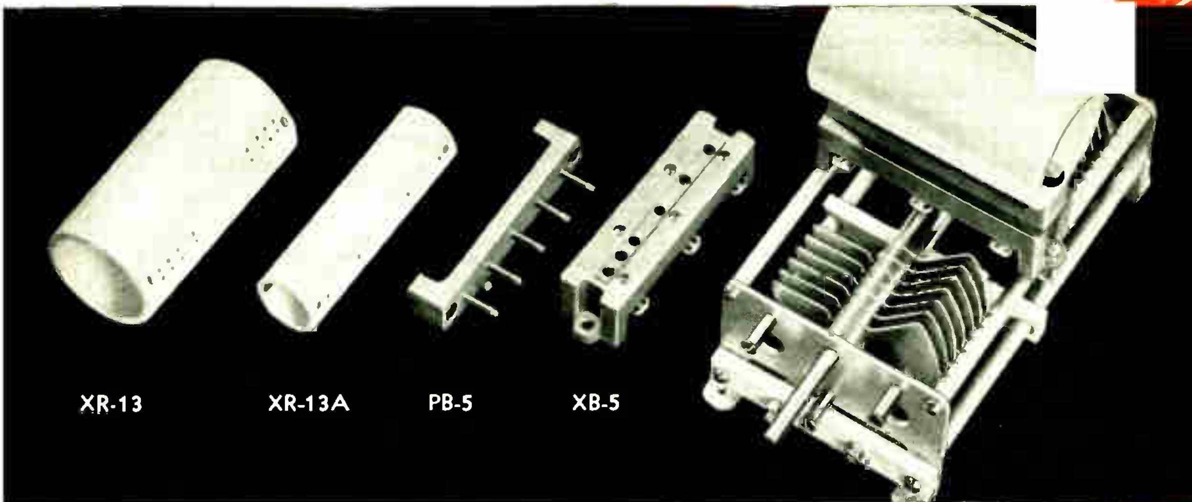
These air-spaced coils are suitable for use in stages where the plate input does not exceed 50 watts and are available in the sizes tabulated below. Capacities listed will resonate the coils at the low frequency end of the band and include all stray circuit capacities. All have separate link coupling coils and all fit the PB-16 Plug and XB-16 Socket.

The XR-16 Coil Form also fits the PB-16 Plug and XB-16 Socket. It has a winding diameter of 1¼″ and a winding length of 1¾″.

Band	End Link	Cap Mmf	Center Link	Cap Mmf	Swinging Link	Cap Mmf
6 meter	AR16-6E	25	AR16-6C	25		
10 meter	AR16-10E	20	AR16-10C	20	AR16-10S	25
20 meter	AR16-20E	26	AR16-20C	26	AR16-20S	40
40 meter	AR16-40E	33	AR16-40C	33	AR16-40S	55
80 meter	AR16-80E	37	AR16-80C	37	AR16-80S	60
160 meter	AR16-160E	65	AR16-160C	65		

XR-16, Coil Form only List $
PB-16, Plug-in Base only List $
XB-16, Plug-in Socket only List $
AR-16, Coils — Any type (see table). Include PB-16 Plug as illustrated Each, List $

Exciter coil on TMK condenser



BUFFER COIL FORMS

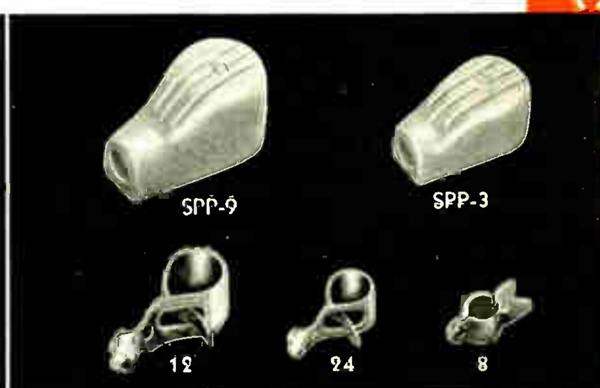
National Buffer Coil Forms are designed to mount directly on the tie bars of a TMC condenser using the PB-5 Plug and XB-5 Socket. Plug and Socket are of molded R-39.

The two coil forms are of Isolantite, left unglazed to provide a tooth for coil dope. The larger form, Type XR-13, is 1¼" in diameter and has a winding length of 2¼". The smaller form, Type XR-13A, is 1" in diameter and provides a winding length of 2¾". Both forms have holes for mounting and for leads.

SINGLE UNITS
XR-13, Coil Form only — **List $**
XR-13A, Coil Form only — **List $**
PB-5, Plug only — **List $**
XB-5, Socket only — **List $**

ASSEMBLIES
UR-13A, Assembly (including small Coil Form, Plug and Socket) — **List $**

UR-13, Assembly (including large Coil Form, Plug and Socket) — **List $**



FIXED TUNED EXCITER TANK

Similar in general construction to National I.F. transformers, this unit has two 25 mmf., 2000 volt air condensers and an unwound XR-2 coil form.

FXT, without plug-in base — **List $**
FXTB-5, with 5 prong base — **List $**
FXTB-6, with 6 prong base — **List $**

PLUG-IN BASE AND SHIELD

The low-loss R-39 base is ideal for mounting condensers and coils when it is desirable to have them shielded and easily removable. Shield can is 2" x 2⅜" x 4⅛".

PB-10-5, (5 Prong Base & Shield) — **List $**
PB-10-6, (6 Prong Base & Shield) — **List $**
PB-10A-5, (5 Prong Base only) — **List $**
PB-10A-6, (6 Prong Base only) — **List $**

SAFETY GRID AND PLATE CAPS

National Safety Grid and Plate Caps have a ceramic body which offers protection against accidental contact with high voltage caps on tubes.

SPP-9 — **List $**
Ceramic insulation. Fits 9/16" diameter.

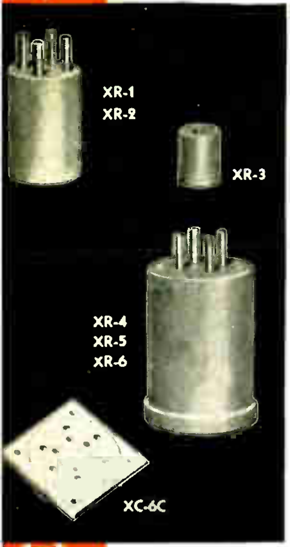
SPP-3 — **List $**
Ceramic insulation. Fits ⅜" diameter.

GRID AND PLATE GRIPS

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.

Type 12, for 9/16" Caps — **List $**
Type 24, for ⅜" Caps — **List $**
Type 8, for ¼" Caps — **List $**

NATIONAL PARTS



COIL FORMS

XR-1, Four prong, **List $**
XR-2, without prongs **List $**

Molded of R-39, permitting them to be grooved and drilled. Coil form diameter 1", length 1½".

XR-3 **List $**

Molded of R-39. Diameter 9/16", length ¾". Without prongs.

XR-4, Four prong, **List $**
XR-5, Five prong, **List $**
XR-6, Six prong, **List $**

Molded of R-39, permitting them to be grooved and drilled. Coil form diameter 1½", length 2¼". A special socket is required for the six-prong form.

XC6C, Special six-prong socket for XR-6 Coil Form, **List $**

OSCILLATOR COIL

OSR **List $**

A shielded oscillator coil which tunes to 100 KC with .00041 Mfd. Two separate inductances, closely coupled. Excellent for interruption-frequency oscillator in super-regenerative receivers.

H. F. COIL FORMS

Symbol	Outside Diameter	Length	List
PRC-1	3/8"	3/8"	$
PRC-2	3/8"	1/2"	
PRC-3	3/8"	3/4"	
PRD-1	1/2"	1/2"	
PRD-2	1/2"	1"	
PRE-1	9/16"	3/4"	
PRE-2	9/16"	1"	
PRE-3	9/16"	2"	
PRF-1	3/4"	3/4"	
PRF-2	3/4"	1 [illegible]"	

COIL SHIELDS

RZ, coil shield **List $**
1⅜" square x 4" high.

RS, coil shield **List $**
1 11/16" x 1⅞" x 3½" high.

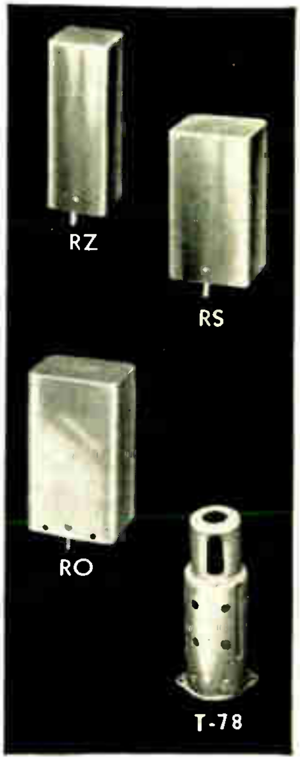
RO, coil shield **List $**
2" x 2⅜" x 4⅛" high.

National coil shields are formed from a single piece of pure aluminum. They are mechanically strong and have ample thickness to mount small parts on the walls.

The RZ, RS and RO coil shields are supplied with two threaded studs extending downward from the open end for attaching to the chassis.

T-78, tube shield complete **List $**

National tube shield type T-78 is a three-piece pure aluminum shield suitable for shielding glass tubes with ST-12 bulb, such as the 6C6 and 6D6 tubes.



JACK SHIELD

JS-1, Jack shield **List $**

For shielding small standard jacks mounted behind a panel, or on the ends of extension cords.

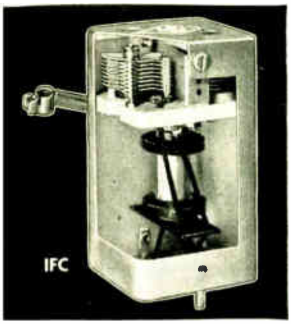
NATIONAL CABINETS

The National Cabinets listed below are the same as those used in National Receivers, except that they are supplied in blank form. They are made of heavy gauge steel, and the paint is unusually well bonded to the metal. Sub-bases and bottom covers are included in the price.

	Width	Height	Depth	List Price
Type C-SW3	9¾"	7"	9"	
Type C-NC100	17¼"	8¾"	11¼"	
Type C-HRO	16¾"	8¾"	10"	
Type C-One-Ten	11"	7"	7¼'	
Type C-SRR	7½"	7"	7½"	

NATIONAL CABINETS

NATIONAL PARTS

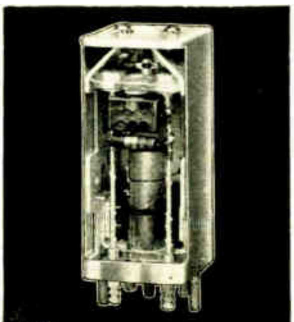


IFC

I. F. TRANSFORMERS

IFC, Transformer, air core List $
IFCO, Oscillator, air core List $

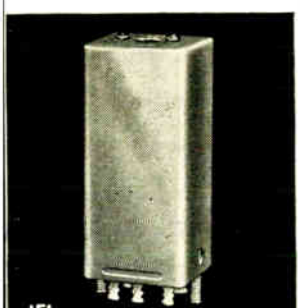
Air dielectric condensers isolated from each other by an aluminum shield. Litz wound coils on a moisture proofed ceramic base. Shield can 4⅛" x 2⅛" x 2". Available for either 175 KC or 450-550 KC. Specify frequency.



IFK

IFG, IF Transformer List $
IFH, Discriminator List $

High frequency IF transformers, similar in construction to the IFC above. They are intended for FM receivers and others requiring a high IF frequency. Frequency is 3 MC. When definite assignment of the bands has been made these transformers will be available in a frequency which gives the minimum images in the FM and television bands.



IFL

15 Mc. IF transformers suitable for ultra high frequency superheterodynes. They are made in two models, with and without variable coupling. Approximate stage gain of 10 is obtained with IFJ or IFK Transformer and 6AB7 tube.

IFJ, with variable coupling List $

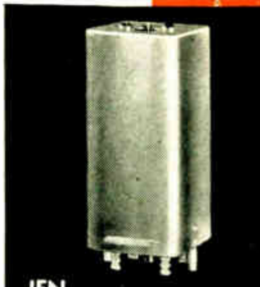
IFK, with fixed coupling List $

IFM

IFL, IFM, IFN and IFO transformers operate at 10.7 Mc. and designed for use in AM or FM Superheterodyne receivers. The transformer cans are 1⅜" square and stand 3½" above the chassis. Two 6-32 spade bolts are provided for mounting.

The IFL transformer is a 10.7 Mc. FM discriminator transformer suitable for use in conventional FM receiver discriminator circuit and is linear over a band of ±100 Kc.

The IFM transformer is a 10.7 Mc. IF transformer with a 150 Kc. bandwidth at 1.5 db attenuation. Approximate stage gain of 30 is obtained with IFM Transformer and 6SG7 tube.



IFN

The IFN transformer is a 10.7 Mc. IF transformer with a 100 Kc. pass band at 1.5 db attenuation. Approximate stage gain of 30 is obtained with IFN Transformer and 6SG7 tube.

The IFO transformer is a 10.7 Mc. FM discriminator transformer of the ratio type and is linear over a band of ±100 Kc.

IFL FM Discriminator List $
IFM IF Transformer List $
IFN IF Transformer List $
IFO FM Ratio Discriminator List $

CHART FRAME

CHART FRAME

The National Chart Frame is blanked from one piece of metal, and includes a celluloid sheet to cover the chart. Size 2¼" x 3¼", with sides ¼" wide.

Type CFA List $

COIL DOPE

COIL DOPE

CD-1, ¼ pint can List $

Liquid Polystyrene Cement — is ideal for windings as it will not spoil the properties of the best coil form.

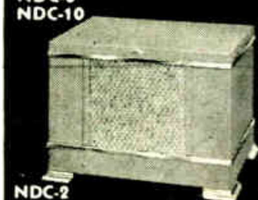
TOUCH-UP PAINT

TOUCH-UP PAINT

A high quality air-drying paint that may be applied with a brush. It is especially suited to touching up places on radio equipment where the paint may have become marred through abrasion.

CP-1, gray List $
CP-2, black List $

NDC-8
NDC-10



NDC-2

SPEAKER CABINETS

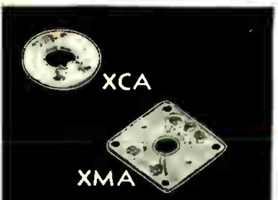
NDC-8 for 8" speaker List $

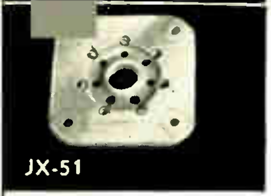
NDC-10 for 10" speaker List $

NDC-2 for 10" speaker List $

These metal speaker cabinets are acoustically correct. They are lined with acoustic felt, and are of welded construction to eliminate rattles. Finish is black wrinkle on NDC-8 and NDC-10. NDC-2 is finished in gray wrinkle to match the NC-2 40D receiver.

NATIONAL LOW-LOSS SOCKETS AND INSULATORS





XLA List $

A low-loss socket for the 6F4 and 950 series acorn tubes for frequencies as high as 600 MC. Conventional by-pass condensers may be compactly mounted between the contact terminals and the chassis. Low contact resistance, short and direct leads and low and constant inductance are features.

XLA-S List $

An internal shield fitting the XLA socket and suitable for tubes such as the 956.

XLA-C List $

This miniature by-pass condenser may be mounted inside the socket, directly below the contact. Capacities of 50 or 100 mmf. available.

XCA List $

A low-loss socket for acorn triodes.

XMA List $

For pentode acorn tubes, this socket has built-in by-pass condensers. The base is a copper plate.

XM-10 List $

A heavy duty metal shell socket for tubes having the XU base.

XM-50 List $

A heavy duty metal shell socket for tubes having the Jumbo 4-pin base ("fifty watters").

JX-50 List $

Without Standoff Insulators

JX-51 List $

A low loss wafer socket for the 813 and other tubes having the Giant 7-pin base.

HX-100S List $

With Standoff Insulators

A low-loss wafer socket suitable for the type 4-125-A, 4-250-A and other tubes using the Giant 5-pin base. Shield grounding clips are supplied which mount on the chassis with the socket mounting screws to ground the tube shield at three points. Air holes are provided in the socket to permit forced air cooling.

GS-1, 1/2" x 1 3/8" List $

GS-2, 1/2" x 2 7/8" List $

GS-3, 3/4" x 2 7/8" List $

GS-4, 3/4" x 4 7/8" List $

GS-4A, 3/4" x 6 7/8" List $

Cylindrical low-loss steatite standoff insulators with nickel plated caps and bases.

GSJ, (not illustrated) List $

A special nickel plated jack top threaded to fit the 3/4" diameter insulators GS-3, GS-4 & GS-4A.

GS-5, 1 1/4" List, each $

GS-6, 2" List, each $

GS-7, 3" List, each $

GS-10, 3/4", package of 10 List $

These cone type standoff insulators are of low-loss steatite. They have a tapped hole at each end for mounting.

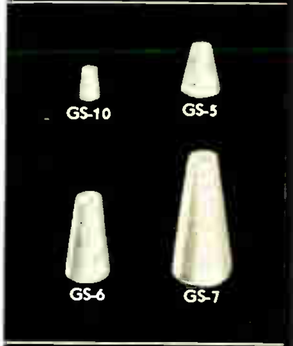
GS-8, with terminal List $

GS-9, with jack List $

These low-loss steatite standoff insulators are also useful as lead-through bushings.

HX-29 List $

A low-loss wafer socket with steatite insulation for the popular 829 and 832 tubes.

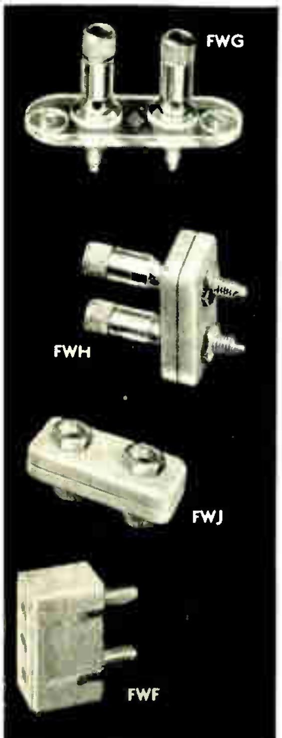


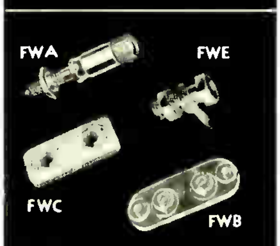
XC Series Sockets

XC-4	List $
XC-5	List $
XC-6	List $
XC-7S	List $
XC-7L	List $
XC-8	List $

National wafer sockets have exceptionally good contacts with high current capacity together with low loss steatite insulation. All types have a locating groove to make tube insertion easy.

NATIONAL LOW-LOSS SOCKETS AND INSULATORS





FWG **List $**

A Victron terminal strip for high frequency use. The binding posts take banana plugs at the top, and grip wires through hole at the bottom, simultaneously, if desired.

FWH **List $**

The insulators of this terminal assembly are molded R-39 and have serrated bosses that allow the thinnest panel to be gripped firmly, and yet have ample shoulders. Binding posts same as FWG above.

FWJ **List $**

This assembly uses the same insulators as the FWH above, but has jacks. When used with the FWF plug (below), there is no exposed metal when the plug is in place.

FWF **List $**

This molded R-39 plug has two banana plugs on ¾" centers and fits FWH or FWJ above. Leads may be brought out through the top or side.

FWA, Post **List, each $**
Brass Nickel Plated

FWE, Jack **List, each $**
Brass Nickel Plated

FWC, Insulator
List, per pair $
R-39 Insulation

FWB, Insulator **List, each $**
Polystyrene insulation

CIR Series Sockets

Any Type **List $**

Type CIR Sockets feature low-loss isolantite or steatite insulation, a contact that grips the tube prong for its entire length, and a metal ring for six position mounting.

AA-3 **List $**

A low-loss steatite spreader for 6 inch line spacing. (600 ohms impedance with No. 12 wire.)

AA-5 **List $**

A low-loss steatite aircraft-type strain insulator.

AA-6 **List $**

A general purpose strain insulator of low-loss steatite.

XS-6 **List, each $**

A low-loss isolantite bushing for ½" holes.

XP-6 Same as above but polysterene.

List, box of ten $

TPB **List, per dozen $**

A threaded polystyrene bushing with removable .093 conductor moulded in, ¼" diam., 32 thread.

XS-7, (⅜" Hole) **List $**

XS-8, (½" Hole) **List $**

Steatite bushings. Prices include male and female bushings with metal fittings.

XS-1, (1" Hole) **List $**

XS-2, (1½" Hole) **List $**

Prices listed are per pair, including metal fittings. Insulation steatite.

XS-3, (2¾" Hole) **List $**

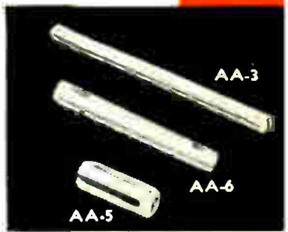
XS-4, (3¾" Hole) **List $**

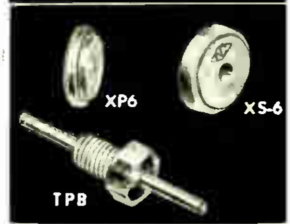
Prices are per pair, including metal fittings. These low-loss steatite bowls are ideal for lead-in purposes at high voltages.

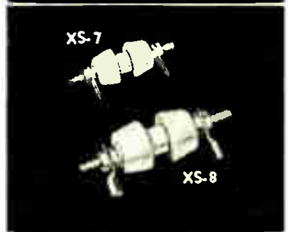
XS-5, Without Fittings
List, each $

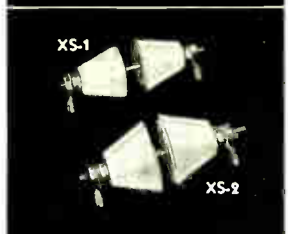
XS-5F, With Fittings
List, per pair $

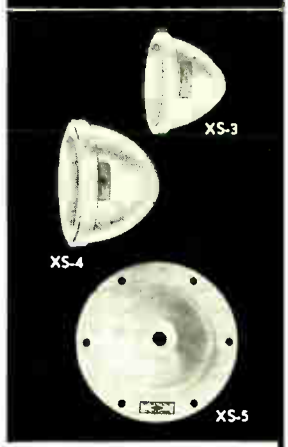
These big low-loss bowls have an extremely long leakage path and a 5¼" flange for bolting in place. Insulation steatite.

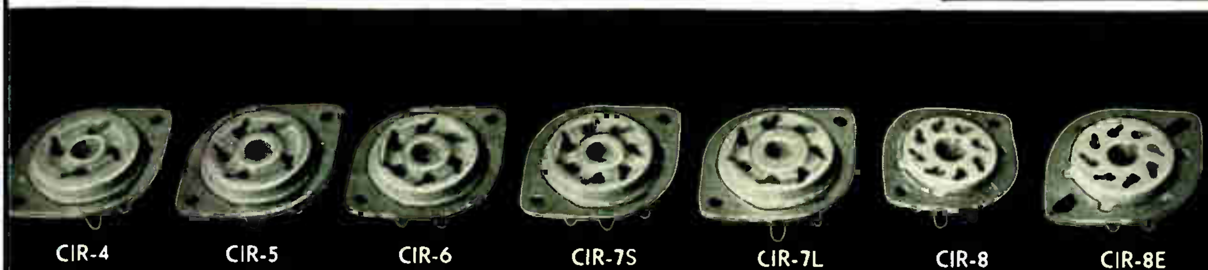




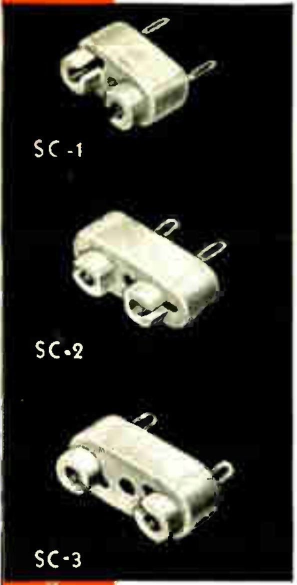








NATIONAL PARTS



The SC-1, SC-2 and SC-3 are crystal mounting sockets for crystal holders with mounting pins spaced 0.500'', 0.486'' and .750'' respectively and pin diameters of 1/16'', 3/32'' and 1/8'' respectively. Steatite Insulation. Single 4-36 or 4-40 screw mounting for CS-1 and CS-2; single 6-32 screw mounting for CS-3.

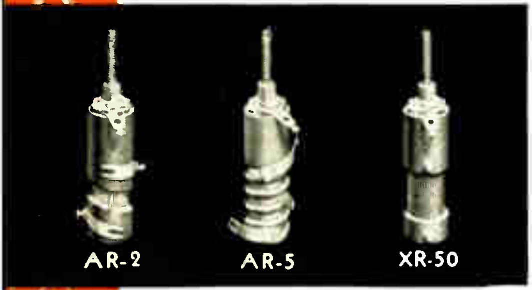
SC-1 List $
SC-2 List $
SC-3 List $

The AR-2 and AR-5 coils are high Q permeability tuned RF coils. The AR-2 coil tunes from 75 Mc. to 220 Mc. with capacities from 100 to 10 micro-micro-farads. The AR-5 coil tunes from 37 Mc. to 110 Mc. with capacities from 100 to 10 micro-micro-farads. The inductive windings supplied may be replaced by other windings as desired to modify the tuning range.

AR-2 High Frequency Coil List $
AR-5 High Frequency Coil List $

The XR-50 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 1/2 inch. The iron slug is 3/8'' dia. by 1/2'' long.

XR-50 List $



The XOA Socket is a socket for the Miniature Button 7 Pin base tubes. Low loss mica filled bakelite insulation. Mounts with two 4-40 screws. Socket contacts extend axially from base of socket.

XOA List $

The XOR Socket is the same as the XOA Socket except that the contacts extend radially from base of socket.

XOR List $

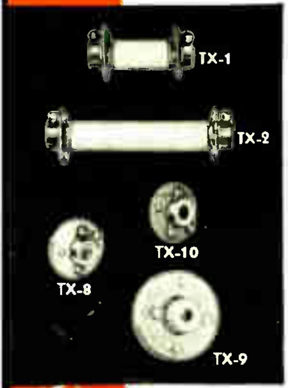
The XOS tube shield is a two piece shield for the Miniature Button 7 Pin base tubes. The shield is available in three sizes corresponding to the 1 5/16'', 1 1/2'' and 2'' tube body heights. The shield contains a spring which centers tube in shield and holds tube and shield firmly in place. The two 4-40 sonde bolts serve to mount the XOA or XOR Socket and the XOS tube shield.

XOS-1 For 1 5/16'' high tube body List $

XOS-2 For 1 1/2'' high tube body List $

XOS-3 For 2'' high body List $

NATIONAL SHAFT COUPLINGS



TX-1, Leakage path 1''. List $

TX-2, Leakage path 2 1/2''. List $

Flexible couplings with glazed steatite insulation which fit 1/4'' shafts.

TX-8 List $

A non-flexible rigid coupling with steatite insulation. 1'' diam. Fits 1/4'' shaft.

TX-9 List $

This small insulated flexible coupling provides high electrical efficiency when used to isolate circuits. Insulation is steatite. 1 5/8'' diam. Fits 1/4'' shaft.

TX-10 List $

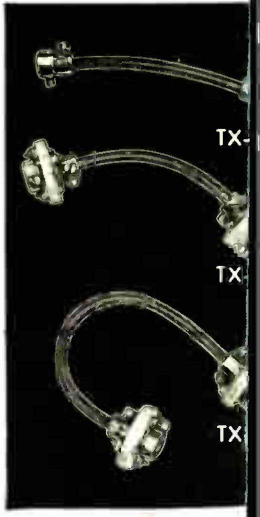
A very compact insulated coupling free from backlash. Insulation is canvas Bakelite. 1 1/16'' diam. Fits 1/4'' shaft.

TX-11 List $

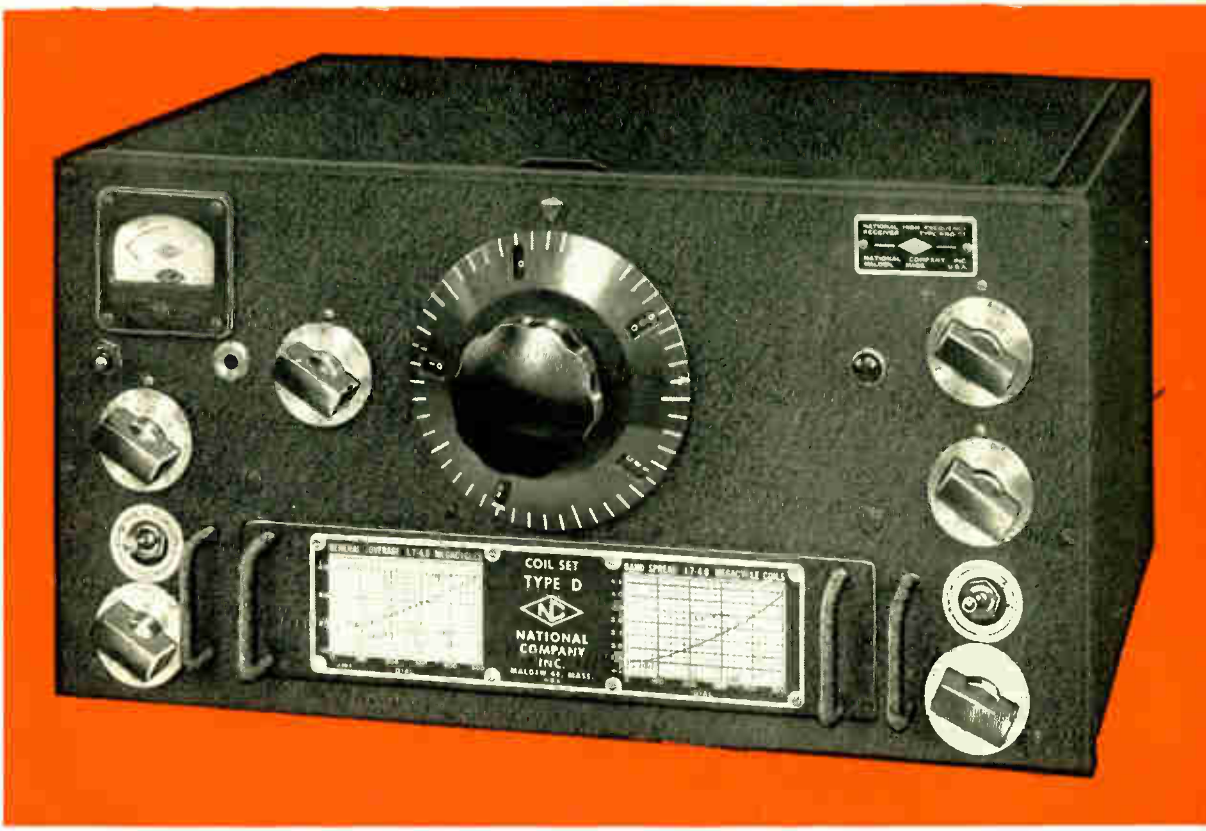
The flexible shaft of this coupling connects shafts at angles up to 90 degrees, and eliminates misalignment problems. Fits 1/4'' shafts. Length 4 1/4''.

TX-12, Length 4 5/8'' List $
TX-13, Length 7 1/8'' List $

These couplings use flexible shafting like the TX-11 above, but are also provided with steatite insulators at each end.



NATIONAL HRO-5A1



DESCRIPTION

The development of the National HRO-5A1 Radio Receiver brings the famous HRO series to a new high in receiver performance.

Items characterizing the HRO-5A1 Receiver are as follows: Two R.F. preselector stages; separate mixer and local oscillator tubes; two I.F. stages with a crystal filter employing phasing and selectivity controls; combined second detector AVC and second audio stage; first audio stage; double action limiter stage; audio output stage; C.W. oscillator with pitch control; and a signal strength meter Metal tubes, first used in the HRO-5, are also employed in the HRO-5A1. The Loud Speaker and Power Unit are separate units. The data listed below indicates the versatility and the extremely high standards of performance to be found in the HRO-5A1.

CONTROLS

Main Tuning Dial: AVC Switch: B+ ON-OFF; Audio Gain; R.F. Gain; C.W. Oscillator Pitch Control; Selectivity Control; Phasing Control; S-Meter Switch; Limiter Control.

SPECIFICATIONS

Frequency Range:

The Frequency Range of the HRO-5A1 with the 4 Coil Sets normally supplied is 1.7 — 30.0 MC. Each Coil Set covers the frequencies listed below:

Coil Set	General Coverage	Bandspread
D	1.7 — 4.0	3.5 — 4.0
C	3.5 — 7.3	7.0 — 7.3
B	7.0 — 14.4	14.0 — 14.4
A	14.0 — 30.0	28.0 — 30.0

NATIONAL Coil Sets to cover the low frequency range of the receiver are available as follows:

Type J 50 — 100 KC.	Type F 480 — 960 KC.
Type H 100 — 200 KC.	Type E 900 — 2050 KC.
Type G 180 — 430 KC.	

SELECTIVITY:

Voltage Ratio	Nominal Bandwidth
Crystal Filter Out	
6 DB.	3.0 KC.
60 DB.	21.5 KC.
Crystal Filter In	
Max. Selectivity 20 DB.	200 Cycles
Min. Selectivity 20 DB.	6.0 KC.

SENSITIVITY:

The sensitivity of the HRO-5A1 is 1. microvolt or better throughout the normal frequency range.

POWER INPUT:

Using Type 697 Power Pack; 75 watts at 115 volts, 50/60 cycles, 1 phase AC.

POWER OUTPUT:

Maximum output 3 watts. Output with negligible distortion 1.5 watts.

PRICES

Table Model (with tubes & A,B,C,D coils)	List $
Rack Model (with tubes & A,B,C,D coils)	List $
Table Model MCS Loud Speaker	List $
Rack Model RFSH Loud Speaker	List $
Table Model 697 Power Unit	List $
Rack Model SPU-697 Power Unit	List $

NATIONAL HRO-5C

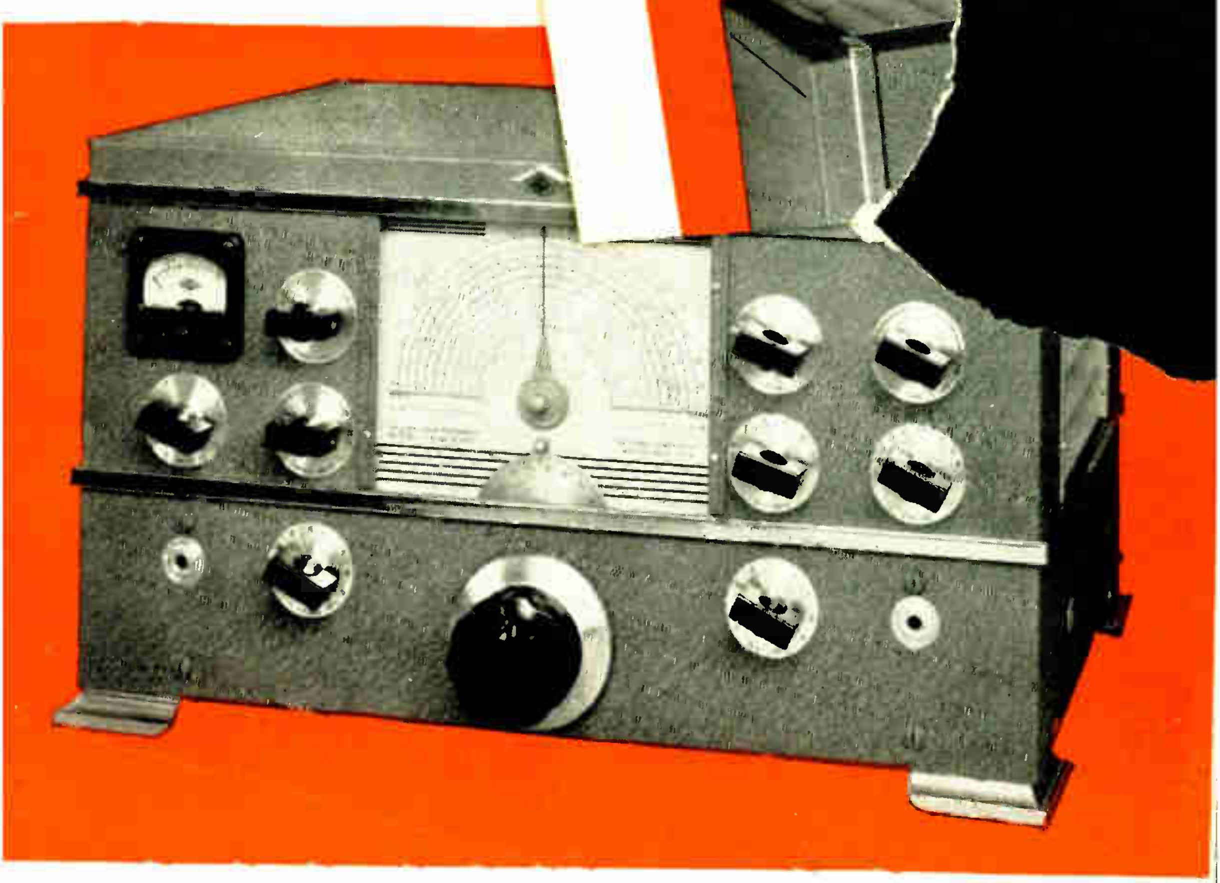
Description

The HRO-5C is a Deluxe Receiver Installation consisting of an HRO-5A1 Receiver with SPC Unit (power unit, coil container and loud-speaker) in a MRR Table Rack. Chromium-plated appearance strips and side trim strips are included.

The HRO series of receivers is an honored product of the National Company. The HRO-5A1, newest and finest of these receivers, features a number of additional refinements among which are a new highly efficient noise limiter and a redesigned flexible crystal filter. Circuit revisions have been made to further improve the performance standards of this outstanding Receiver. For a detailed description of the HRO-5A1 Receiver supplied on the HRO-5C Deluxe Installation, see page 18 in this catalog.

HRO-5A1 Receiver, with tubes and A, B, C, D Coils	**List $**
SPC Unit Combination	**List $**
MRR Table Rack 24½″ Panel Capacity	**List $**
HRO-5C Deluxe Receiver Combination	**List $**

NATIONAL NC-2-40



DESCRIPTION

Designed for the radio amateur, the NC-2-40D series of superheterodyne receivers are also suitable for general communications service in the 490 to 30,000 KC. range. Calibrated electrical bandspread tuning is provided for the 80, 40, 20, and 10 meter radio amateur bands. Features included are a full vision, easy to read, calibrated dial with 6 general coverage and 4 bandspread scales, a single tuning and band switching control knob, a stable high frequency oscillator circuit, a flexible crystal filter, a series valve noise limiter and an auxiliary numerical logging dial. These outstanding features plus conventional items such as a signal strength meter, phonograph or high level microphone pickup jack, an automatic volume control circuit, a beat frequency oscillator for CW reception, a tone control, a phones jack, and a 115-230 volt A.C. change-over switch provide the operator with a means for coping with a wide variety of receiving conditions and requirements.

CONTROLS

Band Tuning and Band Switching; R.F. Gain Control and Signal Strength Meter Switch; Audio Gain; B+ —ON OFF; Selectivity; Limiter; Tone; CW Oscillator; A.V.C.; Phasing.

SPECIFICATIONS

Frequency Range:

General Coverage:

490 KC. to 30 MC.

Band Spread:

28 to 30 MC
14 to 14.4 MC.
7 to 7.3 MC.
3.5 to 4 MC.

Selectivity:

Crystal Filter OFF

Voltage Ratio	**Nominal Bandwidth**
6 DB	4.0 KC
60 DB	22.0 KC

Crystal Filter In — 20 DB Voltage Ratio

Position 1	6.0 KC
2	4.0 KC
3	2.0 KC
4	1.0 KC
5. Max. Selectivity 200. Cycles	

SENSITIVITY

Less than 1 microvolt input produces a 6 DB signal to noise ratio.

POWER INPUT

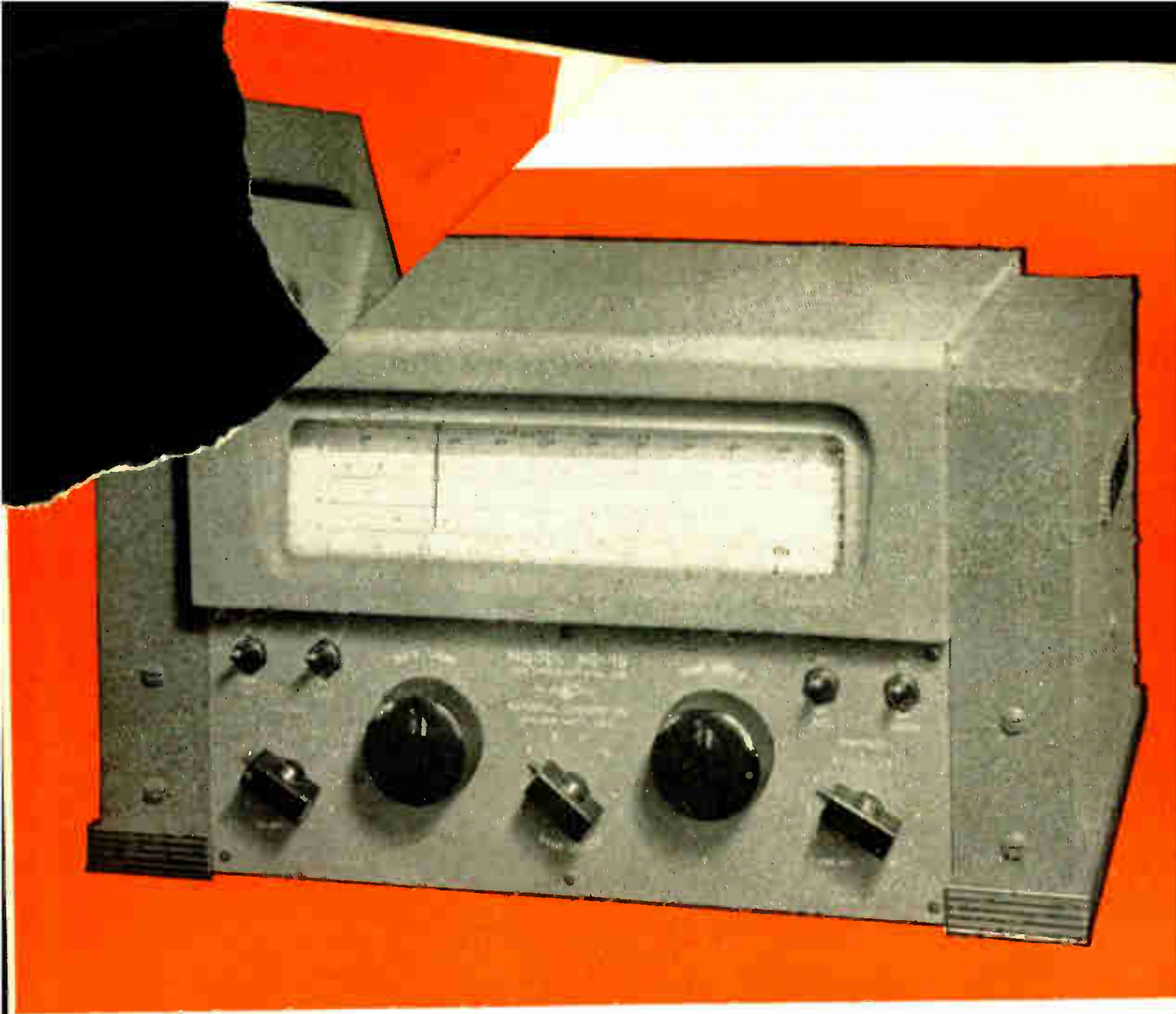
Approximately 70 watts; either 110- 120 or 220-240 volts 50/60 cycle, 1 Phase A.C. A plug and socket is provided for convenient external battery connection as necessary for battery operation.

POWER OUTPUT

A 10,000 ohm output circuit delivers 8 watts with negligible distortion.

PRICES

Rack or Table Model (with tubes)	List $
Rack or Table Model Speaker	List $



DESCRIPTION

The National NC-46 is a 105 to 130 Volt AC-DC receiver which provides 3 watts of audio output. The Receiver tunes the Broadcast and Short Wave bands and employs 10 tubes. Electrical bandspread is provided for vernier tuning. The circuit consists of a 6K8 converter-oscillator stage, two 6SG7 IF stages, 6H6 detector-limiter stage, 6SF7 AVC Amplifier, 6SJ7 CW Oscillator, 6SC7 Audio-Inverter, push-pull audio output stage with two 25L6GT tubes, and a 25Z5 Rectifier.

CONTROLS

Main Tuning Dial; Bandspread Tuning Dial; Sensitivity Control; Volume Control; Tone Switch; C. W. Oscillator Switch; AVC Switch; Limiter Switch; Band Selector Switch; B+ Switch and Power Switch.

TERMINALS

On Rear Panel; Phone Jack; B+ Terminals; 8 Ohm Spkr. terminals; Ant. Terminal; Fuse extractor post.

SPECIFICATIONS

Frequency Range:

The Frequency Range of the NC-46 Receiver is 540. Kc. to 30. Mc. covered in four bands.

Band	General Coverage	Band Spread
A	11 5 -30.0 Mc.	28.0-30.0 Mc, 40 dial div. 14 0-14.4 Mc; 56 dial div.
B	4.4 -12.0 Mc.	7.0- 7 3 Mc; 50 dial div.
C	1.55 - 4.6 Mc.	3.5- 4.0 Mc; 70 dial div.
D	0.540- 1.6 Mc.	

Sensitivity:
Approximately 5 microvolts input provides a 50 Milliwatt output over the entire range.

Selectivity:
The total bandwidth is approximately 4.5 Kc. at 6 db. down and approximately 70 db. attenuation 10 Kc. off resonance is obtained.

Automatic Volume Control:
The Receiver output with AVC operating varies less than ±4 db. with inputs ranging from 10 to 100,000 microvolts.

DIMENSIONS

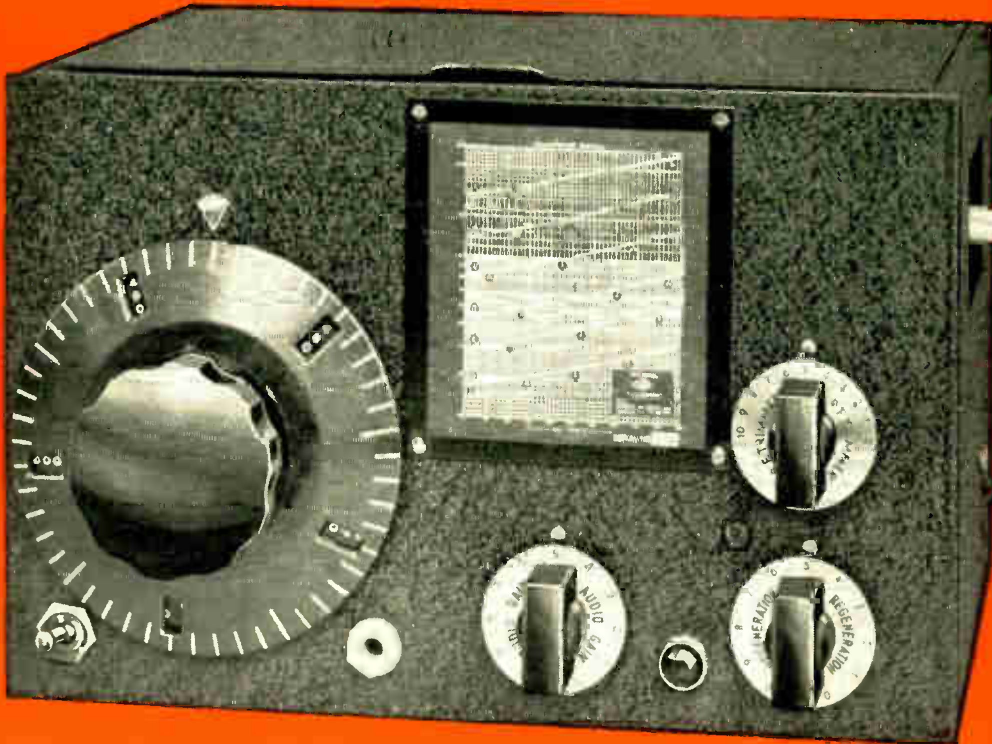
NC-46 Receiver: 9 7/16″ high by 17 3/8″ wide by 12 3/8″ deep.
Weight 32 lbs.
NC-46TS Speaker: 8 7/8″ high x 10 7/16″ wide x 7 1/2″ deep.
Weight 8 lbs.

PRICES

NC-46 Table Model Complete with Tubes — **List $**

NC-46TS Table Model Speaker — **List $**

NATIONAL 1-10A RECEIVER



The 1-10A is an improved superregenerative Receiver covering all wave lengths from 1 to 11 meters. The 1-10A is designed for use in both Amateur and Commercial services and the natural advantages inherent in a superregenerative receiver make this one of the simplest and most reliable receivers for use on these wave lengths. This Receiver is suitable for the reception of voice and tone modulated code signals. The 1-10A is supplied in a table mounting model which through virtue of its compact size can be handily used for portable operations.

The circuit of the 1-10A Receiver employs 4 tubes and consists of one stage of tuned RF, a self-quenching superregenerative detector transformer coupled to a first stage of audio which, in turn, is resistance coupled to a power output stage. Receiver controls are held to a minimum and include Audio Gain, Regeneration, RF Trimmer and Main Tuning Controls. Plug-in coil types are used to tune the frequency range of the Receiver in six tuning bands. The location of these coils in the receiver make them readily accessible for interchanging. Tuning is accomplished by a two-gang variable capacitor geared to a micrometer dial which reads directly from 0 to 500 and has a linear scale length of approximately 12 feet, requiring ten revolutions to cover any one band. The scale length plus the vernier action of the Main Dial gives the operator the equivalent of continuous bandspread tuning on all bands.

The 1–10A Receiver is designed for operation from National type 5886 Power Unit, all voltage dividers, etc., being built in so that but one B voltage lead is necessary. The 5886 Power Unit operates on 105–120 volts, 50–60 cps. This Power Unit furnishes 6.3 volts at 1.6 amperes to the heater circuit and 180 volts at 35 milliamperes to the plate and screen circuits. A 3 volt C battery, mounted in the receiver, is used to supply bias to the RF tube. The 1–10A Receiver may be operated from batteries by connecting suitable batteries to the pins of the 4 prong power plug.

Tubes

RF Amplifier	954
Detector	955
First Audio	6J5
Second Audio	6V6

Price List

1-10A Receiver, table model, complete with tubes and 6 sets of plug-in coils. **List $**

5886 Power Unit, 105-120 volt, 50-60 cps. **List $**

MCS 8" PM loud-speaker with impedance matching transformer. **List $**

NATIONAL CRU OSCILLOSCOPE

NATIONAL·OSCILLOSCOPES

CRU

CRU-P

CRU WITH THE CRU-P PANEL

Description

The CRU Oscilloscope is a compact inexpensive instrument whose capabilities make it outstanding in its field. Amateurs and electronic experimenters will recognize this 2" scope as an indispensible item of equipment to guarantee the efficient operation of their stations. Put the CRU scope to work in your station and watch it:

Measure Percentage Modulation.

Check distortion, excitation, overmodulation, etc., by the Trapezoidal pattern method.

Monitor RF and Audio circuits continuously while you are on the air.

Test Audio and RF circuits where extreme sensitivity is not required.

The circuit of the CRU is simple yet ample having a self contained power supply and controls for brilliancy and focus, a potentiometer for controlling the amplitude of the horizontal deflection as well as a built-in 60 cycle sweep. Approximately 100 volts dc. will give a 1" deflection on the CRU screen.

Tubes

Cathode-Ray	2AP1-A
Rectifier	6X5

Controls

A.C. ON/OFF: the A.C. line switch.

Intensity: A potentiometer controlling the brilliancy of the pattern.

Focus: A potentiometer controlling the clarity of the scope image.

Sweep: A potentiometer controlling the length of the pattern.

"Ext."—"60 cycle": A two position switch, which when on "Ext." connects the horizontal deflection plates to the horizontal terminal strip at the rear of the cabinet. In the "60 cycle" position the 60 cycle A.C. sweep is connected to the horizontal deflection plates.

BSW: A pair of insulated beam switch control terminals permitting connection to a switch or relay so that a trace appears on the screen only during transmission periods.

Prices

CRU-Table Model Oscilloscope, Less tubes **List $**

CRU-P Rack Panel and Control Plate (to rack mount CRU Oscilloscope) **List $**

NATIONAL POWER SUPPLIES

National Power Supplies are specially designed for high frequency receivers, and include efficient filters for RF disturbances as well as for hum frequencies.

686S, Table model (165 V., 50 MA.), for operation from 6.3 volts DC, with vibrator. **List $**

SPU-686S Rack Model **List $**

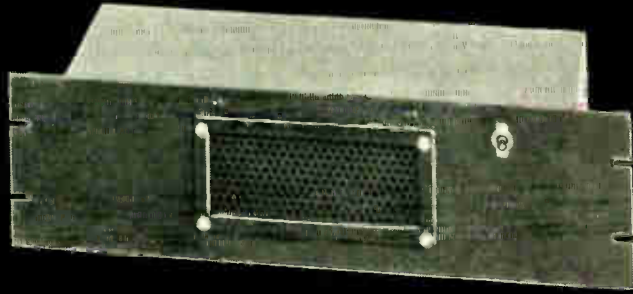
697 Table Model (240V., 70 Ma. and 6.3 V., 3.4 A.), for operation from 115/230 Volts, 50/60 cps. A.C. **List $**

SPU-697 Rack Model **List $**

5886 Table Model (155 V., 50 Ma. and 6.3 V., 2.5 A.) for operation from 115 Volt, 50/60 cps. A.C. **List $**

POWER SUPPLIES

686S, 697, 5886



SPU-686S, SPU-697

MICRO (MS) SWITCH

A DIVISION OF FIRST INDUSTRIAL CORPORATION

Freeport, Illinois

Branch Offices

CHICAGO 6..308 W. Washington Street
NEW YORK 17........101 Park Avenue
CLEVELAND 3.......4900 Euclid Avenue
LOS ANGELES 14...1709 West 8th Street
BOSTON 16........126 Newbury Street

•

The Precise, Small Lightweight, Sensitive Switch for Radio Applications

Micro Switch precision snap-action switches have proved invaluable for applications that call for switching substantial amounts of power by a unit operating in a small space. Micro Switch products are important electrical switching units for electrical mechanisms that make change, package products, control temperatures, heat water, bottle fluids, limit machine tools, record airplane flights, control electronic tubes and perform thousands of other diversified electrical control functions.

MICRO SWITCH Products Meet These Requirements

Small Size . . . No larger than your thumb, the basic, plastic enclosed switch measures 11/16" x 27/32" x 1 15/16".

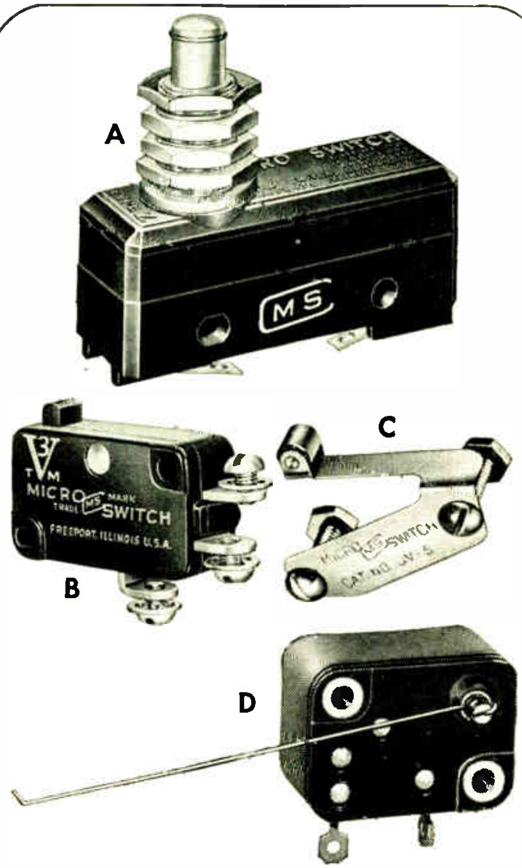
Light Weight . . . With pin-type plunger, the plastic enclosed switch weighs less than one ounce.

Long Life . . . Patented three-bladed beryllium copper spring gives millions of accurate repeat operations.

Small Operating Force . . . Force required to operate the switch may be as little as one ounce . . . or as much as 60 ounces.

Small Operating Movement . . . Movement of the operating plunger may be as little as .0004".

Good Electrical Capacity . . . Switch is Underwriters' listed and rated at 1200 V.A. at 125 to 460 volts a.c.



A. General Purpose Basic Switch with panel mounting. This "MICRO" basic switch is handy and useful as a door switch, or as a manual or mechanical push button switch. The threaded stem, with two thin brass hex nuts and two steel lock nuts aids adjustable location with respect to the panel. The internal switch mechanism is protected from excessive overtravel by a stop ring located near the tip of the plunger. This type switch proves both handy and useful.

B. The "MICRO" V3-1 Small Precise Switch. For a switch that must perform in small quarters the "MICRO" V3-1 switch is of a size to meet these requirements. Small but accurate and dependable the V3-1 is provided with two mounting holes, one elongated to provide greater accuracy in locating. Flat bosses on side add to ease of stacking or grouping when requirements demand they be used that way.

C. JV-5 Actuator for use with V3-1 Switch. The JV-5 Auxiliary Actuator with roller is designed for rapid cam or slide actuation of the V3-1 switch. The frame is stainless steel with the oil-impregnated bronze bearing serving as the roller.

D. The "MICRO" V3-12 Switch. Low torque features this switch which can be actuated with .14 ounce-inches—practically a feather touch. Pretravel of the actuating arm is 20° maximum with overtravel 20° minimum. It also features high resistance to shock, and in addition has clean make and break without contact bounce. Being enclosed keeps out dust and dirt and assures trouble-free operation. Time-tested and proved dependability, based on experience gained in making millions of switches, gives users an assurance of freedom from trouble. Actuating wire not furnished.

CHOOSE FROM

TYPE 30

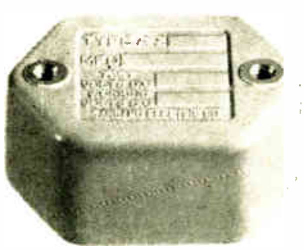
Type 30 Sangamo Tubular Capacitors, molded in a thermo-setting, smooth brown finish plastic material are permanently sealed against moisture resulting in low power factor, long life and successful operation at higher ambient temperatures.

TYPE C

Types C and K plain or silvered mica capacitors, members of the Sangamo quality mica family, insure dependability and life in radio receiver and commercial low voltage applications requiring small capacitance values.

TYPE K

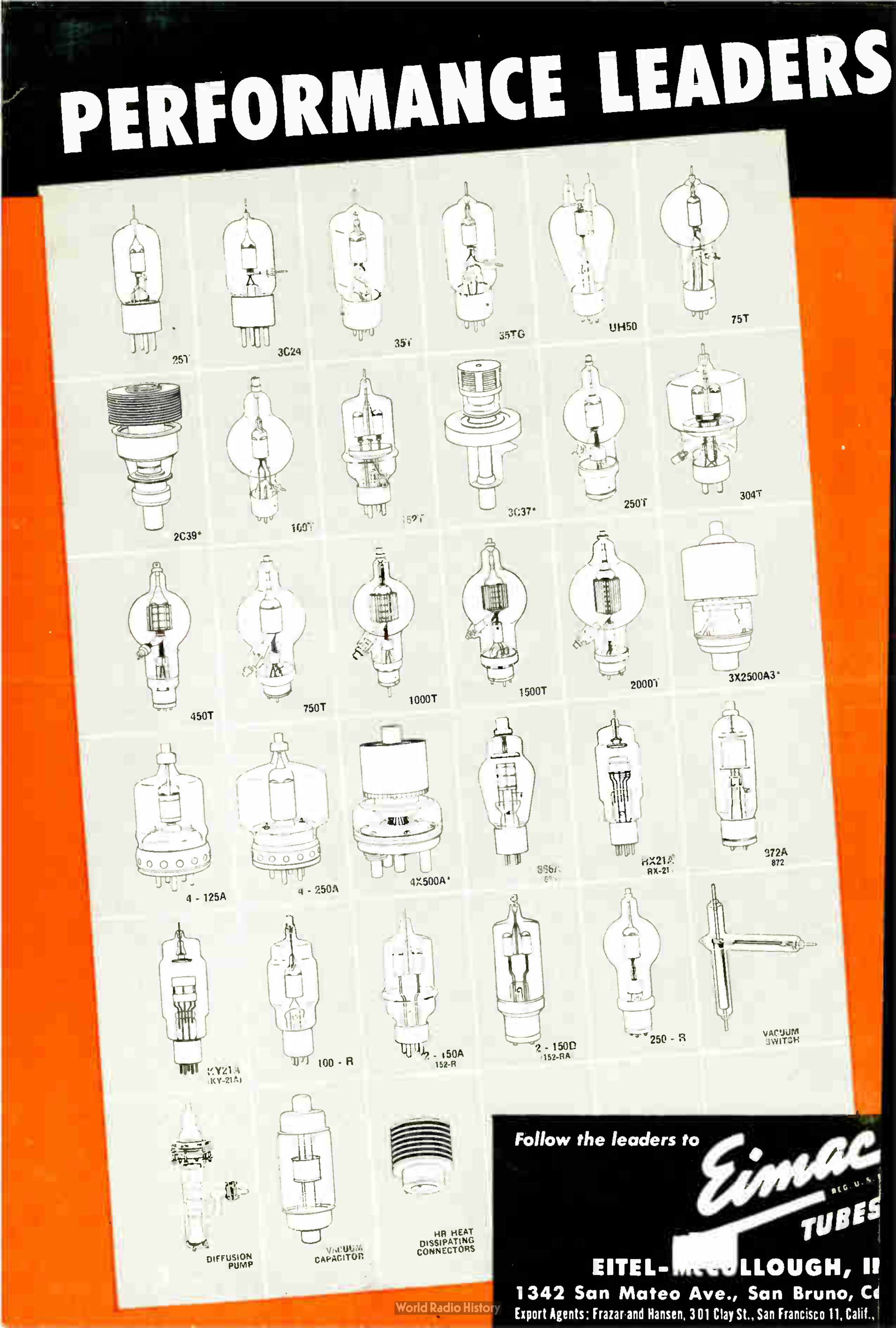
Types A and H famous high quality Sangamo mica capacitors are precision built to provide continuous, dependable service in industrial and transmitting applications for which they are designed.



TYPE A

TYPE H

SANGAMO ELECTRIC COMPANY SPRINGFIELD ILLINOIS


PERFORMANCE LEADERS
25T
3C24
35T
35TG
UH50
75T
2C39*
100T
152T
3C37*
250T
304T
450T
750T
1000T
1500T
2000T
3X2500A3*
4 - 125A
4 - 250A
4X500A*
866A
RX21A
RX-21
872A
872
KY21A
(KY-21A)
100 - R
2 - 150A
152-R
2 - 150D
152-RA
250 - R
VACUUM SWITCH
DIFFUSION PUMP
VACUUM CAPACITOR
HR HEAT DISSIPATING CONNECTORS
Follow the leaders to
Eimac
REG. U. S.
TUBES
EITEL-McCULLOUGH, I
1342 San Mateo Ave., San Bruno, C
Export Agents: Frazar and Hansen, 301 Clay St., San Francisco 11, Calif.,

For over a decade, Eimac tubes have led the field in *performance*—the acid test of electronic equipment. Ultra-modern Eimac tubes provide maximum power and efficiency for today's equipment, and are ready and waiting for the needs of tomorrow.

These pages contain basic data on many Eimac products. Complete information on any of these world-famous Eimac tubes is yours for the asking. Write for it today!

EIMAC TRANSMITTING TUBES

	EIMAC TUBE TYPES	ELECTRICAL							MECHANICAL				MAX. RATINGS						TUBE PRICE	RECOMMENDED HR. HEAT DISSIPATING CONNECTORS	
		FIL. VOLTS	FIL. AMPS.	AMP. FACTOR	GRID-PLATE, UUF	INPUT, UUF	OUTPUT, UUF	TRANSCONDUCTANCE, UMHOS	BASE	BASING	MAX. LENGTH, INCHES	MAX. DIAMETER INCHES	PL. VOLTAGE	PL. CURRENT, MA.	SCREEN VOLTAGE	SCREEN DISSIPATION	GRID DISSIPATION, WATTS	PL. DISSIPATION WATTS		PLATE	GRID
TRIODES	25T	6.3	3.0	29	1.6	2.4	0.4	2500	M8-071	3G	4.38	1.43	2000	75	...	.	7	25	$ 6.00	HR-1	...
	3C24	6.3	3.0	25	1.6	1.8	0.2	2500	M8-071	3G	4.38	1.43	2000	75			8	25	6.00	HR-1	HR-1
	35T	5.0	4.0	30	1.9	4.0	0.2	2850	M8-078	3G	5.5	1.81	2000	150			15	50	7.00	HR-3	
	35TG	5.0	4.0	30	1.9	1.9	0.2	2850	M8-078	2M	5.76	1.81	2000	150			15	50	8.00	HR-3	HR-3
	UH50	7.5	3.25	13	2.4	2.2	0.4	....	M8-078	2M	7.0	2.69	1250	125			13	50	15.00	HR-2	HR-2
	75TH	5.0	6.5	20	2.3	3.5	0.25	4150	M8-078	2M	7.25	2.81	3000	225			16	75	10.50	HR-3	HR-2
	75TL	5.0	6.5	11	2.3	2.2	0.4	3350	M8-078	2M	7.25	2.81	3000	225			13	75	10.50	HR-3	HR-2
	2C39*	6.3	1.1	..	1.95	6.5	0.30	21,000		...	2.75	1.26	1000	100			3	100	30.00	..	...
	100TH	5.0	6.2	40	2.0	2.9	0.4	5500	M8-078	2M	7.75	3.19	3000	225			20	100	15.00	HR-6	HR-2
	100TL	5.0	6.5	12	2.3	2.0	0.4	2300	M8-078	2M	7.75	3.19	3000	225		.	15	100	15.00	HR-6	HR-2
	152TH	5 or 10	13 or 6.5	20	4.7	7.0	0.5	8300	5000B	4BC	7.63	2.56	3000	450			30	150	24.00	HR-5	HR-6
	152TL	5 or 10	13 or 6.5	11	5.0	4.8	0.8	7150	5000B	4BC	7.63	2.56	3000	500			25	150	24.00	HR-5	HR-6
	3C37*	6.3	2.4	.	3.50	4.25	0.60	8000	...	..	3.10	1.50	1000	....				150	45.00	.	
	250TH	5.0	10.5	37	2.9	5.0	0.7	6650	5001B	2N	10.13	3.81	4000	350			40	250	27.50	HR-6	HR-3
	250TL	5.0	10.5	13	3.5	3.0	0.5	2650	5001B	2N	10.13	3.81	4000	350			35	250	27.50	HR-6	HR-3
	304TH	5 or 10	26 or 13	20	9.4	14.0	1.0	16,700	5000B	4BC	7.63	3.56	3000	900			60	300	50.00	HR-7	HR-6
	304TL	5 or 10	26 or 13	11	10.0	10.0	1.5	16,700	5000B	4BC	7.63	3.56	3000	1000	.		50	300	50.00	HR-7	HR-6
	450TH	7.5	12.0	38	4.7	8.1	0.8	6650	5002B	4AQ	12.63	5.13	6000	500	.		80	450	70.00	HR-8	HR-8
	450TL	7.5	12.0	19	5.0	6.6	0.9	6060	5002B	4AQ	12.63	5.13	6000	500			65	450	70.00	HR-8	HR-8
	750TL	7.5	21.0	15	4.5	6.0	0.8	3500	5003B	4BD	17.0	7.13	6000	1000	...		100	750	150.00	HR-8	HR-8
	1000T	7.5	16.0	30	4.0	6.0	0.6	9050	5004B	4AQ	12.63	5.13	6000	750	...	.	80	1000	125.00	HR-9	HR-9
	1500T	7.5	26.0	24	7.0	9.0	1.3	10,000	5005B	4BD	17.0	7.13	6000	1250	..		125	1500	200.00	HR-8	HR-9
	2000T	10.0	26.0	20	9.0	13.0	1.5	11,000	5006B	4BD	17.75	8.13	6000	1750	.		150	2000	250.00	HR-8	HR-9
	3X2500A3*	7.5	48	20	20	48	1.2	20,000		.	9.0	4.25	5000	2000	.	.	125	2500	165.00		
TETRODES	4-125A	5.0	6.2	6.2	0.03	10.3	3.0	2450	5008B	.	5.69	2.72	3000	225	400	30	5	125	25.00	HR-6	
	4-250A	5.0	14.5		0.06	12.7	4.5	4000	5008B		6.38	3.56	4000	350	600	50	5	250	36.00	HR-6	
	4X500A*	5.0	12.2		0.05	11.1	3.75	5200	....		4.32	2.57	4000	300	450	30	5	500	85.00	...	

* External Anode requiring forced-air cooling, Cathode Current

EIMAC RECTIFIERS

	MERCURY VAPOR RECTIFIERS				HIGH VACUUM RECTIFIERS			
	866A (866)	RX21A (RX-21)	872A (872)	KY21A (KY-21 Grid Control)	100-R	2-150A (152-R)	2-150D (152-RA)	250-R
1. Filament Voltage..........	2.5	2.5	5.0	2.5	5.0	5.0	5.0	5.0
2. Filament Current..........	5.0 amperes	10 amperes	7.5 amperes	10 amperes	6.5	13.0	13.0	10.5
3. Peak Inverse Voltage......	10,000	11,000	10,000	11,000	40,000	30,000	30,000	60,000
4. Peak Plate Current........	1.0 amperes	3 amperes	5.0 amperes	3 amperes		....	..	..
5. Average Plate Current	.25 amperes	.75 amperes	1.25 amperes	.75 amperes	.100 amperes	.150 amperes	.150 amperes	.250 amperes
Price	$1.75	$8.00	$7.50	$10.00	$13.50	$15.00	$15.00	$20.00

EIMAC VACUUM CAPACITORS

Type.......	VC6-20	VC12-20	VC25-20	VC50-20	VC6-32	VC12-32	VC25-32	VC50-32
Capacity ...	6-mmfd	12-mmfd	25-mmfd	50-mmfd	6-mmfd	12-mmfd	25-mmfd	50-mmfd
Rating .. RF Peak	20-KV	20-KV	20-KV	20-KV	32-KV	32-KV	32-KV	32-KV
Price.......	$12.00	$13.50	$16.50	$20.00	$14.00	$16.00	$19.00	$22.50

HEAT DISSIPATING CONNECTORS

Type	Hole Dia.	Price	HR-5	125	$ .80
HR-1	.052	$ .60	HR-6	360	.80
HR-2	0625	.60	HR-7	125	1.60
HR-3	070	.60	HR-8	570	1.60
HR-4	.1015	.80	HR-9	570	3.00

EIMAC DIFFUSION PUMP

HV-1 Diffusion Pump	PRICE ON APPLICATION
An air-cooled vacuum pump of the oil-diffusion type. Capable of reaching an ultimate vacuum of 4×10^{-7} mm. of mercury when used with a suitable mechanical forepump. Speed (without baffle) approximately 67 liters/second at 4×10^{-4} to 4×10^{-6} mm.	
Eimac Pump Oil	

EIMAC VACUUM SWITCHES

TYPE	GENERAL DATA	PRICE
VS-2..	Single pole double throw switch within a high vacuum adaptable for high voltage switching. Contact spacing .015". Switch will handle R-f potentials as high as 20 Kv. In DC switching will handle approximately 1.5 Amps at 5 Kv.	$12.00
VS-1..	Same as above except for slightly smaller glass tubulation.	$12.00

MILLEN MODERN PARTS

MILLEN RADIO PRODUCTS are well designed MODERN PARTS for MODERN CIRCUITS, attractively packaged, moderately priced, and fully guaranteed. They have been designed with a view toward easy and practical application as well as efficient performance. For instance, the terminals are located so as to provide shortest possible leads, mounting feet are designed for easy insertion of screws and socket contacts, so that the solder won't run down inside them and make impossible the insertion of the tube, etc. Thus our slogan, "Designed for Application." Our general catalog is available for the asking either from your favorite parts supply house or direct from the factory.

11000, 12000, 13000, 14000 SERIES CONDENSERS
.077″ air gap is for 3000 volt peak rating

MILLEN TYPE

Code	Capacity per side Max.	Capacity per side Min.	Air Gap	Voltage Rating	Net Price
11035	36	4.6	.077″	3000	$6.90
11050	51	6.5	.077	3000	7.14
11070	74	9.5	.077	3000	7.80
13035	35	4.9	.077	3000	4.56
13050	49.5	6.3	.077	3000	5.20
13070	71	7.3	.077	3000	5.88
14200	204	10.7	.077	3000	14.00
14100	90.5	12.9	.171	6000	12.00
14050	50		.171	6000	7.20
14060	60		.265	9000	12.00

CONVENTIONAL SINGLE SECTION TYPE

Code	Capacity per section Min.	Capacity per section Max.	Air Gap	Finish on Plates	Net Price
12935	9	37	.176″	Polished	$4.32
12936	9	37	.176	Plain	3.90
12536	6	43	.077	Plain	2.40
12551	7	55	.077	Plain	2.70
12576	9	76	.077	Plain	3.00
12510	12	101	.077	Plain	3.60
12515	18	151	.077	Plain	4.50

CONVENTIONAL DOUBLE SECTION TYPE

Code	Capacity per section Min.	Capacity per section Max.	Air Gap	Finish on Plates	Net Price
12035	6	43	.077″	Polished	$4.32
12036	6	43	.077	Plain	3.90
12050	7	55	.077	Polished	5.10
12051	7	55	.077	Plain	4.32
12075	9	76	.077	Polished	5.61
12076	9	76	.077	Plain	5.40

Code	Description	Net Price
10000	Worm Drive Unit	$1.50
10001	Drum Meter Dial-0-100	1.85
10007	1[illegible]″ Nickel Silver Inst. Dial-0-100	.60
10008	3½″ Nickel Silver Inst. Dial-0-100	1.00
10050	Dial Lock	.45
10060	Shaft Lock for ¼″ Shafts	.36
10061	Shaft Lock	.36
10065	Vernier Drive Unit	.36
15001	Neutral Condenser 0.7–4.3	.90
15002	Neutral Condenser 0.5–13.5	1.05
15003	Neutral Condenser 1.5–8.5	.90
15005	Neutral Condenser 3.4–14.6	2.00
15006	Neutral Condenser 2.8–9.1	3.00
20015	Steatite Ultra Midget 15 mmfd SS	.75
20035	Steatite Ultra Midget 35 mmfd SS	1.00
20050	Steatite Ultra Midget 50 mmfd SS	1.20
20100	Steatite Ultra Midget 100 mmfd SS	1.50
20920	Steatite Ultra Midget 20 mmfd DS	1.20
20935	Steatite Ultra Midget 35 mmfd DS	1.40
21050	Steatite Ultra Midget 50 mmfd SS	1.75
21100	Steatite Ultra Midget 100 mmfd SS	1.90
21140	Steatite Ultra Midget 140 mmfd SS	2.10
21935	Steatite Ultra Midget 35 mmfd DS	1.90
22075	Steatite Midget 75 mmfd SS	1.32
22100	Steatite Midget 100 mmfd SS	1.3[illegible]
22140	Steatite Midget 140 mmfd SS	1.62
22915	Steatite Midget 15 mmfd DS	1.20
22935	Steatite Midget 35 mmfd DS	1.30
22950	Steatite Midget 50 mmfd DS	1.50
23075	Steatite Dual Midget 75 mmfd per section SS	2.60
23100	Steatite Dual Midget 100 mmfd per section SS	2.50
23925	Steatite Dual Midget 25 mmfd per section DS	2.25
23950	Steatite Dual Midget 50 mmfd per section DS	2.50
24100	100 mmfd per section, Single spaced	2.75
24935	35 mmfd per section, Double spaced	2.75
26025	3.2–25 Air Padder	.96
26050	4–50 Air Padder	1.08
26075	4.3–76 Air Padder	1.20
26100	5–97 Air Padder	1.32
26920	4.5–20 Air Padder	1.40
26935	5.5–36 Air Padder	1.50
27030	30 mmfd Mica Padder	.21
30001	Standoff, ½ x 1[illegible], QuartzQ	.15
30002	Standoff, ½ x 2⅞, QuartzQ	.21
30003	Standoff, [illegible] x 2[illegible], QuartzQ	.55
30004	Standoff, [illegible] x 4⅞, QuartzQ	.65
31001	Standoff, ½ x 1, Isolantite	.20
31002	Standoff, ½ x 2½, Isolantite	.27

JAMES MILLEN MFG. CO., INC.
MAIN OFFICE AND FACTORY
MALDEN, MASSACHUSETTS, U.S.A.

DESIGNED for APPLICATION

Code	Description	Net Price
31003	Standoff, ¾ x 2, Isolantite	$.30
31004	Standoff, ¾ x 3½, Isolantite	.42
31011	Cone, ¾ x ½, Steatite	.10
31012	Cone, 1 x 1, Steatite	.21
31013	Cone, 1½ x 1, Steatite	.27
31014	Cone, 2 x 1, Steatite	.42
31015	Cone, 3 x 1½, Steatite	.45
32100	Steatite Bushing for ⅜" hole	.30
32101	Steatite Bushing for ½" hole	.35
32102	Steatite Bushing for ¼" hole	.20
32103	Steatite Bushing for ¾" hole	.45
32150	Isolantite Thru-bushing, for ¼" hole	.05
32201	Steatite Bushing and Hardware	.75
32300	Isolantite Bushing	1.80
33002	Crystal Socket	.25
33004	4 Prong Socket	.27
33005	5 Prong Socket	.27
33006	6 Prong Socket	.27
33007	7 Prong, Large, Socket	.34
33008	8 Prong, Octal, Socket	.27
33087	Base Clamp for 807 etc.	.30
33102	Crystal Socket	.25
33105	Acorn Socket, QuartzQ	.90
33202	Crystal Socket	.25
33307	Socket, Midget 9000 Series with Shield	.65
33407	Socket, Midget 9000 Series less Shield	.45
33888	Aluminum Shield for 33008	.18
33991	Socket for 991 etc.	.45
34010	Shielded 10 MH receiving	.75
34100	Universal 2.5 MH	.36
34101	Universal 2.5 MH, less Standoff	.30
34102	Commercial type 2.5 MH	.36
34110	Universal air core Transmitting	1.00
34154	Transmitting Choke	1.50
34210	General Purpose RFC 10 MH	.60
34225	General Purpose RFC 25 MH	.75
34240	General Purpose RFC 40 MH	.75
34285	General Purpose RFC 85 MH	1.25
34800	Interruption Frequency Oscillator Coil	1.20
35151	Steatite Antenna Insulators	.30
36001	Ceramic Plate Cap, 9/16" for 866 etc.	.21
36002	Ceramic Plate Cap, ⅜" for 807 etc.	.21
37001	Black Bakelite Safety Terminal	.40
37104	Four Terminal, Black Bakelite	.60
37202	Steatite Plates, Pr.	.30
37211	Bracket	.15
37222	Terminal Posts, Pr.	.40
37302	Two Terminal, Steatite	.60
37303	Three Terminal, Steatite	.70
37304	Four Terminal, Steatite	.80
37305	Five Terminal, Steatite	.90
37306	Six Terminal, Steatite	1.00
37501	Low Loss Mica Bakelite Safety Terminal	.55
38001	Isolantite 3/16" O.D. Beads (Pk of 50)	.30
38501	100 Beads, 5/16" dia., QuartzQ	.60
39001	Truly Flexible Isolantite	.36
39002	Conventional	.36
39003	Solid Brass N.P.	.21
39005	Universal Joint, Non-Insulated	.36
39006	Slide Action	.36
40205	Midget Plug	.24
40305	Intermediate size plug	.45
41205	Midget Socket	.30
41305	Intermediate size socket	.45
43001	QuartzQ blank form and plug	.90
43011		1.25
43021	Midget coils for each	1.25
43041	band. Mounted on No. 40205	1.25

Code	Description	Net Price
43081	plug. No. 1 at end of code means	$1.25
44000	QuartzQ form 1¼" dia. x 3¼"	.75
44001	QuartzQ blank form and plug	1.20
44005		1.50
44010		1.50
44020		1.50
44040	"100 watt" coils	1.50
44080	for each band. Mounted on	1.90
44500	Swinging link and socket	1.75
45000	Coil Form, 1" dia. no p., low loss mica base Phenolic	.35
45004	Coil Form, 1" dia. 4 p., low loss mica base Phenolic	.45
45005	Coil Form, 1" dia. 5 p., low loss mica base Phenolic	.45
45500	Coil Form, ⅝" dia., Steatite	.45
46100	Coil Form, 1½" dia. no p., QuartzQ	.45
47001	Coil Form, ½" dia., QuartzQ	.10
47002	Coil Form, ½" dia., QuartzQ	.15
47003	Coil Form, ¾" dia., QuartzQ	.35
47004	Coil Form, ¾" dia., QuartzQ	.45
55001	Sheet, 3 x 8½ x .1, QuartzQ	.45
58001	Coil Dope, 2 oz., QuartzQ	.50
59001	Panel Marking Decalcomania Kit	1.25
59002	Decalcomania Kit — Black	1.25
69041	Ultra-High Coil Form	.60
74001	Permeability Tuned Shielded Form	1.85
74002	Untuned Shielded Form	1.50
74400	Octal Base and Shield	.75
77083	"83" Hash Filter 250MA	1.00
77866	"866" Hash Filter 500MA	1.25 pr.
77872	"872" Hash Filter	1.40 pr.
79020	14mc Band Wave Trap	.90
79040	7mc Band Wave Trap	.90
79080	3.5mc Band Wave Trap	.90
79160	1.7mc Band Wave Trap	.90
	Air Trimmed	
60454	456 Diode Air Core	4.50
60455	456 Interstage (1) Air Core	4.50
60456	456 Interstage (2) Air Core	4.50
60501	5000 Interstage (2) Air Core	4.50
60502	5000 Diode Air Core	4.50
60503	5000 FM Interstage Air Core	4.50
60504	5000 FM Disc Air Core	4.50
62161	1600 Interstage Iron Core	4.50
62162	1600 Diode Iron Core	4.50
62454	456 Diode Iron Core	4.50
62456	456 Interstage Iron Core	4.50
63163	1600 BFO Air Core	4.50
63456	456 BFO Air Core	4.50
63503	5000 BFO Air Core	4.50
	Permeability Tuned	
64454	456 Diode (2)	2.25
64456	456 Interstage (2)	2.25
65456	456 BFO	2.25
70721	Complete set of four Wavemeters, In case	4.00
90600	Hetrofil	18.00
90605	Range 2.8 to 9.7 mc. Wavemeter	4.50
90606	Range 9.0 to 28 mc. Wavemeter	4.50
90607	Range 26 to 65 mc. Wavemeter	4.50
90608	Range 50 to 140 Wavemeter	4.50

JAMES MILLEN MFG. CO., INC.

MAIN OFFICE AND FACTORY

MALDEN, MASSACHUSETTS, U.S.A.

29 DIFFERENT RECTIFIER TUBES *Now Replaceable*—

WITH THE FEDERAL SELENIUM RECTIFIER

(No. 403D2625)

IT'S SMALLER . . .

COSTS LESS . . .

LASTS LONGER . . .

RUNS COOLER . . .

READILY INSTALLED!

—FIRST OF ITS KIND IN THE INDUSTRY!

RADIO HAMS: Here's a new component for your sets—a dry-plate rectifier to use in place of the conventional rectifier tube. It will give your radio receivers new performance, instant starting, longer life. In every way, Federal's miniature Selenium Rectifier makes the ideal power supply for all AC-DC sets and it has been adopted as a standard component by many leading radio manufacturers.

Among its many other diversified applications are: —vibrator power supplies, television sets, phonograph combinations, heating lamps, door chimes, electric train accessories, scientific research apparatus, stethoscopic and bacteriological equipment, measuring, intercommunications and electrical musical instruments and other electronic devices.

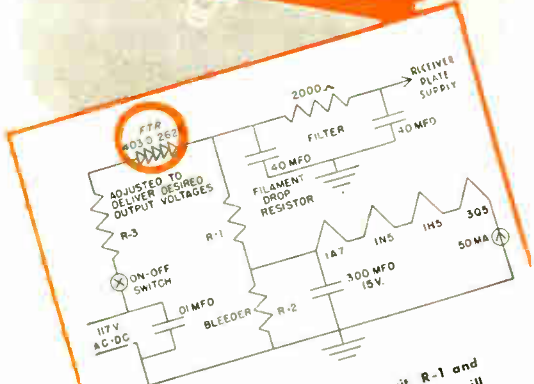
These rectifiers can be obtained from your dealer or direct from Federal Telephone and Radio Corporation, Newark 1, New Jersey. Price $1.60 each net. Send $12 for standard package of twelve units. Write to Dept. F1065 for complete technical literature on how to apply rectifiers.

Replacement for these Tubes:

5T4	5Z3	5Y3	6X5	6Y5	7Y4	25Z6	35Z4	50Y6	117Z6
5U4	5W4	5Y4	0Z4	6Z5	12Z3	35W4	35Z5	50Z7	0Y4
5V4	5X4	5Z4	80	12Z5	25Z5	35Z3	35Z6	117Z3	

Electrical Characteristics:

Maximum RMS Voltage	130 Volts
Maximum Inverse Voltage	380 Volts
Maximum Peak Current	1200 ma.
Maximum RMS Current	325 ma.
Maximum DC Output	100 ma.
Approximate Rectifier Drop	5 Volts



Recommended simplified circuit. R-1 and R-2 resistors existing in receiver. R-3 will be 25 to 150 ohms as required for proper voltage in set.

Federal Telephone and Radio Corporation

In Canada:—Federal Electric Manufacturing Company, Ltd., Montreal.
Export Distributors:—International Standard Electric Corp. 67 Broad St., N.Y.C.

An IT&T Associate

Newark 1, New Jersey

TURNER *Microphones*

Faithfully Reproduce Your Complete Messages

These famous microphones, priced within the range of every amateur, amplify all vibrations received by the diaphragm without adding any of the harmonics to assure clear, sharp communications without distortion. You can rely on Turner under all climatic and acoustic conditions.

211 DYNAMIC

No. 211 is a Rugged Dynamic utilizing a new type magnet structure and acoustic network. The high frequency range has been extended and the extreme lows have been raised 2 to 4 decibels to compensate for overall deficiencies in loud speaker systems. Unique diaphragm structure results in extremely low harmonic and phase distortion without sacrificing high output level. Tilting head, balanced line output connection. Chrome or gunmetal finish.

33D Dynamic 33X Crystal

The full satin chrome finish of 33D Dynamic adds class to any rig. 90° tilting head gives semi- or non-directional pick-up. Output level –54DB. Range 40–9,000 cycles. Ruggedly built for recorder or P.A. work. Built in transformer free from hum pick-up. Available in 200 or 500 ohms or high impedance. 33X Crystal same appearance as 33D. Level –52DB. Range 30–10,000 cycles.

22X 22D

22X Crystal is tops in performance. Reproduces clean and sharp. Smart engineering cuts feedback to minimum. Tilting head and removable 7-foot cable set. Built-in wind-gag permits outdoor operation. Crystal impregnated against moisture. Automatic barometric compensator. Chrome type finish. Level –52 DB. Range 30–7,000 cycles.

22D Dynamic is identical in appearance with 22X but has high level dynamic cartridge. Dependable indoors or out. Output –54 DB. Range 30–8,000 cycles. 200 or 500 ohms or high impedance.

NEW TURNER CHALLENGERS

Plus Performance at Low Cost

Model CX

Crystal, in rich brushed chrome finish, with 7 foot removable cable set using Amphenol connectors. Level –52 DB. Range 50–7,000 cycles.

Model CD

Dynamic, same style and finish as CX, with removable 7 foot cable set. In 200–250 ohms, 500 ohms or hi-impedance. Level –52 DB. Range 50–7,000 cycles.

Model BX

Crystal mike for recording, P.A. and ham work. Bronze enamel finish. Level –52 DB. Range 50–6,000 cycles. An excellent unit. With 7 foot cable.

Model BD

Dynamic, same finish as BX. Works indoors or out. Level –52 DB. Range 50–6,000 cycles. 200–250 ohms, 500 ohms or high impedance with 7 foot cable.

Han-D

Hang it, hold it, use it on desk or floor stands. Han-D does the job of several mikes. Available as 9X Crystal, in brushed chrome finish, Level –48DB, or 9D Dynamic in brushed chrome or gunmetal. Level –50DB, 200 or 500 ohms or hi-impedance.

FREE Turner Microphone Catalog with complete information and prices on Turner Microphones. Write for your free copy.

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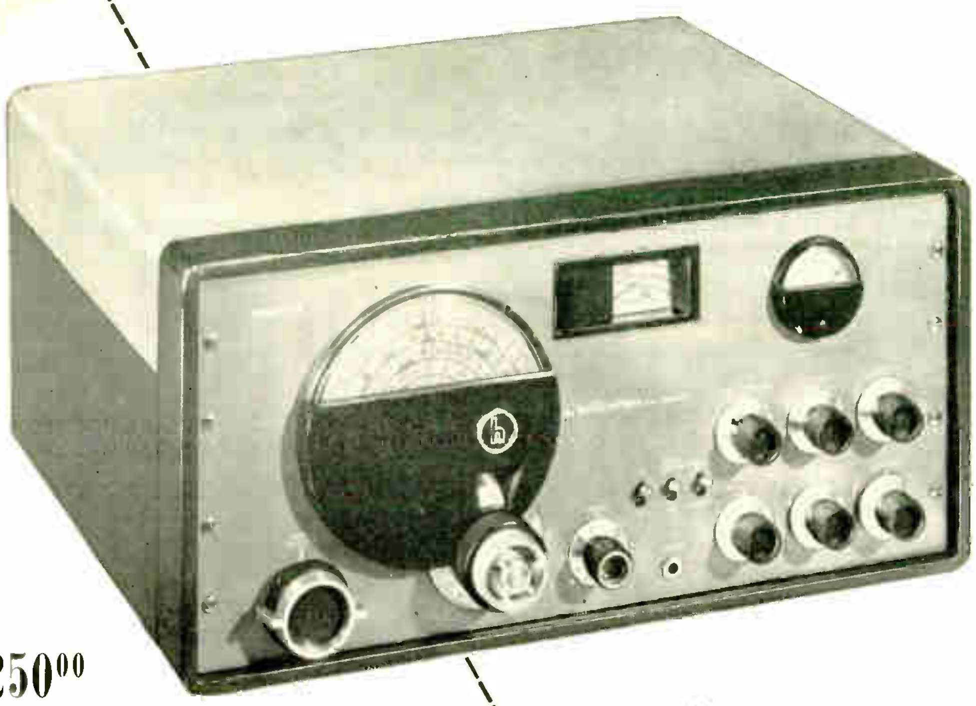
THE TURNER CO.

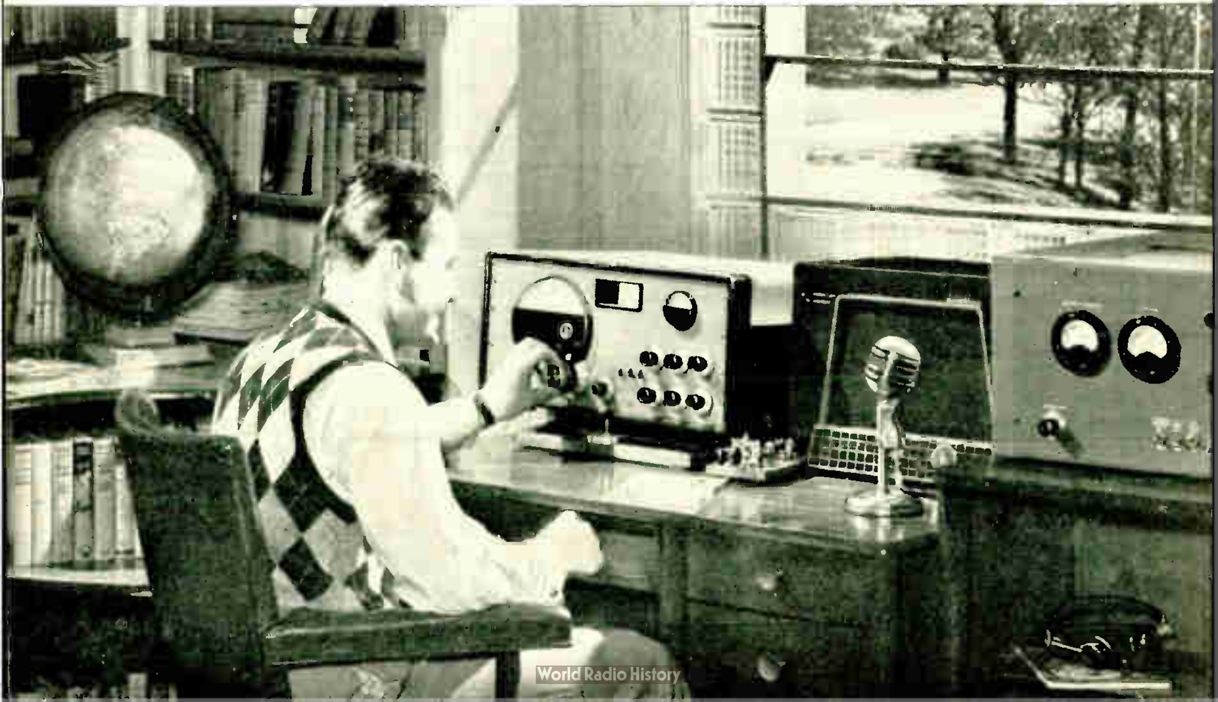
CEDAR RAPIDS, IOWA



$250.00
Amateur Net

Model **SX·42**

Greatest continuous frequency coverage of any communications receiver



From 540 kc. to 110 Mc. AM • FM • CW

In the Model SX-42 Hallicrafters sets a new high standard of receiver performance and versatility. Covering all frequencies from 540 kilocycles to 110 megacycles, the SX-42 combines in one superbly engineered unit a top-flight standard and VHF communications receiver, standard, short-wave and FM broadcast receiver, and high fidelity phonograph amplifier.

The tremendous frequency range of the SX-42, greater continuous coverage than has ever before been available in a receiver of this type, is made possible by the development of a new "split-stator" tuning system and the use of dual intermediate frequency transformers. Reception of amplitude modulated and continuous wave telegraph signals is provided for throughout the entire range of the SX-42. In addition, a discriminator and two limiter stages are available on bands 5 and 6 (27 to 110 megacycles) to permit the reception of frequency modulated signals. Musical reproduction of true high fidelity is assured by an audio system with a response curve essentially flat from 60 to 15,000 cycles and an undistorted output of eight watts.

The controls of the SX-42 are arranged for maximum convenience and simplicity of operation. MAIN TUNING and BANDSPREAD knobs are mounted coaxially, focusing the tuning functions in a single precision-built unit. BANDSWITCH and VOLUME are located at either side of the main dial. Auxiliary controls such as CRYSTAL PHASING, SENSITIVITY, etc., are logically placed so that those most frequently used are in the most accessible positions. Hallicrafters new system of color coding makes it possible for the entire family to enjoy this fine receiver. The normal control positions for standard broadcast reception are indicated by tiny red dots while FM adjustments are in green.

The main tuning knob is provided with a precision vernier scale which is separately illuminated through a small window in the one-piece Lucite main dial housing. The main tuning dial is calibrated in megacycles and is marked with the numbers in the new FM band of 88 to 108 megacycles. The bandspread dial is calibrated for the amateur 3.5, 7, 14, 28, and 50 megacycle bands. An additional logging scale is provided on this dial for use in other ranges. The small locking knob mounted coaxially with the main and bandspread tuning knobs permits either to be rotated freely while holding the other firmly in position.

The many new and ingenious circuit features which make possible the amazing versatility of the SX-42 stem directly from Hallicrafters long experience in the design and production of VHF and UHF communications equipment. The newly developed "split-stator" tuning system used on the three higher bands provides a far greater gain per stage than is possible with older methods. Each IF transformer contains windings for both 455 kilocycles and 10.7 megacycles and the changeover is accomplished automatically between bands 4 and 5. As band 4 runs to 30 megacycles and band 5 starts at 27 megacycles it is possible to use either narrow-band standard communications receiver performance or wide-band FM performance on the amateur frequencies from 28 to 29.7 megacycles. A type 7A4 tube functions as a beat frequency oscillator for CW reception. When the receiver is switched to FM, however, this tube becomes a direct current amplifier to operate the FM tuning meter. This meter performs as a normal carrier level indicator for AM reception. A four position switch on the panel selects the desired mode of operation—PHONO, FM, AM or CW.

In addition to its many new features the SX-42 continues all of the time-tried advantages characteristic of Hallicrafters top models. Freedom from "drift" and maximum stability are provided by temperature compensation and the use of a type VR-150 voltage regulator tube.

Designed to function at a new high peak of high frequency efficiency

A crystal filter circuit combined with variable intermediate frequency channel width offers six different degrees of selectivity on the four lower bands (to 30 megacycles). CRYSTAL PHASING, CW PITCH, SENSITIVITY, and four position TONE control for LOW, MED, HI FI, and BASS, are all conveniently placed on the front panel as are RECEIVE STAND-BY, NOISE LIMITER, and AVC switches.

The beauty and modern functional styling of this new receiver are self evident. Without in any way detracting from the "precision instrument" appearance which characterizes fine communications equipment, Hallicrafters designers have succeeded in creating a receiver which is not out of place in the most luxurious surroundings. The rich deep gray of the panel, satin chrome "airodized" top, and light gray lettering with touches of red and green combine with the precision-tooled controls and light translucent green of the illuminated dials and meter in a harmoniously integrated whole.

Note in closeups at left the compact efficiency of the concentrically mounted main tuning and bandspread controls and the precise, logical grouping of the other dials.

A finishing touch is furnished by the instrument type adjustable base, available as an accessory. By simply turning the knurled rim of the front support the receiver can be tilted to provide an "eye-angle" view of the dials for maximum accuracy and ease of tuning.

Extraordinary versatility . . .
Features every ham wants

CONTROLS: BAND SELECTOR, MAIN TUNING, BANDSPREAD, and selective DIAL LOCK, VOLUME and POWER OFF, AVC, NOISE LIMITER, RECEIVE/STANDBY, SELECTIVITY, TONE, SENSITIVITY, CRYSTAL PHASING, RECEPTION, CW PITCH. "S" meter adjustment on rear of chassis.

EXTERNAL CONNECTIONS: Antenna connections for doublet or single wire antenna. Input impedance matches 300 ohm line except on broadcast band which is designed for use with ordinary single wire antenna. Output terminals to match 500 or 5000 ohm speaker. Phone jack on front panel. Phonograph input connector on rear of chassis. Socket for use of external power supply. Remote standby switch connections provided for in power socket. Power cord and plug.

PHYSICAL CHARACTERISTICS: The Model SX-42 is housed in a steel cabinet of true functional design. Panel and chassis are assembled as a unit and may be removed for servicing or for mounting in a relay rack. Panel is finished in deep gray, top of cabinet is of "airodized" steel finished in satin chrome and swings open on a full length piano hinge for maximum accessibility. Main dial housing is a single piece of Lucite fabricated by an injection molding process. Panel lettering is in light gray with incidental red and green markings for standard AM and FM reception. Dials are a light translucent green and are indirectly illuminated.

FIFTEEN TUBES: 1—6AG5 1st RF amplifier; 1—6AG5 2nd RF amplifier; 1—7F8 converter; 1—6SK7, 1st IF amplifier; 1—6SG7, 2nd IF amplifier; 1—6H6 AM rectifier and noise limiter; 1—7H7 1st FM limiter amplifier; 1—7H7 2nd FM limiter; 1—6H6 FM discriminator; 1—6SL7 audio inverter; 2—6V6 audio output tubes; 1—7A4 beat frequency oscillator and FM tuning meter amplifier; 1—VR-150 voltage regulator; 1—5U4G high voltage rectifier.

OPERATING DATA: The standard Model SX-42 is designed for operation on 105-125 volts 50/60 cycle alternating current. The universal Model SX-42U may be operated on 110, 130, 150, 220 or 250 volts, 25 to 60 cycle, alternating current. The standard model draws 0.93 amperes at 117 volts. When operated from batteries through the auxiliary power supply socket it requires 5 amperes at 6 volts DC for heater current and 150 milliamperes at 270 volts DC for plate current. Total battery current when operating from a 6 volt battery and using a vibrapack as a source of plate power is 16 amperes.

DIMENSIONS: Model SX-42. Cabinet only, 20 inches wide by 9¾ inches high by 16 inches deep. Overall, 20 inches wide by 10¼ inches high by 18 inches deep.

WEIGHT: Model SX-42. Receiver only, approximately 52 pounds. Packed for shipment, approximately 65 pounds. Model B-42. Adjustable base, packed for shipment, approximately 5 pounds.

SX-42 FEATURES

1. Continuous frequency range — 540 kilocycles to 110 megacycles in six bands.
   - Band 1—540 to 1620 kilocycles.
   - Band 2—1.62 to 5 megacycles.
   - Band 3—5 to 15 megacycles.
   - Band 4—15 to 30 megacycles.
   - Band 5—27 to 55 megacycles.
   - Band 6—55 to 110 megacycles.

   Adequate overlap is provided at the ends of all bands.
2. Wide vision main tuning dial accurately calibrated.
3. Separate electrical bandspread dial calibrated for amateur 3.5, 7, 14, 28, and 50 megacycle bands.
4. Beat frequency oscillator functions throughout entire range of receiver. CW pitch adjustable from panel.
5. Four-position switch selects mode of operation, PHONO, FM, AM, or CW.
6. RECEIVE/STANDBY switch.
7. Series type automatic noise limiter.
8. Push-pull final audio stage delivers over 8 watts with less than 8% harmonic distortion.
9. Audio amplifier response curve is essentially flat from 60 to 15,000 cycles.
10. Red markings for broadcast reception and green markings for FM reception simplify operation for general use.
11. Connections for coordinated operation with Hallicrafters transmitters.
12. Separate SENSITIVITY (RF) and VOLUME (AF) controls.
13. Four-position tone control provides LOW, MED, HI FI, and BASS.
14. Special socket for use of external power supply.
15. High frequency oscillator temperature compensated to reduce drift.
16. "Micro-set" permeability adjusted coils in RF section.
17. AVC switch.
18. "Airodized" steel top provides full ventilation and swings open on full length piano hinge for greatest accessibility.
19. Wide band FM, AM or CW available from 27 to 110 megacycles.
20. Six-position selectivity switch with crystal filter operates on frequencies between 540 kilocycles and 30 megacycles.
21. Combination carrier level meter and FM tuning indicator. BFO tube performs dual function as FM tuning indicator amplifier.
22. New FM band marked with channel numbers in addition to megacycle calibration.
23. Dual intermediate frequency transformers. 455 kilocycle IF for standard operation. 10.7 megacycle IF for VHF and FM operation.
24. "Split-stator" tuning makes possible superior performance in VHF range.
25. Chassis and panel can be removed as a unit for rack mounting.
26. Crystal phasing control.
27. Antenna input impedance matches 300 ohm line.
28. New Hallicrafters Type HA-6 crystal used in crystal filter circuit. Holder of Mycalex, non-hygroscopic and unaffected by temperature.
29. Two limiter stages for maximum quieting on FM.
30. Two tuned RF stages using miniature tubes for superior VHF performance.
31. Phonograph input connections on rear of chassis.
32. Type VR-150 voltage regulator tube provides maximum stability in high frequency oscillator, converter, BFO, and FM tuning meter circuits.
33. MAIN and BANDSPREAD tuning controls and dial lock are mounted coaxially as a single precision-built unit.
34. Main tuning knob provided with precision vernier scale, separately illuminated through small window in one-piece Lucite dial housing.

NEW SPEAKERS TO MATCH THE NEW MODELS

The R-42 and the R-45 (the rack mounting version of the R-42) represent one of the greatest innovations in speaker design in recent years. This is the first speaker of its size to offer the splendid advantages of the bass reflex principle. Now in this sleek, highly functional design, matching the new line of Hallicrafters receivers, the bass reflex feature is available in a compact speaker that offers a new high quality of reproduction. The speaker size is [illegible] inches. Two position switch on front panel for communications or high fidelity reception. Terminals on rear for 500/600 ohm line. R-42, size: 12½" deep, 11¾" high, 17" wide. R-45, size: 12¾" deep, 17½" high, 19" wide.

R-45 Speaker . . [illegible] R-42 Speaker . . [illegible]

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hallicrafters RADIO

Model S·40

Function, beauty combined in an outstanding value...

FEATURES

Many circuit refinements never before available in this price class.

1. Overall frequency range — 540 kilocycles to 43 megacycles in 4 bands.
   - Band 1—540 to 1700 kilocycles.
   - Band 2—1.7 to 5.35 megacycles.
   - Band 3—5.35 to 15.7 megacycles.
   - Band 4—15.7 to 43 megacycles.

   Adequate overlap is provided at the ends of all bands.
2. Wide vision main tuning dial accurately calibrated.
3. Separate electrical bandspread dial, inertia flywheel tuning.
4. Beat frequency oscillator, pitch adjustable from front panel.
5. CW/AM switch.
6. Standby/receive switch.
7. Automatic noise limiter.
8. Maximum audio output—2½ watts.
9. Internal PM dynamic speaker held in rubber shock mounts.
10. Red markings for broadcast reception simplify operation for general use.
11. Connections for coordinated operation with Hallicrafters transmitters.
12. Separate SENSITIVITY (RF) and VOLUME (AF) gain controls.
13. Three-position tone control.
14. Special socket permits use of external auxiliary power supply.
15. High frequency oscillator temperature compensated to reduce drift.
16. "Micro-set" permeability adjusted coils in RF section.
17. AVC switch.
18. Exceptional accessibility of all parts due to new cabinet design.
19. Socket for connection of Model SM-40 "S" meter.

Model SM-40 "S" Meter

This new external "S" meter is available as an accessory and can be easily connected through a special socket on the rear of the receiver chassis. May also be used with other Hallicrafters models such as the S-20R, S-18, etc.

540 kc. to 43 Mc.

★ 540 to 1700 kc. ★ 5.35 to 15.7 Mc.
★ 1.7 to 5.35 Mc. ★ 15.7 to 43 Mc.

The sensational new S-40 with the finest performance ever presented in the popular price field is housed in a cabinet of true functional design — a completely new conception of receiver beauty and styling. Full use is made of newly developed materials and techniques. Maximum ventilation is assured by a multitude of tiny openings in the upper section of the cabinet which also impart a smart and pleasing appearance. The entire top of the cabinet opens on a full length piano hinge for complete accessibility. Panel and chassis may be removed from the cabinet as a unit without disturbing any controls or connections. All controls are clearly identified and the normal positions for standard broadcast reception are marked in red, making it easy for the whole family to use this fine receiver.

The Model S-40 incorporates many circuit refinements and features never before available in this price class. The RF section uses permeability adjusted "micro-set" inductances, identical with those in the most expensive Hallicrafters receivers. Automatic noise limiter, temperature compensated RF oscillator, beat frequency oscillator, separate RF and AF gain controls, three-position tone control, separate electrical bandspread, with inertia flywheel tuning, and many other features make this beautiful new receiver an outstanding value.

$79.50

Amateur Net

CONTROLS: SENSITIVITY (including "S" meter on/off switch), BAND SELECTOR, VOLUME, TUNING, BANDSPREAD, AVC ON/OFF, CW/AM, NOISE LIMITER ON/OFF, TONE AC OFF, PITCH CONTROL, STANDBY/RECEIVE.

NINE TUBES: 1—6SG7 RF amplifier; 1—6SA7 converter; 1—6SK7 1st IF amplifier; 1—6SK7 2nd IF amplifier; 1—6SQ7 2nd detector and 1st audio amplifier; 1—6F6G output audio amplifier; 1—6H6 automatic noise limiter and gas gate; 1—6J5GT beat frequency oscillator; 1—80 rectifier.

OPERATING DATA: The standard Model S-40 is designed for use on 105-125 volts, 50 to 60 cycle alternating current. The universal Model S-40U can be used on 110, 130, 150, 220 or 250 volts, 25 to 60 cycle, alternating current. The standard model draws .76 amperes at 117 volts. When used with external batteries the heater current is 5 amperes at 6 volts and plate current is 70 milliamperes at 270 volts. If a vibrapack is used for plate supply the total current demand for both plate and heaters is 10 amperes at 6 volts.

DIMENSIONS: Model S-40, Cabinet only, 18½ inches wide by 8½ inches high by 9⅝ inches deep. Overall, 18½ inches wide by 9 inches high by 11 inches deep.

WEIGHT: Model S-40, Receiver only, approximately 28 pounds. Packed for shipment, approximately 33 pounds. Model SM-40, Meter only, approximately 1¾ pounds. Packed for shipment approximately 3 pounds.

See what you hear with the SKYRIDER PANORAMIC SP-44

Hallicrafters new Skyrider Panoramic adaptor, Model SP-44, offers all the advantages of panoramic reception in an unusually compact and inexpensive unit. With this adaptor connected to a Hallicrafters receiver it is possible to monitor up to 200 kilocycles of the radio spectrum visually and to analyze the characteristics of radio signals from your own or other transmitters.

This new adaptor may be used with any receiver having an IF frequency between 450 and 470 kilocycles. Ten tubes.

$99.50

Amateur Net

hallicrafters RADIO

$39^{50}

Amateur Net

FEATURES

1. Overall frequency range — 540 kilocycles to 32 megacycles in 4 bands.

   Band 1—540 to 1650 kc.
   Band 2—1.65 to 5 Mc.
   Band 3—5 to 14.5 Mc.
   Band 4—13.5 to 32 Mc.

   Adequate overlap is provided at the ends of all bands.

2. Main tuning dial accurately calibrated.
3. Separate electrical bandspread dial.
4. Beat frequency oscillator, pitch adjustable from front panel.
5. AM/CW switch. Also turns on automatic volume control in AM position.
6. Standby/receive switch.
7. Automatic noise limiter.
8. Maximum audio output—1.6 watts.
9. Internal PM dynamic speaker mounted in top.
10. Controls arranged for maximum ease of operation.
11. 105-125 volt AC/DC for operation. Resistor line cord for 210-250 volt operation available.
12. Speaker/phones switch.

CONTROLS: SPEAKER/PHONES, AM/CW, NOISE LIMITER, TUNING, CW PITCH, BAND SELECTOR, VOLUME, BANDSPREAD, RECEIVE/STANDBY.

EXTERNAL CONNECTIONS: Antenna terminals for doublet or single wire antenna. Ground terminal. Tip jacks for headphones. Line cord and plug.

PHYSICAL CHARACTERISTICS: The Model S-38 is housed in a sturdy steel cabinet finished in rich satin black. Speaker grille in top is of airodized steel. Chassis is cadmium plated. Lettering is in light gray and switch knobs are red.

For hams, beginning hams and all who want the finest receiver available at a low price

The Model S-38 meets the demand for a truly competent communications receiver in the low price field. Styled in the post-war Hallicrafters pattern and incorporating many of the features found in its more expensive brothers, the S-38 offers performance and appearance far above anything heretofore available in its class. Four tuning bands, CW pitch control adjustable from the front panel, automatic noise limiter, self-contained PM dynamic speaker and "Airodized" steel grille, all mark the S-38 as the new leader among inexpensive communications receivers.

The S-38 is an especially fine receiver for younger people just beginning to find the unending fascination offered by radio as a hobby. In addition to being a good standby receiver for any amateur, the S-38 has unlimited uses. Its compact functional design, its high performance on both short waves and standard broadcast reception makes it an ideal receiver for use in den or library, in college dormitory, at camp or cottage or in any room around the house wherever a good extra receiver at a low cost is desired.

Model S·38

SIX TUBES: 1—12SA7 converter; 1—12SK7 IF amplifier; 1—12SQ7 second detector, AVC, first audio amplifier; 1—12SQ7 beat frequency oscillator, automatic noise limiter; 1—35L6GT second audio amplifier; 1—35Z5GT rectifier.

OPERATING DATA: The Model S-38 is designed to operate on 105-125 volts AC or DC. A special external resistance line cord can be supplied for operation on 210 to 250 volts AC or DC. Power consumption on 117 volts is 29 watts.

DIMENSIONS: Model S-38. Cabinet only, 12⅞ inches wide by 6⅞ inches high by 7⅞ inches deep. Overall, 12⅞ inches wide by 7⅜ inches high by 8⅝ inches deep.

WEIGHT: Model S-38. Receiver only, 11 pounds. Packed for shipment, 13½ pounds.

First of its kind . . . a low power, high quality, low price transmitter

$50.00 *Amateur Net (Approximate)*

Model HT-17

The Model HT-17 offers real Hallicrafters transmitter performance with maximum convenience and economy. No larger than a small receiver and styled to match the postwar Hallicrafters line, this new transmitter provides an honest ten watts of crystal-controlled CW output on the amateur 3.5, 7, 14, 21, and 28 megacycle bands.

A pi-section matching network is an integral part of the plate circuit and, together with an adjustable link, provides coupling to any type of antenna or permits the HT-17 to be used as an exciter for a high power final amplifier. The oscillator stage uses a type 6V6-GT tube and is automatically switched to a Tritet circuit when coils for the three higher bands are plugged in. Full output on the 14, 21, and 28 megacycle bands is obtained with 7 megacycle crystals. A type 807 tube is used in the final amplifier, and the self-contained power supply, for 105-125 volt AC operation, employs a 5Z3 rectifier. Connections are provided for an external modulator. The "airodized" steel top opens on a full length piano hinge for maximum accessibility and ease in changing coils and crystals. A pilot lamp is provided on the front panel for tuning. Coil sets extra.

CONTROLS: PLATE, LOADING, TRANSMIT/STANDBY, METER OSC/PWR AMP, AC ON/OFF (all on front panel). Oscillator plate tuning, Tritet tuning (easily accessible by raising top).

EXTERNAL CONNECTIONS: Antenna terminals for single wire, using pi-section network tuning, or two-wire low impedance line, using link coupling. Ground terminal. Connections for key and external modulator. AC line cord and plug. Special socket for use of external power supply. Fuse.

PHYSICAL CHARACTERISTICS: The Model HT-17 is enclosed in a sturdy steel cabinet with all operating components mounted on a strong cadmium plated chassis. Top is of "Airodized" steel and opens on a full-length piano hinge for maximum accessibility. Dials are of the slide rule type. Finish is rich satin black. Trim and lettering match the new Hallicrafters receivers.

THREE TUBES: 1—6V6-GT crystal oscillator; 1—807 power amplifier; 1—5Z3 rectifier.

OPERATING DATA: The Model HT-17 is designed for operation on 105-125 volts 50/60 cycle alternating current. Connections are provided for use with external batteries or other emergency power source. When operated on 117 volts the total current is 1.07 amperes (125 watts). Heater current needed for auxiliary power supply operation is 1.35 amperes at 6 volts, plate current is 135 milliamperes at 400 volts. Total demand when used with a vibrapack on six volt battery is 18 amperes.

DIMENSIONS: Model HT-17. Cabinet only, $12\frac{7}{8}$ inches wide by $6\frac{7}{8}$ inches high by $7\frac{7}{8}$ inches deep. Overall, $12\frac{7}{8}$ inches wide by $7\frac{3}{8}$ inches high by $8\frac{5}{8}$ inches deep.

WEIGHT: Model HT-17. Transmitter only 21 pounds. Packed for shipment 25 pounds. Coils, packed for shipment, per set, approximately $1\frac{1}{2}$ pounds.

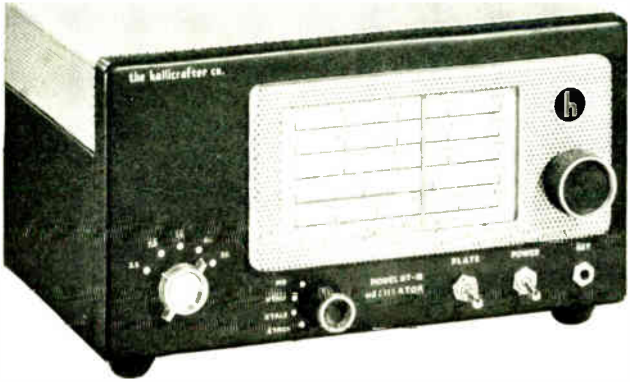
SM-2 plate milliampere meter. Range 0 to 150 ma. Supplied for quick installation in HT-17 transmitter in place of tuning pilot lamp, at extra cost.

FEATURES

1. Frequency range — amateur bands from 3.5 to 30 megacycles.
2. Power output—10 watts minimum on all bands.
3. Pi-section matching network plus coupling link permits use with any antenna.
4. May be easily coupled to drive a high power final amplifier.
5. "Airodized" steel top for maximum ventilation.
6. Full-length piano hinge permits entire top to swing open for ease in changing coils, crystals.
7. All operating and tuning controls easily accessible.
8. Self-contained power supply for 105-125 volt 50/60 cycle AC operation.
9. Special socket for use of external auxiliary power supply.
10. Oscillator circuit automatically switched from Pierce to Tritet, on three higher bands.
11. Full output at highest frequency with 7 megacycle crystal.
12. Terminals for connection of external modulator.
13. Panel switch to connect tuning pilot lamp in exciter or amplifier circuits.
14. New styling harmonizes with Hallicrafters postwar receivers.
15. Plug-in provision for SM-2 plate milliampere meter.

$75.00 *Amateur Net (Approximate)*

A variable master oscillator combining excellent stability and ease of operation



Model HT-18

Here is another new and welcome addition to the Hallicrafters line, a variable master oscillator. It is specifically designed to provide the amateur operator with a continuously variable exciter unit which is as easy to tune or shift to a new frequency as a modern receiver. Outstanding features never before available in a unit of this kind include excellent stability, negligible frequency drift, voltage regulator and complete simplicity of operation. It is accurately calibrated for the five ham bands. The heart of the unit is the variable master oscillator which employs a 6BA6 tube in an electron coupled circuit with plate and screen voltage regulation. This circuit is scientifically temperature-compensated and is tuned by one section of an air dielectric variable condenser, another section of which tunes the 6V6 frequency multiplier amplifier. Power output of the HT-18 is fed through a six foot 72 ohm coaxial line which may be connected to any commonly used crystal circuit of a transmitter. The RF output at the line end is not less than 2½ watts and it can therefore be used to drive a high power class "C" amplifier. For example the unit will provide ample driving power to two 813's which will supply over 500 watts of CW power and about 300 watts of phone carrier.

In addition to variable frequency operation, the Model HT-18 provides for three crystals for spot frequency use. These crystals may be switched into the circuit from the front panel.

CONTROLS: BAND SELECTOR, TUNING, VARIABLE FREQUENCY—CRYSTAL SELECTOR, POWER ON/OFF SWITCH, CARRIER ON/OFF SWITCH, BEAT FREQUENCY SWITCH.

EXTERNAL CONNECTIONS: R-F output terminals. Power line cord, carrier switch terminal connectors for receiver and transmitter control. Shorting type key jacks on front panel.

PHYSICAL CHARACTERISTICS: The cabinet of the Model HT-18 is styled to match the new Hallicrafters models and is finished in rich satin black. Airodized steel top swings open on a full-length piano hinge for maximum accessibility. Panel lettering is light gray and dial scale is green indirectly illuminated.

FIVE TUBES: 1—6BA6 electron coupled oscillator or crystal oscillator; 1—6V6 amplifier or frequency multiplier; 1—VR-105 voltage regulator; 1—VR-150 voltage regulator; 1—5Y3GT power rectifier.

OPERATING DATA: The Model HT-18 is designed for operation on 105-125 volts 50/60 cycle, alternating current. Crystals used if desired are in the 3.5 megacycle band but are not supplied with unit.

FEATURES

1. Frequency range. Five amateur bands.
2. Wide vision tuning dial accurately calibrated.
3. 2½ watts measured output at end of 6 foot 72 ohm transmission line.
4. Negligible drift.
5. Scientifically temperature-compensated.
6. Oscillator and amplifier keyed.
7. Built in crystal sockets for spot frequency operation.
8. Two voltage regulators.
9. Complete band switching.
10. Ganged tuning.
11. All coils self-contained, no plug in coils.
12. Tubes and circuit components carefully selected for maximum stability.
13. Oscillator operates on lowest frequency range only.
14. Higher frequency bands reached by means of high efficiency frequency multiplier.

Hallicrafters Model HT-9 is an ideal medium power transmitter. Designed for maximum flexibility and convenience. In addition to coils and crystals it requires only a microphone or key, antenna and a source of AC power to go on the air.

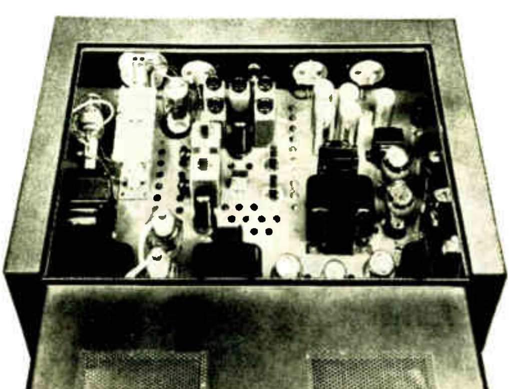
Five individual plug-in tuning units and crystals may be accommodated in the exciter section simultaneously. Band switching is easily accomplished by changing one coil in the final amplifier and selecting the desired exciter frequency by means of a panel switch. Exciter units are pre-tuned and the only additional operation needed is a slight adjustment of the final tank tuning capacitor.

Separate meters are provided for the power amplifier plate and grid circuits and a third meter may be switched into either the exciter or modulator cathode circuits. All controls are conveniently arranged on the panel and a safety interlock switch is provided for protection against accidental shock when the cabinet is opened.

A real ham rig
Medium power
Maximum flexibility

$250.00

Less Coils and Crystals
Amateur Net



Model HT·9 TRANSMITTER

FEATURES

1. Frequency range 1500 kilocycles to 18 megacycles and amateur 28 megacycle band.
2. Power output 100 watts on CW, 75 watts on phone.
3. Antenna coil will match any resistive load from 10 to 600 ohms.
4. Maximum ventilation provided by louvers on sides, cutouts at rear.
5. Hinged top permits access to interior for changing coils and crystals.
6. All operating controls on front panel.
7. Self contained power supply for 105-125 volts, 50/60 cycle AC operation.
8. Input for any medium level, high impedance microphone
9. Metering of cathode current of exciter or modulator, power amplifier grid and power amplifier plate.
10. 100 per cent modulation with low distortion.
11. Carrier hum more than 40 db. below 100% modulation.
12. Frequency response flat within 3 db. from 100 to 5000 cycles.
13. Five operating frequencies may be pre-set in the oscillator and buffer doubler stages and selected at will by means of the band switch.
14. Line fuses mounted on rear of chassis.
15. Convenient table mounting.
16. Rugged construction and oversize components assure dependable operation.

CONTROLS: AUDIO GAIN, (SPEECH AMPLIFIER) OFF/ON, CATHODE CURRENT EXC. MOD., PLATE PWR. ON/OFF, FIL. PWR. ON/OFF, C.W. PHONE, BAND-SWITCH, TRANSMIT/STANDBY, PLATE TUNING.

METERS: Cathode current, P.A. grid, P.A. plate.

EXTERNAL CONNECTIONS: Antenna terminals. Terminal strip for key, antenna relay, and remote control of receiver. Line cord and plug. Two line fuses. Microphone input connector (on left end of cabinet). All connections except microphone are located on rear of chassis.

PHYSICAL CHARACTERISTICS: The Model HT-9 is constructed on a heavy cadmium plated steel chassis. Cabinet is of steel finished in gray wrinkle enamel and is provided with heavy rubber mounting feet. Ventilating openings in top and sides assure adequate cooling. Interlock switch under lid cuts high voltage supply when cabinet is opened.

TUNING UNITS: Final amplifier coils and exciter tuning units are available for the 1.75, 3.5, 7, 14, and 28 Mc. amateur bands. General coverage coils and units for all frequencies between 1.5 and 18 Mc. may be obtained on special order.

FOURTEEN TUBES: 1—6L6 crystal oscillator (used above 8 Mc. only); 1—6L6 crystal oscillator or doubler; 1—814 final RF amplifier; 1—6SJ7 1st speech amplifier; 1—6J5 2nd speech amplifier; 4—6L6 push-pull parallel modulator stage; 2—5Z3 rectifiers; 1—80 rectifier; 2—866 rectifiers.

OPERATING DATA: The model HT-9 is designed for operation on 105-125 volts, 50/60 cycle alternating current. In normal operation it draws approximately 3.5 amps. (400 w.).

DIMENSIONS: Model HT-9 overall clearance: 29⅛ inches wide by 12½ inches high by 20½ inches deep.

WEIGHT: Model HT-9 transmitter, 120 pounds. Packed for shipment, 125 pounds.

BUILDERS OF Skyfone
AVIATION RADIOTELEPHONE

hallicrafters RADIO

THE HALLICRAFTERS CO., MANUFACTURERS OF RADIO AND ELECTRONIC EQUIPMENT • CHICAGO 16, U. S. A.

Sole Hallicrafters Representatives in Canada: Rogers Majestic Limited, Toronto - Montreal

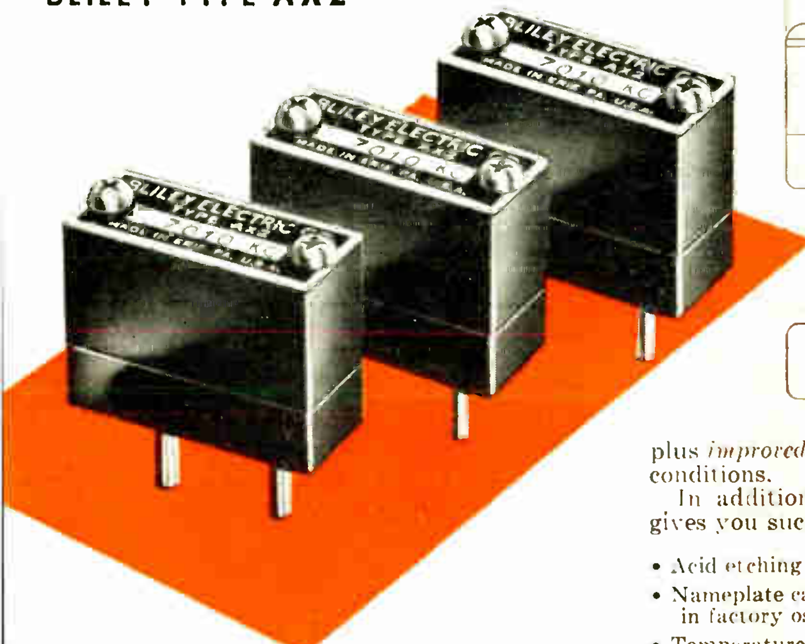
Bliley CRYSTALS..

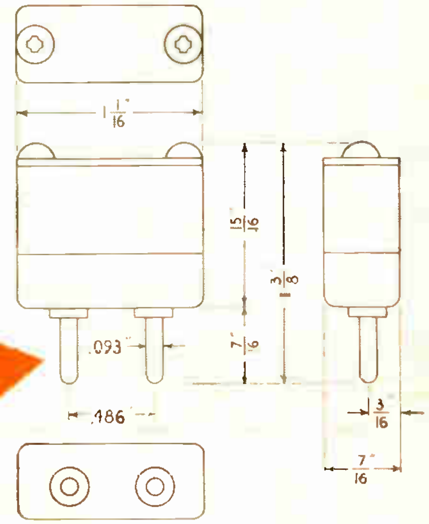
First for amateur frequencies

Type **AX2** Units, **80**-meter band **$2.80** Each
Type **AX2** Units, **40**-meter band **2.80** Each
Type **AX2** Units, **20**-meter band **3.95** Each

PLATED CRYSTALS

BLILEY TYPE AX2





We are justly proud of the technical accomplishments represented in the AX2 *plated* crystal. Its advanced development and pace-setting design again demonstrate Bliley's leadership in the manufacture of crystals for amateur frequencies.

Primary electrodes in the AX2 *plated* crystal unit consist of a micro-thin metal film which is deposited directly on the major surfaces of the quartz crystal by evaporation under high vacuum. This film exhibits extremely high adhesion to the crystal and can almost be considered as a chemical bond to the quartz.

Secondary electrodes, under spring pressure are used to clamp the crystal in position and to provide a medium for thermal dissipation.

Bliley's *plated* crystal gives you *better grid current stability* over a wide temperature range plus *improved frequency stability* under high drive conditions.

In addition to the *plating* feature, the AX2 gives you such famous Bliley qualities as:

- Acid etching to frequency to prevent aging.
- Nameplate calibration accurate to ± .002% at 25° C in factory oscillator.
- Temperature stability better than ± .02% between −10° and +60° C.
- Activity level tested between −10° C and +60° C.
- Solid, stainless steel pins.
- Welded contact between pins and contact plates.
- Neoprene gasket seal.
- Moisture resistant, molded phenolic case and cover.
- Small, compact size permits easy stacking. Two units may be mounted back to back in standard octal socket.
- All nomenclature on top of holder for easy identification.

Not a thing has been overlooked to insure top performance under any conditions encountered in amateur equipment. All our wartime experience is reflected in this new model, engineered specifically for amateur frequencies.

BLILEY ELECTRIC COMPANY · UNION STATION BUILDING, ERIE, PENNSYLVANIA

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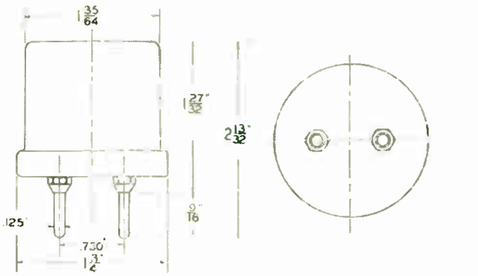
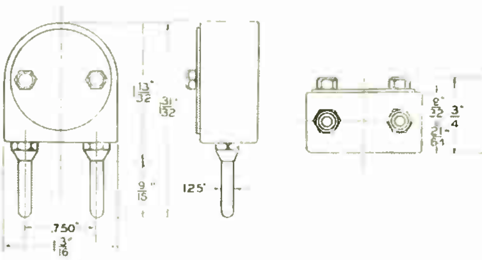


TYPE **FM6-S** 100 kc.

Primarily for use as a freq. standard. Plated precision crystal, rigidly clamped between resonant pins, provides exceptional electrical and mechanical stability. Freq. is adjustable to exactly 100kc. at 25° C when unit is used in recommended oscillator circuit. Freq. stability ±.005% at any temp. 0° C to 50° C.

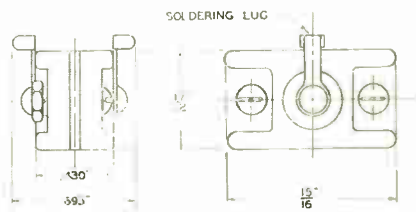
PRICE **$18.75**



TYPE **CF3** 455 kc.

Single signal filter crystal unit. Exceptionally low holder capacity permits sharp signal discrimination in filter network of general communications receivers. Frequency 455 kc. free from spurious responses within ±7 kc.

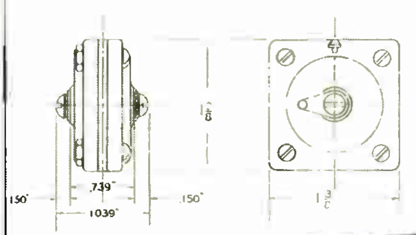
PRICE **$5.00**



TYPE **CF6** 455 kc.

Single signal filter crystal unit. Frequency 455 kc., ± 5 kc. — free from spurious responses within ± 7 kc. of fundamental. Designed for intermediate frequency filter in general communications receivers.

PRICE **$4.50**



TYPE **SMC100** 100–1000 kc.

Dual frequency crystal provides either 100 kc. or 1000 kc. frequency source. When used in recommended oscillator circuit 1000 kc. frequency is within ±.05% at 25° C and 100 kc. frequency can be adjusted to zero beat at 25° C. Suggested for signal generators used in alignment of radio receivers.

PRICE **$8.75**

The New Bliley CCO

CRYSTAL CONTROLLED OSCILLATOR

for Radio Service Technicians

For instant channel selection and frequency accuracy, radio service technicians use this Bliley test instrument. It provides direct crystal control for i-f alignment. Write for descriptive Bulletin 32.

AMPHENOL "SIGNAL SQUIRTER" ROTARY BEAM ANTENNA

Amphenol now offers this world famous rotary beam antenna developed by M. P. Mims, W5BDB. High forward gain, high front to back ratio, a rugged rotary drive system and a simplified direction indicator characterize this fine antenna which has been the standard of comparison for many years. Available for the 10 and 20 meter bands and in a combination covering both bands.

TWIN-LEAD TRANSMISSION LINE

Combining convenience and efficiency, Amphenol Twin-Lead is the first choice of amateurs for construction of antennas and transmission lines. Type 14-023 Transmitting Twin-Lead, with an impedance of 75 ohms, is the favorite for transmitter applications. Conservatively rated at 1 kw. Three impedance values:—300 ohms, 150 ohms and 75 ohms are available in Amphenol Receiving Twin-Lead for receiving and low power transmitting applications.

In addition to the three new products described, Amphenol is the world's largest single source of:

COAXIAL CABLES AND CONNECTORS
ANTENNAS
RADIO COMPONENTS
PLASTICS FOR ELECTRONICS

All are available from your distributor. See him tomorrow.

AMPHENOL "EASY-TO-DRILL" CLEAR POLYSTYRENE WINDOW PANE

This clear polystyrene window pane ends the problem of bringing in lead-ins through glass. It is easy to drill and cut to size. Ordinary wood-working tools will do the job. Offering the high dielectric strength of polystyrene, this window pane ends broken glass and drilling through sash. Ordinary putty holds it in place. Available in 12" x 16" panes of 3/32" thickness, and in other sizes to order.

AMPHENOL

AMERICAN PHENOLIC CORPORATION
CHICAGO 50, ILLINOIS
In Canada • Amphenol Limited • Toronto

HAMMARLUND

"HQ-129-X" AMATEUR RECEIVER

The Hammarlund "HQ-129-X" amateur communications receiver is designed to meet the demands of the most critical amateurs. Its design includes every feature essential to finest performance.

The "HQ-129-X" has a continuous range from .54 to 31 megacycles in six separately calibrated bands with continuous bandspread throughout the entire range. In addition, the bandspread dial is calibrated for each of the four most important amateur bands—3.5–4 mc, 7–7.3 mc, 14–14.4 mc and 28–30 mc.

The "HQ-129-X" has the Hammarlund patented variable wide-band crystal filter which works exceptionally well on phone or short wave broadcast signals.

There are many other features: Variable antenna compensator, beat oscillator, voltage regulator, series noise limiter, send-receive switch, automatic volume control, calibrated "S" meter, audio gain control, sensitivity control—plus all that goes into a receiver built by engineers who have spent a lifetime designing commercial communication equipment.

The "HQ-129-X" is available complete in a two-tone gray finish including tubes and a 10 inch P. M. dynamic speaker.

"HQ-129-X".......... **Amateur Net Price $168.00**

SC-10—Speaker cabinet finished to match.
Amateur Net Price **5.25**

Send for twenty-page technical booklet

SERIES 400 "SUPER-PRO"

The Series 400 Commercial "Super-Pro" receiver covers a new and wider range of frequencies. The SP-400-X model covers from .54 to 30 megacycles taking in all of the standard and short wave broadcast bands as well as amateur bands down to 30 megacycles. The "Super-Pro" has become standard equipment with many engineers in the radio press and broadcast fields. During the recent war, "Super-Pros" were standard equipment in practically every Army Airways Communications System installation throughout the world. Many letters from the men who operated them attest to the soundness of design and ability to withstand the most gruelling operating conditions.

The "Super-Pro" has continuous variable selectivity from razor-sharp "single-signal" to wide band high fidelity for broadcast reception. This feature together with the high power high fidelity 8 watt audio amplifier makes this an ideal receiver for use in entertainment installations as well as for home use. In addition the SP-400-X has AVC, continuous bandspread, calibrated "S" meter, BFO, noise limiter, send-receive switch, ear phone terminals, phono-input and separate heavy duty power supply.

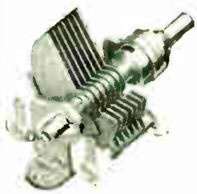
	Net Price
SPC-400-X • Receiver (Table Model) with P.M. speaker unit only..................................	**$342.00**
SPR-400-X • Receiver (Rack Model) less speaker.........	**344.55**
SC-46 • Speaker Cabinet only.........................	**5.25**

Prices subject to change without notice

HAMMARLUND MANUFACTURING CO., INC., 460 W. 34th Street, New York 1, N. Y.

HAMMARLUND

"MC" MIDGET CAPACITORS

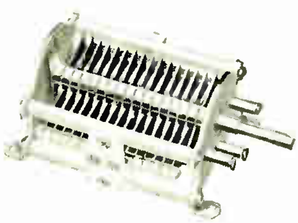


Ideal variable for high and very high frequency tuning, laboratories, etc. Isolantite Insulation. All contacts riveted or soldered. Vibration proof. New improved Hammarlund split type rear bearing, and noiseless wiping contact. Cadmium plated soldered brass plates. Shaft—¼".

Code	Capacity	List
MC-20-S	20 mmf.	$2.55
MC-35-S	35 mmf.	2.65
MC-50-S	50 mmf.	2.80
MC-50-M	50 mmf.	2.80
MC-75-S	80 mmf.	3.00
MC-75-M	80 mmf.	3.00
MC-100-S	100 mmf.	3.25
MC-100-M	100 mmf.	3.25
MC-140-S	140 mmf.	3.50
MC-140-M	140 mmf.	3.50
MC-200-M	200 mmf.	3.80
MC-250-M	260 mmf.	4.15
MC-325-M	320 mmf.	4.65

"M"—Midline Plates. "S"—Straight Line Cap. Plates.

"MTC" TRANSMITTING CAPACITORS



Compact types. Isolantite insulation. Base or panel mounting. Polished aluminum plates. Stainless steel shaft. Size of 150 mmf. with .070" plate spacing only 4⅝" behind panel. "B" models have rounded plates. "C" types have plain plate edges. Self-cleaning wiping contact.

Code	Capacity	List
MTC-20-B	20 mmf.	$4.75
MTC-100-B	100 mmf.	6.15
MTC-150-C	150 mmf.	7.10
MTC-250-C	260 mmf.	6.15
MTC-350-C	365 mmf.	6.75

FLEXIBLE COUPLINGS

These flexible couplings are designed for both insulated and non-insulated applications. The FC-46-S is insulated for 6000 volts with silicone treated ceramic, will compensate for considerable shaft misalignment, but will not give springy action. Overall depth 13 16", diameter 1¼". The FNC-46-S is a non-insulated coupling for use where insulation is unnecessary. The general design is the same as the FC-46-S but has a heavy metal body instead of ceramic. Overall depth 23 32", diameter 1¼".

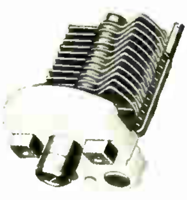
Code	List
FC-46-S—Insulated	$.90
FNC-46-S—Non-insulated	.90

BUTTERFLY CAPACITOR

The new butterfly capacitor is designed for use in VHF and UHF applications where the butterfly design is indispensable. Can be used as a single series unit or as a split stator with grounded rotor. This new butterfly capacitor is ideal for use in transmitters as well as receivers. Has soldered rotor and stator assembly; is plated to resist corrosion; silver plated rotor contact; sleeve type bearing, low-loss ceramic end panel. Approximately 1⅜" square. Depth behind panel depends on number of plates. Insulated mounting studs prevent rotor from being grounded when mounted to metal.

Code	MMF. Cap. per Sec. Max.	MMF. Cap. per Sec. Min.	Series Cap. Max.	Series Cap. Min.	
BFC-12	14.5	3.5	7.9	2.2	PRICE
BFC-25	27.5	5.0	14.5	3.0	NOT
BFC-38	40.5	6.3	21.0	3.7	AVAILABLE

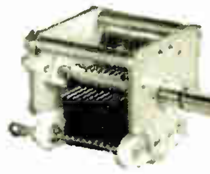
"APC" MICRO CAPACITORS



For H.F. and very H.F. For I.F. tuning, trimming R.F. Coils or gang capacitors, general padding, etc. Constant capacity under any condition of temperature or vibration. Size 100 mmf. 1 7 32" x 1 5 16" x 1 7 32". Isolantite base. Cadmium plated soldered brass plates.

Code	Capacity	List
APC-25	25 mmf.	$1.50
APC-50	50 mmf.	1.75
APC-75	75 mmf.	1.95
APC-100	100 mmf.	2.20
APC-140	140 mmf.	2.60

"RMC" CAPACITOR

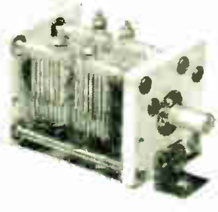


The new "RMC" is designed for applications where strength and solid construction is as important as electrical design. Its frame consists of 3 32" aluminum end plates reinforced by three horizontal bars or pillars which hold the assembly rigid.

Two low loss silicone treated ceramic insulated bars support the stator. Bearings are hand-fitted sleeve in the front and single ball thrust in the rear —torque is smooth and uniform. Contact to the rotor is made through a silver-plated beryllium forked spring. Brackets are provided for mounting either side down, or to a front panel with spacing pillars. Voltage rating—1000 V.

Code	Capacity	List
RMC-50-S	50. mmf.	$3.75
RMC-100-S	105. mmf.	4.25
RMC-140-S	143.5 mmf.	4.50
RMC-325-S	327. mmf.	5.65

"VU" UHF CAPACITOR



The capacitors listed below are available for use by manufacturers, engineers and amateurs for all types of communications equipment having tuned circuits operating as high as 500 mc. The many advantages of these new capacitors are of course due to the silent electrical operation made possible through the use of pyrex glass ball bearings.

Elimination of the rotor contact further precludes the possibility of noise and permits a more symmetrical design of the capacitor itself and consequently allows better circuit layout. Two sets of contacts are provided, so that the vacuum tube can be mounted on one side and the inductor on the other side of the capacitor. Voltage rating—700 V.

Code	Capacity	List
VU-20	22.5 mmf.	$10.75
VU-30	31.5 mmf.	11.50
VU-45	45.0 mmf.	12.70

Prices subject to change without notice

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Since 1923 we have been pioneers in the design and manufacture of the finest in radio hardware: Ceramic, Porcelain and Steatite Insulators . . . Antennas . . . Hook-Up Wires . . . Shielded Microphone Cable and S. J. Cable . . . and other top-quality "Birnbach" products.

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In keeping with its policy of providing all services within its power, The American Radio Relay League makes available to amateurs and would-be amateurs literature properly prepared to present in the best form all available information pertaining to amateur radio. The fact that its offices are the national and international headquarters of radio amateurs, makes League publications authoritative, complete, up-to-the-minute; written from a thoroughly practical amateur's point-of-view. These publications are frequently revised and augmented to keep abreast of the fast-changing field. At this time they will be found particularly adaptable for radio training purposes. All are printed in the familiar *QST* format which permits thorough but economical presentation of the information. Most of the publications and supplies described in the following pages are handled by your dealer for your convenience.

•

THE AMERICAN RADIO RELAY LEAGUE, INC.

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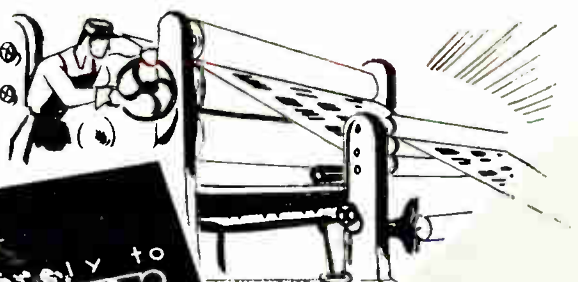
THE
OFFICIAL MAGAZINE
OF THE
AMERICAN RADIO RELAY LEAGUE

QST faithfully and adequately reports each month the rapid development which makes Amateur Radio so intriguing. Edited in the sole interests of the members of The American Radio Relay League, who are its owners, *QST* treats of equipment and practices and construction and design, and the romance which is part of Amateur Radio, in a direct and analytical style which has made *QST* famous all over the world. It is essential to the well-being of any radio amateur. *QST* goes to every member of The American Radio Relay League and membership costs $2.50 in the United States and Possessions, $3.00 in the Dominion of Canada, $4.00 in all other countries. Elsewhere in this book will be found an application blank for A.R.R.L. membership.

For thirtytwo years (and thereby the oldest American radio magazine) *QST* has been the "bible" of Amateur Radio.



QST
BINDERS

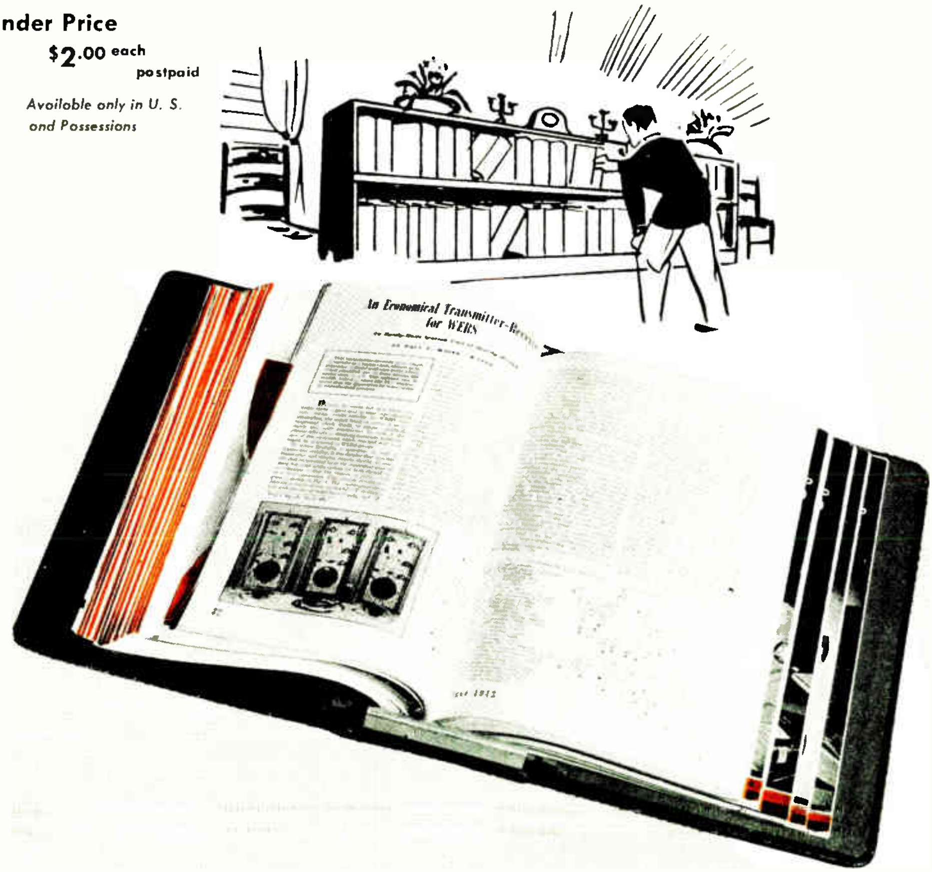
Those who take pride in the appearance of their lay-out and wish to keep their reference file of ***QST***'s in a presentable manner, appreciate the ***QST*** binder. It is stiff-covered, finished in beautiful and practical fabrikoid. Cleverly designed to take each issue as received and hold it firmly without mutilation, it permits removal of any desired issue without disturbing the rest of the file. It accommodates 12 copies of ***QST***. Opens flat at any page of any issue.

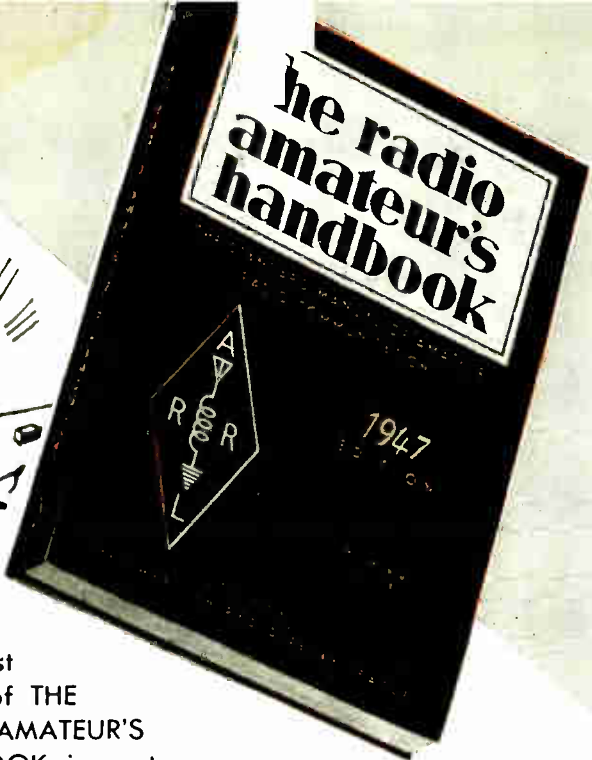
With each Binder is furnished a sheet of gold and black gummed labels for years 1928 through 1948. The proper one can be cut from the sheet and pasted in the space provided for it on the back of the binder. The back copies of ***QST*** contain the record of development of modern amateur technique. They are invaluable as technical references. Back copies are generally available—list will be sent upon request.

Binder Price

$2.00 each postpaid

Available only in U. S. and Possessions





The latest
edition of THE
RADIO AMATEUR'S
HANDBOOK is postwar
in content, containing the
kind of information which has made the
HANDBOOK world-famous. To maintain the
high standard of practical usefulness set by previous editions, a new treatment of the constructional sections of the HANDBOOK has been accomplished. The theory and design sections cover every subject encountered in practical radio communication. Completely sectionalized by topics with abundant cross-referencing, and fully indexed. The HANDBOOK continues to be the world's most valuable and widely-used radio book.

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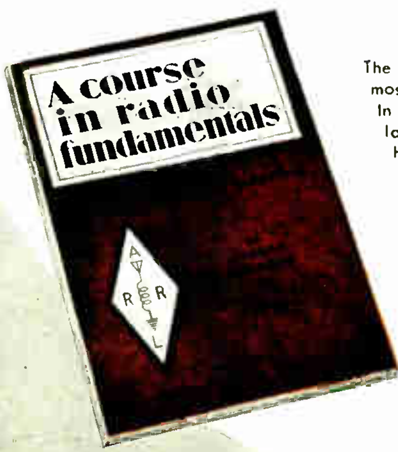

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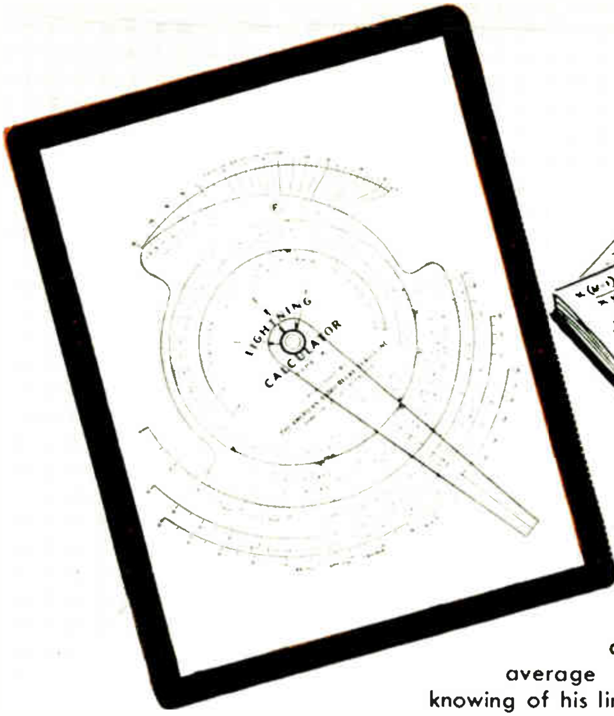
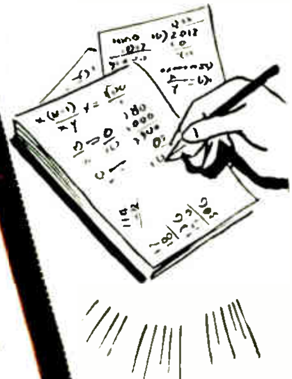
The objective in preparing this course was to accent those principles most frequently applied in actual radio communication. "A Course In Radio Fundamentals" is a study guide, examination book and laboratory manual. Its text is based on the "Radio Amateur's Handbook." Either the special edition prepared during the war for training purposes or the Standard Edition may be used. References contained in the "Course" are in both editions. As a text, this book greatly smooths the way for the student of the technicalities of radio. It contains interesting study assignments, experiments and examination questions for either class or individual instruction. It describes in detail 40 experiments with simple apparatus giving a complete practical knowledge of radio theory and design.





Aware of the practical bent of the average amateur and knowing of his limited time, the League, under license of the designer, W. P. Koechel, has made available these calculators to obviate the tedious and sometimes difficult mathematical work involved in the design and construction of radio equipment. The lightning calculators are ingenious devices for rapid, certain and simple solution of the various mathematical problems which arise in radio and allied work. They make it possible to read direct answers without struggling with formulas and computations. They are tremendous time-savers for amateurs, engineers, servicemen and experimenters. Their accuracy is more than adequate for the solution of practical problems, and is well within the limits of measurement by ordinary means. Each calculator has on its reverse side detailed instructions for its use; the greatest mathematical ability required is that of dividing or multiplying simple numbers. They are printed in several colors. You will find lightning calculators the most useful gadgets you ever owned.

RADIO CALCULATOR

TYPE A

This calculator is useful for the problems that confront the amateur every time he builds a new rig or rebuilds an old one or winds a coil or designs a circuit. It has two scales for physical dimensions of coils from one-half inch to five and one-half inches in diameter and from one-quarter to ten inches in length, a frequency scale from 400 kilocycles through 150 megacycles; a wavelength scale from two to 600 meters; a capacity scale from 3 to 1,000 micro-microfarads; two inductance scales with a range of from one microhenry through 1,500; a turns-per-inch scale to cover enameled or single silk covered wire from 12 to 35 gauge, double silk or cotton covered from 0 to 36 and double cotton covered from 2 to 36. Using these scales in the simple manner outlined in the instructions on the back of the calculator, it is possible to solve problems involving frequency in kilocycles, wavelength in meters, inductance in microhenrys and capacity in microfarads, for practically all problems that the amateur will have in designing—from high-powered transmitters down to simple receivers. Gives the direct reading answers for these problems with accuracy well within the tolerances of practical construction.

$1.00
Postpaid

OHM'S LAW CALCULATOR

TYPE B

This calculator has four scales.

A power scale from 10 microwatts through 10 kilowatts.

A resistance scale from .01 ohm through 100 megohms.

A current scale from 1 microampere through 100 amperes.

A voltage scale from 10 microvolts through 10 kilovolts.

With this concentrated collection of scales, calculations may be made involving voltage, current, and resistance, and can be made with a single setting of a dial. The power or voltage or current or resistance in any circuit can be found easily if any two are known. This is a newly-designed Type B Calculator which is more accurate and simpler to use than the justly-famous original model. It will be found useful for many calculations which must be made frequently but which are often confusing if done by ordinary methods. All answers will be accurate within the tolerances of commercial equipment.

$1.00
Postpaid

The standard elementary guide for the prospective amateur. Features equipment which, although simple in construction, conforms in every detail to present practices. The apparatus is of a thoroughly practical type capable of giving long and satisfactory service—while at the same time it can be built at a minimum of expense. The design is such that a high degree of flexibility is secured, making the various units fit into the more elaborate station layouts which inevitably result as the amateur progresses. Complete operating instructions and references to sources of detailed information on licensing procedure are given.

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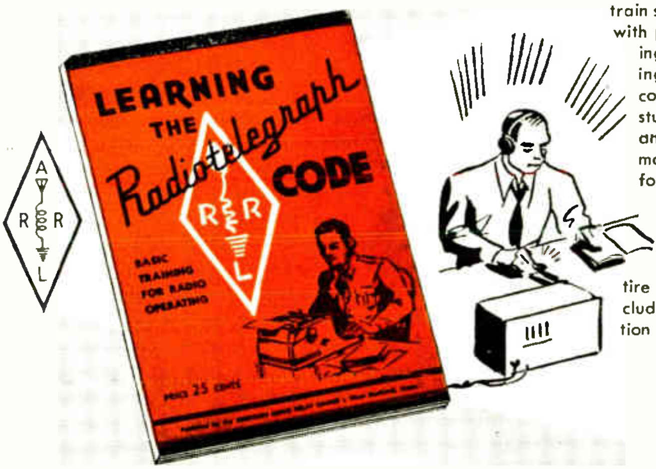

This booklet is designed to train students to handle code skillfully and with precision, both in sending and in receiving. Employing a novel system of code-learning based on the accepted method of *sound* conception, it is particularly excellent for the student who does not have the continuous help of an experienced operator or access to a code machine. It is similarly helpful home-study material for members of code classes. Adequate practice material is included for classwork as well as for home-study. There are also helpful data on high-speed operation, typewriter copy, general operating information—and an entire chapter on tone sources for code practice, including the description of a complete code instruction table with practice oscillator.

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To obtain an amateur operator's license you must pass a government examination. The License Manual tells how to do that—tells what you must do and how to do it. It makes a simple and comparatively easy task of what otherwise might seem difficult. In addition to a large amount of general information, it contains questions and answers such as are asked in the government examinations. If you know the answers to the questions in this book, you can pass the examination without trouble.

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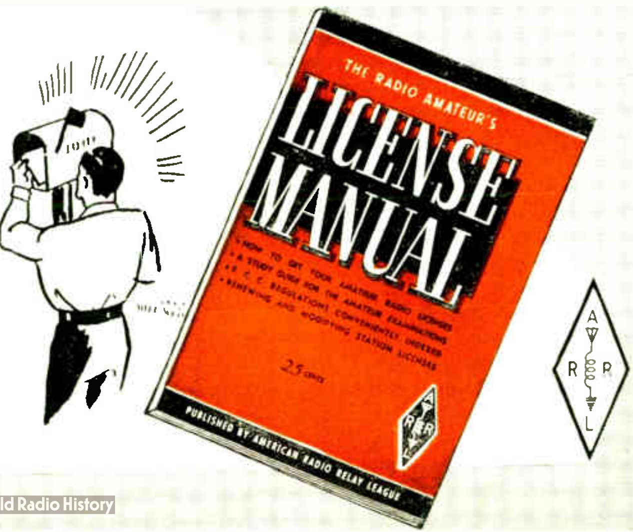

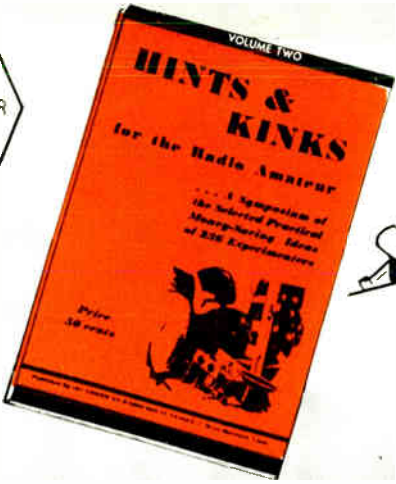

Amateurs are noted for their ingenuity in overcoming by clever means the minor and major obstacles they meet in their pursuit of their chosen hobby. An amateur must be resourceful and a good tinkerer. He must be able to make a small amount of money do a great deal for him. He must frequently be able to utilize the contents of the junk box rather than buy new equipment. Hints and Kinks is a compilation of hundreds of good ideas which amateurs have found helpful. It will return its cost many times in money savings—and it will save hours of time.

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Postpaid

A comprehensive manual of antenna design and construction, by the headquarters staff of the American Radio Relay League. Eighteen chapters, profusely illustrated. Both the theory and the practice of all types of antennas used by the amateur, from simple doublets to multi-element rotaries, including long wires, rhomboids, vees, phased systems, v.h.f. systems, etc. Feed systems and their adjustment. Construction of masts, lines and rotating mechanisms. The most comprehensive and reliable information ever published on the subject.

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The Story of Amateur Radio, by Clinton B. DeSoto—a detailed, accurate presentation in full book length of all the elements that have served to develop the most unique institution of its kind in the history of the world. A book of history but not a history-book, TWO HUNDRED METERS AND DOWN: The Story of Amateur Radio tells in spirited, dramatic fashion the entire chain of significant events in the development of the art.

OUT OF PRINT

Approximately 200 pages, 90,000 words, with durable imitation leather red paper cover

$1.00 *Postpaid*

Deluxe edition bound in blue cloth

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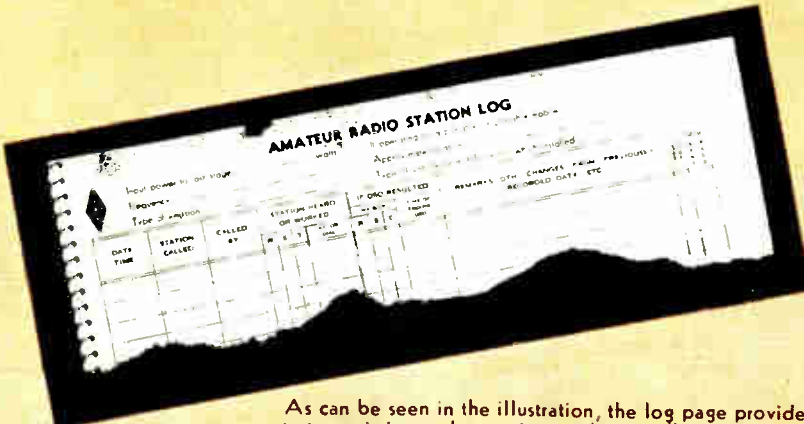
STATION OPERATING Sup



LOG BOOK

As can be seen in the illustration, the log page provides space for all facts pertaining to transmission and reception, and is equally as useful for portable or mobile operation as it is for fixed. The log pages with an equal number of blank pages for notes, six pages of general log information (prefixes, etc.) and a sheet of graph paper are spiral bound, permitting the book to be folded back flat at any page, requiring only the page size of 8½ x 11 on the operating table. In addition, a number sheet, with A.R.R.L. Numbered Texts printed on back, for traffic handlers, is included with each book. **35c** per book

OFFICIAL RADIOGRAM FORMS

The radiogram blank is designed to comply with the proper order of transmission. All blocks for fill-in are properly spaced for use in typewriter. It has a heading that you will like. Radiogram blanks, 8½ x 7¼, lithographed in green ink, and padded 100 blanks to the pad, 25c per pad, postpaid.

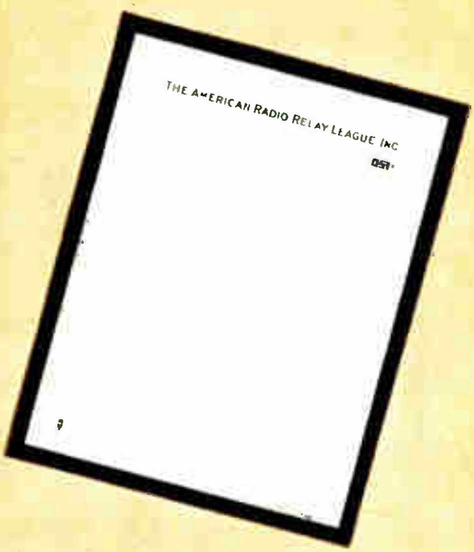
MESSAGE DELIVERY CARDS

Radiogram delivery cards embody the same design as the radiogram blank and are available in two styles — on stamped government postcard, 2c each; unstamped, 1c each.

The operating supplies shown on this page have been designed by the A.R.R.L. Communications Department.

lies

MEMBERSHIP



STATIONERY

Members' stationery is standard 8½ x 11 bond paper which every member should be proud to use for his radio correspondence. Lithographed on 8½ x 11 paper.

100 Sheets, **50c** *250 Sheets,* **$1.00**

500 Sheets, **$1.75**

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In the January, 1920 issue of *QST* there appeared an editorial requesting suggestions for the design of an A.R.R.L. emblem — a device whereby every amateur could know his brother amateur when they met, an insignia he could wear proudly wherever he went. There was need for such a device. The post-war boom of amateur radio brought thousands of new amateurs on the air, many of whom were neighbors but did not know each other. In the July, 1920 issue the design was announced — the familiar diamond that greets you everywhere in Ham Radio — adopted by the Board of Directors at its annual meeting. It met with universal acceptance and use. For years it has been the unchallenged emblem of amateur radio, found wherever amateurs gathered, a symbol of the traditional greatness of that which we call Amateur Spirit — treasured, revered, idealized.

THE LEAGUE EMBLEM, with both gold border and lettering, and with black enamel background, is available in either pin (with safety clasp) or screw-back button type.

In addition, there are special colors for Communications Department appointees.

- Red enameled background for the SCM.
- Blue enameled background for the ORS or OPS.

THE EMBLEM CUT: A mounted printing electrotype, ⅝" high, for use by members on amateur printed matter, letterheads, cards, etc.

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Stationery and Emblems are available only to A.R.R.L. Members.

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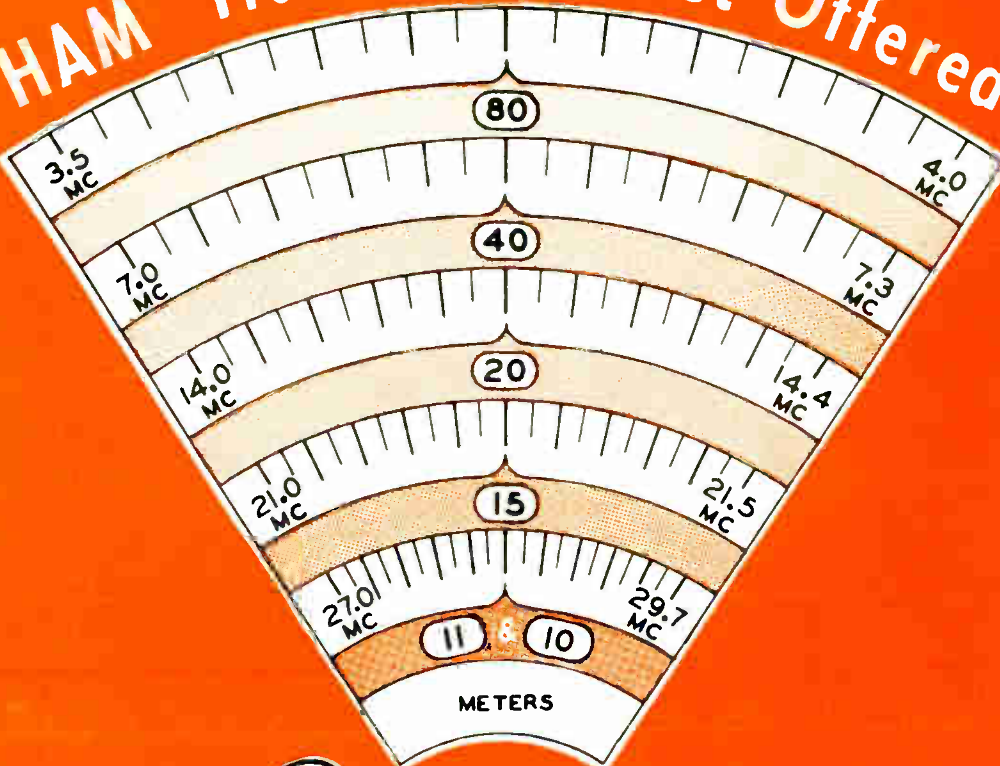

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ᐳREME TRANSMITTER Model AF-100, 6-Band, 100 Watt tput) Desk Type Transmitter. Embodies ALL the features st desired by the majority of the amateurs. Designed to er the amateur bands most frequently used: 10, 11, 15, 40 and 80 meters for CW, ICW, AM and FM Phone ısmission. This is the very first transmitter offered to the ateur which has the new feature of Frequency Modulation the band of frequencies assigned for this purpose, namely 185 to 27.455 and 29 to 29.7 megacycles. Model AF-100 :ontinuously tunable throughout the range of each of the ateur bands. A highly stable variable oscillator followed slug-tuned buffer and doubler stages which are ganged to oscillator dial simplifies the problem of working through severe QRM and further enhances the pleasures of easily establishing and retaining QSOs. Band changing is easily accomplished in the exciter by a band selector switch and in the final by the plugging in of a coil for the particular band selected. This unit is one of the simplest to operate—and highly efficient on all bands, for all types of emission.

Front Panel Controls: Oscillator Dial; Final Amplifier Dial; Oscillator Selector Dial; Modulation Selector Dial; Microphone Gain Control, Band Selector Switch; Filament Power Switch; Plate Power Switch; Emission Selector Switch; Standby Control.

Metering: PA Plate Current; PA Grid Current; Modulator Plate Current.

TUBE COMPLEMENT

›e	Function
-6AC7	Reactance Tube Modulator
6J5	Variable Frequency Oscillator
6AC7	Class "A" Amplifier or Crystal Oscillator
6L6	80 meter Buffer or 40 meter Doubler
6L6	20 meter Doubler
6L6	15 meter Tripler
6L6	10 meter Doubler
3D23	Final Amplifier
807	Class AB_2 Modulators
6J5	Modulator Driver
6SJ7	Speech Amplifier
866A	High Voltage Rectifiers
5R4GY	Low Voltage Rectifier
5R4GY	Modulator Rectifier
80	Speech Rectifier
6X5GT	Bias Rectifier
VR 150	Voltage Regulator
6SN7GT	Audio Oscillator

ELECTRICAL CHARACTERISTICS

Frequency Range:	Amateur Bands—10, 11, 15, 20, 40, 80 meters
Output Power:	100 watts on CW, ICW and Frequency Modulation 100 watts Amplitude Modulation
Method of Modulation:	AM—High Level Class AB_2 FM—Reactance Tube Modulation
Modulation Capabilities:	AM—100% FM—100% = ± 75 kilocycles
Input Audio Source:	High Impedance Crystal or Dynamic Microphone. Level 60 DB down
Audio Frequency Response:	AM—±2DB, 200 to 6000 cps FM—±1DB, 100 to 7500 cps
Noise Level:	AM—Minus 45DB below 100% modulation FM—Minus 60DB below 100% modulation (± 75 kilocycles)
Audio Frequency Distortion:	AM—5% at 85% modulation for 100 watt output FM—1.5% at 100% modulation
Frequency Control Elements:	Stabilized Variable Frequency Oscillator or two (2) crystal controlled positions.

MPACT: 29 ⅜" long, 11 ¾" wide, 18⅜" deep. Power Source: 110-117 volts 50/60 cycles AC.
ınomical — Power Consumption: 325 watts. Approximate Weight: 125 lbs.

COMPLETE: The only items needed to get "on the air" are a key, a mike and two crystals.

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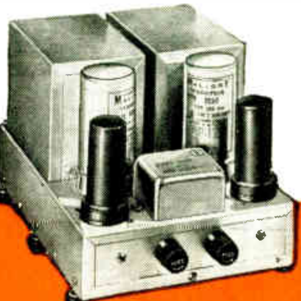

The Vibrapack line includes models for input voltages of 6, 12, and 32 volts DC and nominal output voltages from 125 to 400. Models available with switch for four output voltages in 25-volt stages. Hermetically sealed vibrators. High efficiency—low battery drain.

Vibrapacks* Provide Dependable Plate Power for Portable Equipment

Mallory Vibrapacks provide economical, efficient and dependable plate power for operating radio receivers, transmitters, PA systems, direction finders and other electronic equipment on vehicles, farms, portable equipment, or wherever commercial AC power is unavailable.

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Built with adequate safety factor for long life. Round and square can styles. Available in 20 stock sizes, working voltages from 600 to 6,000.

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Mallory fixed and adjustable power resistors provide maximum efficiency in operation with excellent temperature and humidity characteristics. Available in rated capacities from 10 to 200 watts, resistances from 1 to 100,000 ohms.

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Ceramic insulation provides low losses at high frequencies. The Mallory Ham Band Switch is rated for use in transmitter plate circuits using up to 1,000 volts DC with power up to 100 watts, inclusive.

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- Volume Controls

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. . . feature hard-surfaced materials, patented design and skilled workmanship. They are manufactured for long service under extreme conditions.

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. . . are interchangeable with air trimmers for most applications and have advantages in space requirements and mechanical stability.

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. . . have self-cleaning double-wipe contact jaws, positive index, adjustable stop and bakelite or steatite insulation

CUSTOM CERAMICS

. . . at Centralab include Steatite, Cordierite, Zirconite and a wide range of dielectric materials. Available to manufacturers only.

TUBULAR CERAMIC CAPACITORS

. . . are manufactured for high frequency circuits where the ultimate in stability and retrace characteristics is essential.

HI-ACCURACY CERAMIC CAPACITORS

. . . represent a new development in which the order of accuracy is suitable for rigid frequency control applications.

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. . . include off-on switches, station selectors in inter-communication systems, talk-listen switches and simplified band change switches.

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. . . are suitable for transmitters, power supply converters and special industrial and electronic uses. Sections and indexes also available.

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. . . are particularly adapted to broadcasting, receiving, public address, test instruments and industrial uses. They are available in ten combinations.

CERAMIC TRANSMITTING CAPACITORS

. . . feature extremely low power factor, stable retrace characteristics and maximum flashover commensurate with physical size.

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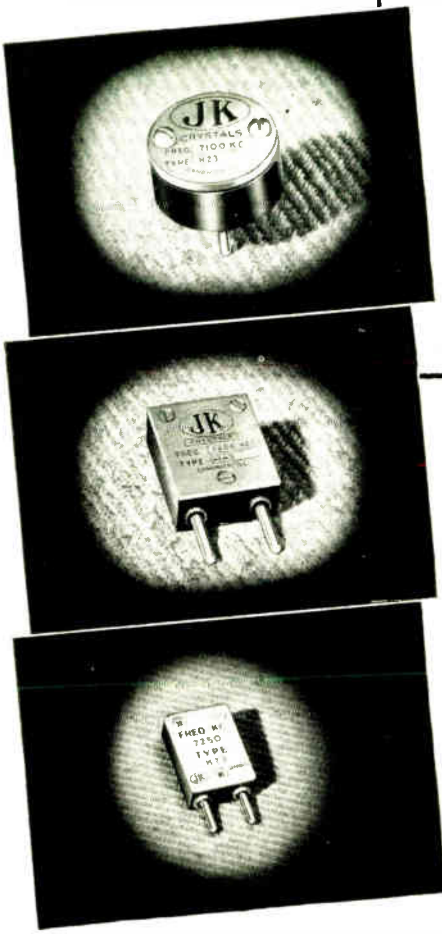
"Stabilized"*

Crystals



Precision Quality for the Amateur!

When building your new rig or revamping the old one to conform to the new Ham bands, be sure to use JK "Stabilized" Crystals—the crystals that are so carefully made and so perfectly mounted that you can install them and forget about them. JK "Stabilized" Crystals will stay "right on" at any normal operating temperature, and they're not affected by dust, moisture or vibration. Ask your jobber about JK "Stabilized" Crystals or write direct for our new illustrated folder.

*JK "Stabilized" Crystals are made by the latest known methods of precision crystal manufacture. Our process known as "Stabilizing" absolutely prevents frequency shifts due to aging, either in operation or on the shelf.

FS344 Frequency Standard $79.50

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"A Radio Researcher Since the Early Days"

The men of the James Knights Company have grown up with Ham Radio. Because of their work with piezo quartz since it first came into use as a frequency control, they know what is expected of a good Ham Crystal. You can depend on JK "Stabilized" Crystals.

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W9DAX—"An Active Ham Since 1913"

The JAMES KNIGHTS Co.

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Sixty Miles Southwest of Chicago

Crystals for the Critical

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The many AlSiMag insulators in the average rig must meet several different insulation requirements. For that reason, AlSiMag insulators are made of many different technical ceramic compositions. All these compositions are sold under the trade name AlSiMag. Each one has definite electrical and mechanical properties which have been carefully determined. The new Property Chart gives the properties of the more frequently used AlSiMag compositions. A copy will be sent you free on request.

● All AlSiMag technical ceramics are custom made. Sales are made only to manufacturers or to supply houses. No sales can be made direct to hams. However, the Research Division of American Lava Corporation welcomes suggestions or criticisms and is always on the alert for suggestions for improvements in AlSiMag products.

Address requests for Property Chart to:

AMERICAN LAVA CORPORATION

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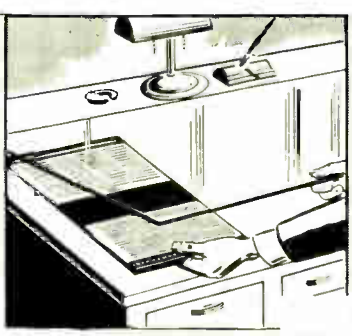
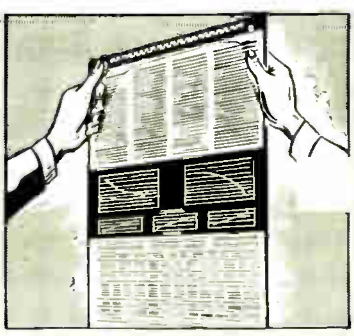
44TH YEAR OF CERAMIC LEADERSHIP

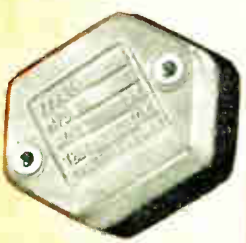
The new AlSiMag Property Chart is conveniently designed so that it can be tacked on the wall, placed under your desk glass or ploced in your regular business file. It is easy to reod and mokes it easy to find exactly the AlSiMag composition which is best suited for each of your electronic or electrical applications.





it all depends

MICA • DYKANOL



TYPE 9: Moulded mica capacitor for r.f. by-pass, grid and plate blocking. Mechanically strong, well-insulated, moisture-resistant, with short heavy terminals.

TYPE 59: Mica transmitter capacitor used as grid, plate blocking, coupling, tank and by-pass applications. In low-loss, white glazed ceramic case and wide-path end terminals.

TYPE 6: Mica capacitor with series-stack construction. For aircraft, portable equipment, high-power transmitters, etc. Especially suited for grid, plate, coupling, tank and by-pass applications.

. . . and you radio "hams" know it . . . capacitor performance "makes" or breaks" the rig. Now that you're getting back on the air you want to stay on it — without a letdown or a break.

You make no mistakes when you depend on C-D Capacitors. Cornell-Dubilier has built a name and a reputation that stand for consistently *dependable* capacitor design and manufacture. And, if uninterrupted pleasure in your hobby depends on capacitor performance, you'll be wise to depend on a high-quality, low-cost product that has enjoyed the loyal choice of wireless and radio amateurs for more than thirty-seven years.

ASK YOUR LOCAL C-D DISTRIBUTOR FOR CATALOG #195A, *containing complete listings of C-D Capacitors for Amateur Equipment, Capacitor Test Instruments and Quietone Interference Filters. Or write direct to Cornell-Dubilier Electric Corporation, South Plainfield, New Jersey.*

FIVE MODERN PLANTS • SKILLED PERSONNEL

The better to serve you, Cornell-Dubilier now operates five huge modernly-equipped plants centrally located: at South Plainfield, N. J.; New Bedford, Worcester, and Brookline, Mass., and at Providence, R. I. At each plant skilled capacitor specialists and trained technicians build C-D Capacitors for many uses, check every process carefully, test and cautiously inspect every capacitor to guarantee perfection.

CORNELL-DUBILIER

world's largest manufacturer of

1910

CAPACITORS

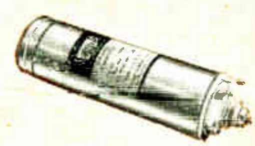
1947

P A P E R • E L E C T R O L Y T I C

TYPE YAB: Dykanol impregnated and filled, non-inflammable, non-oxidizable and remains unaffected by extensive variations in temperature, humidity or voltage stresses. For by-pass, audio frequency coupling, etc.

TYPE TQ: Dykanol transmitting filter capacitor with universal mounting. Designed primarily for amateur low-power broadcast and commercial transmitters; also adopted for high power, high fidelity public address systems and portable power amplifiers. In sturdy, non-corrosive aluminum case.

TYPE TJU: Dykanol transmitting filter capacitor, compact, lightweight, furnished with universal mounting clamp, well-insulated terminals. Hermetically sealed in non-corrosive, steel container.



TYPE TLA: Inverted Dykanol Filter capacitor of extra high dielectric strength, lower-r.f. resistance, light weight, high insulation resistance, and higher voltage breakdown. Hermetically sealed in sturdy aluminum containers.

"Cardyne" Cardioid Dynamic
Models 731 and 726

"Cardax" Cardioid Crystal
Model 950

Bi-Directional Velocity
Models V-3, V-2, V-1

Differential Carbon
Hand-Held Model 205-S

Model 610 Dynamic
Model 910 Crystal

No finer choice than

Electro-Voice

TODAY'S MOST COMPLETE LINE OF

MICROPHONES

Differential Dynamic
Model 606

WITH important E-V developments such as the *Mechanophase* Principle,* the *Differential† Principle,* the *Acoustalloy* Diaphragm... ELECTRO-VOICE brings you the widest selection of quality microphones, for both general and specific applications. In the complete E-V line of uni-directional, bi-directional, and non-directional types, you can easily find exactly the right microphone to give you the sound pick-up and reproduction you want A few models are shown here. ELECTRO-VOICE engineers are glad to help you on any special problems. *Send now for new E-V Catalog No. 101.*

*Patents Pending †Patent No. 2,350,010
Crystal Microphones Licensed under Brush patents.

High Fidelity Dynamic
Model 630

Hand-Held Dynamic Model 600-D
Differential Dynamic Model 602

Versatile Dynamic
Model 640

General-Purpose Dynamic
Model 605

Differential Handset
Model 260 Carbon, Model 660 Dynamic

Comet Crystal Model 902

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THERE'S AN RCA TUBE

RCA POWER TRIODES

810—Class B or class C. 750 watts input in class C telegraphy up to 30 Mc. at 2500 plate volts.

811—Class B or class C. 220 watts audio output in class B at 1500 plate volts.

812—Class C r-f amplifier. 225 watts input up to 60 Mc. at 1500 plate volts.

833-A—Class C r-f amplifier. A single tube takes a full KW at comparatively low plate voltage.

8005—Class B or class C. 300 watts input in class C telegraphy up to 60 Mc. at 1500 plate volts.

RCA MINIATURES

3A4—Power amplifier pentode. AF output, 700 milliwatts. RF output, 1.2 watts at 10 Mc.

3A5—HF twin triode. Class C telegraphy, output of about 2 watts at 40 Mc.

6AK5—HF amplifier pentode with sharp cut-off, for frequencies up to 400 Mc.

6J4—UHF amplifier triode. Primarily for use as grounded-grid amplifier up to 500 Mc.

9001—Detector amplifier pentode with sharp cut-off. For UHF applications.

9002—UHF triode detector, amplifier, or oscillator in superheterodyne receivers at frequencies up to 500 Mc.

RCA RECTIFIERS

5R4-GY—Full-wave, high-vacuum type. 250 ma. output at 700 volts.

816—Half-wave, mercury-vapor type. In full-wave circuit, 250 ma. output at 1500 volts.

866A/866—Half-wave, mercury-vapor type. In full-wave circuit, 500 ma. output at 3100 volts.

RCA BEAM POWER

807—Oscillator or amplifier 75 watts input in class C telegraphy up to 60 Mc. at 750 plate volts.

813—Class C amplifier. 500 watts CW input up to 30 Mc. at 2250 plate volts.

2E26—Class C amplifier. 40 watts CW input up to 125 Mc. and 30 watts input up to 160 Mc. at 600 plate volts.

RCA DUAL BEAM POWER

815—Class C amplifier. 68 watts input in class C telegraphy up to 150 Mc.

829-B—Class C amplifier. 120 watts input in class C telegraphy up to 200 Mc. and 105 watts up to 250 Mc.

832-A—Class C amplifier. 36 watts input in class C telegraphy up to 200 Mc. and 32 watts up to 200 Mc.

RCA ACORNS

6F4—UHF triode oscillator for frequencies up to 1200 Mc.

954—UHF pentode detector or amplifier for frequencies up to 430 Mc.

955—UHF triode detector, amplifier or oscillator. For receivers or transmitters up to 600 Mc.

958-A—UHF triode amplifier or oscillator for low-power UHF transceivers.

FOR EVERY AMATEUR SERVICE

RCA UHF POWER TRIODES

2C43—"Lighthouse" triode. 20 watts input up to 1500 Mc. Useful as keyed or modulated oscillator as high as 3000 Mc.

826—Oscillator, r-f amplifier or frequency multiplier. 50 watts input in class C telegraphy up to 250 Mc.

8025-A—Oscillator, r-f amplifier or frequency multiplier. 50 watts input in class C telegraphy up to 500 Mc.

RCA RECEIVING TYPES

NOW—CW RATINGS ON RCA RECEIVING TUBES

Strictly for the benefit of radio amateurs, Class C CW transmitting ratings have been established on the following receiving types: 6AG7, 6AK6, 6AQ5, 6F6, 6L6, 6N7, 6V6-GT and 12AU7.

Detailed information on these new ratings will be found in the October-November 1946 issue of Ham Tips. A copy may be obtained on request.

Have you seen HAM TIPS?
Get a free copy from your local
RCA Tube Distributor

RCA has an amateur type tube for every service, every power and every active band. A few of the most popular types in each classification are listed.

In addition, there are special types, such as voltage regulators, thyratrons, and the well-known receiving types in metal, glass, and miniature.

Your local RCA Tube Distributor has complete technical data on all RCA tube types. Contact him for further information on the types in which you are interested, or write RCA, Commercial Engineering, Section A-1K, Harrison, New Jersey.

RCA LABORATORIES,
PRINCETON, N. J.

THE FOUNTAINHEAD OF MODERN TUBE DEVELOPMENT IS RCA

TUBE DEPARTMENT

RADIO CORPORATION of AMERICA

HARRISON, N. J.

The CONNEAUT Model 600-S
DYNAMIC Model DN-HZ-S
Model N-80
Crystal and Dynamic
MICROPHONES
FOR EVERY AMATEUR APPLICATION
Model D-104
Model G Stand
Model T-3
Model JT-40
FROM away back in 1933, when Astatic introduced Model D-104, the first practical crystal microphone ever developed, veteran amateurs the world over, have long used and enjoyed Astatic microphones Many models with desired voice range characteristics, including new streamlined designs, are now available. For grand performance and long, dependable service . . . it's an "Astatic" . . . every time.
See your Radio Parts Jobber, or write for Catalog
THE Astatic CORPORATION
ASTATIC
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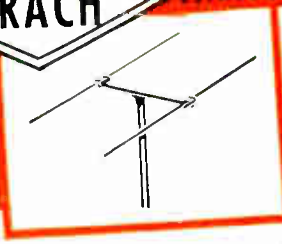
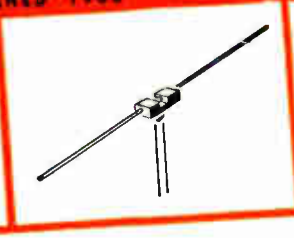
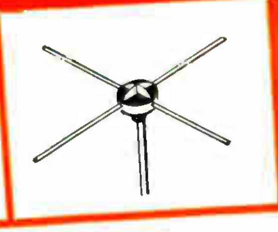
Astatic Crystal Devices manufactured under Brush Development Co patents

BRACH ESTABLISHED 1906

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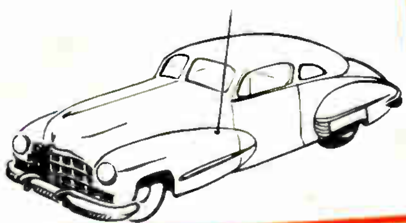
for USERS of PRIVATE BRANDS

This Company, which has devoted its major interests to the manufacture of Antenna Systems since the inception of radio broadcasting, now offers its expanded facilities to users of private brands. Our engineering service will cooperate with you on your own designs.

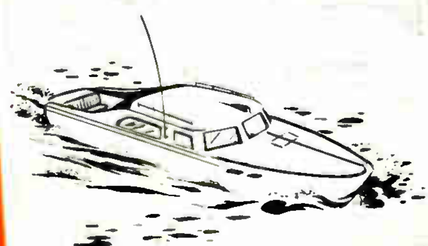
We manufacture the following, under private labels and trademarks:



AUTOMOBILE ANTENNAS . . . every variety, including types that can be raised and lowered from inside the car.

RESIDENTIAL ANTENNAS . . . AM and FM, for homes, stores and multi-family buildings. Complete lines of noise-reducing systems incorporating latest patented developments of coupling transformers.

FM, AM and TELEVISION . . . Dipoles with or without reflectors, folded dipoles, turnstile, radiating types and other combinations for roof, sidewall and other mountings.



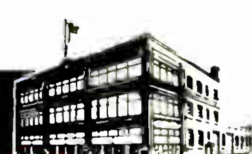
MARINE ANTENNAS . . . Collapsible and transmitting types for every purpose.

For POLICE and other mobile units . . . roof-top antennas for ultra-high frequencies.

WE INVITE INQUIRIES AND CONSULTATIONS.

SOME OF THE OTHER BRACH PRODUCTS

Lightning Protective Devices • Junction Boxes • Pot Heads • Gas Relays • Arrester Housings • Protective Panels • Terminals and Housings • High Tension Detectors • Test-O-Lite for Circuits 100-550 AC or DC • Solderall



L. S. BRACH MFG. CORP.
200 CENTRAL AVENUE NEWARK, 4 N. J.

WORLD'S OLDEST AND LARGEST MANUF… OF RADIO ANTENNAS AND ACCESSORIES

An Outstanding Success!

Raytheon's

RC-11 STUDIO CONSOLE

for AM or FM

The Most Versatile Unit of its Kind . . . Easily Controlling Two Studios, Announcer's Booth and Nine Remote and Two Network Lines.

THIS REMARKABLE Raytheon Console commands the attention of studio engineers and managers as few items of broadcast equipment ever have!

It provides *complete* high-fidelity speech-input facilities with all the control, amplifying and monitoring equipment contained in a single compact cabinet. It easily handles any combination of studios, remote lines or turntables—broadcasting and auditioning simultaneously, if desired, through two high quality main amplifier channels. It makes it a simple matter to cue an oncoming program and pre-set the volume while another program is on the air.

Note the sloping front and backward-sloping top panel, giving maximum visibility of controls and an unobstructed view into the studio. Note the telephone-type, lever action, three-position key switches, *eliminating nineteen controls.*

The beauty of this console, in two-tone metallic tan . . . the efficient, functional look of it . . . will step up the appearance of any studio, yet blend easily with other equipment.

RAYTHEON MANUFACTURING COMPANY

Broadcast Equipment Division

7517 No. Clark Street, Chicago 26, Ill.

Devoted to Research and Manufacture for the Broadcasting Industry

Compare

THESE OUTSTANDING FEATURES WITH ANY OTHER CONSOLE

1. **SEVEN** built-in pre-amplifiers—*more than any other console*—making possible 5 microphones and 2 turntables, or 7 microphones, on the air simultaneously.
2. **NINE** mixer positions—*more than any other console*—leading to 5 microphones, two turntables, one remote line and one network line.
3. **NINE** remote and two network lines—*more than any other console*—may be wired permanently.
4. **TELEPHONE-TYPE** lever-action key switches used throughout—the most dependable, trouble-free switches available. *No push buttons.*
5. **FREQUENCY RESPONSE** 2 db's from 30 to 15,000 cycles. Ideal speech input system for either AM or FM.
6. **DISTORTION** less than 1%, from 50 to 10,000 cycles.
7. **NOISE LEVEL** minus 65 db's or better. Airplane-type four-way rubber shock mounting eliminates outside noise and operational "clicks."
8. **ALL FCC REQUIREMENTS** for FM transmission are met.
9. **DUAL POWER SUPPLY** provides standby circuit instantly available for emergency use.
10. **POWER SUPPLY** designed for mounting on desk, wall or relay rack.
11. **INSTANT ACCESS** to all wiring and components. Top hinged panel opens at a touch. Entire cabinet tilts back on sturdy full-length rear hinge.

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Excellence in Electronics

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Reliable Service

FOR AMATEURS ALL OVER THE WORLD

WHY do hams in Casablanca, Buenos Aires, Athens, Shanghai, and Helsinki seek out "The Shack" — by mail, cable, and sometimes in person? BECAUSE since 1924 RADIO SHACK has meant RELIABLE SERVICE to amateurs in the far corners of the world . . . Service made possible by our real understanding of the problems of the amateur and by an enormous stock of the products of all the world-famous manufacturers.

The service that has made Radio Shack customers in Cavite and Constantinople was established for you — the amateurs of America. Use it to build your better, bigger "Ham Shack" — faster, for less cost.

You'll find our Catalog H-47 a handy aid. Write today — it's free.

Your Microphone or Pickup is Here!

"55"

The Unidyne is the most widely publicized microphone for public address. A super-cardioid dynamic that is used where perfect sound reproduction is an absolute *must*.

The "556" Dynamic has all the essentials for high quality broadcasting. Has Super-Cardioid pickup pattern that reduces random noise by 73%.

"556"

"100"

High quality carbon microphone specially designed for military and police equipment where ruggedness and dependability are vital factors.

The Uniplex Cardioid Crystal offers the advantages of high-priced performance at low cost. Excellent for public address, recording, and similar applications.

"730B"

"508"

The Stratoliner Dynamic is a rugged microphone with unusually smooth response — good for police transmitters, airports, and industrial paging systems.

The Stratoliner Crystal is good for use when low cost is important factor. Clear reproduction of voice or music. High output with wide range response.

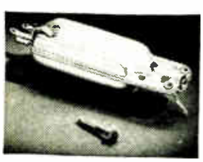
"708"

"76B"

The "76B" Lapel is designed for public address, lecturing in large halls, broadcasting from portable transmitters. High quality speech reproduction.

"707A" Crystal has been designed to give quality performance at low cost. Suitable for P. A. systems, call systems, ham use, and similar applications.

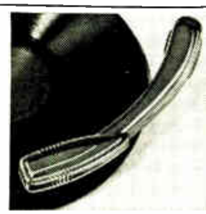
"707A"



LEVER-TYPE CARTRIDGE

Shure Cartridges are good for replacement of crystal cartridges in current use. Flexible needles, high output voltages. Pin-tip terminals — no soldering necessary.

Lightweight crystal pickup with needle force of 1⅛ ounces. Replaces heavy pickups on old phonographs and can be installed on new record players. Clearer, fuller tone qualities.



"GLIDER"

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Microphones and Acoustic Devices

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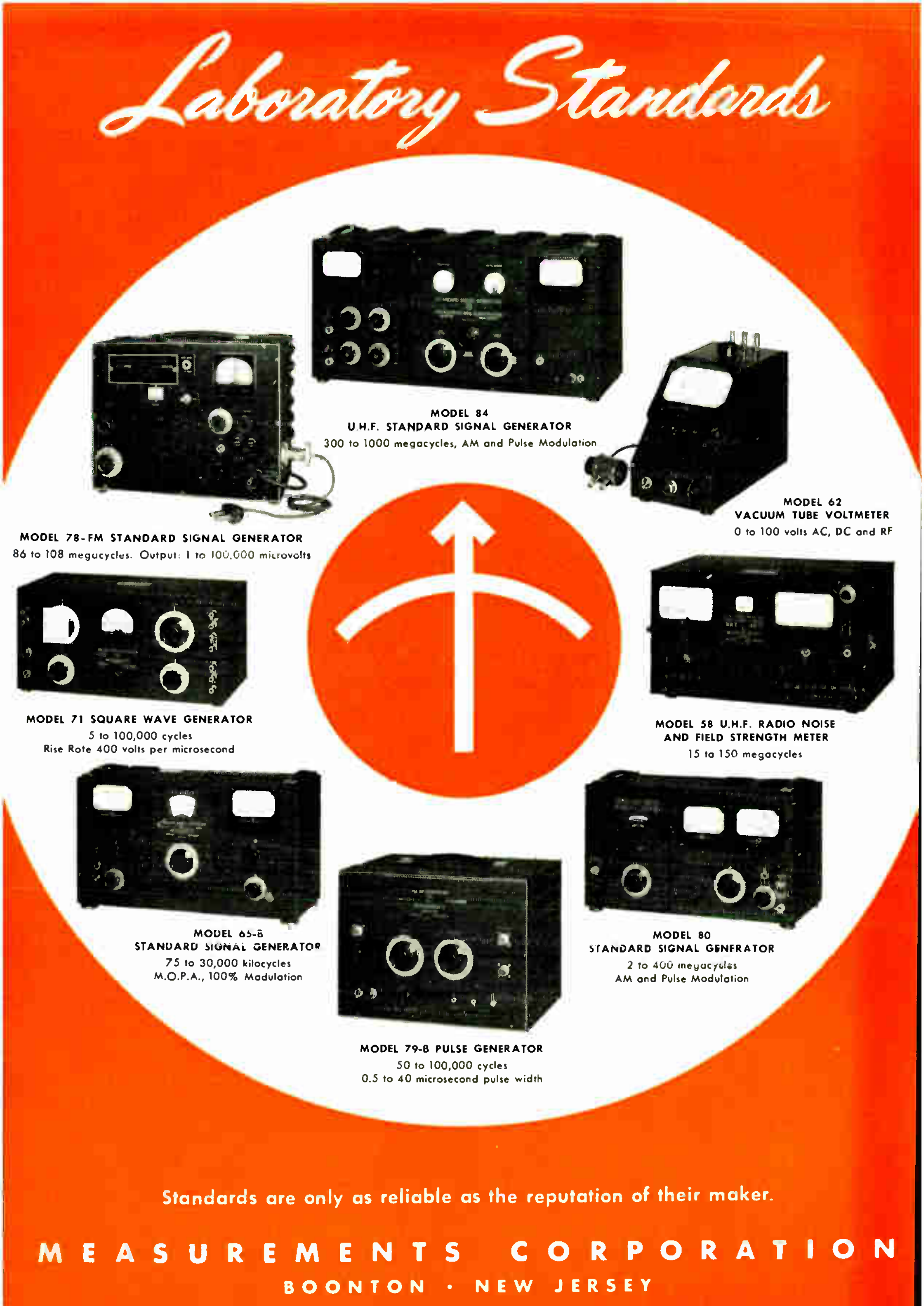
Laboratory Standards
MODEL 84
U.H.F. STANDARD SIGNAL GENERATOR
300 to 1000 megacycles, AM and Pulse Modulation
MODEL 78-FM STANDARD SIGNAL GENERATOR
86 to 108 megacycles. Output: 1 to 100,000 microvolts
MODEL 62
VACUUM TUBE VOLTMETER
0 to 100 volts AC, DC and RF
MODEL 71 SQUARE WAVE GENERATOR
5 to 100,000 cycles
Rise Rate 400 volts per microsecond
MODEL 58 U.H.F. RADIO NOISE
AND FIELD STRENGTH METER
15 to 150 megacycles
MODEL 65-B
STANDARD SIGNAL GENERATOR
75 to 30,000 kilocycles
M.O.P.A., 100% Modulation
MODEL 80
STANDARD SIGNAL GENERATOR
2 to 400 megacycles
AM and Pulse Modulation
MODEL 79-B PULSE GENERATOR
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0.5 to 40 microsecond pulse width
Standards are only as reliable as the reputation of their maker.
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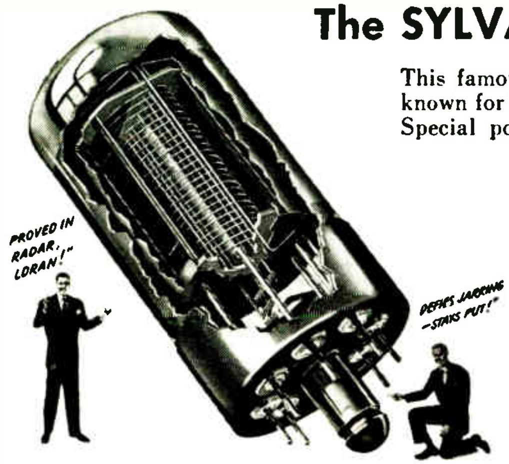
The SYLVANIA LOCK-IN TUBE



This famous product of Sylvania research is well known for its electrical and mechanical superiority. Special points of merit, of interest to you, are:

Lock-In locating plug . . . also acts as shield between pins.

Short, direct connections . . . fewer welded joints—less loss.

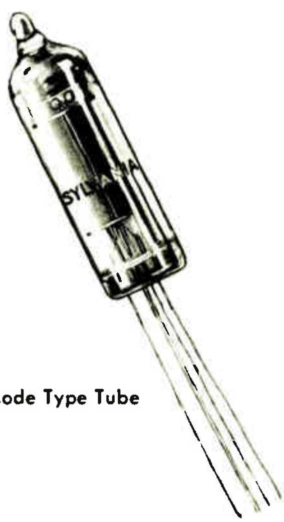
Getter located on top . . . shorts eliminated by separation of getter material from leads.

No top cap connection . . . overhead wires eliminated.

Reduced overall height . . . space saving.

Stays put in mobile and portable rigs.

THE NEW HIGH-FREQUENCY OSCILLATOR



6K4 Cathode Type Tube

The 6K4 cathode type tube—another development of special interest to hams. This high-frequency oscillator, in the new T-3 size, is ideally suited for your use. It's compact, rugged—developed from the famous proximity fuze type tube that was made to be shot from a gun.

SYLVANIA

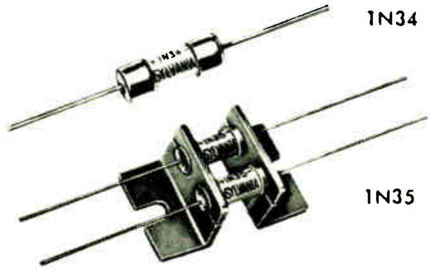
MAKERS OF RADIO TUBES; CATHODE RAY TUBES; ELECTRONIC DEVICES;

SOME OUTS ANDING PRODUC S
OF SYLVANIA RESEARCH

The 1N34 GERMANIUM CRYSTAL DIODE *and* 1N35 DUO-DIODE

FEATURES

1. Smoll size.
2. Eliminotion of heoter supplies. Removes possible source of hum.
3. Pigtoil construction —con be soldered into ploce (1N34).
4. Greot resistonce to vibrotion ond shock.
5. Low forword resistonce volue.
6. Low shunt copocitonce (obout 3 micromicroforods for unit mounted in ploce in circuit).



1N34

1N35

The 1N34 and 1N35 are ideal for use in lightweight and portable equipment. Fields of application include: field strength meters, detectors, clippers, discriminators, series noise limiters, demodulators, meter rectifiers.

The LOW-COST, EASY-TO-USE MODULATION MONITOR

Now you can monitor your modulation percentage and speech quality with this new Sylvania Model X-7018 Modulation Meter. Compactly styled. Economical. Of great assistance in complying with FCC regulations on overmodulation. Helps keep your average percentage up between 60% and 90%. Indicates carrier shift.

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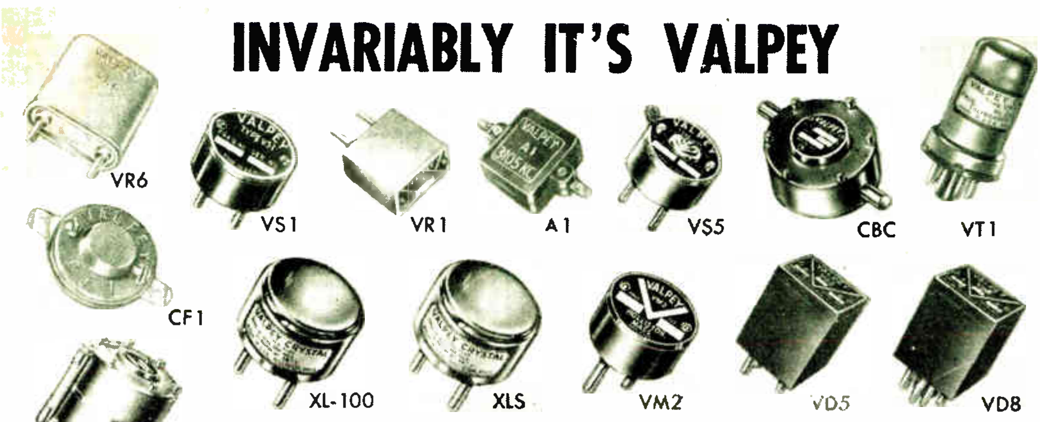
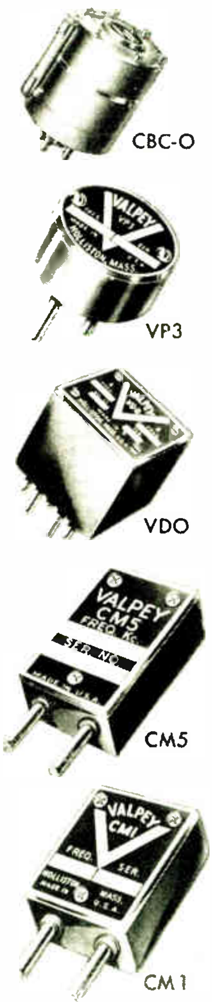
FLUORESCENT LAMPS, FIXTURES, WIRING DEVICES; ELECTRIC LIGHT BULBS

INVARIABLY IT'S VALPEY





TYPE	FREQUENCY RANGE	PINS	DESCRIPTION	USE
CBC-O	60-10000KC	Standard 5-Pin Mount	6, 8, 10 Volt Oven Variable Air Gap ±½°C. Accuracy	Broadcast, Fixed Stations and Freq. Standards.
CBC	60-10000KC	Special	Micrometer Adjust. Variable Air Gap	Broadcast, Fixed Stations and Freq. Standards.
VDO	1000-10000KC	Standard 5-Pin Mount	Single or Dual 6 Volt Oven Gasket Sealed ±½°C. Accuracy	Fixed and Mobile for Transceiver Equipment. Railroad Communications.
VS5	1000-4000KC	.125 Dia. Pins ¾" Spacing	Variable Air Gap Horizontal Mount	Police and Fixed Stations.
VS1	1000-4000KC	.125 Dia. Pins ¾" Spacing	Fixed Air Gap Pressure Clamped Horizontal Mount	Police and Fixed Stations.
VD5	1000-6000KC	Special 3-Pin Mount 5/32" Dia.	Single or Dual Crystals Gasket Sealed	Marine, Aircraft or Police.
VD8	1000-6000KC	Octal 1, 8-4, 5 Xtal A—Xtal B	Single or Dual Crystals Gasket Sealed	Marine, Aircraft or Police.
XLS	80-1000KC	.125 Dia. Pins ¾" Spacing	Clamped Crystal Mount. Hermetically Sealed	Radar and Fixed Stations in the Low Frequency Range.
XL-100	100KC	.125 Dia. Pins ¾" Spacing	Clamped Crystal Mount. Hermetically Sealed	Frequency Standards.
VT1	1000-10000KC	Octal 2, 3-7, 8	Vacuum Sealed Metal Tube Type Unit	Frequency Meters, Standards and General Applications.
VM2	1000-4000KC	.125 Dia. Pins ¾" Spacing	Fixed Air Gap Horizontal Mount Gasket Sealed	Fixed and Mobile Applications.
VP3	2000-60000KC	.125 Dia. Pins ¾" Spacing	Fixed Air Gap Horizontal Mount Gasket Sealed	Marine, Police, Amateur, Fixed and Mobile Stations.
CM1	1000-4000KC	.125 Dia. Pins and G.R. Pins ¾", ⅝", ⅞", .850 Spacing	Gasket Sealed Fixed Air Gap Vertical Mount	Marine, Police, Aircraft and General Applications.
CM5	2000-60000KC	.094 Dia. Pins .486" Spacing	Gasket Sealed Fixed Air Gap Vertical Mount	Marine, Police, Amateur, Fixed and Mobile Stations.
A1	1000-4000KC	Solder Lugs	Flat Compact Gasket Sealed	Aircraft
VR1	2000-10000KC	.125 Dia. Pins .486" Spacing	Fixed Air Gap Vertical Mount Gasket Sealed	Marine, Police, Aircraft.
CF1	455, 456, 465 KC	Solder Lugs	Small, Flat, Compact	Filter Applications.
VR6	4000-60000KC	.050 Dia. Pins .486" Spacing	Vacuum Sealed Metal Case	Mobile, Fixed Stations, VHF, Experimental.

For every crystal application, VALPEY invariably gives outstanding performance. Select your VALPEY unit from the above chart, or send your specific crystal requirements to VALPEY. In every field where accurate crystal control is the aim — invariably it's VALPEY.

Valpey CRYSTALS
HOLLISTON, MASS.
Craftsmanship in Crystals Since 1931

For CW Only or CW and FON Operation

The HARVEY
100-T TRANSMITTER

You can get the HARVEY 100-T Transmitter for CW operation only (without Modulator) or complete for radio, telephone and telegraph operation. These sturdy, efficient, thoroughly dependable units will meet your highest expectations of operating ease and performance.

The HARVEY 100-T without modulator a complete CW transmitter capable of 175 watts input on 5 bands.

The HARVEY 100-T Modulator may be added to the 100-T Transmitter at any time

SPECIFICATIONS

Frequency Range: 1.7-30 mc.
Power Input: 130 Watts Phone
175 Watts CW
Power Output: 100 Watts Phone
130 Watts CW
Radio Frequency Tubes: 6V6 Crystal oscillator
6L6 Doubler
814 Final amplifier
Audio Frequency Tubes: 6SJ7 1st Audio amplifier
6SF5 2nd Audio amplifier
6F6 Class B driver
2-807 Class B modulators

Rectifier Tubes: 2-866 Final amplifier supply
83V Oscillator-Doubler supply
5Z3 Speech amplifier supply
RK-60 Modulator supply
Power Source: 115 volts 50/60 cycles
Power Drain: 730 watts
Microphone: Single cell crystal type
Cabinet Size: 20½" high, 19½" wide, 13½" deep
Net Weight: 150 Lbs. 68.04 Kilos
Shipping Weight: 225 Lbs. 102.06 Kilos

SEE The Performance Of Your Circuits

The HAR-CAM Visual Alignment Signal Generator provides the swiftest, surest method for the adjustment and alignment of tuned circuits. With it the complete radio frequency response curve of any circuit can be presented on an oscilloscope screen. Variations in circuit alignment can be accurately evaluated and necessary changes made in a matter of seconds.

Left: The HAR-CAM Visual Alignment Signal Generator instantly shows up the misalignment in I-F circuit. Right: With the HAR-CAM Signal Generator, the performance of the circuit is easily, correctly adjusted.

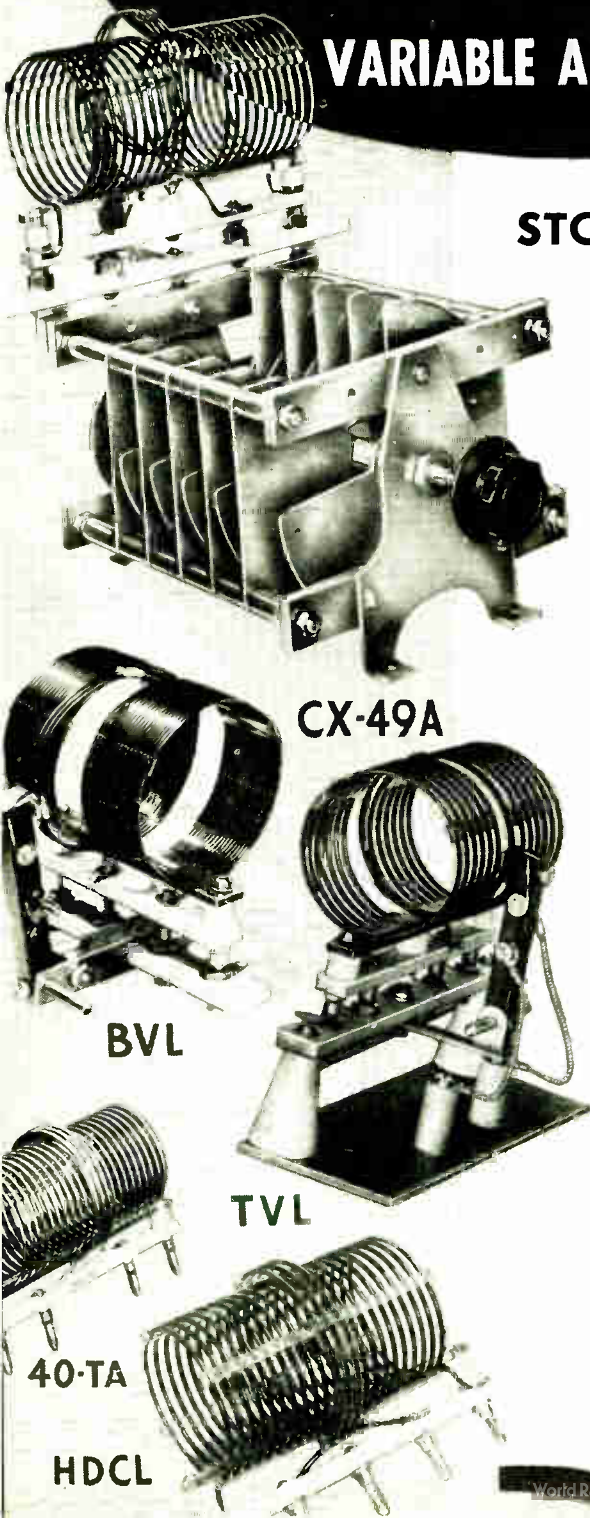
Write for Bulletins containing information on the latest HARVEY Transmitter developments.

HAR-CAM
HARVEY OF CAMBRIDGE

HARVEY RADIO LABORATORIES, INC.
436 Concord Ave., Cambridge 38, Mass.

Makers of Marine Radio Telephones, Regulated Power Supplies and Emergency Communications Equipment.

B & W
AIR INDUCTORS
AND
VARIABLE AIR CONDENSERS



STOP TANK CIRCUIT LEAKS

. . . with this complete B & W Coil and Capacitor assembly

B & W Type CX Variable Capacitors provide for direct mounting of B & W Air Inductors. Wiring is eliminated. Circuit lead lengths are reduced to an absolute minimum. Opposed stator sections in the capacitors provide short r-f path. Butterfly rotor construction permits grounding rotor at the center r-f voltage point with respect to stators. Built-in neutralizing capacitors can be mounted on end plate. Standard types rated at 500, 750 and 1,000 watts. Treat your new rig to *real* tank circuit efficiency! Write for catalog.

Neutralizing Plates Available in 4 Types

B & W B, T AND HD INDUCTORS

100-WATT, 500-WATT AND 1 KW TYPES

- MINIMUM DIELECTRIC IN THE FIELD OF THE COIL
- EXTREMELY LOW LOSSES — RUGGED CONSTRUCTION
- EXCELLENT APPEARANCE — LOW COST

Type "B" inductor is for use on oscillator and buffer-doubler stages developing up to 100 watts. Available in center tapped models without link; end link; center link, center tapped; and variable link — center tapped. For 5, 10, 15, 20, 40 and 80 meter bands.

Type "T" is specially suited for high powered neutralized buffer and final tank stages where powers of 500 watts are developed. Available in center tapped models without link; center linked with center tap and variable linked with center tap. Made for 10, 15, 20, 40 and 80 meter bands.

Type "HD" is for maximum power and handles a kilowatt with ease. Available in center tapped models without link; center linked with center tap and variable linked with center tap. Made for 10, 15, 20, 40 and 80 meter bands.

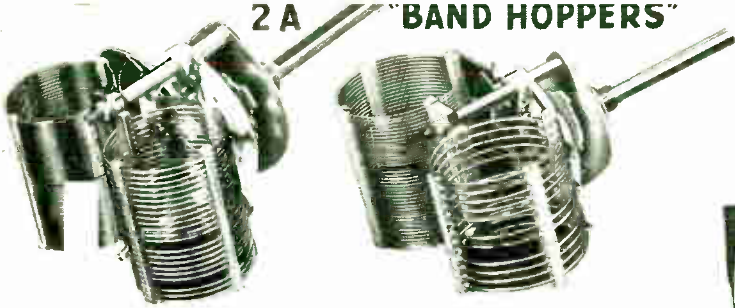
B & W TVH INDUCTORS

for powers up to 500-watts input

Here is a special group of units designed for greater flexibility through use of an eight plug jack bar. With these inductors it is possible to connect automatically, a fixed padding capacitor when using the low frequency coil. Available for 10, 15, 20, 40 and 80 meter bands.

SEE B & W PRODUCTS AT YOUR JOBBER'S

Write for Complete Amateur Catalog



B & W TURRET ASSEMBLIES

Fast, positive band switching for your rig! Moderate in cost — easy to install — adaptable to 80, 40, 20, 15 and 10 meter bands. These turrets eliminate absorption effects through use of a unique switching assembly which shorts unused coils.

B & W — 75-Watt 2A "BAND HOPPERS" — A compact and panel controlled unit which may be used for interstage coupling between two beam power tubes or between beam power tubes and triodes.

B & W 75-WATT TURRETS — for link coupling single ended or push-pull low power stages. Mounted on a positive action switch arranged for panel mounting through a single ⅜" hole.

Type JTCL — Center linked, center tapped coils.
Type JTEL — End linked, untapped coils.

B & W 150-WATT TURRETS — for single- and double-ended circuits. These mount the same as 75-watt turrets and are used with tubes operating at voltages up to 1000 volts.

Type BCL — Center linked, center tapped coils.
Type BEL — End linked, untapped coils.

B & W BABY TURRETS—35-WATTS

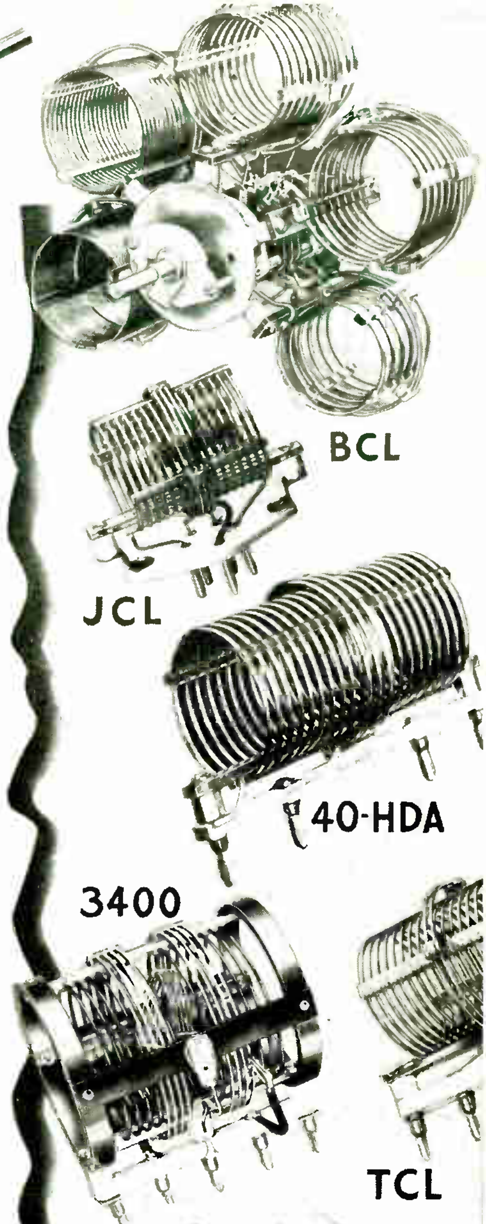
Rated at 35 watts, these compact, 5-band switching units cover amateur bands from 10 to 80 meters. They are suitable for all services with any of the 50 mmfd. midget condensers. Sturdy construction and unusual design assures permanent coil alignment and maximum efficiency with the minimum number of tubes. Available in four types: BTM straight untapped; BTCT — center tapped; BTEL — end linked; and BTCL — center linked. All provide vastly improved band switching efficiency in low power transmitters and exciter stages.

ANTENNA INDUCTORS TA AND HDA

These coils are wound with tinned copper wire for ease in tapping feeders and have fixed center links for coupling to either fixed or variable linked final tank circuits through low impedance line. Available for 10, 15, 20, 40 and 80 meter bands. Type TA for power input up to 500 watts and Type HDA for power inputs of one kilowatt.

B & W 3400 SERIES INDUCTORS

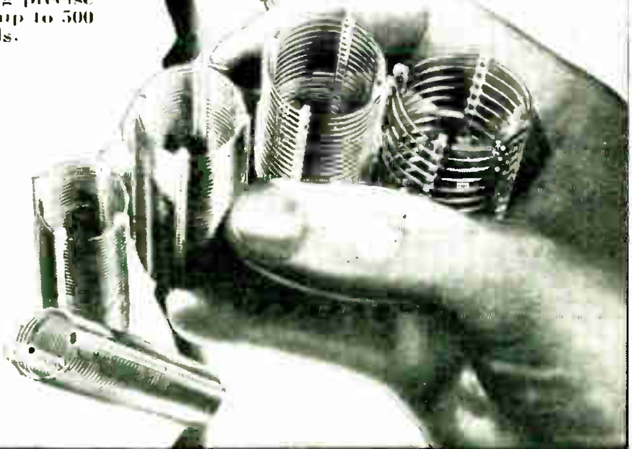
Presenting the utmost in sturdy construction and electrical flexibility, these coils are built with an individual internal center coupling, adjustable over 360° — permitting precise impedance matching up to 600 ohms. For powers up to 500 watts. Available for 10, 15, 20, 40 and 80 meter bands.



THE MIDGET R-F COILS
of dozens of uses

Goodbye to homemade high-frequency coils! B & W Miniductors cost little, are beautifully constructed — and do the job right. Every day, amateurs, experimenters and equipment manufacturers tell us of new applications where Miniductors have replaced homemade coils with a big boost of efficiency. Use them for receivers, transmitters and test equipment — in tank circuits as r-f chokes, high-frequency i-f transformers and loading coils and for dozens of other purposes.

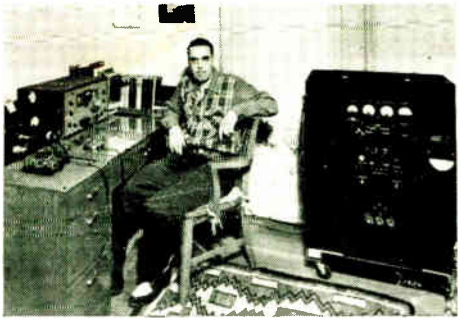
B & W "Air Wound" construction permits small but sturdy supports with the absolute minimum of insulating material in the electrical field. Q factor is amazingly high. Standard Miniductor diameters are ½", ⅝", ¾" and 1", each available in four different winding pitches. Ask your jobber. He can supply these coils, individually packaged, in standard 2" or 3" lengths.



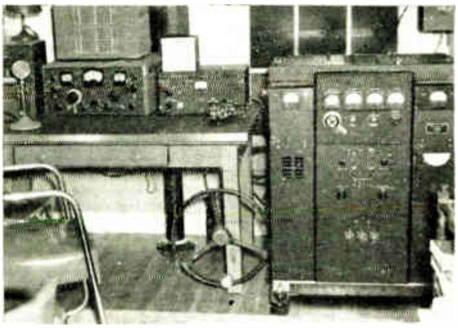
BARKER & WILLIAMSON

235 FAIRFIELD AVE., UPPER DARBY, PA.

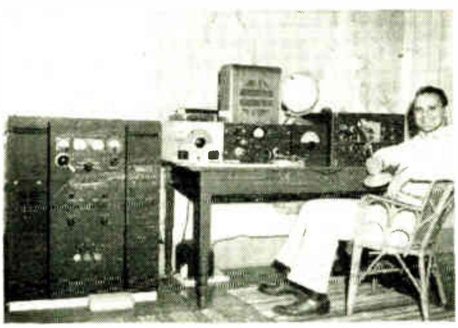
"Equipped by ALLIED"



WØFBS—Tom Atherstone—Denver, Colorado
Equipped by ALLIED. Listen for him on 10 and 20 meter phone.



W5DZ—Colonel W. P. Clarke—Waco, Texas
Equipped by ALLIED. Listen for him on 20 and 75 meter phone.

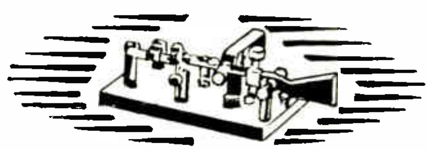


W5SH—Edwin C. Shaw—Fort Worth, Texas
Equipped by ALLIED. Listen for him on 10 and 20 meter phone.

"Equipped by ALLIED" is a famous by-word in the Amateur Radio world. From the early days of 200 meters and spark transmitters to the micro-waves of today, ALLIED-equipped rigs have been writing radio history.

The three post-war stations illustrated here are typical of hundreds of modern Ham Shacks "equipped by ALLIED." Each has its own story—a friendly, personal story, because into each has gone the full interest of ALLIED'S amateurs working with fellow amateurs. There's more to each than just the equipment.

Whether you're starting from "scratch," rebuilding, or newly licensed, you can count on ALLIED . . . where hams understand ham problems . . .



IN AMATEUR RADIO FOR OVER 25 YEARS

ALLIED RADIO

833 W. JACKSON BLVD., DEPT. 10-7, CHICAGO 7, ILL.

WRITE FOR YOUR COPY OF THE LATEST ALLIED CATALOG

PR CRYSTALS

SINCE 1934

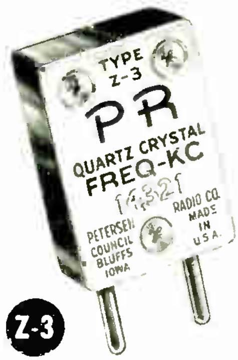
ACCURACY — STABILITY ACTIVITY — HIGH OUTPUT DEPENDABILITY *at Low Cost*

For years PR Precision Crystals have set performance standards in all types of service . . . amateur, commercial, marine, broadcast, mobile, police, aircraft. PRs are the foremost choice of amateurs . . . the most critical users of crystals today. PR Crystals have earned this reputation by LOW DRIFT characteristics, less than 2 cycles per MC per degree Centigrade . . . HIGH OUTPUT AND DEPENDABILITY even at highest permissible crystal currents . . . ACCURACY within .01 per cent of specified frequency . . . HIGH ACTIVITY especially desirable for break-in CW operation . . . X-ray orientation . . . CONTAMINATION AND MOISTURE-PROOF through permanent gasket seal . . . ½-inch pin spacing. Every PR is UNCONDITIONALLY GUARANTEED. Your EXACT FREQUENCY (Integral Kilocycle) AT NO EXTRA COST. See your jobber for PRs. His stock is complete for ALL BANDS. Accept no substitute. — Petersen Radio Company, 2800 West Broadway, Council Bluffs, Iowa. (Telephone 2760)

COMMERCIAL PR Type Z-1	Frequency range 1.5 to 10.5 MC. Designed for rigors of all types of commercial service. Calibrated .005 per cent of specified frequency. Weight less than ¾ ounce. Sealed against moisture and contamination. Meets FCC requirements for all types of service.
80 and 40 METERS PR Type Z-2	Rugged. Low drift fundamental oscillators. High activity and power output. Stands up under maximum crystal currents. Stable, long-lasting, permanently sealed . . . . $2.65 Net
20 METERS PR Type Z-3	Harmonic oscillator. Low drift. High activity. Can be keyed in most circuits. Stable as fundamental oscillators. Fine for doubling to 10 and 11 meters or "straight through" 20 meter operation. $3.50 Net
10 METERS PR Type Z-5	Harmonic oscillator for "straight through" mobile operation and for frequency multiplying to VHF. Heavy output in our special circuit . . . . . $5.00 Net

Z-1

Z-2



Z-3

Z-5

PR *Precision* CRYSTALS

In Service Everywhere!..

DUMMY ANTENNA RESISTORS

To check R. F. power, determine transmission line losses, check line to antenna impedance match. Helps tune up to peak efficiency. Non-inductive, non-capacitive, constant in resistance. 100 and 250 watt sizes in various resistances.

BROWN DEVIL RESISTORS

Small, extra sturdy, wire wound vitreous enameled resistors for voltage dropping, bias units, bleeders, etc. Proved right in vital installations the world over. 10 and 20 watt sizes in resistances up to 100,000 ohms.

PARASITIC SUPPRESSOR

Small, light, compact non-inductive resistor and choke, designed to prevent u.h.f. parasitic oscillations which occur in the plate and grid leads of push-pull and parallel tube circuits. Only 1¾" long overall and ⅝" in diameter.

CENTER-TAPPED RESISTORS

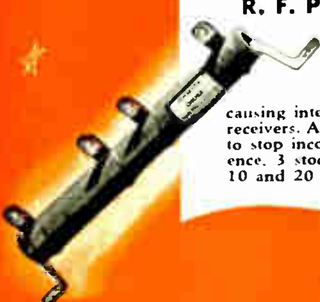
For use across tube filaments to provide an electrical center for the grid and plate returns. Center tap accurate to plus or minus 1%. Wirewatt (1 watt) and Brown Devil (10 watt) units, in resistances from 10 to 200 ohms.

R. F. PLATE CHOKES

Single layer wound on low power factor steatite or bakelite cores, with moistureproof coating. Nine stock sizes for all ham bands from 1.8 mc to 460 mc. Small, high frequency chokes mount by wire leads. Larger sizes mount on brackets. All sizes rated 1000 ma or more.

ADJUSTABLE DIVIDOHMS

You can quickly adjust these handy Dividohms to the exact resistance you want, or put on one or more taps wherever needed. 7 sizes from 10 to 200 watts. Many resistance values up to 100,000 ohms.



R. F. POWER LINE CHOKES

Keep R.F. currents from going out over the power line and causing interference with radio receivers. Also used at receivers to stop incoming R.F. interference. 3 stock sizes, rated at 5, 10 and 20 amperes.

FIXED RESISTORS

Resistance wire is wound over a porcelain core, permanently locked in place, insulated and protected by Ohmite vitreous enamel. Available in 25, 50, 100, 160 and 200 watt stock sizes, in resistances from 1 to 250,000 ohms.

Be Right with OHMITE

RHEOSTATS ★ RESISTORS ★ CHOKES ★ TAP SWITCHES

Ohmite Vitreous Enamel is unexcelled as a protective and bonding covering for resistors and rheostats.

OHMITE Rheostats ★ Resistors ★ Chokes ★ Switches

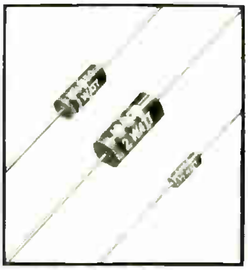
CLOSE-CONTROL RHEOSTATS

Insure permanently smooth, close control of communications, electronic and electrical devices. Widely used in industry and in war equipment. All ceramic, vitreous enameled. 25, 50, 75, 100, 150, 225, 300, 500, 750 and 1000 watt sizes. Approved Army and Navy types.

HIGH-VOLTAGE SWITCH

For general use where high voltage insulation is required. Suitable for circuits up to 1 K.W. rating. Used for band changing, meter switching, tapped transformer circuits, etc. Ceramic construction.

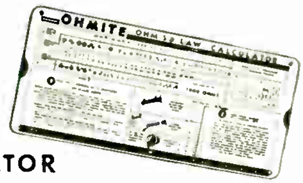
LITTLE DEVIL INDIVIDUALLY MARKED INSULATED COMPOSITION RESISTORS



New, tiny, molded fixed resistors each marked with resistance and wattage rating. ½ Watt, 1 watt and 2 watt sizes, 10% tolerance. Meet Army-Navy Specification JAN-R-11. Can be used at full wattage rating at 70 C (158 F) ambient temperature. Dissipate heat rapidly. Low noise level. Low voltage coefficient. Stocked in standard RMA values from 10 ohms to 22 megohms.

Available only from Ohmite Distributors

HANDY OHMITE OHM'S LAW CALCULATOR



Very useful in training schools, in laboratories and in industry. Figures ohms, watts, volts, amperes — quickly, easily. Solves any Ohm's Law problem with one setting of the slide. All values are direct reading. No slide rule knowledge is necessary. Scales on two sides cover the range of currents, resistances, wattages and voltages commonly used in radio and electronic applications. Size only 4⅛" x 9". Send only 10c in coin to cover handling cost.

HIGH-CURRENT TAP SWITCHES

Compact, all ceramic, multipoint rotary selectors for A.C. use. Silver to silver contacts. Rated at 10, 15, 25, 50 and 100 amperes with any number of taps up to 11, 12, 12, 12, and 8 respectively. Single or tandem assemblies.

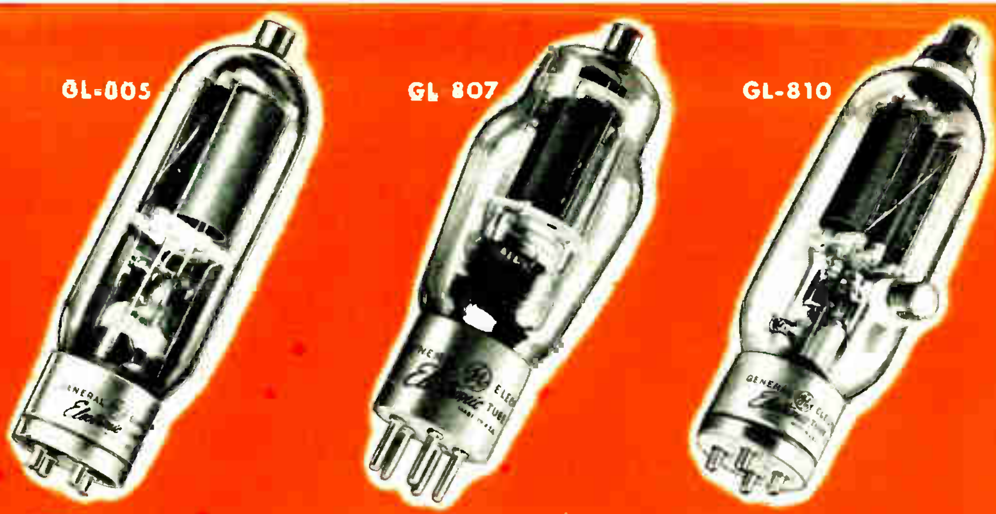
RB-2 DIRECTION INDICATOR POTENTIOMETER

A compact, low cost unit used in a simple potentiometer circuit as a transmitting element to indicate, remotely, the position of a rotary beam antenna. Used with an 0-1 milliameter.

SEND FOR FREE STOCK CATALOG—Gives helpful information and data on Ohmite stock units for essential applications—lists hundreds of stock values. Very handy for quick reference.

OHMITE MANUFACTURING COMPANY

4844 Flournoy St., Chicago 44, U. S. A. Cable "Ohmiteco"



STANDARD TUBES...
BRAND-NEW TYPES...

..both are included in G.E.'s complete line!

YOUR G-E tube distributor will be glad to give you, on request, a copy of Booklet ETX-19, listing all G-E ham tubes with their prices and ratings. Or write for this booklet—as well as any special circuit information or facts you need about tube applications—direct to *Electronics Department, General Electric Company, Schenectady 5, New York.*

Read G.E.'s "Ham News" regularly, and keep posted on up-to-the-minute circuit developments. Your distributor has this publication for you—**FREE.**

GENERAL ELECTRIC

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ELECTRONIC TUBES OF ALL TYPES FOR THE RADIO AMATEUR

HAMS for HAMS by HAMS of HAMS

LET'S GET THIS STRAIGHT!

What we mean by the slogan on the left is just this: Newark has HAMS in all branches. Being hams, they are naturally FOR HAMS. And they know how to fill all requests BY HAMS. The huge stocks they draw from will fill all the needs OF HAMS in their pursuit of the world's most fascinating hobby.

ONLY ONE WAY TO GET THE LATEST DOPE!

You naturally want to be in on the latest dope in ham merchandise. We keep everyone informed so far as possible by advertising. BUT — magazines go to press long before they appear. And today's fast-moving markets find new products and tremendous surplus bargains coming out between issues. So we publish:

BIG BARGAIN BULLETINS--GET 'EM!

Chock full of real ham values, these are mailed direct to hams who want them and who take a moment to let us know they want them. So drop us a card or fill in the coupon below and paste it on a card. Do it now!

X3

NEWARK ELECTRIC COMPANY, INC.
(Send to Chicago or New York, whichever is nearer)

Gentlemen: By all means put my name on your list of those who receive the bargain bulletins.

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150 MORE APPROVED TYPES

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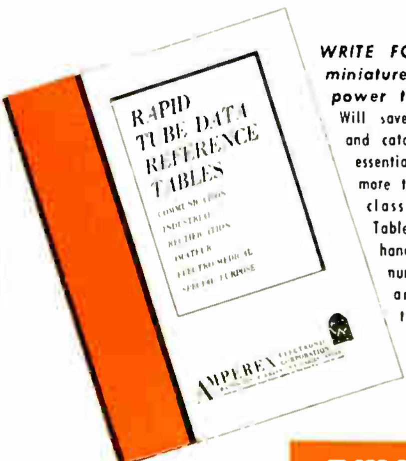
AMPEREX POWER TUBES

For a quarter century AMPEREX has been closely identified with creative research, laboratory approach, precision manufacture and helpful service in its chosen field — power tubes. This specialized concentration has earned AMPEREX a unique position of leadership.

Make Good Equipment Better With Better AMPEREX Tubes



WRITE FOR THIS 8-page miniature encyclopedia of power tube information. Will save you hours of text and catalog exploring. Gives essential characteristics of more than a hundred tubes classified by function. Tables keyed to output with handy cross index of type numbers. Truly useful and practical. Real time-saver. Write for your copy today — IT'S FREE.

YOUR PLANS—YOUR PROBLEMS

As tube specialists deeply concerned with all modern developments, Amperex engineers are in a position to give you detached advice and information. YOU ARE INVITED to write the *Amperex Application Engineering Department* for answers to specific questions.

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...HEADSETS

Your next pair of headsets will be a TRIMM. . . . Perhaps your station is known 'round the world; the best is none too good; TRIMM Featherweights or Commercials are your choice. Perhaps yours is a more modest rig; TRIMM Professionals, Dependables, "B" or "E" may fill your needs. Or perhaps you're just about to learn code and pennies count. . . . You'll want a TRIMM Acme or Rex. Twenty-five years' experience assure the best in each price class.

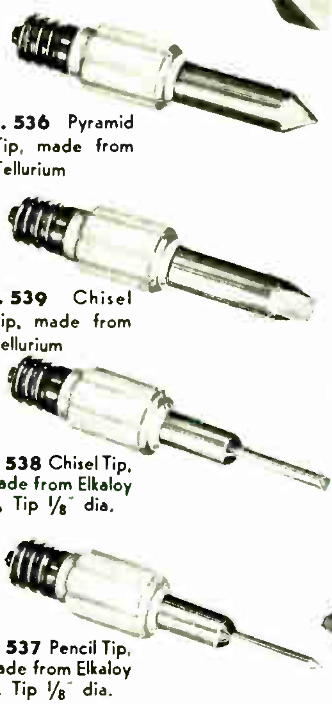
Available at all leading distributors of amateur supplies throughout the world. Literature available by addressing Department A.

TRIMM, INC.
1770 W. Berteau Ave., Chicago, Ill.

Light & Versatile

FOR BETTER SOLDERING

Built for speedy accurate soldering, this powerful little pencil iron will perform rugged heavy duty jobs as well as those requiring intricate exactness.



No. 536 Pyramid Tip, made from Tellurium

No. 539 Chisel Tip, made from Tellurium

No. 538 Chisel Tip, made from Elkaloy A, Tip 1/8" dia.

No. 537 Pencil Tip, made from Elkaloy A, Tip 1/8" dia.

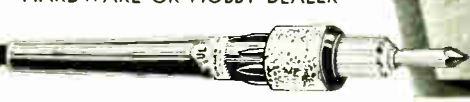
Check These Points!

- √ QUICK HEATING . . . 90 SECONDS
- √ LIGHTWEIGHT . . . . 3.6 OUNCES
- √ HANDY SIZE . . . . 7 INCHES
- √ SAVES ELECTRICITY . . 20 WATTS

√ Perfect Balance

These 4 INTERCHANGEABLE tips give you a point for each particular job.

SEE YOUR NEAREST RADIO, HARDWARE OR HOBBY DEALER



UNGAR Electric Tools

BOX 2255, TERMINAL ANNEX · LOS ANGELES 54, CALIF.

The Collins 32V-1 150 watt Transmitter

Medium Power in Small Space

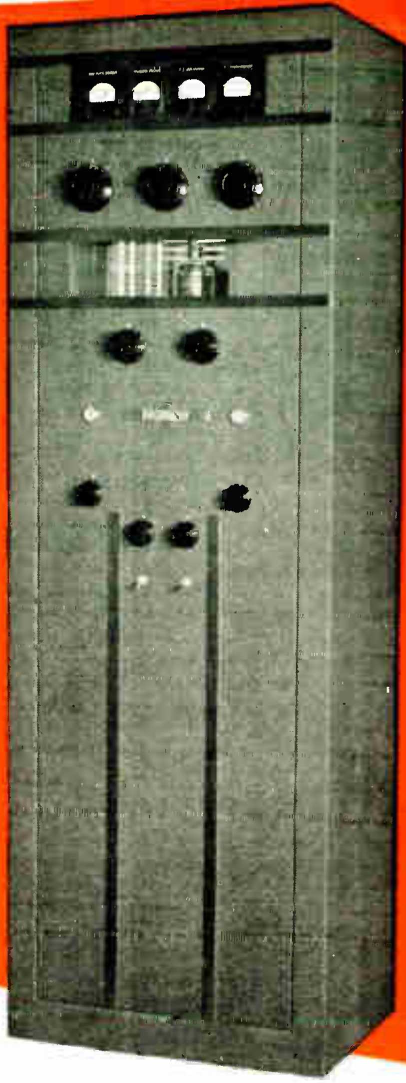
The 32V-1 is a natural for those who want medium power in a small cabinet. It is complete in one package. All you need to put it on the air are a key or microphone, antenna, and a 115 volt a-c power source. Its convenient size allows either permanent installation or portable use.

The 32V-1 is rated at 150 watts input on CW or 120 watts on phone. A receiver type cabinet houses the entire unit—power supply, audio, and r-f. All r-f stages except the final are permeability tuned and ganged with the v.f.o. The dial is calibrated directly in frequency. Bandswitching is employed, thus eliminating plug-in coils.

The output network will match impedances of 50 ohms to 600 ohms. Balanced or unbalanced open wire lines and antennas, and concentric transmission lines, can be used with good efficiency. Other transmission systems can be used with link coupling to the transmitter output.

In brief, here are the features of the 32V-1:

- bandswitching
- v.f.o. control
- 150 watts input on CW
- 120 watts input on phone
- 80, 40, 20, 15, 11, 10 meters
- push-to-talk
- table model
- 21⅛" w, 12 7/16" h, 13⅞" d
- ganged tuning
- pi output network
- clean keying
- direct frequency calibration



The Collins 30K
500 watt Transmitter

Designed Specifically for Amateur Radio

Every detail of the 30K is thoroughly engineered to assure the best performance for amateurs—it is not modified military equipment. Operating convenience and reliability are provided in the design and construction. The v.f.o. controlled exciter unit is in a receiver type cabinet that can be set right on the operating desk. Bandswitching in both the exciter unit and the transmitter itself facilitates multi-band operation. Three sets of antenna terminals are provided, with provision for switching antennas.

The speech clipper and low pass audio filter in the speech amplifier enable the operator to maintain a high average modulation, yet keep a narrow signal and prevent overmodulation.

Compare the following features—see how they fit your desires:

- bandswitching
- v.f.o. control
- 500 watts input on CW
- 375 watts input on phone
- 100% modulation
- speech clipper
- push-to-talk
- smooth, modern styling
- clean, sharp keying
- 80, 40, 20, 15, 11, 10 meters
- break-in operation
- 115 volts a-c power source

Attractive in appearance, efficient in operation, the 30K will make a satisfying nucleus for your ham shack.

FOR RESULTS IN AMATEUR RADIO, IT'S . . . COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

11 West 42nd Street, New York 18, N.Y. • 458 South Spring Street, Los Angeles 13, California

The Collins 75A Receiver

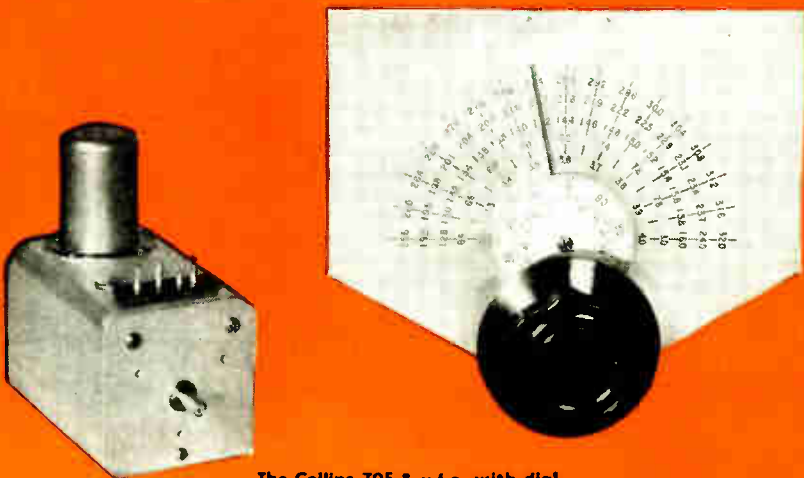
A New Standard for Amateur Receivers

The 75A was engineered specifically for amateurs. It covers six ham bands, with straight line tuning on all bands. The calibration is accurate to within one kilocycle on 15 meters, and to within two kilocycles on the 11 and 10 meter bands. Double conversion is utilized. The overall stability is within one dial division under all normal operating conditions.

The 75A is permeability tuned. It performs equally well on all amateur bands. Image rejection is a minimum of 50 db *on all bands*. The thoroughly engineered crystal filter circuit operates smoothly in providing a bandwidth variable in five steps from 4 kc to 200 cps. There is no loss in gain.

Here are some of its many desirable features:

- double conversion
- straight line tuning
- direct frequency calibration
- 80, 40, 20, 15, 11, 10 meters
- 50 db image rejection
- variable selectivity
- high sensitivity
- self-contained power supply
- signal strength meter
- permeability tuned
- receiver disabling circuit
- 10 db signal to noise ratio
- three IF amplifiers
- very high stability
- accurate calibration
- amplified avc



The Collins 70E-8 v.f.o. with dial

Know Your Frequency with this v. f. o.

The overall accuracy and stability of the 70E-8 are within 0.015% under all normal operating conditions. That means that you can set it to within ½ kc of any desired frequency on the 80 meter band, *and know that it will stay there.*

Sixteen turns of the vernier dial vary the frequency from 1600 kc to 2000 kc. The following table shows the relation between the oscillator frequency and various amateur bands:

Band (meters)	Freq. (mc)	No. of dial divisions	kc/dial division
80	3.5-4.0	1000	0.5
40	7.0-7.3	300	1.0
20	14.0-14.4	200	2.0
15	21.0-21.5	166	3.0
11	27.185-27.455	67.5	4.0
10	28.0-29.7	425	4.0
6	50.0-54.0	533.3	7.5
2	144.0-148.0	200	20.0
1¼	220.0-225.0	833.3	30.0
⅔	420.0-450.0	500	60.0

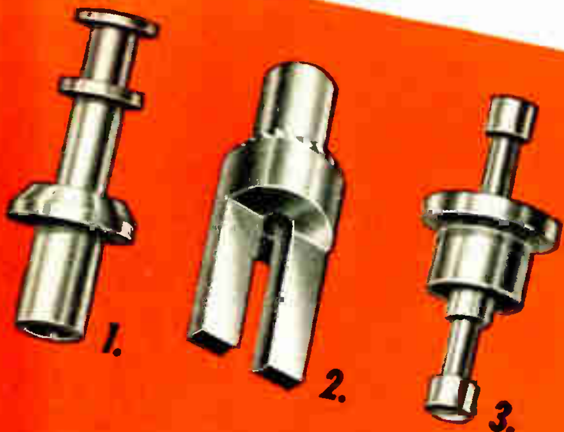
The 70E-8 is permeability tuned. The dial is calibrated directly in frequency up to and including the 10 meter band. 10 volts r-f output are available for use in an exciter, band edge spotter, heterodyne frequency meter, or other applications.

FOR RESULTS IN AMATEUR RADIO, IT'S . . . COLLINS

COLLINS RADIO COMPANY, Cedar Rapids, Iowa

11 West 42nd Street, New York 18, N. Y. • 458 South Spring Street, Los Angeles 13, California

COMPONENTS THAT *Click*



1. 2. 3.

4.

5.

6.

1. C.T.C. TURRET TERMINAL LUGS. Heavily silver plated brass lugs. A short cut to speedy assembly. Firmly anchored to terminal boards by simple swaging operation. Lugs heat quickly, assuring neat, positive wiring. Two soldering spaces. Stocked to fit 1/32", 1/16", 3/32", 1/8", 3/16" and 1/4" terminal board thicknesses. Also available with single soldering space.

2. C.T.C. SPLIT TERMINAL LUGS are being enthusiastically received by manufacturers of transformers and other potted units that require soldering after potting. A .050" hole through the lugs makes them ideal for this type of application. Perfect for terminal boards, too, because wiring can be done from top or bottom of the board without drilling. Made of brass, heavily silver plated, to fit 3/32" terminal boards.

3. C.T.C. DOUBLE END TERMINAL LUGS. Twin terminal posts in a single swaging operation. Perfect electrical contact because both posts are part of the same lug. Neat, positive wiring from either top or bottom. These heavily silver plated brass lugs are stocked to fit 3/32" terminal boards.

4. C.T.C. ALL-SET TERMINAL BOARDS are proving a time-saver in the laboratory and on the assembly line. Just select proper width board and go to work.

All-Set Terminal Boards are made in 3/32", 1/8" and 3/16" linen bakelite in 4 widths — 1/2"; 2" (lug row spacing 1 1/2"); 2 1/2" (lug row spacing 2") and 3" (lug row spacing 2 1/2"). Fit all standard resistors and condensers. Boards may be broken in fifths by bending on scribed line or used full length. Available in sets of any of the 4 widths or in any single width in lots of 6 or multiples of 6.

5. C.T.C. HAND PRESSURE SWAGER for quick, firm, uniform swaging of terminal lugs to terminal boards. Adjustable to fit all thicknesses of boards. Lugs are put in board *right side up* and may be swaged as far as 1 7/8" from edge. Adjustable pressure assures uniform swage. Unit pictured swages all C.T.C. standard Turret Lugs. Can be furnished with additional anvils and punches to fit C.T.C. Double End and Split Lugs.

6. C.T.C. MATHEMATICALLY DIMENSIONED CRYSTALS. A new C.T.C. development, mathematical dimensioning achieves greater accuracy. It assures consistent performance—guarantees frequency stability, high activity and long life in every C.T.C. Crystal.

WRITE FOR C.T.C. CATALOG NO. 100

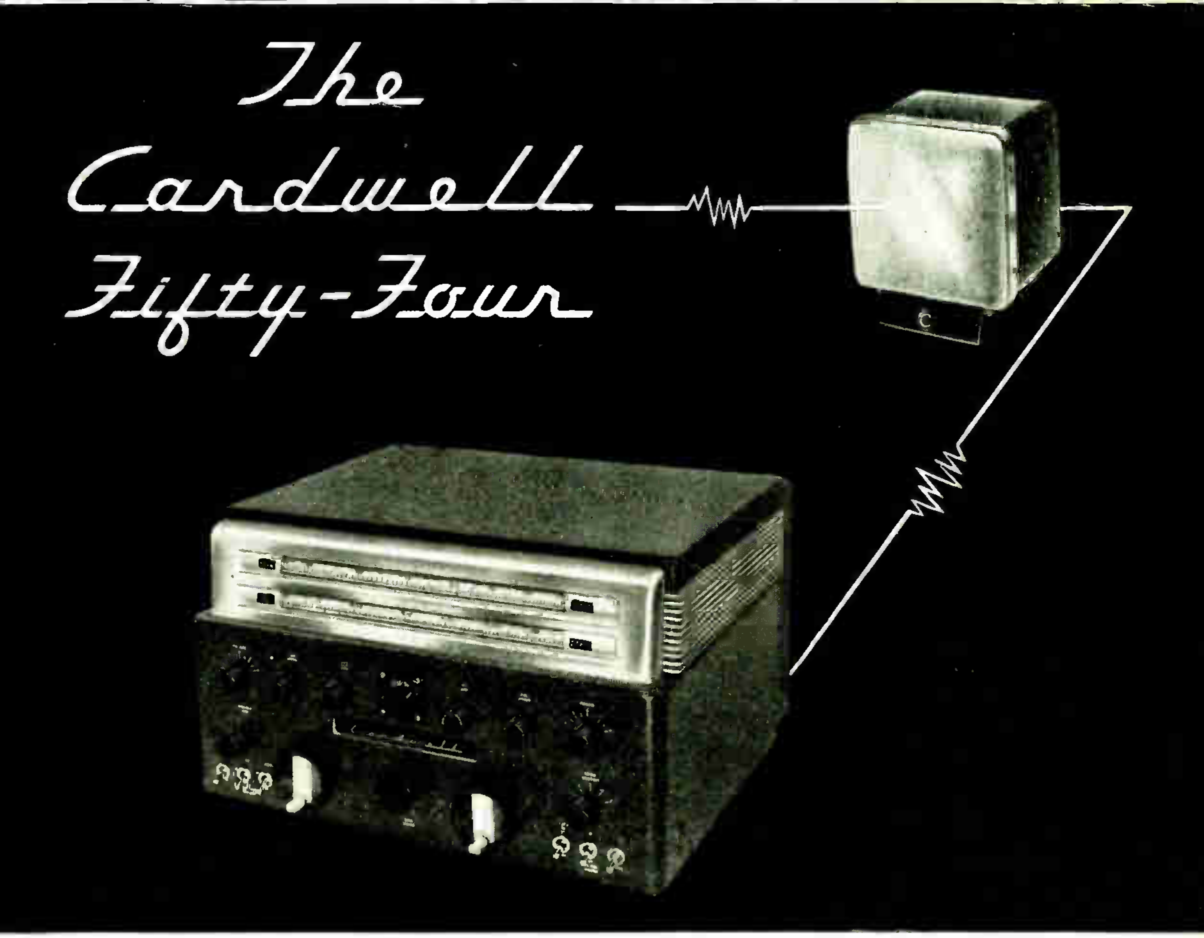
It contains complete information on these and other C.T.C. radio and electronic components you should know about. It's yours for the asking.

CTC

CAMBRIDGE THERMIONIC CORPORATION

455 CONCORD AVENUE • CAMBRIDGE 38, MASSACHUSETTS

BUILT TO CARDWELL STANDARDS



for more and better QSOs

Here is a communications receiver that is engineered to satisfy exacting commercial requirements. Yet it is ideal for the amateur who wants better QSOs and more of them. Remember, it is engineered by Cardwell and built to Cardwell standards . . . then read the following outstanding features:

1. Full Turret Type R. F. Section.
(Sturdy cast aluminum construction.)

2. Wide Frequency Coverage.
(Range .54 to 54.0 mcs. Basic turret covers .54 through 40 mcs. in six bands. Extra coil strip, supplied on special order, extends range to 54 mcs.)

3. Secondary Frequency Standard.
(Unique type crystal calibrator provides check points of either 100 or 1000 kcs.)

4. Variable Selectivity Crystal Filter.
(Choice of 5 degrees of selectivity—three with crystal, two without.)

5. Exceptional Signal to Noise Ratio
(Receiver noise less than 6 db above thermal!)

6. New Type Noise Limiter.
(A really effective aid in reducing local ignition interference and similar noises.)

7. Electrical Band Spread.
(Band spread scales calibrated directly. Arbitrary scale 0-100 also visible on each setting.)

8. Direct Reading Precision Dials.
(Excellent visibility—pointer travel better than 10½ inches—velvet smooth dial action.)

9. Temperature Compensated Oscillator.
(Stability is better than 25 parts per million per degree centigrade. V.R. tube maintains maximum frequency stability against line voltage fluctuations.)

10. Mechanical Coupling Provisions.
(Control shafts are brought out at rear for linkage to other units such as a transmitter exciter.)

11. Aluminum Unit Construction.
(Receiver and power supply combined in one sturdy, lightweight unit 18¼" wide x 16" deep x 11" high. Weight approximately 70 lbs.)

12. Heavy Duty Speaker.
(Compact tilting unit 9¼" wide x 8¼" deep x 11" high for wall or table mounting.)

13. Eight Watts Audio Output.
(Push-pull class AB—with four output impedances. Connections provided for phono-pickup or high level microphone input.)

14. 18 Tubes—All Miniature.

15. Threshold Squelch.

16. Panoramic Adaptor Jack.

17. Rack Mounting Model.
(Will also be available.)

WRITE FOR COMPLETE TECHNICAL BULLETIN

THE ALLEN D. CARDWELL MANUFACTURING CORP.

MAIN OFFICE & FACTORY: 97 WHITING STREET PLAINVILLE, CONN.

YOUR JOBBER HAS A CARDWELL CAPACITOR FOR EVERY POWER AND PURPOSE

NOW AVAILABLE

FROM *Raytheon*

OVER 125 AMATEUR AND SPECIAL PURPOSE TUBES!

HIGHLIGHTS OF THE RAYTHEON LINE

RAYTHEON SUBMINIATURE TUBES

TYPE NO.	CONSTRUCTION	APPLICATION	FILAMENT Volts	FILAMENT Amps.	FILAMENT Type	Max. Plate Volts	Max. Plate Cur. Ma.	CAPACITANCES—Mmfds. G-P	Input	Output	TYPE NO.
CK502AX	Pentode	Output Stage	1.25	0.030	Oxide	45	0.6	.14	3.0	5.7	CK502AX
CK503AX	Pentode	Output Stage	1.25	0.030	Oxide	45	0.8	.1	3.7	6.3	CK503AX
CK505AX	Pentode	Voltage Amplifier	0.625	0.030	Oxide	30	0.15	.07	2.7	4.4	CK505AX
CK506AX	Pentode	Output Stage	1.25	0.050	Oxide	45	1.25	.09	3.5	6.2	CK506AX
CK510AX	Double Space Charge Tetrode	Voltage Amplifier	0.625	0.050	Oxide	45	0.06	0.6	2.4	2.1	CK510AX
CK515BX	Triode	Voltage Amplifier	0.625	0.030	Oxide	45	0.15	.65	1.5	2.5	CK515BX
CK556AX	Triode	U.H.F. Oscillator	1.25	.125	Filament	135	4.0	2.0	1.3	4.0	CK556AX
CK569AX	Pentode	Amplifier	1.25	.050	Filament	67 5	1.8	.01	3.3	3.8	CK569AX
CK605CX	Pentode	U.H F. Amplifier	6.3	0.2	Heater	120	7.5	0.015	4.4	3.8	CK605CX
CK606BX	Diode	U.H.F. Rectifier	6.3	0.15	Heater	420	9			2.1	CK606BX
CK608CX	Triode	Oscillator-Amp.	6.3	0.2	Heater	120	9				CK608CX
CK619CX	Triode	Oscillator-Amp	6.3	0.2	Heater	250	4				CK619CX

RAYTHEON TRANSMITTING TUBES

TYPE NO.	CONSTRUCTION	SPECIAL APPLICATIONS	FILAMENT Volts	FILAMENT Amps.	MAX. VOLT. Plate	MAX. VOLT. Screen	MAX. CUR.—MA. Plate	MAX. CUR.—MA. Screen	POWER—WATTS Dissipation	Drive	Output	CAPAC. Mmfds. G-P	TYPE NO.
2C34/RK34	Dual Triode	H-F Oscillator-Amp.	6.3	0.8	300		80		10	1.8	16	2.4	2C34 RK34
RK4D22	Beam Tetrode	R-F Oscillator-Amp.	25.2 12.6	0.8 1.6	750	350	300	35	50	1.25	100	0.27	RK4D22
RK4D32	Beam Tetrode	R-F Oscillator-Amp.	6 3	3.75	750	350	300	35	50	1.25	100	0.27	RK4D32
5D23 RK65	R-F Tetrode	R-F Amplifier	5.0	14.0	3000	500	250	80		15.0	565	0.42	5D23 RK65
RK6D22	Tetrode	R-F, A-F Amplifier	5.0	28.5	3500	500	500	165	450	22.0	1000	0.5	RK6D22
RK20A	R-F Pentode	Suppressor Mod.	7.5	3.25	1250	300	92	36	40	1.6	84	0.01	RK20A
RK28A	R-F Pentode	Suppressor Mod.	10.0	5.0	2000	400	175	60	125	2 2	250	0.02	RK28A
RK38	Triode	R-F, A-F Amplifier	5.0	8.0	3000		160		100	10.0	225	4.3	RK38
RK48A	Beam Tetrode	R-F Oscillator Amp.	10.0	5.0	2000	400	180	40	100	1.2	250	0.2	RK48A
814 RK47	Beam Tetrode	R-F Oscillator-Amp.	10.0	3.25	1250	300	150	14	50	1.0	120	0 12	814/RK47

RAYTHEON RECTIFIER TUBES

TYPE NO.	CONSTRUCTION	FILAMENT Volts	FILAMENT Amps.	MAX. PEAK INVERSE VOLTS	MAX. PEAK CURRENT	AVERAGE CURRENT D.C.	AVERAGE TUBE DROP	TYPE NO.
BH	Full Wave—Gas			1,000	400 Ma.	125 Ma.	90	BH
RK3B24	Half Wave—High Vacuum	2.5 5.0	3.0 3.0	20,000 20,000	150 Ma. 300 Ma.	30 Ma. 60 Ma.		RK3B24
RK3B29	Half Wave—High Vacuum	2.5	4.75	16,000	250 Ma.	65 Ma.	130	RK3B29
RK4B31	Clipper Diode—High Vacuum	5.0	5.25	16,000	16 Amp.	60 Ma.	150	RK4B31
RK72	Half Wave—High Vacuum	2.5	3.0	20,000	150 Ma.	30 Ma.	200	RK72
RK705A	Half Wave—High Vacuum	2.5 5.0	5.0 5.0	35,000 35,000	375 Ma. 750 Ma.	50 Ma. 100 Ma.		RK705A
RK866A/866	Half Wave—Mercury	2.5	5.0	10,000	1.0 Amp.	250 Ma.	15	RK866A 866
RK872A/872	Half Wave—Mercury	5.0	7.5	10,000	5.0 Amp.	1.25 Amp.	10	RK872A 872
1005 CK1005	Full Wave—Gas	6.3	0.1	450	210 Ma.	70 Ma.	20	1005 CK1005
1006 CK1006	Full Wave—Gas	1.75	2.0	1,600	600 Ma	200 Ma.	20	1006 CK1006
CK1007	Full Wave—Gas	1.0	1.2	980	330 Ma.	110 Ma.	24	CK1007
CK1012	Full Wave—Gas	1.75	2.0	1,200 1,200	900 Ma. 900 Ma.	300 Ma. 300 Ma.	25 25	CK1012
1641/RK60	Full Wave—High Vacuum	5.0	3.0	4,500 2,500	150 Ma. 330 Ma.	50 Ma. 250 Ma.	61	1641 RK60
5517 CK1013	Half Wave—Gas			2,800	50 Ma.	6 Ma.	100	5517 CK1013

RAYTHEON SPECIAL PURPOSE TUBES

TYPE NO.	CONSTRUCTION	SPECIAL APPLICATIONS	FILAMENT Volts	FILAMENT Amps.	FILAMENT Type	Max. Plate Volt.	Max. Plate Cur. Ma.	CAPACITANCES—Mmfds. G-P	Input	Output	TYPE NO.
6AK5	R-F Pentode	U.H.F. Amplifier	6.3	0.175	Heater	180	7	0.01	4.3	2.1	6AK5
6N4	Triode	U.H.F. Osc.-Amp.	6.3	0.2	Heater	180	12	2.3	3.1	0.55	6N4
RK61	Gas Triode	Radio Control	1.4	0.05	Oxide	45	1.5	2.5	2.7	2.8	RK61
2050	Gas Tetrode	Control Thyratron	6.3	0.6	Heater	650	100	0.2			2050

FREE! Get your copy of Raytheon's new "Characteristics Chart" giving all important characteristics of Raytheon's line of over 125 amateur and special purpose tubes from your dealer or by writing us direct.

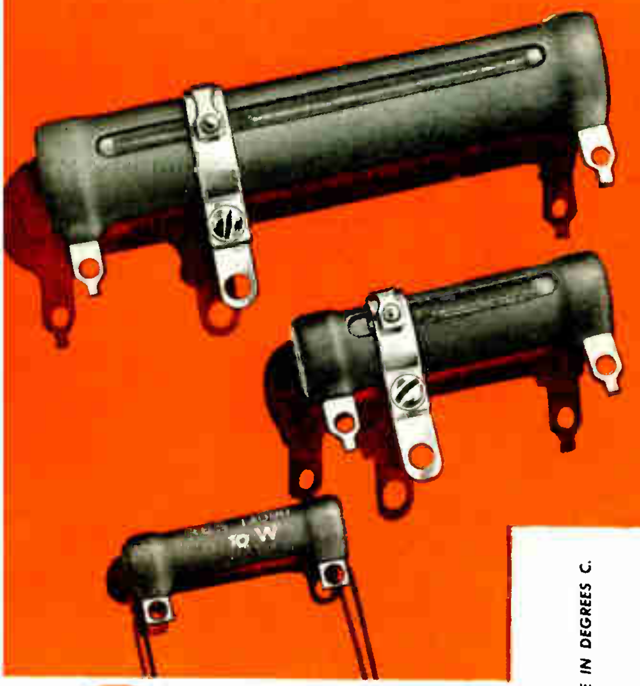
RAYTHEON MFG. CO.

55 CHAPEL STREET

NEWTON 58, MASSACHUSETTS

RAYTHEON

HERE'S HELPFUL INFORMATION ON IRC POWER WIRE WOUNDS!

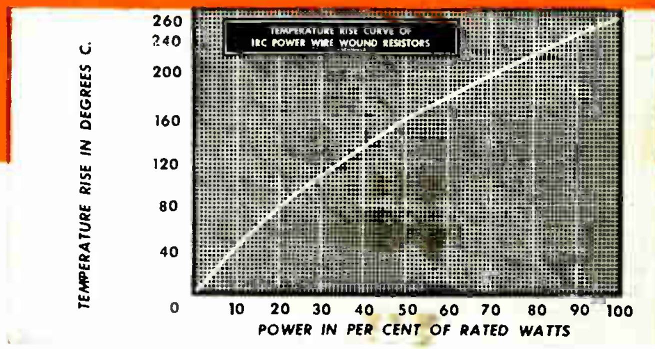


FIXED POWER WIRE WOUND RESISTORS

Type	Watts	Length	Dia.	Resistance Range Ohms
AB	10	1¾"	5/16"	1 to 25,000
DG	20	2"	9/16"	1 to 50,000
EP	50	4½"	¾"	5 to 0.1 meg.
ES	80	6½"	¾"	5 to 0.1 meg.
HA	100	6½"	1⅛"	25 to 0.1 meg.
HO	200	10½"	1⅛"	25 to 0.1 meg.

ADJUSTABLE POWER WIRE WOUND RESISTORS

Type	Length	Dia.	Resistance Range Ohms
ABA	1¾"	5/16"	1 to 10,000
DHA	2½"	9/16"	1 to 25,000
EPA	4½"	¾"	5 to 0.1 meg.
ESA	6½"	¾"	5 to 0.1 meg.
HAA	6½"	1⅛"	100 to 0.1 meg.
HOA	10½"	1⅛"	100 to 0.1 meg.



To guard against the harmful action of atmospheric moisture and corrosion, IRC Fixed and Adjustable Power Wire Wound Resistors have a special cement coating.

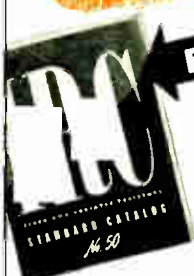
This coating is dark and rough, dissipates heat rapidly, does not deteriorate under any reasonable overload. It guards the winding against the inroads of moisture and corrosive action, contains no chemically active ingredients, no salts, to attack the wire. The cement is crack proof, is cured and hardened at low temperature to prevent the temper from being baked out of winding and terminals.

IRC Fixed and Adjustable Power Wire Wound Resistors are wound on tough, non-porous ceramic forms, have extreme mechanical strength. They are available from 10 to 200 watts.

The many exclusive construction features of IRC Volume Controls, Rheostats, BT & BW Resistors, and Precision Wire Wound Resistors have made them the proven favorite of radio amateurs. IRC manufactures a resistance unit for every ham-rig requirement.

NEW! RESISTO-GUIDE
— AIDS IN RESISTOR IDENTIFICATION!

Here's practical aid in resistor range identification. Just turn the three wheels to correspond with the color code and the standard RMA range is automatically indicated. 10c at all IRC distributors.



FREE CATALOG!

IRC Catalog 50 lists a resistance unit for every ham-rig requirement. It's available at your local IRC Distributor.

PREFERRED FOR PERFORMANCE · IRC · FIXED & VARIABLE RESISTORS

INTERNATIONAL RESISTANCE CO.

401 N. BROAD ST., PHILADELPHIA 8, PA.

In Canada: International Resistance Co., Ltd., Toronto, Licensee

FOR QUICK REFERENCE

HYTRON TRANSMITTING AND SPECIAL PURPOSE TUBES
CONTINUOUS COMMERCIAL SERVICE RATINGS

Description	Type No.	Filament Ratings Volts	Filament Ratings Amps	Type	Max Plate Volts	Max Plate Ma	Max Plate Dis	Amateur Net Price
LOW AND MEDIUM MU TRIODES	3A5	1.4 / 2.8	0.22 / 0.11	Oxide	150	30*	2*	$1.35
	10Y	7.5	1.25	Thor	450	65	15	1.50
	HY24	2	0.13	Oxide	180	20	2	1.50
	801A/801	7.5	1.25	Thor	600	70	20	2.50
	841	7.5	1.25	Thor	450	60	15	2.25
	864	1.1	0.25	Oxide	135	5	—	1.20
	1626	12.6	0.25	Cath	250	25	5	1.60
HIGH-MU TRIODES	HY31Z	6	2.55	Thor	500	150*	30*	3.95
	HY1231Z	6 / 12	3.2 / 1.6	Thor	500	150*	30*	4.50
	5514	7.5	3	Thor	1250	175	65	3.95
VHF TRIODES	2C26A	6.3	1.15	Cath	3500	NOTE	10	7.75
	HY75A	6.3	2.6	Thor	450	90	15	3.95
	HY1148	1.4	0.155	Oxide	180	12	1.8	2.25
	HY615	6.3	0.175	Cath	300	20	3.5	2.25
	955	6.3	0.15	Cath	200	8	1.8	2.85
	9002	6.3	0.15	Cath	200	8	1.8	2.00
BEAM TETRODES AND PENTODES	2E25*	6	0.8	Thor	450	75	15	3.95
	2E30	6	0.65	Oxide	250	60	10	2.25
	3D21A	6.3	1.7	Cath	3500	NOTE	15	7.50
	HY69	6	1.6	Thor	600	100	30	3.95
	807	6.3	0.9	Cath	600	120	25	2.30
	837	12.6	0.7	Cath	500	80	12	4.15
	HY1269	6 / 12	3.2 / 1.6	Thor	750	120	30	4.50
	1625	12.6	0.45	Cath	600	120	25	2.30
	5516	6	0.65	Oxide	600	80	15	5.95
ACORNS MINIA-TURES	6AK5	6.3	0.175	Cath	Sharp cutoff pentode			2.60
	6AL5	6.3	0.3	Cath	Full-wave det., f-m discrim.			.95
	954	6.3	0.15	Cath	Sharp cutoff pentode			4.50
	9001	6.3	0.15	Cath	Sharp cutoff pentode			2.50

	Type No.	Filament Ratings Volts	Filament Ratings Amps	Type Rect	Peak Plate Ma	Max D-C Ma†	Inv Peak Pot.	Amateur Net Price
RECTIFIERS	816	2.5	2.0	Mer	500	250	5000	$1.25
	866A/866	2.5	5.0	Mer	1000	500	10000	1.75
	872A	5.0	7.5	Mer	5000	2500	10000	7.50
	1616	2.5	5.0	Vac	800	260	6000	7.50

	Type No.	Average Operating Voltage	Operating Ma Min	Operating Ma Max	Av Volts Reg	Min Starting Voltage	Amateur Net Price
GASEOUS VOLTAGE REGU-LATORS	OA2	150	5	30	2	185	$2.30
	OB2	108	5	30	1	133	2.30
	OC3/VR105	108	5	40	2	133	1.20
	OD3/VR150	150	5	40	3.5	185	1.20

*Both sections of twin triode. NOTE: Special pulse tube, not recommended for c-w; consult Hytron Commercial Engineering Dept. for data. *2E25 supersedes and replaces HY65. †Current for full wave.

For better reception, it's also Hytron — oldest manufacturer specializing in radio receiving tubes—originator of the popular Bantam GT.

For more complete data, write for your free copy of the Hytron transmitting and special purpose tube catalogue.

HYTRON RADIO AND E EC

Five NEW

To improve upon the HY75 was not easy. But the new HY75A does the trick. Maximum plate current of the HY75A is increased to 90 ma. Grid-to-plate capacitance is sharply reduced to 2.6 $\mu\mu$fd. An HY75A substituted for an HY75 in a 144-mc quarter-wave line oscillator raises the resonant frequency by 20-30 mc. Efficiency is up; 25% more power output at 144 mc. How was this accomplished? By a shorter mount, smaller elements, special high-voltage processing of the lava insulators, redesigned vertical bar grid, and zirconium-coated graphite anode. All at no extra cost to you. Substituted for the HY75, the HY75A requires only pruning of the tank circuit and a higher value of grid resistor. For replacement or new vhf equipment, the new HY75A is your logical choice.

HY75A — IMPROVED VERSION OF HY75 VHF TRIODE

$3.95

The new 5514 supplants the HY30Z, HY40, HY40Z, HY51A, HY51B, and HY51Z. Economies of standardization give you the low price of $3.95. A tube to grow with — the 5514 is efficient at plate potentials from 400-1250 volts. Associated components are economical and still usable as power is increased. At conservative CCS ratings, two 5514's handle 435 watts class C input; deliver 300 watts class B output. One HY69 or 807 can overdrive at maximum input two 5514's in class C. No costly protective fixed bias is needed for this all-purpose, zero-bias 5514. Features: zirconium-coated graphite anode, low loss lava insulators, dual plate connections, ceramic bushing for plate cap, grid leads to pins 2 and 3, convenient 4-pin medium low-loss base and 7.5-volt filament.

5514 — ECONOMICAL VERSATILE ZERO-BIAS TRIODE

$3.95

Designed for frequencies beyond the capabilities of the 2E25, the new 5516 plate-modulated delivers useful power outputs of 21, 16, and 12 watts at 75, 125, and 165 mc respectively. No neutralization is needed in properly designed circuits. All electrode potentials may be applied simultaneously for minimum battery drain in mobile and aircraft use. A dish-pan stem gives short leads with low inductance and capacitance. The zirconium-coated plate and specially treated grids permit higher power outputs. Three separate base-pin connections to the filament center tap provide for lowest possible cathode lead inductance. Excellent r-f screening, high power sensitivity, conservative CCS ratings make the 5516 ideal for powering all stages — r-f and a-f — of your mobile equipment, thus simplifying the spares problem.

5516 — INSTANT-HEATING 165-MC BEAM AMPLIFIER

$5.95

Best illustration of the 2E30's versatility is Ed. Tilton's article beginning on page 31 of QST for June, 1946. Mr. Tilton uses the 2E30 as crystal oscillator, frequency multiplier, speech amplifier, and class AB_2 modulator. Primarily for mobile and aircraft vhf equipment, the 2E30 is an excellent driver for h-f or vhf fixed stations. Designed, manufactured, and tested for transmitting, the 2E30 has a husky, instant-heating filament and generous maximum plate dissipation (10 watts). It develops high efficiency at only 250 volts plate and screen. Imagine doubling to 144 mc with 4 watts output and 0.5 watt drive. The miniature bulb is compact, has low base losses, lead inductance, and capacitance. You can find many uses for the economical 2E30 — a peanut for size, a power-house for output.

2E30 — INSTANT-HEATING VHF MIN. BEAM DRIVER

$2.25

With this HY-Q 75 linear oscillator kit, you can be on 1¼ or 2 meters in an hour. Features are: carefully engineered for easily duplicated results, micrometric tuning (140-250 mc), silver-plated tank, precision-machined shorting bar, special filament, grid, and plate chokes, non-inductive coaxial plate blocking condenser, quick band changing, chart for frequency determination, peak performance for HY75A or HY75 (useful power output with HY75A is 17.5 w on c-w, 13.5 w on phone), easy pictorial instruction manual.

HY-Q 75 — 1¼—2 METER VHF KIT

Unassembled **$9.95**
Assembled **$11.95**
(without tube)

SEE THESE FIVE NEW HYTRON PRODUCTS AT YOUR JOBBER'S.

NICS CORP., SALEM, MASS.

UNIVERSAL MICROPHONES

ready to help you pin the meter!

As America's oldest independent manufacturer of microphones, the Universal Microphone Company has always been at the service of radio amateurs everywhere. You were our first customers "away back when," and we are ready, now and in the future, to help you make your rig pin the meter with a Universal Microphone!

NEW D20 SERIES DYNAMIC MICROPHONES

This postwar model is especially suited for recording, public address, transmitters, or wherever a full-ranged dynamic microphone is desired. It combines modern appearance and rugged stability. The built-in cable connection is easily accessible without interference with the microphone, and "stand and cord" noises are minimized because the internal element is mechanically isolated. The D20 Dynamic Microphone is designed for indoor and outdoor use with a frequency range of 50 to 8000 cycles. Its special "Micro-Adjust Swivel" assures smooth, easy adjustment and steady, positive positioning anywhere throughout a 60° angle. Finished in satin chrome. Complete with 25′ low loss cable and detachable connector. Available in models of 50, 200, 500, and 40,000 Ohms. An exceptional value at only $32.50.

A HOME RECORDING HEAD OF PROFESSIONAL QUALITY!

Universal's design and engineering skill, long experienced in the manufacture of precision studio recorders, has produced this superior home recording head. It outperforms similar recording heads of magnetic design since it purposely accentuates the high frequency range in amount and degree to compensate for high frequency losses common to home recording records and phonograph circuits. This assures a "sparkling" tone quality. Its sensitivity and impedance are keyed to match standard home recorders, thus eliminating special adjustments. Finished in deep brown enamel. Complete with solder terminals, spring tempered phosphor bronze knife edge, steel attachment plate, mounting screws and long styli set screw. Only $11.50.

A132 RECORDING HEAD

"KD" DYNAMIC MICROPHONE

This popular low-cost microphone now is available in a new and improved design. Excellent for home recording, amateur applications. Complete with 10 ft. rubber covered cable. Impedance 40,000 Ohms. Only $17.75.

D61 CONSTANT VELOCITY FREQUENCY RECORD

Here's a handy tool for direct checking of response characteristics of phonograph pick-ups. Also for indirect checking of recording heads, loud speakers, loud speaker installations, theater sound equipment, public address equipment and audio frequency equipment. Complete with data sheet. Only $3.00 each.

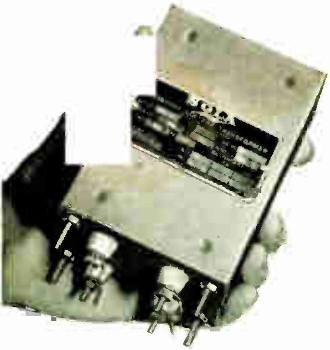
OTHER UNIVERSAL MODELS ARE AVAILABLE, including dynamic, velocity, and carbon in standard, hand, communications and cartridge types. For a complete catalog on these as well as Universal Recording Components, see your radio parts distributor or write direct to us.

YOUR LOCAL RADIO PARTS DISTRIBUTOR CARRIES UNIVERSAL PRODUCTS

UNIVERSAL MICROPHONE CO.

Inglewood, California

Export Division, 301 Clay Street, San Francisco 11, California. Canadian Division, 560 King Street West, Toronto 2, Ontario, Canada

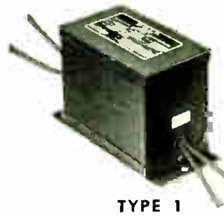


TYPE 11

SMALL, LOW-COST, SOLA CONSTANT VOLTAGE TRANSFORMERS FOR CHASSIS MOUNTING

Reliable communications equipment must have stabilized voltage—and the right place to provide for it is in the equipment itself. These three types of SOLA Constant Voltage Transformers have been specifically designed for "built-in" applications. They are low in cost and their use will often permit the elimination of other components. For complete information consult Bulletin 34CV-102, available on request.

TYPE 12



TYPE 1

	Catalog Number	Output Capacity in VA	Input Volts	Output Volts	Dimensions in Inches A	B	C	E	F	Approx. Shipping Weight	List Price Each
TYPE 1	30488	15	95-125	6.0	5 11/16	2 5/8	3 7/16	5 1/16		6	$15.00
	30492	15	95-125	6.3	5 11/16	2 5/8	3 7/16	5 1/16		6	15.00
	30498	15	95-125	115.0	5 11/16	2 5/8	3 7/16	5 1/16		6	15.00
TYPE 11	30785	17	95-125	6.3	5 [illegible]/16	3 21/32	2 19/32	3	2	5 1/2	20.00
	30955	17	95-125	115.0	5 13/16	3 21/32	2 19/32	3	2	5 1/2	20.00
*TYPE 12	301002	15	95-125	6.3	5 9/16	3 1/2	2 1/4	3	1 1/2	2 1/2	18.50
	301003	15	95-125	115.0	5 9/16	3 1/2	2 1/4	3	1 1/2	2 1/2	18.50

*Condenser supplied as separate unit.

DIMENSIONS:

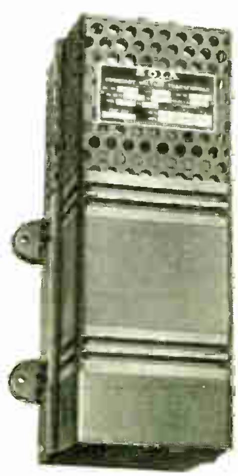
A: Overall Length
B: Overall Width
C: Overall Height
E & F: Mounting Dimensions

Prices subject to change without notice.

TYPE 2

FOR COMMUNICATIONS EQUIPMENT NOW IN SERVICE

Where provision for constant voltage protection has not been made within the equipment itself, these standard SOLA Constant Voltage Transformers can be easily installed. They require no supervision or maintenance, are instantaneous in operation and they protect both themselves and the equipment against short-circuit. Other capacities ranging from 10VA to 15KVA fully described in Bulletin 34CV-102, available on request.



TYPE 3

	Catalog Number	Output Capacity in VA	Input Volts	Output Volts	Dimensions in Inches A	B	C	E	F	Approx. Shipping Weight	List Price Each
TYPE 2	30804	30	95-125	115.0	8 9/16	4 [illegible]/16	4 3/8	7 15/16	2 3/8	12	$17.00
	30805	60	95-125	115.0	8 15/16	4 [illegible]/16	4 3/8	8 1/16	2 3/8	13	24.00
	30806	120	95-125	115.0	9 11/16	4 [illegible]/16	4 3/8	8 15/16	2 3/8	17	32.00
TYPE 3	30807	250	95-125	115.0	11 5/8	6 15/16	5 5/8	3 1/4	6 1/8	30	52.00
	30M807	250	190-250	115.0	11 5/8	6 15/16	5 5/8	3 1/4	6 1/8	30	52.00
	30808	500	95-125	115.0	14 1/2	6 15/16	5 5/8	5	6 1/8	40	75.00
	30M808	500	190-250	115.0	14 1/2	6 15/16	5 5/8	5	6 1/8	40	75.00

SOLA Constant Voltage Transformers

SOLA ELECTRIC COMPANY, 2525 CLYBOURN AVE., CHICAGO 14, ILL.

BURGESS BATTERIES
For Ham Operators

RECOGNIZED BY THEIR STRIPES • REMEMBERED BY THEIR SERVICE

The batteries on this page only illustrate a few of the many popular types of Burgess Batteries for Ham Operators. Your local Burgess distributor has fresh stocks for all your needs.

No. 4FA

No. 5308

No. F4BP

No. 4FA LITTLE SIX—1 1/2 volts—replaces one round No. 6 cell. Radio "A" type; is recommended for the filament lighting of vacuum tubes. Size, 4 1/4" x 2 9/16" x 2 9/16". Weight, 1 lb. 5 oz.

No. 5308—45 volt "B" battery equipped with insulated junior knobs. Taps at — +22 1/2, +45 volts. Size, 5 7/8" x 4 3/16" x 2 9/16". Weight, each—2 lbs. 15 oz.

No. F4BP—A 6 volt, heavy-duty portable battery, designed for Burgess X109 headlight. Contains four F cells connected in series. Screw terminals and brass knurled nuts. Size, 2 21/32" x 2 21/32" x 4 7/32". Weight, 1 lb. 6 oz.

No. 2308—A 45 volt super-service, standard size radio "B". Designed for receivers with plate current drain of 10 to 15 milliamperes. Size, 7 1/8" x 8" x 2 7/8". Weight, 7 lbs. 6 oz.

No. Z30NX 45 volt "B" battery. Improved small size. Adapted to radio, portable receivers and transmitters. Screw terminals. Size, 3" x 1 7/8" x 4 31/32". Weight, 1 lb. 4 oz.

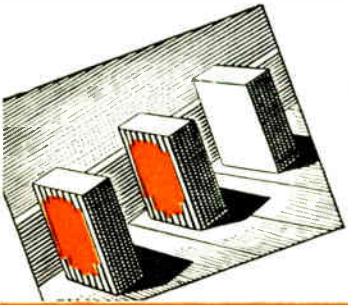
No. 2F2H—A 3 volt radio "A" battery used with portable radios, amplifiers, and special instruments. Size, 2 5/8" x 2 5/8" x 4 3/8". Weight, 1 lb. 6 oz.

No. W30BPX—45 volts. Extremely small and light in weight. Very suitable for personal transceivers used by amateur clubs and radio stations. Equipped with insulated junior knobs. Size, 1 7/32" x 2 21/32" x 4 1/16". Weight, 10 oz.

No. Z30NX

No. 2F2H

No. W30BPX



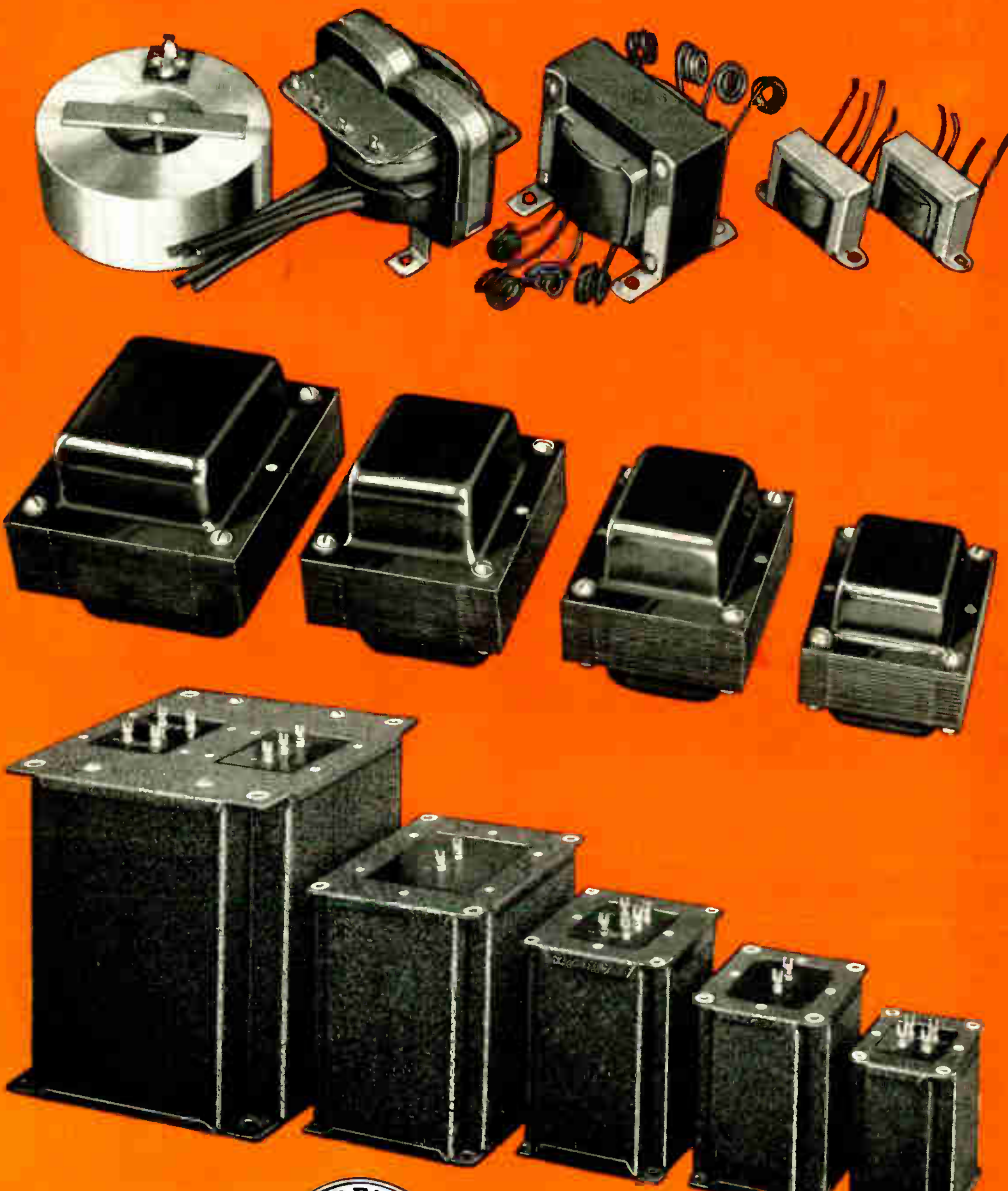
VOTED FIRST CHOICE
IN NATIONAL POLL OF ELECTRONIC ENGINEERS

2 out of 3! That's the way electronic engineers voted for Burgess Batteries in a recent survey. Use the brand the experts choose...BUY BURGESS!

Write For The Name of Your Local Distributor

BURGESS BATTERY COMPANY
DEPT. R46 FREEPORT, ILLINOIS

COMPONENTS Of *Quality* For TELEVISION And ELECTRONIC APPLICATIONS



FREED TRANSFORMERS

FREED TRANSFORMER CO.

FREED TRANSFORMER CO., INC.

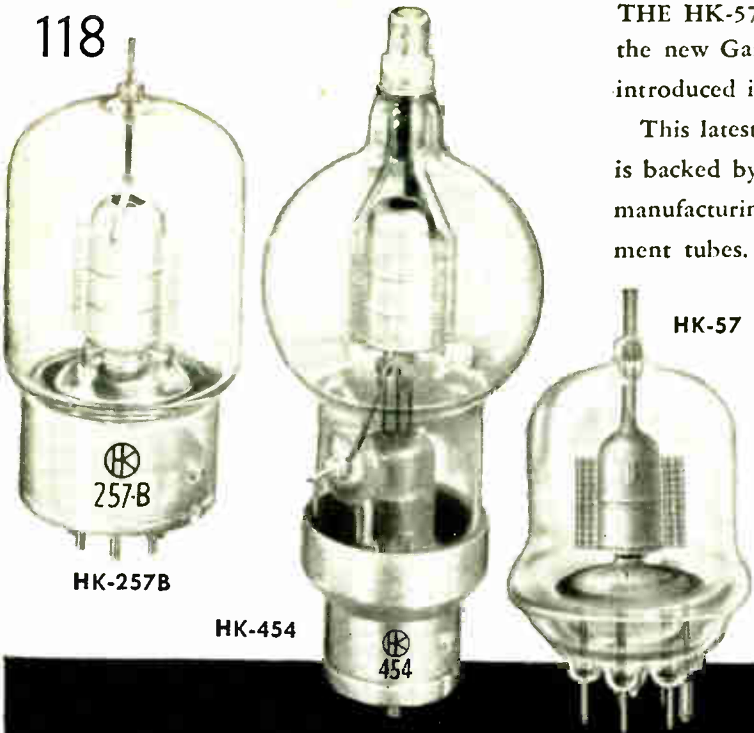
72H SPRING ST., NEW YORK 12, N. Y.

THE HK-57 BEAM PENTODE is the first of the new Gammatron transmitting tubes to be introduced in 1947.

This latest addition to the Gammatron line is backed by twenty years of engineering and manufacturing experience with tantalum element tubes.

Gammatrons are known throughout the world for their electrical and mechanical ruggedness, for their long life, and ability to take it.

Other types, in addition to those listed below, will be announced soon. Watch for them!



HEINTZ AND KAUFMAN LTD.
SOUTH SAN FRANCISCO
CALIFORNIA

Gammatrons

...SEND FOR DATA SHEET ON ANY OF THESE TUBES

TYPE NO.	24	24G	54	57†*	254	257B*	304L	304H	354C	354E	454L	454H	654	854L	854H	1054L	1554	2054A
MAX. POWER OUTPUT: Class 'C' R.F.	90	90	250	250	500	400	1220	1220	615	615	900	900	1400	1800	1820	3000	3600	2000
PLATE DISSIPATION: Watts	25	25	50	75 (1) 50 (2)	100	125 (1) 100 (2)	300	300	150	150	250	250	300	450	450	750	1000	1200
AVERAGE AMPLIFICATION FACTOR	25	25	27	—	25	—	10	19	14	35	14	30	22	14	30	13.5	14.5	10
MAX. RATINGS: Plate Volts Plate M.A. Grid M.A.	2000 75 25	2000 75 25	3000 150 30	3000 150 15	4000 225 40	4000 225 25	3000 1000 150	3000 1000 150	4000 300 60	4000 300 70	5000 375 60	5000 375 85	4000 600 100	6000 600 80	6000 600 110	6000 1000 125	5000 1000 250	3000 800 200
MAX. FREQUENCY, Mc.: Power Amplifier	200	300	200	200	175	200	175	175	50	50	150	150	50	125	125	100	30	20
INTERELECTRODE CAP: C g—p.u.u.f. C g—f u.u.f. C p—f u.u.f.	1.7 2.5 0.4	1.6 1.8 0.2	1.8 2.1 0.5	.049 7.29 in 3.13 out	3.6 3.3 1.0	0.08 10.5 In 4.7 out	9 12 0.8	10.5 14 1.0	3.8 4.5 1.1	3.8 4.5 1.1	3.4 4.6 1.4	3.4 4.6 1.4	5.5 6.2 1.5	5 6 0.5	4 8 0.5	5 8 0.8	11 15.5 1.2	18 15 7
FILAMENT: Volts Amperes	6.3 3	6.3 3	5.0 5	5.0 5.0	5.0 7.5	5.0 7.5	5-10 26-13	5-10 26-13	5 10	5 10	5 11	5 11	7.5 15	7.5 12	7.5 12	7.5 21	11 17.5	10 22
PHYSICAL: Length, Inches Diameter, Inches Weight, Oz. Base	4 1/4 1 3/8 1 1/2 Small UX	4 1/4 1 3/8 1 1/2 Small UX	5 7/16 2 2 1/2 Std. UX	4 1/16 2 3/8 2 1/4 #247 Johnson	7 2 5/8 6 1/2 Std. 50 Watt	6 3/16 2 11/16 5 1/2 Giant 7 Pin	7 3/4 3 1/2 9 Johnson #213	7 3/4 3 1/2 9 Johnson #213	9 3 1/8 6 1/2 Std. 50 Watt	9 3 3/8 6 1/2 Std. 50 Watt	10 3 3/4 7 Std. 50 Watt	10 3 3/4 7 Std. 50 Watt	10 1/8 3 3/4 14 Std. 50 Watt	12 1/2 5 14 Std. 50 Watt	12 1/2 5 14 Std. 50 Watt	16 1/2 7 42 Johnson #214	18 6 56 HK 255	21 1/4 6 66 W.E. Co.

*Beam Pentode.

† Type No. 57 tentative specifications   Note 1: Intermittent telegraph rating   Note 2: Constant key down rating

Now.. A NEW BROWNING HAM FREQUENCY METER!

CHECKS XMTR FREQ. IN ANY HAM BAND FROM 3.5 TO 148 MC. ON FM OR AM

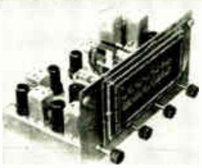
This latest Browning unit designed especially for hams — the Model MJ-9 Frequency Meter — is a high sensitivity job that checks your operating frequencies accurately If you place it near your xmtr, you may not even need a pickup wire for usable signals! Can be used for measuring frequency of remote transmitters and for calibration of receivers within ham bands or ham band harmonics. Truly a necessity for every modern shack

MODEL MJ-9 FREQ. METER

- **Direct, frequency-reading dial on seven ham bands.**
- **.05% accuracy at all frequencies.**
- **Audio detection of zero beat.**
- **Low power consumption.**
- **All operating controls on front panel.**

**PLUS MANY OTHER FEATURES.
WRITE US FOR LITERATURE.**

And.. A NEW BROWNING TUNER! this time for FM-AM



MODEL RJ-12 FM-AM TUNER

- **Separate RF and IF systems on both bands.**
- **One antenna serves both FM and AM.**
- **Tuning eye shows correct tuning.**
- **2-stage cascade limiting on FM.**
- **Phono position on channel selector switch; phono input connector in back.**
- **Armstrong circuit employed on FM.**

**PLUS MANY OTHER FEATURES.
WRITE US FOR LITERATURE.**

FOR HI-FI RECEPTION IN THE NEW FM BAND AND IN STANDARD BC. BAND

Ask any old-timer about the Browning-Drake tuner, and he'll warm right up with a recitation of how much that unit meant to him. Now put an RJ-12 in your house, and the xyl and all the neighbors will decide that you, after all, are the king pin of radio in your community! For here's a hi-fi, hi-sensitivity unit which provides distortion-free reception. Put it in special cabinet, desk drawer, shelf — wherever it will look best. Or for use with rock-mounted amplifier. Diagram of HI-FI AMPLIFIER included with every unit

BROWNING LABORATORIES, INC. WINCHESTER, MASS.

CANADIAN REPRESENTATIVES: MEASUREMENT ENGINEERING 61 DUKE STREET, TORONTO, CANADA

SELECT *Sigma Relays*

FOR SENSITIVITY, STURDINESS AND RELIABILITY

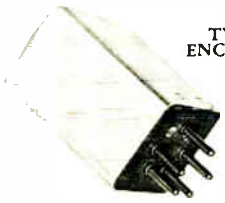
SIGMA Instruments, Inc., fully guarantee the performance of any relay they manufacture. You can buy SIGMA relays in confidence knowing they will serve long and well.

TYPE 4F
1 3/8" x 1 3/4" x 1 5/32"
Weight: 2 3/16 oz.

Recommended for:
- Light weight
- High speed
- Moderately severe environment (temperature and vibration)
- Moderate sensitivity
- Minimum input power 10 milliwatts
- Coil resistance up to 14,000 ohms.

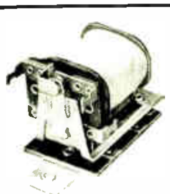
TYPE 5F
1 3/8" x 1 3/8" x 1 5/8"
Weight: 3 1/2 oz.

Recommended for:
- Exceptionally severe environmental conditions
- Maximum sensitivity
- Minimum input power 0.5 milliwatts
- Coil resistance up to 18,000 ohms.



TYPE R ENCLOSURE

— Mounted on a standard 5-pin base and enclosed in an aluminum can.

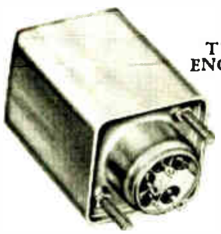


TYPE 41F
1" x 1 7/32" x 2 3/16"
Weight: 2 5/16 oz.

Recommended for:
- Critical performance requirements at a low price in high volume applications. Features of other SIGMA Relays not needed in these applications have been omitted.
- Minimum input power 20 milliwatts
- Coil resistance up to 14,000 ohms.

SMALL COMPACT TYPE 41RO
with Octal Base
Only 1 1/4" x 1 1/4" x 1 13/16"

In addition to their regular mounts, all series 41 Relays are now available in an octal (8 prong) mount for standard radio sockets.



TYPE RJ ENCLOSURE

— Mounted in a hermetically-sealed enclosure of cadmium-plated brass. Also available with octal base, type RJO.

TYPE A ENCLOSURE

— Mounted on a standard 5-pin molded bakelite base with snap-on aluminum enclosure.

ADAPTABLE

All relays are obtainable unenclosed or in a variety of enclosures, including hermetically-sealed. Sample orders filled promptly—components are on hand to be assembled to your requirements.

SPECIALIZED TYPES AVAILABLE

SIGMA can supply extra sensitivity, DC sensitivity on AC current, time delay, and other special features.

SEND SIGMA YOUR RELAY PROBLEMS FOR COMPETENT SOLUTIONS.

SIGMA

Sigma Instruments, INC.

Sensitive **RELAYS**

54 CEYLON ST., BOSTON 21, MASS.

You Can Depend on This Famous Name

Cinaudagraph Speakers

The enviable reputation which Cinaudagraph Speakers have attained for high fidelity, superior performance and ruggedness is deserved through exceptional engineering and production facilities. Cinaudagraph PM Speakers use Alnico 5, the "miracle metal" which gives you four times the performance without size or weight increase. Acceptance of Cinaudagraph Speakers by leading radio manufacturers is assurance of their high fidelity performance.

A complete line of speakers from 2-inch units for portable radios to 15-inch models for commercial phonographs and P. A. systems is available. Write us about your speaker requirements.

Aireon Cinaudagraph Speaker for small radios — remarkable fidelity reproduction within a two-inch cone.

Cinaudagraph Speakers, INC.

SLATER, MISSOURI

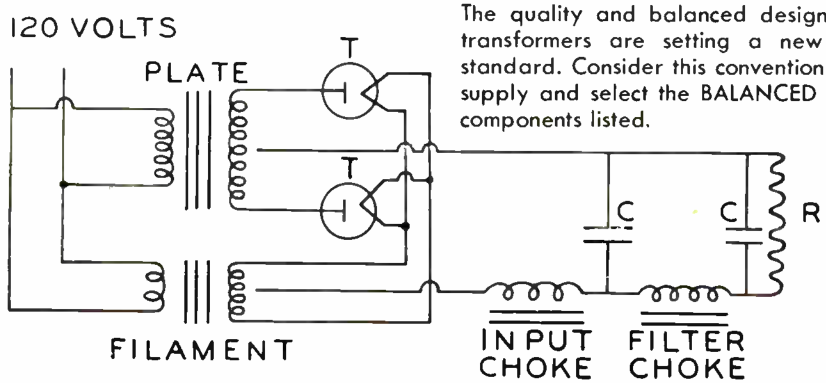
A Subsidiary of Aireon Manufacturing Corporation

CINAUDAGRAPH • SPEAKERS •

SNC TRANSFORMERS

Audio • Bias • Chokes • Universal Driver • Filament
Universal Modulation • Plate • Power • Output

QUALITY *is important . . . so is* BALANCE



The quality and balanced design of SNC transformers are setting a new industry standard. Consider this conventional power supply and select the BALANCED QUALITY components listed.

BALANCED COMPONENTS

DC Volts	*DC Amps*	*Plate*	*Input Choke*	*Filter Choke*
600–750	0.2	7P530	2P148	2P147
600–750	0.3	7P535 7P536*	2P152	2P151
1000–1250	0.3	7P542 7P543*	2P152	2P151
1500–1750	0.3	7P551 7P552*	2P152	2P151
2000–2500	0.3	7P557 7P558*	2P152	2P151
2000–2500	0.5	7P563 7P564*	2P156	2P151

* Compound filled continuous duty types

BALANCED FILAMENT COMPONENTS

4P226 for 866's or other 2.5 V. @ 5A Rectifiers
4P242 for 872's or other 5.0 V. @ 10A Rectifiers

THE POWER SUPPLY IS THE HEART OF THE RIG—COMPLETE IT WITH BALANCED QUALITY SNC TRANSFORMERS FOR GREATER ENJOYMENT!

Write for our interesting catalogue giving valuable information and specifications on all the types of transformers you need.

INbuilt Quality *for* OUTstanding PERFORMANCE

SNC MANUFACTURING COMPANY, INC.

WEST LAKE AVENUE, AT LEHIGH
GLENVIEW, ILLINOIS

INbuilt Quality *for* OUTstanding PERFORMANCE

CAMERADIO

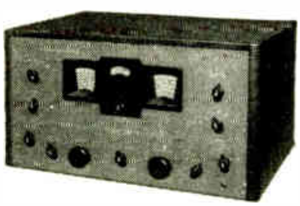
SINCE  1919

ALWAYS THE LATEST — ALWAYS THE FINEST

PIONEERS IN RADIO AND ELECTRONICS

NATIONAL HRO-5TA-1 • Net $274.35
697 Power Unit $20.36 • MCS Speaker $12

A Receiver, designed for the Amateur, to provide dependable communications, under most severe conditions. New Noise Limiter incorporated. Band spread on all "Ham" Bands—Coverage 1.7 to 30 MC.



HAMMARLUND HQ-129-X • Net $173.25
Price Includes Cabinet Speaker

A "Ham" Receiver with the professional touch, providing flexibility of operation. Six bands for easy location of stations. Has all "Ham" features with sensitive and stable performance.

HALLICRAFTER SX-42 • Net $250.00
R-42 TABLE SPEAKER • Net $25.00

The SX-42 has many new and ingenious circuit features. Precision tuning for all AM and FM Broadcast and Amateur Bands. Temperature compensated oscillator. High fidelity phono-amplifier—8 watts. Continuous coverage 540 KC to 110 MC.

CARDWELL MODEL CR-54 RECEIVER
RACK MODEL AVAILABLE LATER

This completely new communications receiver uses 18 miniature tubes. Large direct reading precision dials—electrical band spread. New type noise limiter. Variable selectivity crystal filter. Temperature compensated oscillator. Wide coverage—540 KC to 54 MC, with extra coil strip.

QUALITY MERCHANDISE • PROMPT SERVICE

CAMERADIO CO., 963 LIBERTY AVE., PITTSBURGH 22, PA.

Capacitors FOR THE AMATEUR

MOLDED OIL-PAPER CAPACITORS

TOBE Molded oil paper capacitors, hermetically sealed against moisture, provide dependable service under the conditions which cause repeated paper tubular failure. For r.f. use up to 40 m.c., and for a.f. by-pass, and filter circuits.

Type	*Voltage Range*	*Capacitance Range*	*Length*	*Width*	*Thickness*
AP	120- 400 VDC	1000- 10000 Mmfd	11/16"	29/64"	7/32"
DP	120- 600 "	1000- 10000 "	13/16"	13/16"	19/64"
MP	120- 800 "	1000-100000 "	1 11/64"	41/64"	17/64"
FP	120-1000 "	1000-100000 "	1 15/32"	49/64"	21/64"
EP	120-1000 "	5000-200000 "	1 15/32"	49/64"	17/64"

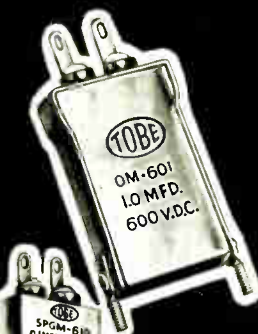
OIL-IMPREGNATED FILTER CAPACITORS

TOBE "Oil-Mites" for filter and by-pass applications involving high voltages and extremes of temperature and humidity. Mineral oil impregnant achieves more compact size without impairing performance. Power factor and capacitance are stable through temperatures from —55 to 185 F. Either upright or inverted mounting is provided by the reversible hold-down bracket.

***Voltage range:* 100 to 1000 VDC. *Capacitance range:* .01 to 4.0 Mfd.**

"Midget Oil-Mites," to meet small space requirements, incorporate design and structural features which permit these capacitors to be contained in smaller cases than the conventional "Oil-Mites." Seamless metal cases are hermetically sealed, phenolic-insulated solder-lug terminals are leakproof.

***Voltage range:* 100 to 1000 VDC. *Capacitance range:* Single sections, .01 to 1.0 Mfd. Multiple sections, 2 x .05, 2 x .10 Mfd.**



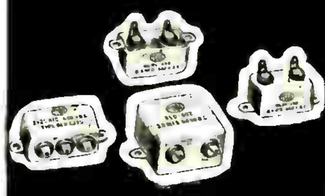
OIL-IMPREGNATED BY-PASS CAPACITORS

TOBE Compact "bathtub" design combined with the operating advantages—stability of capacitance and low power factor—of mineral oil-impregnated and oil-filled constructions.

Capacitance ranges:

Single sections: 600 VDC—.01 to 2.0 Mfd.
1000 VDC—.01 to 1.0 Mfd.

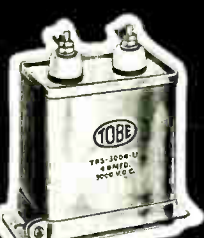
Multiple Sections	600 VDC	2 x .05	2 x .1	3 x .25	3 x .05	3 x .1	2 x .25	3 x .5
	1000 VDC	2 x .05	2 x .1	3 x .25		3 x .1	2 x .25	



TRANSMITTING CAPACITORS

TOBE Type TRS for use in filter, transmitting, and timing circuits are designed to handle high transient voltages and temperatures encountered in such service. Impregnated and filled with mineral oil, these units are contained in hermetically sealed steel cases. For convenient mounting, permanently attached mounting feet, hook hold-down brackets, or adjustable wrap around brackets for upright or inverted mounting are available. The adjustable wrap-around bracket permits mounting with terminals at any desired distance above the mounting surface.

***Capacitance and voltage ranges:* 600 to 6000 VDC and .03 to 50 Mfd.**



TELEVISION CAPACITORS

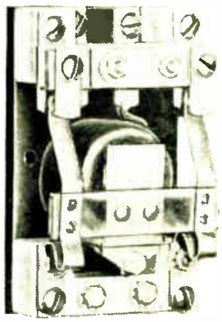
TOBE Type VRC capacitors, for employment in television circuits are built to withstand high voltages at high temperatures—a chief consideration in such applications. Typical ratings are:

Type VRC-8011—8000 VDC, 0.1 Mfd. 2 insulated terminals
Type VRC-16005—16000 VDC, 0.05 Mfd. { 1 terminal insulated, 1 terminal grounded.

If your jobber cannot supply these Tobe long-life capacitors, write directly to—

TOBE DEUTSCHMANN *Corporation*
CANTON, MASSACHUSETTS

3 Basic Ways TO IMPROVE THE EFFICIENCY AND STABILITY OF YOUR RIG

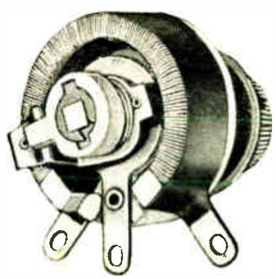


1 WARD LEONARD RELAYS

Proper relays have many practical uses in amateur rigs. They provide convenient circuit control and greater operating efficiency. They help reduce the length of connecting leads, prevent unnecessary transfer of stray or transient currents to critical parts of the transmitter circuit. Designed and built for rugged service, WARD LEONARD Amateur Relays are available from stock in the following types: Antenna Change-Over — Antenna Grounding — Band Switching — RF Break-In — Keying — Overload — Underload — Latch-In — Remote Control — Safety — Sensitive — Time Delay.

2 WARD LEONARD RESISTORS

The installation of VITROHM wire-wound resistors insures that *extra* performance you want in critical circuits. With exclusive features developed in the WARD LEONARD laboratories . . . these resistors meet the most rigid specifications. Fixed resistors are available in 8 stock sizes from 5 watts to 200 watts. Adjustable resistors, in 7 stock sizes from 10 watts to 200 watts. Wide range of resistance values. Stripohm, Discohm, and Plaque resistors are also available.



3 WARD LEONARD RHEOSTATS

These Vitrohm Ring-Type units are widely used as potentiometers or rheostats, for fixed or variable close control, in commercial as well as amateur equipment. Made of highest grade ceramic materials. Resistance wire is held in place and protected by special WARD LEONARD vitreous enamel. Many other features. Available from stock in 25 watt, 50 watt, 100 watt and 150 watt sizes, in wide range of resistance values.

SEND FOR FREE CATALOGS

For helpful data and information, write for Relay Catalog No. D-111 and for Resistor-Rheostat Catalog No. D-2.

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RELAYS • RESISTORS • RHEOSTATS

Electric control devices since 1892

WARD LEONARD ELECTRIC CO., *Radio and Electronic Distributor Division,* **53K West Jackson Blvd., Chicago 4, U. S. A.**

LABORATORY INSTRUMENTS FOR

INSTRUMENT	FUNCTION	FREQUENCY	CALIBRATION
200-A	Resistance-Tuned Audio Oscillator	35 cps to 35 kc	Dial Scale 35-350 cps Calibration Points 48 Ranges 3 (1. 10. 100 times dial calibration)
200-B	Resistance-Tuned Audio Oscillator	20 cps to 20 kc	Dial Scale 20-200 cps Calibration points 41 Ranges 3 (1. 10. 100 times dial calibration)
200-C	Resistance-Tuned Oscillator	20 cps to 200 kc	Dial Scale 20-200 cps Calibration Points 41 Ranges—4 (1. 10. 100. 1000 times dial calibration)
200-D	Resistance-Tuned Oscillator	7 cps to 70 kc	Dial Scale 7-70 cps Calibration Points 78 Ranges 4 (1. 10. 100. 1000 times dial calibration)
202-D	Resistance-Tuned Oscillator	2 cps to 70 kc	Dial Scale 7-70 cps Calibration Points 78 Ranges 4 (1. 10. 100. 1000 times dial calibration) Dial Scale A 2.50 cps Calibration Points 81
200-I	Resistance-Tuned Spread Scale Audio Oscillator	6 cps to 6 kc	Dial Scale A 6-20 cps Calibration Points 168 Ranges 3 (1. 10. 100 times dial calibration) Dial Scale B 20-60 cps Calibration Points 109 Ranges 3 (1. 10. 100 times dial calibration)
201-B	Resistance-Tuned Spread-Scale Audio Oscillator	20 cps to 20 kc	Dial Scale 20-200 cps Calibration Points 95 Ranges 3 (1. 10. 100 times dial calibration)
205-A	Resistance-Tuned Audio Signal Generator	20 cps to 20 kc	Dial Scale 20-200 cps Calibration Points 80 Ranges 3 (1. 10. 100 times dial calibration)
205-AG	Resistance-Tuned Audio Signal Generator with 49 db Voltmeter	20 cps to 20 kc	Dial Scale 20-200 cps Calibration Points 80 Ranges—3 (1. 10. 100 times dial calibration)
205-AH	Resistance-Tuned Signal Generator	1 kc to 100 kc	Dial Scale 1-10 kc Calibration Points 130 Ranges 2 (1. 10 times dial calibration)

INSTRUMENT	FUNCTION	FREQUENCY	ACCURACY
100-A	Low Frequency Standard	Output 100 kc. 10 kc. 1 kc. 100 cps	±0.01% over room temperature variation of 33 C
100-B	Low Frequency Standard	Output 100 kc. 10 kc. 1 kc. 100 cps	±0.001% from −10° C to +50° C
210-A	Square Wave Generator	Input 20 cps to 100 kc	Square within ±1% from 20 cps to 10 kc
300-A	Harmonic Wave Analyzer	Measurement Range— 30 cps to 16 kc	Frequency ±3% Voltage overall ±5%
320-A	Distortion Analyzer	Measures at 400 cps and 5 kc	Less than ±5% [at distortions of 30% or less]
320-B	Distortion Analyzer	Measures at — 50 cps 100 cps, 100 cps. 1 kc, 5 kc and 7.5 kc	Less than ±5% [at distortions of 30% or less]
325-B	Noise and Distortion Analyzer	Measures at 30 cps, 50 cps. 100 cps. 400 cps; 1 kc, 5 kc. 7.5 kc. 15 kc	Voltmeter Overall ±3% Distortion Less than ±5% [at distortion of 30% or less]
330-B	Distortion Analyzer	Measurement Range— 20 cps to 20 kc	Voltmeter overall ±3 Distortion ±5 for distortion levels as low as 0.5
350-A	Attenuator	Max. input 100 kc	Each Resistor ±0.5% Response Accumulative Error at 100 kc approx. 1 db in 100 db.
400-A	Vacuum Tube Voltmeter	Measurement Range— 10 cps to 1 mc	10 cps to 100 kc ±3% 100 kc to 1 mc ±5%
410-A	High Frequency Vacuum Tube Voltmeter, DC Voltmeter, Ohmmeter	Measurement Range— 20 cps to 700 mc	±3% AC and DC Frequency Response flat within 1 decibel 20 cps to 700 mc
500-A	Electronic Frequency Meter	Measurement Range— 5 cps to 50 kc in 10 ranges	±2% of full scale
505-A	Electronic Tachometer	An Electronic Frequency Meter and a Tachometer Assembly calibrated to measure speeds up to 3.000.000 RPM.	
710-A	Regulated Power Supply		

Resistance-Tuned Audio Oscillator
-hp- 200C

Vacuum Tube Voltmeter
-hp- 400A

Vacuum Tube Voltmeter
-hp- 410A

Fast, easy measurements of pin-point accuracy

Swift, sure measurements—so necessary in amateur radio, commercial broadcasting, laboratory research, or industrial electronics—are made possible and practical by -hp- precision instruments.

Condensed specifications of many -hp- instruments appear on these pages. -hp-'s new catalog—yours for the asking—gives more information on electronic measurements and the instruments which make them. Send for your copy now. No obligation.

Distortion Analyzer
-hp- 330B

SPEED AND ACCURACY

FREQUENCY RESPONSE	STABILITY	ACCURACY OF CALIBRATION	POWER OUTPUT INTO RATED LOAD	LOAD IMPEDANCE	DISTORTION AT RATED OUTPUT	HUM LEVEL BELOW RATED OUTPUT	POWER REQUIREMENTS	SIZE
±1 decibel, 20 cps to 15 kc	±2%	±2%	1 watt	500 ohms	less than 1%	60 db	115 volts 50-60 cyc 60 watts	Length 16 ins. Height 8 ins. Depth 9 ins. Weight 32 lbs.
±1 decibel, 20 cps to 15 kc	±2%	±2%	1 watt	500 ohms	less than 1%	60 db	115 volts 50-60 cyc 60 watts	Length 16 ins. Height 8 ins. Depth 9 ins. Weight 32 lbs.
±1 decibel, 20 cps to 150 kc	±2%	±2%	100 milliwatts	1000 ohms	less than 1% 20 cps to 20 kc	60 db	115 volts 50-60 cyc 60 watts	Length 16 ins. Height 8 ins. Depth 9 ins. Weight 30 lbs.
±1 decibel, 7 cps to 70 kc	±2%	±2%	100 milliwatts	1000 ohms	less than 1% 10 cps to 20 kc	60 db	115 volts 50-60 cyc 60 watts	Length 17 ins. Height 8¾ ins. Depth 11 ins. Weight 32 lbs.
±1 decibel, 7 cps to 70 kc ±2 decibels, 2 cps to 7 cps	±2%	±2%	100 milliwatts	1000 ohms	less than 2% 7 cps to 70 kc	60 db	115 volts 50-60 cyc 60 watts	Length 17 ins. Height 8¾ ins. Depth 11 ins. Weight 32 lbs.
±1 decibel, 6 cps to 6 kc	±2% or ±1% with Standardization	±2%	100 milliwatts	1000 ohms	less than 1% 10 cps to 6 kc	60 db	115 volts 50-60 cyc 60 watts	Length 17 ins. Height 8¾ ins. Depth 11 ins. Weight 28 lbs.
±1 decibel, 20 cps to 20 kc	±2% or 1% with Standardization	±2%	3 watts	600 ohms	Less than 1% at 3 watts (Less than ½% at 1 watt	60 db	115 volts 50-60 cyc 75 watts	Length 17 ins. Height 8½ ins. Depth 11 ins. Weight 32 lbs.
Down 2.0 decibels at 20 cps Down 1.0 decible at 20 kc at full output	±2% or ±1% with Standardization	±2%	5 watts	50, 200, 500, 5000 ohms (all ct)	less than 1% 30 cps to 20 kc at rated output	60 db below output or 90 db below zero level whichever is larger	115 volts 50-60 cyc 125 watts	Length 21⅝ ins. Height 11⅞ ins. Depth 14⅝ ins. Weight 70 lbs.
enerator down 2.0 db at 20 cps wn 1.0 db at 20 kc at full output. meter within ±0.2 db of 400 cps ref. from 20 cps to 20 kc	±2% or ±1% with Standardization	±2%	5 watts	Generator 50, 200, 500, 5000 ohms (all ct) Voltmeter 5000 ohms input impedance	less than 1% 30 cps to 20 kc at rated output	60 db below output or 90 db below zero level whichever is larger	115 volts 50-60 cyc 125 watts	Length 21⅝ ins. Height 11⅞ ins. Depth 14⅝ ins. Weight 73 lbs.
db from 10 kc ref. 1 kc to 100 kc at full output	±1% after ½ hour warm-up	±2%	5 watts	50, 200, 500, 5000 ohms (all ct)	less than 1% at 1 watt 3% at 5 watts	65 db below output or 65 db below zero level whichever is larger	115 volts 50-60 cyc 125 watts	Length 21⅝ ins. Height 11⅞ ins. Depth 14⅝ ins. Weight 63 lbs.

VOLTAGE	IMPEDANCE	MISCELLANEOUS CHARACTERISTICS	POWER REQUIREMENTS	SIZE
Output 5 volts into 1000 ohms	Load Not less than 1000 ohms	Wave Shape Sinusoidal total distortion not more than 4% on open circuit	115 volts 50-60 cyc 100 watts	Length 21⅝ ins. Height 11⅞ ins. Depth 14 ins. Weight 53 lbs.
Output 5 volts into 1000 ohms	Load Not less than 1000 ohms	Wave Shape Sinusoidal total distortion not more than 4% on open circuit	115 volts 50-60 cyc 105 watts	Length 21⅝ ins. Height 11⅞ ins. Depth 14 ins. Weight 53 lbs.
Input min. 2; max. 200 utput 60 v peak to peak on open circuit	Input 25,000 ohms Internal Each side, 500 ohms to ground	Wave Shape Square 1 microsecond to 90% of maximum) Attenuator 70 db in 5 db steps	115 volts 50-60 cyc 85 watts	Length 16 ins. Height 8 ins. Depth 9 ins. Weight 30 lbs.
Input 1 mv to 500 v	Input 200,000 ohms	Variable Selectivity at 40 db down from resonance max. selectivity is 30 cps; min. selectivity is 145 cps Dial Calibration Points 62	115 volts 50-60 cyc 105 watts	Length 21⅝ ins. Height 24 ins. Depth 14⅛ ins. Weight 78 lbs.
Max. Input 100 v	Analyzer Input 20,000 ohms Detector Input Should be not less than 100,000 ohms	Max. Attenuation: Fundamental more than 60 db (1%) Second and higher harmonics less than 5% Filters Tuned to nominal frequencies within ±5% (non-adjustable) Attenuator 70 db in 1 db steps		Length 13 ins. Height 9 ins. Depth 8 ins. Weight 15 lbs.
Max. Input 100 v	Analyzer Input 20,000 ohms Detector Input Should be not less than 100,000 ohms	Max. Attenuation: Fundamental more than 60 db 1% Second and higher harmonics less than 5% Filters Tuned to nominal frequencies within ±5% (non-adjustable) Attenuator 70 db in 1 db steps		Length 13 ins. Height 9 ins. Depth 8 ins. Weight 17½ lbs.
Voltmeter Measurement Range .01 v to 300 v in 9 ranges tortion min. input 1 v for .1% distortion oise min. input .003 volts for full scale	Amplifier Input 200,000 ohms shunted by approx. 24 mmfd Voltmeter Input 1 megohm (min.) shunted by approx. 32 mmfd	Max. Attenuation: Fundamental more than 60 db 1% Second and higher harmonics less than 5% Filters Tuned to nominal frequencies within ±5% (adjustable 1%) Voltmeter Average Reading calibrated in rms volts and in db above a 1 mw. 600 ohm level	115 volts 50-60 cyc 65 watts	Length 21⅝ ins. Height 11⅞ ins. Depth 14 ins. Weight 56 lbs.
Voltmeter measurements .01 v to 300 v in 9 ranges tortion min. input 1 v for .1% distortion e minimum input 0.0003 v for full scale	Amplifier input 200,000 ohms shunted by approx. 24 mmfd Voltmeter Input 1 megohm (min.) shunted by approx. 32 mmfd	Max. Attenuation: Fundamental more than 60 db 0.1% Second and higher harmonics less than 10% Voltmeter Average reading calibrated in rms volts and db above a 1 mw. 600 ohm level)	115 volts 50-60 cyc 90 watts	Length 19 ins. Height 10½ ins. Depth 13 ins. Weight 50 lbs.
Maximum Input 50 v	Input 500 ohms one side grounded Output 500 ohms one side grounded	Attenuation 110 db in 1 db steps		Length 8 ins. Height 5 ins. Depth 4½ ins. Weight 4 lbs.
Measurement Range - .03 v to 300 v in 9 ranges	Input 1 megohm (min.) shunted by approx. 16 mmfd	Voltmeter—Average Reading calibrated in rms volts and in db above a 1 mw. 600 ohm level	115 volts 50-60 cyc 40 watts	Length 7½ ins. Height 9½ ins. Depth 10¼ ins. Weight 15 lbs.
Measurement Range 1 to 300 VAC in 6 ranges 1 to 1000 VDC in 7 ranges	Input AC 8 megohms in parallel with 1.5 mmf at frequencies below 10 mc Input DC 100 Megohms	AC Voltmeter Peak reading instrument will indicate voltage to 3000 mc Ohmmeter, 0.2 ohms to 500 megohms in 7 ranges	115 volts 50-60 cyc 40 watts Two 1.5 v flashlight cells supply ohmmeter	Length 7¼ ins. Height 12¼ ins. Depth 6¼ ins. Weight 16 lbs.
Input 0.5 v to 200 v	Input 50,000 ohms	Separate External Attachments 1. Photocell Input (jack provided) 2. Esterline-Angus 1 mil, 1400 ohm Automatic Recorder (jack provided)	115 volts 50-60 cyc 65 watts	Length 17½ ins. Height 8¼ ins. Depth 11¾ ins. Weight 28 lbs.
180 to 360 VDC (regulated) 6.3 VAC ct (unregulated)		Output constant within approx. 1% for loads from 0 to 75 ma and for line-voltage variations of ±10%. Noise and hum less than 0.005 v	115 volts 50-60 cyc 90 watts full load	Length 7½ ins. Height 8 ins. Depth 11¾ ins. Weight 18 lbs.

Audio Signal Generator

hp 205AG

Regulated Power Supply

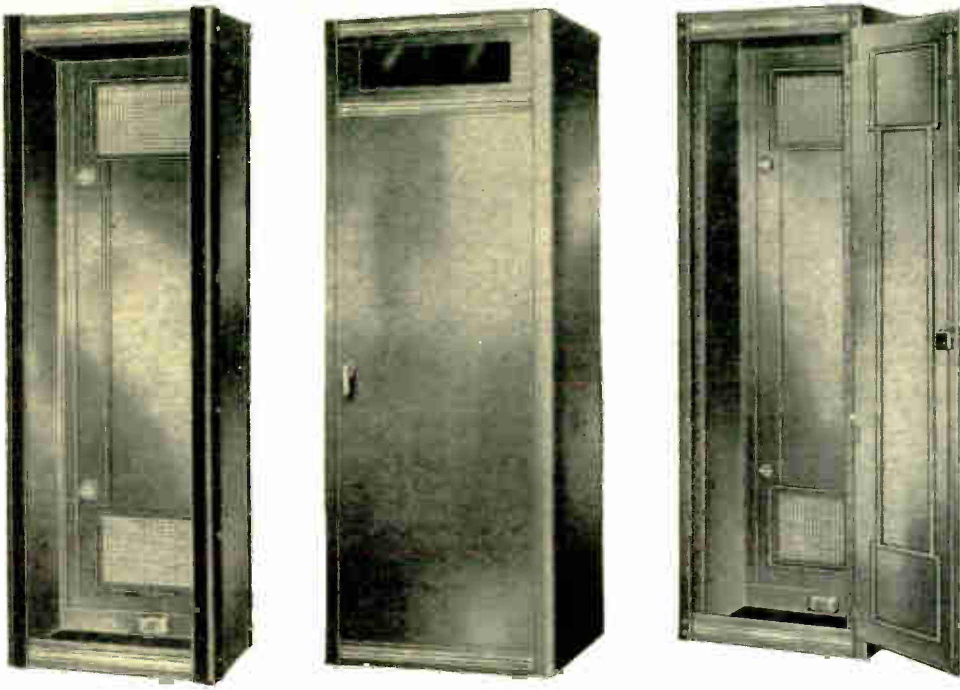
hp 710A

HEWLETT-PACKARD COMPANY

1325 PAGE MILL ROAD
PALO ALTO, CALIFORNIA

PAR-METAL *Standard* RACKS, CABINETS CHASSIS, PANELS

ADAPTABLE FOR EVERY REQUIREMENT



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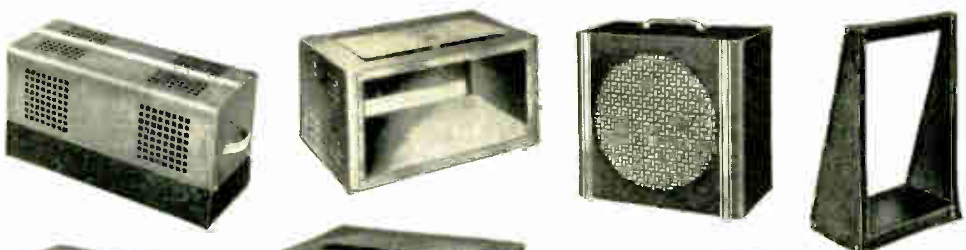

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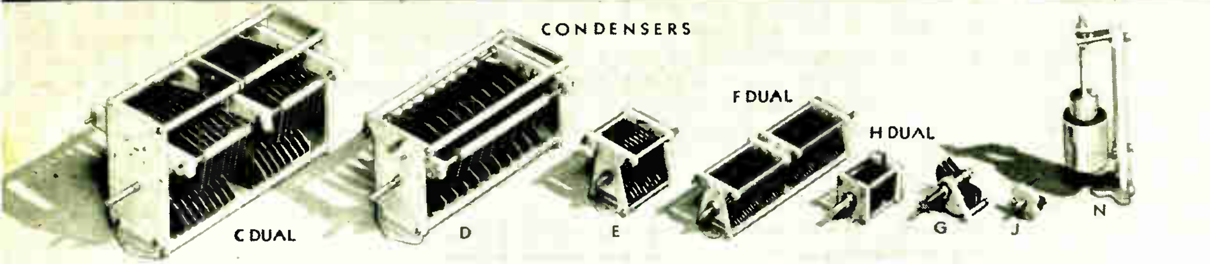
CONDENSERS

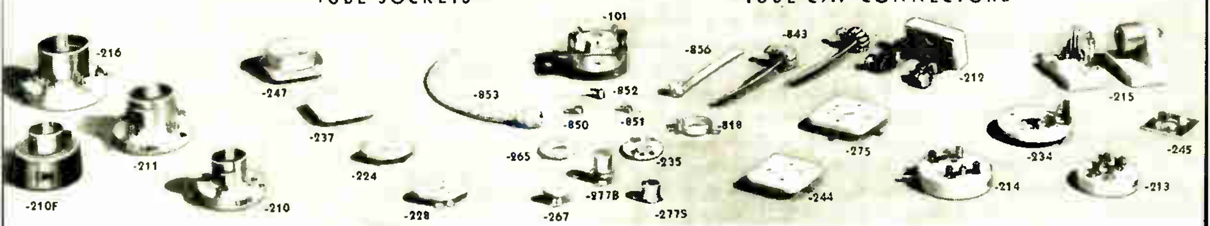


Cat. No.*	Length	Min. Cap.	Cat. No.*	Length	Min. Cap.	Cat. No.*	Length	Min. Cap.	Cat. No.*	Length	Min. Cap.
250E20	2 25/32	12	70ED30	4 15/32	8	70FD20	3 27/32	7	70HD15	2 1/2	8
350E20	3 17/32	15	100ED30	5 5/16	10	100FD20	4 11/16	9	100HD15	3 3/32	10
500E20	4 15/32	19	150ED30	7 1/16	13	150FD20	6	11	35HD30	4 13/32	12
35E30	1 11/16	8	200ED30	8 3/8	15	200FD20	7 7/32	14	50HD30	6	15
50E30	1 29/32	9	50ED45	6 5/32	10	50FD30	4 27/32	8	7J12	9/16	2.6
70E30	2 5/32	9	70ED45	7 7/16	12	70FD30	5 23/32	10	15J12	11/16	3.3
100E30	2 9/16	11	100ED45	9 9/32	15	100FD30	7 7/16	13	25J12	1/8	3.6
150E30	3 7/16	14	35F20	1 1/2	7	25H15	11/16	4	50J12	1 7/32	4.9
250E30	4 15/16	20	50F20	1 5/8	8	35H15	13/16	4	75J12	1 1/2	6
350E30	6 7/16	25	70F20	1 25/32	8	50H15	15/16	4	100J12	1 15/16	7
35E45	2 5/16	9	100F20	2 1/4	10	70H15	1 5/32	6	25G20	29/32	4
50E45	2 31/32	11	150F20	2 7/8	12	100H15	1 7/16	7	50G20	1 5/32	5
70E45	3 9/16	13	250F20	4 3/32	17	150H15	2 13/32	9	8G45	7/8	3.6
100E45	4 17/32	16	35F30	1 7/8	8	250H15	3 13/32	13	13G45	15/32	4.7
150E45	6 3/32	20	50F30	2 5/16	9	25H30	2 3/16	7	23G45	1 13/16	6.4
250E45	9 9/16	32	70F30	2 3/4	11	35H30	2 19/32	8	6G70	1 1/16	3.5
200ED20	5 1/8	10	100F30	3 19/32	14	50H30	3 12/32	11	12G70	2 5/8	6
300ED20	6 21/32	13	150F30	4 7/8	18	70H30	4 13/32	13	N125	Neutralizing Max. 11	—Min. 1.1
50ED30	4 3/32	8	50FD20	3 1/2	7	35HD15	1 11/16	6	N250	Neutralizing Max. 10.6	—Min. 1.4
						50HD15	2 3/32	7	N375	Neutralizing Max. 10.7	Min. 1.7

For "C" and "G" listings see complete catalog

* Numerals preceding letter indicates maximum capacity. Numerals following letter plus two ciphers indicates approximate peak voltage. Second letter "D" indicates two section condenser.

TUBE SOCKETS   TUBE CAP CONNECTORS



Socket	Base or Typical Tube	Socket	Base or Typical Tube
123 209	Med. 4	122 217	Small 7 pin
123 210	Pin Bayonet	122 224	4 pin
123 211	Standard Jumbo 4 pin	122 225	5 pin
123 216	Giant 5 pin Bayonet	122 226	6 pin
124 212	833A	122 227	7 pin med.
124 213	152TL	122 228	Octal
124 214	1500TH	122 234	RK72
124 215	204A	122 237	Giant 7 pin
120 267	9000 series	122 244	Super Jumbo
120 277B	Miniature	122 247	826
121 235		122 248	826
121 215	Acorn	122 275	Giant 4 pin
121 265		124 220	899R
122 101	829		

Cat. No.	Tube Cap Diameter	Type	Cat. No.	Tube Cap Diameter	Type
TUBE CAP CONNECTORS					
119 843	.500	Radiator	119 855	.566	Safety
119 850	.250	Receiver	119 856	.360	With 6" strap
119 851	.360	Receiver	119 857	.566	With 6" strap
119 852	.360		133 817	Clamp for 1.165" tube	
119 853	.360	Safety	133 818	Clamp for 1.275" tube	
119 854	.566		133 820	Clamp for 1.377" tube	

COMING SOON—WORTH WAITING FOR!
★ *The New* JOHNSON CABINET LINE
★ *The New* JOHNSON ROTARY BEAM ANTENNA

COUPLINGS • CHOKES • INDUCTORS • PILOT LIGHTS • CONNECTORS

Cat. No.	Description		
COUPLINGS			
	V	A	B
104 250	1000	1/4	1/4
104 2503	1000	1/4	3/8
104 251	5000	3/8	3/8
104 251A	5000	1/4	1/4
104 251B	5000	1/4	3/8
104 252	1000	1/4	1/4
104 258		1/4	1/4
104 259	8000	1/4	1/4
104 2593	5000	1/4	1/4
104 260	2500	1/4	1/4
104 261	7500	3/8	3/8
104 262	5000	1/4	1/4
104 263	2000	1/4	1/4

Dimension symbols: "V" modulated Peak Voltage, "A" hub i. d., "B" 2nd hub i. d. All insulation steatite.

Cat. No.	Band (Meters)	Cap. to tune (mmf.)
INDUCTORS — Tube Socket "Hi-Q"		
230 640	10	24
230 641	20	33
230 642	40	37
230 643	80	71
230 644	160	130
230 645	14	27
230 650	10	36
230 651	20	58
230 652	40	70
230 653	80	75
230 654	160	110
230 655	14	40
235 646	Form only, 4-prong	
235 647	Form only, 5 prong	

Cat. No.	Description
"Q" ANTENNAS	
137 2Q	2 Meter Band
137 6Q	6 Meter Band
137 10Q	10 Meter Band
137 20Q	20 Meter Band
137 40Q	40 Meter Band

Cat. No.	Frequency (mc.)	Rating (ma.)	Length
R.F. CHOKES			
102 750	1.7 30	150	1 1/2
102 752	1.7 30	500	2 7/8
102 754	1.7 30	750	1 5/16
101 760	UH	250	1 1/2
101 762	UH	1500	2 7/8

Cat. No.	No. of Contacts	Connector Type
MULTI-WIRE CONNECTOR RECEPTACLES		
111 614	12	Chassis
111 615	12	Cord
111 644	7	Chassis
111 645	7	Cord
PLUGS		
111 617	12	Chassis
111 625	12	Cord
111 631	7	Chassis
111 635	7	Cord
111 680	7 contact pin plate bkt. mtd.	
111 682	12 contact pin plate bkt. mtd.	
111 6002	Mtg. yoke for 7 wire conctr.	
111 6003	Mtg. yoke for 12 wire conctr.	
141 7	7 wire cable	
141 12	12 wire cable	
147 3101	Screw base pilot light, Choice of jewel colors	
147 3081	Bay. base pilot light, Choice of jewel colors	
147 330	Screw base panel light	
147 329	Bay. base panel light	

SEE ILLUSTRATION AT RIGHT

STANDARDS

FOR COMPLETE CATALOG SEE YOUR DISTRIBUTOR OR WRITE DEPT. P.
E. F. J[illegible]N CO., Waseca, Minn.

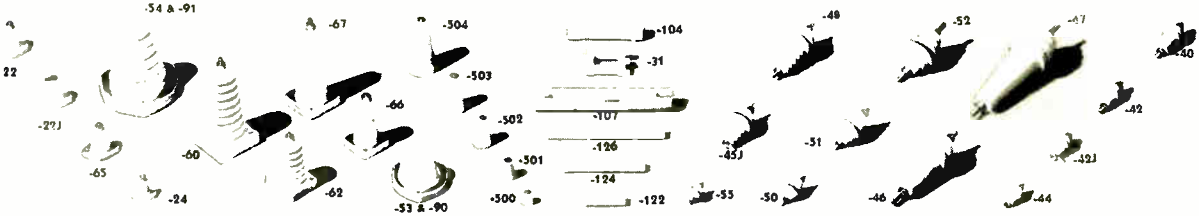
PLUGS & JACKS — RADIO HARDWARE

Cat. No.	Description	
	TINNED COPPER SOLDERING TERMINALS	
	Size Hole	Length
110 880	6-32	9/16″
110 881	1/4″	13/16″
110 882	3/8″	1″
110 883	10-32	15/16″
110 884	10-32	1″
110 885	1/4″	1 1/16″
110 886	.180″	7/8″
110 887	25/64″	1 9/32″
110 888	33/64″	1 9/32″
110 889	25/64″	1 1/32″
110 890	33/64″	1 1/32″
115 840	Clip for 3/8″ dia. fuse	
110 112	Screw terminal	
235 803	Inductor Clip (LC4S)	
235 804	Inductor Clip (LC4)	
235-860	Inductor Clip (860)	

Cat. No.	Description
	PLUGS
105-15	Long Solderless, tip plug
105 14	Above with sharp point
105 415	Short Solderless, tip plug
108 75	Banana, 3/8″, 6 32 shank
108 75A	Banana, 3/4″, 6 32 shank
108 75BB	Insulated Black Handle
108 75BR	As above, red
108 75C	Banana, tapped 6 32
108 75D	Banana, for riveting
108 77	Banana, 5/8″, 1/4 28 shank
108 77A	Banana, 5/8″, 10 32 shank
108 77BB	Insulated Black Handle
108 77BR	As above, red
106 71	Spring sleeve, 1/2″, 1/4 28 screw
106 73	Spring sleeve, 3/8″, 10 32 screw
106 73A	Spring sleeve, 10 32 tapped
	JACKS
108 74	7/32″ x 1/4 28 thread
108 76	15/16″ x 3/8 24 thread
108 76A	1/4 20 tapped
106 70	1/4 20 screw, for No. 106-71

Cat. No.	Description
106 72	10 32 screw, takes 73, 73A
105-520	Removable Round head, tip jack
	For 3/8″ hole, 5/32″ panel
	Choice of colors
105 418	Molded Round head tip, Red
105 419	Molded Round head tip, Black
105 420	Insulated combination, Red
105 421	Insulated combination, Black
105 16	Large Round Head
105 416	Small Round Head
105 417	Small Hex Head
105 1	Headless, tip
105 401	Twin tip, 7/8″ mtg. centers
105 4012	As above, marked "Speaker"
105 4015	As above, marked "Phono"
105 432	Shorting twin tip, Black
105 433	As above, red
115 255	Panel Bearing
115 256	Bearing & 3″ shaft
115 2562	Bearing & 6″ shaft
115 253	3″ flexible shaft
115-254	6″ flexible shaft

INSULATORS

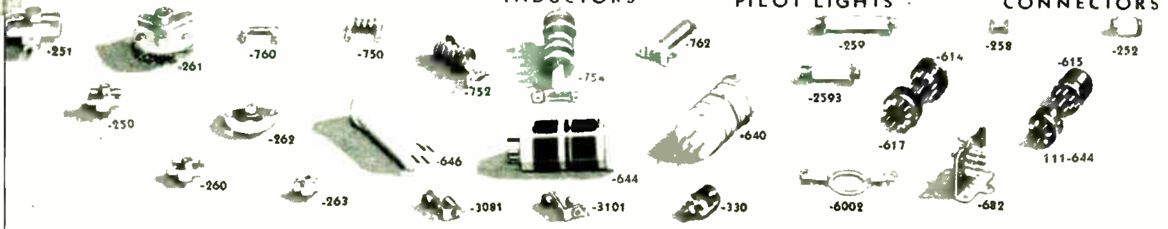


Cat. No.	Description		
Dimension symbols: "H" height, "M" mounting centers			
	STAND-OFF		
	M	H	Hdwe.
135 20	1 5/16	1 9/16	10 32
135 20J	1 5/16	1 9/16	74 jack
135 22	1 3/16	1	8 32
135 22J	1 3/16	1	74 jack
135 60	1 7/8	1 1/2	1/4 20
135 62	1 3/8	2 3/4	1/4 20
	METAL BASE TYPE		
135 65	1 1/2	1 3/8	10 32
135 65B	1 1/2	1 3/8	10 32
135 65J	1 1/2	1 3/8	74 jack
135 66	1 3/8	2 3/4	1/4 20
135 66B	1 3/8	2 3/4	1/4 20
135 66J	1 3/8	2 3/4	76 jack
135 67	1 3/4	4 1/2	1/4 20
135 67B	1 3/4	4 1/2	1/4 20
135 67J	1 3/4	4 1/2	76 jack
135 68	1 3/8	2	10 32
135 68B	1 3/8	2	10 32
135-68J	1 3/8	2	74 jack

Cat. No.	Description			
	STEATITE CONES			
135 500		5/8		6 32
135 501		1		8 32
135 502		1 1/2		8 32
135 503		2		10 32
135 504		3		10 32
Dimension symbols: "B" outside dia. at panel, "D" dia. of panel hole, "H" height of top portion.				
	THRU-PANEL			
	B	D	H	Hdwe.
135-40	15/16	7/16	1 1/4	10-32
135-40J	15/16	7/16	1 1/4	74 jack
135-42	3/4	.100	7/8	10-32
135-42J	3/4	.100	7/8	74 jack
135-44	5/8	.305	5/8	6-32
135-45	1 1/4	1/2	1 3/8	10-32
135-45J	1 1/4	1/2	1 3/8	74 jack
135-46	1 5/8	11/16	2 3/4	1/4-20
135-46J	1 5/8	11/16	2 3/4	76 jack
135-47	2 1/8	31/32	1 1/2	1/4-20
135-47J	2 1/8	31/32	1 1/2	76 jack
135-48	1 5/8	11/16	2	10-32
135-48J	1 5/8	11/16	2	74 jack

Cat. No.	Description			
	LEAD-IN BUSHINGS			
	B	D	H	Hdwe.
135 50	3/4	15/32	1/2	6 32
135 51	1 1/4	27/32	13/16	10 32
135 52	1 3/4	1 7/32	1 1/8	1/4 20
135 53	2 1/2	1 27/32	1 3/4	
135 54	3 1/2	2 11/16	4	
135 55	3/4	15/32	1/4	6 32
135 90	Alum. Mtg. flange for No. 135 53			
135 91	Alum. Mtg. flange for No. 135 54			
	ANTENNA INSULATORS			
136 104	5/8″ x 4″	400 lb. test		
136 107	1″ x 7″	800 lb. test		
136 112	1″ x 12″	800 lb. test		
	FEEDER INSULATORS			
136 122	3/8″ x 1/2″ x 2″ Silicone impregnated			
136 124	3/8″ x 1/2″ x 4″ Silicone impregnated			
136 126	3/8″ x 1/2″ x 6″ Silicone impregnated			
136 31	Transposition insulator			

COUPLINGS — R. F. CHOKES — INDUCTORS — PILOT LIGHTS — CONNECTORS



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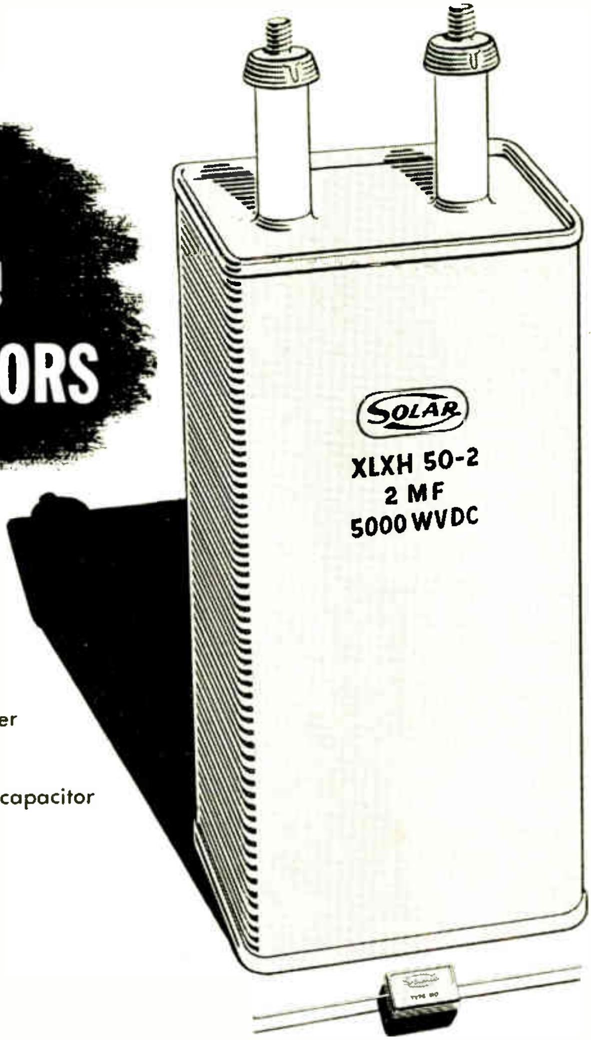

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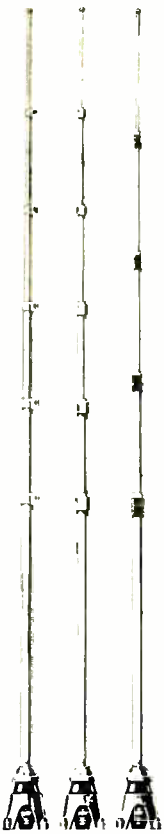

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Weather Resistant Steel Antennas

No.	Description	Extended Length	Collapsed Length	Base O.D.	Base I.D.	Weight Each
112-M	2-sec., telescoping	11'8"	6'1"	.656"	.556"	4 lbs.
318-M	3-sec., telescoping	17'3"	6'2"	.875"	.775"	7 lbs.
224-M	4-sec., telescoping	22'9"	6'3"	1.063"	.963"	11 lbs.
130-M	5-sec., telescoping	28'3"	6'4"	1.250"	1.150"	15 lbs.
136-M	6-sec., telescoping	33'9"	6'5"	1.500"	1.400"	20 lbs.

Light-Weight Aluminum Antennas

No.	Description	Extended Length	Collapsed Length	Base O.D.	Base I.D.	Weight Each
AL-106	1-pc., tapered rod	6'3"	6'3'	.313'	.....	1/4 lb.
AL-312	2-sec., telescoping	12'4"	6'4'	.500'	.334"	1½ lbs.
AL-518	3-sec., telescoping	18'5'	6'4"	.750"	.584"	3 lbs.
AL-324	4-sec., telescoping	24'4"	6'4"	1.000"	.834"	5 lbs.
AL-530	5-sec., telescoping	30'0"	6'5"	1.250"	1.084"	7 lbs.
AL-535	6-sec., telescoping	35'8"	6'5"	1.500"	1.310"	12 lbs.

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No.	Description	Extended Length	Minimum Length	Base O.D.	Base I.D.	Weight Each
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AM-035	2-sec., tapered	35'0"	17'9"	2.000"	1.732"	19 lbs.

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No.	Description	Extended Length	Collapsed Length	Base O.D.	Base I.D.	Weight Each
MM-313	2-sec., telescoping	about 13'	6'9"	.625'	.555"	2¾ lbs.
MM-419	3-sec., telescoping	about 19'	6'9"	.750"	.666"	5 lbs.
MM-425	4-sec., telescoping	about 25'	6'9"	.875"	.777"	8 lbs.
MM-430	5-sec., telescoping	about 30'	6'9"	1.063"	.935"	13 lbs.
MM-435	5-sec., telescoping	about 35'	7'8"	1.063"	.935"	15 lbs.

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108-M	2-sec., telescoping	8'2"	4'7"	.750"	10-meter	2 lbs.
113-M	3-sec., telescoping	12'4"	4'8"	.875"	double zepp	3½ lbs.
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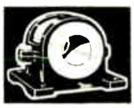

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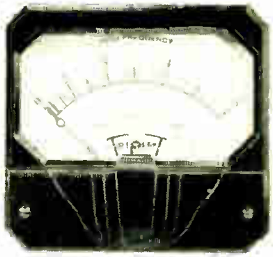
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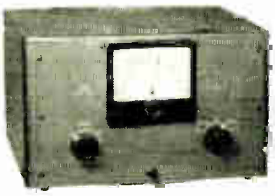

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Handy pocket-size volt ohm-milliammeter. AC-DC voltage ranges to 5000 volts, self contained, plus complete facilities for direct current and resistance analyses.

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It provides four separate circuits for measuring amplitude modulation: (1) per cent modulation (average); (2) peak flash per cent modulation; (3) carrier shift; (4) audio output for headphones. These methods may be used separately, concurrently, or in any combination. ↓

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← Model 3276 Field Radiation Meter

A small, compact portable field radiation meter with power supply and antenna self contained.

Meter calibrated in db per meter from 0 to plus 25 with a set adjustment providing a means for measuring losses and as increments in radiation power.

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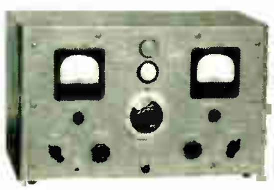

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ASG 8	.01		19/32 x 1 3/16	1.60
ASG 9	.02		19/32 x 1 3/16	1.70
ASG 10	.05	1500	3/4 x 1 3/4	1.85
ASG 11	.1		3/4 x 2 1/4	2.15
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ASG 13	.002		19/32 x 1 3/16	$1.90
ASG 14	.005		19/32 x 1 3/16	2.05
ASG 15	.01		19/32 x 1 3/16	2.25
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ASG 17	.05		3/4 x 1 3/4	2.80
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ASG 19	.25		1 3/8 x 2 3/4	3.70
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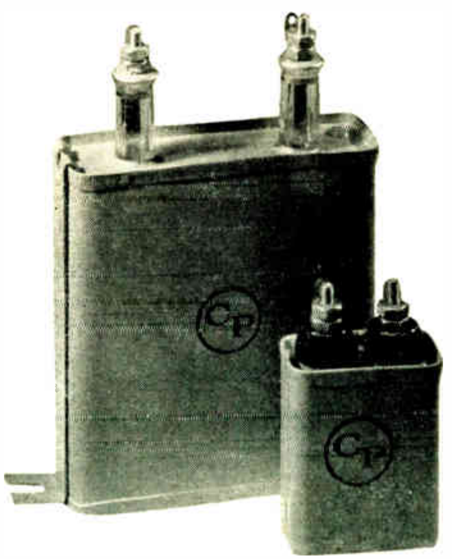
ASG 21	.002		19/32 x 1 3/16	5.25
ASG 22	.005		19/32 x 1 3/16	5.40
ASG 23	.01	3000	19/32 x 1 9/16	5.60
ASG 24	.02		3/4 x 1 3/4	5.85
ASG 25	.05		29/32 x 2 1/4	6.15
ASG 26	.1		1 3/8 x 2 3/4	6.50
ASG 27	.001		19/32 x 1 3/16	$6.50
ASG 28	.002		19/32 x 1 3/16	6.70
ASG 29	.005		19/32 x 1 9/16	6.95
ASG 30	.01	5000	3/4 x 1 3/4	7.25
ASG 31	.02		3/4 x 2 1/4	7.65
ASG 32	.05		1 3/8 x 2 3/4	8.15
ASG 33	.1		1 3/8 x 3 1/2	9.10
ASG 37	.01		3/4 x 2 1/4	8.15
ASG 38	.02	7500	29/32 x 2 3/4	9.25
ASG 39	.05		1 3/8 x 2 3/4	11.50
ASG 40	.0005		19/32 x 1 9/16	$7.30
ASG 41	.001	10,000	19/32 x 1 9/16	7.50
ASG 46	.03		1 3/8 x 3 1/2	15.00
ASG 47	.0005		29/32 x 2 3/4	$14.50
ASG 48	.001	15,000	29/32 x 2 3/4	14.80
ASG 50	.0005		1 3/8 x 3 1/2	$19.50
ASG 51	.001	20,000	1 3/8 x 3 1/2	20.50
ASG 52	.0005	30,000	1 3/8 x 3 1/2	$22.50



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Smaller, lighter, more economical, greater safety factor, longer life

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Type Nos.	Mfds.	D.C.-W.V.	Dimensions	List Price
AOCO6C4	4	600	4" x 2" x 1 1/4"	$5.28
AOC1M1	1	1000	2 3/8" x 1 3/4" x 1"	4.02
AOCO1M2	2	1000	3 1/2" x 2" x 1 1/4"	5.17
AOC1M4	4	1000	4 5/8" x 2 1/2" x 1 3/16"	6.44
AOC1M8	8	1000	4 5/8" x 3 3/4" x 1 3/4"	9.24
AOCO2M1	1	2000	2 3/8" x 2 x 1 1/4"	5.72
AOC2M2	2	2000	3 1/2" x 2 1/2" x 1 3/16"	6.71
AOC2M4	4	2000	3 1/2" x 3 3/4" x 1 3/4"	9.24
AOC3M1	1	3000	4" x 2 1/2" x 1 3/16"	12.10
AOC3M2	2	3000	4" x 3 3/4" x 1 1/4"	15.40
AOC3M4	4	3000	4 5/8" x 3 3/4" x 1 3/4"	21.29
AOC4M1	1	4000	4" x 3 3/4" x 1 1/4"	27.50
AOC4M2	2	4000	4 5/8" x 3 3/4" x 1 3/4"	33.00
AOC5M1	1	5000	4" x 3 3/4" x 1 3/4"	33.00
AOC5M2	2	5000	3 1/2" x 3 3/4" x 4 9/16"	41.25
AOC75C1	1	7500	3 1/2" x 3 3/4" x 4 9/16"	49.50
AOC10M1	1	10,000	4" x 3 3/4" x 4 9/16"	88.00

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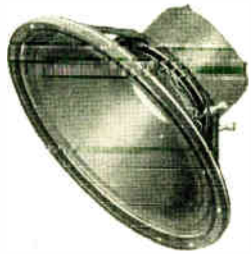
603 DIA-CONE 15-inch*



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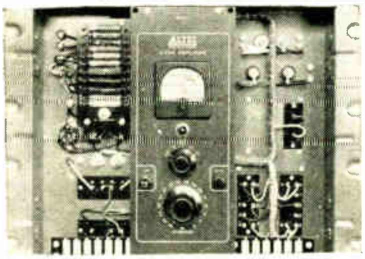
SPEAKERS • AMPLIFIERS • TRANSFORMERS



A-255 AMPLIFIER

TRANSFORMERS

MODEL 604 DUPLEX: Here's the speaker that's the standard of FM studios! The Duplex is a full 2-way multicellular speaker that reproduces the entire FM range, from 50 to 15,000 cycles, without intermodulation or distortion.

MODEL 603 DIA-CONE: Follows close at the heels of the Duplex in quality, but easier on your budget. This speaker incorporates all the advantages of a 2-way multicellular speaker system.

MODEL 600 DIA-CONE: Typical Altec Lansing high fidelity in a smaller-sized, lower-priced speaker. Manufacturers who design sets for critical ears are selecting this model for their finer combinations.

AMPLIFIERS: A complete line of quality equipment ranging from pre-amplifiers to quarter kilowatt amplifiers. For specific information, write for bulletins on the various sizes and types available.

TRANSFORMERS: Altec Lansing has developed a complete line of transformers, especially designed for high quality use. Input, interstage and output transformers available with a transmission range of 20 - 20,000 cycles (± 1 db).

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LANSING CORPORATION
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**Models 603 and 600 both employ the exclusive Dia-Cone principle, reproducing high frequencies and low frequencies from separate diaphragms.*

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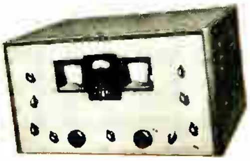
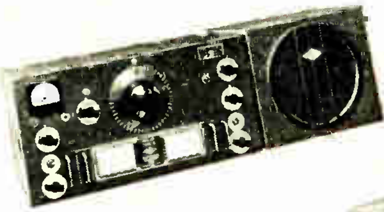


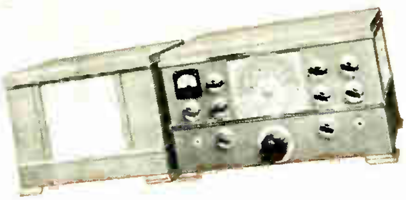
HAMMARLUND HQ-129-X
Professional type communications receiver built to meet the needs of the most discriminating professional. Frequency range .54—31 mc.

HALLICRAFTER SX-42
A new high standard of performance and versatility, covering all frequencies from 540 kc to 110 mc.

HAMMARLUND SP-400-X SUPER PRO
Outstanding communications receiver for professional and amateur use, covering frequency range .54—30.0 mc.



NATIONAL HRO-5A1
Brings the famous HRO series to a new high in receiver performance. Frequency range normally supplied 1.7—30.0 mc.



NATIONAL NC-2-40D
General communications receiver covering 490 to 30,000 kc with calibrated electrical bandspread for the 80, 40, 20 and 10 meter radio amateur band.

Recent completion of our new modern building more than quadruples our capacity to serve amateurs, servicemen, and radio electronics industries.

Special note to radio amateurs: Our amateur department is being constantly expanded in order to render even better service to hams by hams.

We wish to thank our legion of satisfied customers who have made our building expansion possible.

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Better than ±3% accuracy on all ranges. Less than 7 uuf input at all frequencies.

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FULL RANGE

Will measure voltages from 0.2 to 300 volts at 20 cycles to beyond 200 megacycles.

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NEW LOW COST*

Priced at less than half the cost of comparable laboratory instruments.

SPECIFICATIONS

Laboratory Kit Model LKV-300

VOLTAGE RANGE: 0.2 to 300 volts a-c, in five ranges.
ACCURACY: ±3% of full scale on all five ranges for sinusoidal voltages.
LOADING EFFECT: Input circuit has less than 7 uuf capacitance at all frequencies.
FREQUENCY RESPONSE: Within full rated accuracy from 50 cycles to 50 megacycles. Down 1 db at 20 cycles, and up less than 2.5 db at 150 megacycles.
STABILITY: Little if any resetting will be required even over long periods of time, once the initial zero setting is made on the lowest range.
INDICATING INSTRUMENT: Specially designed physically and electrically. The linearity and clarity of the scale permit simple and accurate readings to be made. It is impossible to burn out the movement due to overvoltage applied to the input terminals.
CIRCUIT: Twin diode probe working into a bridge type feed back stabilized amplifier. The circuit was designed for this application by the Alfred W. Barber Laboratories and is fully licensed under their patents.
TUBES SUPPLIED: 1—6AL5, 2—6J5, 1—6ZY5G, all aged, tested and matched.
POWER SUPPLY: Works directly from 105-125 volt a-c line.
RATING: Continuous duty.
MOUNTING: Reverse etched aluminum panel in a black crackle finished steel cabinet.
DIMENSIONS: 6½" x 11" x 5¾
WEIGHT: 10 pounds.

Build it yourself! That's part of the serious Ham's creed. Build it better? That goes without saying. Here is your opportunity to do both. You get fully illustrated wiring and assembly instructions. Every part you need is included—laced cables—screws—soldering lugs—mounting brackets—all the parts integrated to give you an instrument of which you will be proud to build, own and use.

Conservatively rated parts, engineered into a finished unit assure you an instrument of unquestionable laboratory accuracy. The entire kit is composed of parts chosen or specified to a particular performance requirement.

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The American Radio Relay League

THE American Radio Relay League, Inc., is a non-commercial association of radio amateurs, bonded for the promotion of interest in amateur radio communication and experimentation, for the relaying of messages by radio, for the advancement of the radio art and of the public welfare, for the representation of the radio amateur in legislative matters, and for the maintenance of fraternalism and a high standard of conduct.

It is an incorporated association without capital stock, chartered under the laws of Connecticut. Its affairs are governed by a Board of Directors, elected every two years by the general membership. The officers are elected or appointed by the Directors. The League is non-commercial and no one commercially engaged in the manufacture, sale or rental of radio apparatus is eligible to membership on its board.

"Of, by and for the amateur," it numbers within its ranks practically every worth-while amateur in the nation and has a history of glorious achievement as the standard-bearer in amateur affairs.

Inquiries regarding membership are solicited. A bona fide interest in amateur radio is the only essential qualification; ownership of a transmitting station and knowledge of the code are not prerequisite.

Membership Application Blank →

Application for Membership

AMERICAN RADIO RELAY LEAGUE

***Administrative Headquarters:* West Hartford 7, Conn., U. S. A.**

....194...

AMERICAN RADIO RELAY LEAGUE,
West Hartford, Conn., U. S. A.

Being genuinely interested in Amateur Radio, I hereby apply for membership in the American Radio Relay League, and enclose $2.50* in payment of one year's dues, $1.25 of which is for a subscription to QST for the same period. Please begin my subscription with the .... issue.

The call of my station is ....

The class of my operator's license is ....

I belong to the following radio societies ....

....

Send my Certificate of Membership or Membership Card indicate which to the address below:

Name ....

....

....

A bona fide interest in amateur radio is the only essential requirement but full voting membership is granted only to licensed radio amateurs of the United States and Canada. Therefore, if you have a license, please be sure to indicate it above.

*$2.50 in the United States and Possessions, $3.00 in the Dominion of Canada, $4.00 in all other countries. Foreign remittances must yield the above amounts in U. S. funds.

To Readers

WHO ARE ALREADY MEMBERS

HERE IS A FORM for your convenience in expressing to your SCM interest in any Communications Department appointment. Read "Leadership and Station Appointments," Chapter XXI. Select the appointment which best fits your operating interests and qualifications. The SCM will be happy to consider your application for Official Relay Station, Official Experimental Station, Official Phone Station, Official Broadcasting Station, or Official Observer. Appointments as Section Emergency Coordinator, Emergency Coordinator, Phone Activities Manager, and Route Manager also are available to amateurs of proven ability. The SCM is particularly interested to know of your interest in any of the leadership appointments. Copy this form, or cut it out. Send direct to your Section Communications Manager (address on page 6, each *QST*). The Communications Department field organization includes the United States and its territories, Canada, Newfoundland, Labrador, Cuba, the Isle of Pines, and the Philippine Islands. Applications from outside these areas cannot be handled.

APPLICATION FOR APPOINTMENT

To: Section Communications Manager .... ***Section, ARRL***

From .... *Call* ....

Street and Number ....

City .... *County* .... *State* ....

I am interested in appointment as ....
(ORS, OES, OPS, OBS)

I would like to be considered for the following leadership appointment if or when a vacancy exists.

....
(SEC, EC, RM, PAM)

My station is operative in the following bands .... *Mc.*

My ARRL membership expires ....

I understand that each ARRL appointment requires annual endorsement, and may be suspended or cancelled at the discretion of the Section Communications Manager for inactivity, lack of interest or failure to report regularly each month. Please send me detailed forms or further information necessary in connection with this application.

Date .... *Signed* ....