RADIO

Handbook

THIRTEENTH EDITION

THE RADIO HANDBOOK

This book is revised and brought up to date (at irregular intervals) as necessitated by technical progress.

Thirteenth Edition

The Standard of the Field—

for practical radiomen
practical engineers
practical technicians
advanced amateurs

by

Editors and Engineers
LIMITED

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THE RADIO HANDBOOK

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BRAN'S VADE MECUM (THE WORLD'S RADIO TUBES)

THE RADIO HANDBOOK

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CHAPTER ONE

Introduction to Radio

The field of radio is a division of the much larger field of electronics. Radio itself is such a broad study that it is still further broken down into a number of smaller fields of which only shortwave or high-frequency radio is covered in this book. Specifically the field of communication on frequencies from 1.8 to 450 megacycles is taken as the subject matter for this work.

The largest group of persons interested in the subject of high-frequency communication is the more than 100,000 radio amateurs located in nearly all countries of the world. Strictly speaking, a radio amateur is anyone interested in radio non-commercially, but the term is ordinarily applied only to those hobbyists possessing transmitting equipment and a license from the government.

It was for the radio amateur, and particularly for the serious and more advanced amateur, that most of the equipment described in this book was developed. However, in each equipment group simple items also are shown for the student or beginner. The design principles behind the equipment for high-frequency radio communication are of course the same whether the equipment is to be used for commercial, military, or amateur purposes, the principal differences lying in construction practices, and in the tolerances and safety factors placed upon components.

With the increasing complexity of high-frequency communication, resulting primarily from increased utilization of the available spectrum, it becomes necessary to delve more deeply into the basic principles underlying radio communication, both from the standpoint of equipment design and operation and from the standpoint of signal propagation. Hence, it will be found that this edition of the RADIO HANDBOOK has been devoted in greater proportion to the teaching of the principles of equipment design and signal propagation. It is in response to requests from schools and agencies of the Department of Defense, in addition to persistent requests from the amateur radio fraternity, that coverage of these principles has been expanded.

Amateur Radio Amateur radio is a fascinating hobby with many phases. So strong is the fascination offered by this hobby that many executives, engineers, and military and commercial operators enjoy amateur radio as an avocation even though they are also engaged in the radio field commercially. It captures and holds the interest of many people in all walks of life, and in all countries of the world where amateur activities are permitted by law.

Amateurs have rendered much public service through furnishing communications to and from the outside world in cases where disaster has isolated an area by severing all wire communications. Amateurs have a proud record of heroism and service in such occasion. Many expeditions to remote places have been kept in touch with home by communication with amateur stations on the high frequencies. The amateur's fine record of performance with the "wireless" equipment of World War I has been surpassed by his outstanding service in World War II.

By the time peace came in the Pacific in the summer of 1945, many thousand former amateur operators were serving in the allied armed forces. They had supplied the army, navy, marines, coast guard, merchant marine, civil service, war plants, and civilian defense organizations with *trained* personnel for radio, radar, wire, and visual communications and for teaching. Even now, at the time of this writing, amateurs are being called back into the expanded defense forces, are returning to defense plants where their skills are critically needed, and are being organized into communication units as an adjunct to civil defense groups.

Station and Operator Licenses Every radio transmitting station in the United State no matter how low its power must have a license from the federal government before being operated; some classes of stations must have a permit from the government before even being constructed. And every operator of a transmitting station must have an operator's license before operating a transmitter. There are no exceptions. Similar laws apply in practically every major country.

Classes of Amateur Operator Licenses There are at present *six* classes of amateur operator licenses which have been authorized by the Federal Communications Commission. These classes differ in many respects, so each will be discussed briefly.

(a) *Amateur Extra Class.* This class of license will become available on January 1, 1952, to any U. S. citizen who at any time has held for a period of two years or more a valid amateur license, issued by the FCC, excluding licenses of the Novice and Technician Classes. The examination for the license includes a code test at 20 words per minute, the usual tests covering basic amateur practice and general amateur regulations, and an additional test on advanced amateur practice. All amateur privileges are accorded the holders of this operator's license.

(b) *Advanced Class.* This class of license is equivalent to the old Amateur Class A license, and accords to the holders all amateur privileges except those to be reserved to holders of the Amateur Extra Class license. New Advanced Class licenses will not be issued after December 31, 1952, although existing licenses may be renewed as long as the holder to whom the license was issued meets the renewal requirements current at the time of application for renewal. The Advanced Class license is available to any U. S. citizen who at any time has held for a period of one year or more a valid amateur license issued by the FCC, excluding licenses of the Novice and Technician Classes. The examination for the license includes a code test at 13 words per minute, the usual tests covering basic amateur practice and general amateur regulations, plus an additional test on advanced amateur radiotelephony.

(c) *General Class.* This class of amateur license is equivalent to the old Amateur Class B license, and accords to the holders all amateur privileges except those reserved for holders of the Advanced Class license, and those which may be set aside for holders of the Amateur Extra Class license. This class of amateur operator's license is available to any U. S. citizen. The examination for the license includes a code test at 13 words per minute, and the usual examinations covering basic amateur practice and general amateur regulations.

(d) *Conditional Class.* This class of amateur license and the privileges accorded by it are equivalent to the General Class license. However, the license can be issued only to those whose residence is more than 125 miles airline from the nearest location at which FCC examinations are held at intervals of not more than three months for the General Class amateur operator license, or to those who for any of several specified reasons are unable to appear for examination.

(e) *Technician Class.* This is a new class of license which is available to any citizen of the United States. The examination is the same as that for the General Class license, except that the code test is at a speed of 5 words per minute. The holder of a Technician class license is accorded all authorized amateur privileges in the amateur frequency bands above 220 megacycles.

(f) *Novice Class.* This is a new class of license which is available to any U. S. citizen who has not previously held an amateur license of any class issued by any agency of the U. S. government, military or civilian. The examination consists of a code test at a speed of 5 words per minute, plus an examination on the rules and regulations essential to beginner's operation, including sufficient elementary radio theory for the understanding of those rules. The

THE AMATEUR BANDS

(1) 1800 to 2000 kc.

Portions of this band 25 kc. in width (1800-1825 and 1875-1900 or 1900-1925 and 1975-2000) have been assigned on a regional basis with day and night power limitations and subject to no interference to the use of Loran. The assignments, for A1 and A3 only, are as follows:

AREA	BAND, KC.	POWER IN WATTS DAY	NIGHT
Mississippi River to East Coast U.S. (except Florida and states bordering the Gulf of Mexico)	1800-1825 1875-1900	500	200
Mississippi River to West Coast U.S. (except states bordering Gulf of Mexico)	1900-1925 1975-2000	*500	*200
Florida and states bordering Gulf of Mexico	1800-1825 1875-1900	200	no operation
Puerto Rico and Virgin Islands	1900-1925 1975-2000	500	50
Hawaiian Islands	1900-1925 1975-2000	500	200

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

(2) 3500 to 4000 kc. A1, entire band.
A3 telephony, Extra Class or Advanced Class licensees only, 3800 to 4000 kc.
A3 narrow-band FM telephony, Extra Class or Advanced Class licensees only, 3800 to 3850 kc.

(3) 7000 to 7300 kc. A1, entire band.

(4) 14000 to 14400 kc. A1, entire band.
A3 telephony, Extra Class or Advanced Class licensees only, 14,200 to 14,300 kc.
A3 narrow-band FM telephony, Extra Class or Advanced Class licensees only, 14,200 to 14,250 kc.

(5) 21.0 to 21.45 Mc. Proposed only.

(6) 26.96 to 27.23 Mc. A0, A1, A2, A3, A4, and FM.

(7) 28.0 to 29.7 Mc. A1, entire band.
A3 telephony and narrow-band FM telephony, 28,500 to 29,700 kc.
FM, 29.0 to 29.7 Mc.

(8) 50.0 to 54.0 Mc. A1, A2, A3, A4, and narrow-band FM telephony.
FM, 52.5 to 54.0 Mc.

(9) 144 to 148 Mc. A0, A1, A2, A3, A4, and FM.

(10) 220 to 225 Mc. A0, A1, A2, A3, A4, and FM.

(11) 235 to 240 Mc. A0, A1, A2, A3, A4, and FM, as an alternate band to the 220 to 225 Mc. band in those areas where use of 220 to 225 Mc. is not permitted by FCC order.

(12) 420 to 450 Mc. A0, A1, A2, A3, A4, A5, and FM. Peak antenna power is limited to 50 watts.

(13) 1215 to 1300 Mc. A0, A1, A2, A3, A4, A5, and FM.

(14) 2300 to 2450 Mc., 3300 to 3500 Mc., 5650 to 5925 Mc., 10,000 to 10,500 Mc., 21,000 to 22,000 Mc., and any frequency or frequencies above 30,000 Mc. may be used for A0, A1, A2, A3, A4, A5, FM, and pulse emission.

Bands 1 to 7 are termed *high frequency;* bands 8, 9, and 10, *very high frequency;* bands to 3000 Mc., *ultra high frequency;* above 3000 Mc., *super high frequency*.

A0 means unmodulated carrier, A1 means c-w telegraphy, A2 is modulated c.w., A3 is radiotelephony with amplitude modulation, A4 is facsimile, A5 is television, FM is frequency modulation either for telegraphy or telephony.

Novice Class of license affords severely restricted privileges, is valid only for a period of one year (as contrasted to all other classes of amateur licenses which run for a term of five years), and is not renewable.

The holder of the Novice Class of license is restricted to a power input to the output stage of not more than 75 watts, must use a crystal for frequency control, and may use only the following sub-bands and types of emission:

(1) 3700 to 3750 kc.—c-w telegraphy (A1 emission).

(2) 26.96 to 27.23 Mc.—c-w telegraphy (A1 emission).

(3) 145 to 147 Mc.—telegraphy or telephony using any type of emission authorized for use by amateurs in this frequency band.

Starting Your Study When you start to prepare yourself for the amateur or other examination you will find that the circuit diagrams, tube characteristic curves, and formulas appear confusing and

A	· —	N	— ·	1	· — — — —
B	— · · ·	O	— — —	2	· · — — —
C	— · — ·	P	· — — ·	3	· · · — —
D	— · ·	Q	— — · —	4	· · · · —
E	·	R	· — ·	5	· · · · ·
F	· · — ·	S	· · ·	6	— · · · ·
G	— — ·	T	—	7	— — · · ·
H	· · · ·	U	· · —	8	— — — · ·
I	· ·	V	· · · —	9	— — — — ·
J	· — — —	W	· — —	Ø	— — — — —
K	— · —	X	— · · —		
L	· — · ·	Y	— · — —		
M	— —	Z	— — · ·		

Ø MEANS ZERO, AND IS WRITTEN IN THIS WAY TO DISTINGUISH IT FROM THE LETTER "O". IT OFTEN IS TRANSMITTED INSTEAD AS ONE LONG DASH (EQUIVALENT TO 5 DOTS)

PERIOD (.)	· — · — · —	WAIT SIGN (AS)	· — · · ·
COMMA (,)	— — · · — —	DOUBLE DASH (BREAK)	— · · · —
INTERROGATION (?)	· · — — · ·	ERROR (ERASE SIGN)	· · · · · · · ·
QUOTATION MARK (")	· — · · — ·	FRACTION BAR (/)	— · · — ·
COLON (:)	— — — · · ·	END OF MESSAGE (AR)	· — · — ·
SEMICOLON (;)	— · — · — ·	END OF TRANSMISSION (SK)	· · · — · —
PARENTHESIS ( )	— · — — · —	INTERNAT. DISTRESS SIG. (SOS)	· · · — — — · · ·

Figure 1.

The Continental (or International Morse) Code is used for substantially all non-automatic radio communication. DO NOT memorize from the printed page; code is a language of SOUND, and must not be learned visually; learn by listening as explained in the text.

difficult of understanding. But after a few study sessions one becomes sufficiently familiar with the notation of the diagrams and the basic concepts of theory and operation so that the acquisition of further knowledge becomes easier and even fascinating.

As it takes a considerable time to become proficient in sending and receving code, it is a good idea to intersperse technical study sessions with periods of code practice. Many short code practice sessions benefit one more than a small number of longer sessions. Alternating between one study and the other keeps the student from getting "stale" since each type of study serves as a sort of respite from the other.

When you have practiced the code long enough you will be able to follow the gist of the slower sending stations. Many stations send very slowly when working other stations at great distances. Stations repeat their calls many times when calling other stations before contact is established, and one need not have achieved much code proficiency to make out their calls and thus determine their location.

The Code The applicant for any class of amateur operator license must be able to send and receive the Continental Code (sometimes called the International Morse Code). The speed required for the sending and receiving test may be either 5, 13, or 20 words per minute, depending upon the class of license, assuming an average of five characters to the word in each case. The sending and receiving tests run for five minutes, and one minute of errorless transmission or reception must be accomplished within the five-minute interval.

If the code test is failed, the applicant must wait at least one month before he may again appear for another test. Approximately 30% of amateur applicants fail to pass the test. It should be expected that nervousness and excitement will at least to some degree temporarily lower the applicant's code ability. The best prevention against this is to master the code at a little greater than the required speed under ordinary conditions. Then if you slow down a little due to nervousness during a test the result will not prove "fatal."

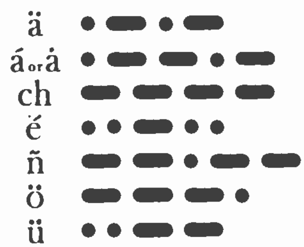


Figure 2.

These code characters are used in languages other than English. They may occasionally be encountered so it is well to know them.

Memorizing the Code There is no shortcut to code proficiency. To memorize the alphabet entails but a few evenings of diligent application, but considerable time is required to build up speed. The exact time required depends upon the individual's ability and the regularity of practice.

While the speed of learning will naturally vary greatly with different individuals, about 70 hours of practice (no practice period to be over 30 minutes) will usually suffice to bring a speed of about 13 w.p.m.; 16 w.p.m. requires about 120 hours; 20 w.p.m., 175 hours.

Since code reading requires that individual letters be recognized instantly, any memorizing scheme which depends upon orderly sequence, such as learning all *"dah"* letters and all *"dit"* letters in separate groups, is to be discouraged. Before beginning with a code practice set it is necessary to memorize the whole alphabet perfectly. A good plan is to study only two or three letters a day and to drill with those letters until they become part of your consciousness. Mentally translate each day's letters into their sound equivalent wherever they are seen, on signs, in papers, indoors and outdoors. Tackle two additional letters in the code chart each day, at the same time reviewing the characters already learned.

Avoid memorizing by routine. Be able to sound out any letter immediately without so much as hesitating to think about the letters preceding or following the one in question. Know *C,* for example, apart from the sequence *ABC.* Skip about among all the characters learned, and before very long sufficient letters will have been acquired to enable you to spell out simple words to yourself in *"dit dahs."* This is interesting exercise, and for that reason it is good to memorize all the vowels first and the most common consonants next.

Actual code practice should start only when the entire alphabet, the numerals, period, comma, and question mark have been memorize so thoroughly that any one can be sounded without the slightest hesitation. Do not bother with other punctuation or miscellaneous signals until later.

Sound Not Sight Each letter and figure *must* be memorized by its *sound* rather than its appearance. Code is a system of sound communication, the same as is the spoken word. The letter *A,* for example, is one short and one long sound in combination sounding like *dit dah,* and it must be remembered as such, and not as "dot dash."

Practice Time, patience, and regularity are required to learn the code properly. Do not expect to accomplish it within a few days.

Don't practice too long at one stretch; it does more harm than good. Thirty minutes at a time should be the limit.

Lack of regularity in practice is the most common cause of lack of progress. Irregular practice is very little better than no practice at all. Write down what you have heard; then forget it; *do not look back.* If your mind dwells even for an instant on a signal about which you have doubt, you will miss the next few characters while your attention is diverted.

While various automatic code machines, phonograph records, etc., will give you practice, by far the best practice is to obtain a study companion who is also interested in learning the code. When you have both memorized the alphabet you can start sending to each other. Practice with a key and oscillator or key and buzzer generally proves superior to to all automatic equipment. Two such sets operated between two rooms are fine—or between your house and his will be just that much better. Avoid talking to your partner while practicing. If you must ask him a question, do it in code. It makes more interesting practice than confining yourself to random practice material.

When two co-learners have memorized the code and are ready to start sending to each other for practice, it is a good idea to enlist the aid of an experienced operator for the first practice session or two so that they will get an idea of how properly formed characters sound.

During the first practice period the speed should be such that substantially solid copy can be made without strain. Never mind if this is only two or three words per minute. In the next period the speed should be increased slightly to a point where nearly all of the characters can be caught only through conscious effort. When the student becomes proficient at this new speed, another slight increase may be made, progressing in this manner until a speed of about 16 words per minute is attained if the object is to pass the amateur 13-word per minute code test. The margin of 3 w.p.m. is recommended to overcome a possible excitement factor at examination time. Then when you take the test you don't have to worry about the "jitters" or an "off day."

Speed should not be increased to a new level until the student finally makes solid copy with ease for at least a five-minute period at the old level. How frequently increases of speed can be made depends upon individual ability and the amount of practice. Each increase is apt to prove disconcerting, but remember "you are never learning when you are comfortable."

A number of amateurs are sending code practice on the air on schedule once or twice each week; excellent practice can be obtained after you have bought or constructed your receiver by taking advantage of these sessions.

If you live in a medium-size or large city, the chances are that there is an amateur radio club in your vicinity which offers free code practice lessons periodically.

Practice Material At the start use plain English, sending from a book, newspaper, or anything handy. Also practice disconnected words from a newspaper or magazine, and groups of mixed letters and numerals.

More detailed instructions on code learning and practice may be obtained from several textbooks which are written to cover this subject exhaustively.*

*THE RADIO AMATEUR NEWCOMER gives further information on code learning and in addition gives a complete list of study questions for the radio amateur operator license examinations with reference data. A number of transmitters and receivers suitable for the beginner in amateur radio are described in detail. The price is $1.00 from our book department (add tax in Calif.).

Skill When you listen to someone speaking you do not consciously think how his words are spelled. This is also true when you read. In code you must train your ears to read code just as your eyes were trained in school to read printed matter. With enough practice you acquire skill, and from skill, speed. In other words, it becomes a *habit*, something which can be done without conscious effort. Conscious effort is fatal to speed; we can't think rapidly enough; a speed of 25 words a minute, which is a common one in commercial operations, means 125 characters per minute or more than two per second, which leaves no time for conscious thinking.

Perfect Formation of Characters When transmitting on the code practice set to your partner, concentrate on the *quality* of your sending, *not* on your speed. Your partner will appreciate it and he could not copy you if you speeded up anyhow.

If you want to get a reputation as having an excellent "fist" on the air, just remember that speed alone won't do the trick. Proper execution of your letters and spacing will make much more of an impression. Fortunately, as you get so that you can send evenly and accurately, your sending speed will automatically increase. Remember to try to see how *evenly* you can *send*, and how *fast* you can *receive*. Concentrate on making signals properly with your key. Perfect formation of characters is paramount to everything else. Make every signal right no matter if you have to practice it hundreds or thousands of times. Never allow yourself to vary the slightest from perfect formation once you have learned it.

If possible, get a good operator to listen to your sending for a short time, asking him to criticize even the slightest imperfections.

Timing It is of the utmost importance to maintain uniform spacing in characters and combinations of characters. Lack of uniformity at this point probably causes beginners more trouble than any other single factor. Every dot, every dash, and every space must be correctly timed. In other words, accurate timing is absolutely essential to intelligibility, and timing of the spaces between the dots and dashes is just as important as the lengths of the dots and dashes themselves.

The characters are timed with the dot as a "yardstick." A standard dash is three times as long as a dot. The spacing between parts of the same letter is equal to one dot; the space between letters is equal to three dots, and that between words equal to five dots.

The rule for spacing between letters and words is not strictly observed when sending slower than about 10 words per minute for the benefit of someone learning the code and desiring receiving practice. When sending at, say, 5 w.p.m., the individual letters should be made the same as if the sending rate were about 10 w.p.m., except that the spacing between letters and words is greatly exaggerated. The reason for this is obvious. The letter *L*, for instance, will then sound exactly the same at 10 w.p.m. as at 5 w.p.m., and when the speed is increased above 5 w.p.m. the student will not have to become familiar with what may seem to him like a new sound, although it is in reality only a faster combination of dots and dashes. At the greater speed he will merely have to learn the identification of the *same* sound without taking as long to do so.

Be particularly careful of letters like *B*. Many beginners seem to have a tendency to leave a longer space after the dash than that which they place between succeeding dots, thus making it sound like *TS*. Similarly, make sure that you do not leave a longer space after the first dot in the letter *C* than you do between other parts of the same letter; otherwise it will sound like *NN*.

Sending vs. Receiving Once you have memorized the code thoroughly you should concentrate on increasing your *receiving* speed. True, if you have to practice with another newcomer who is learning the code with you, you will both have to do some sending. But don't attempt to practice *sending* just for the sake of increasing your sending *speed*.

When transmitting on the code practice set to your partner so that he can get receiving practice, concentrate on the *quality* of your sending, not on your speed.

Because it is comparatively easy to learn to send rapidly, especially when no particular care is given to the quality of sending, many operators who have just received their licenses get on the air and send mediocre or worse code at 20 w.p.m. when they can barely receive good code at 13. Most oldtimers remember their own period of initiation and

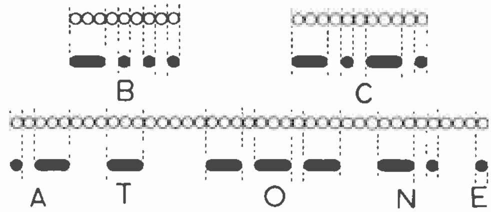


Figure 3.

Diagram illustrating relative lengths of dashes and spaces referred to the duration of a dot. A dash is exactly equal in duration to three dots; spaces between parts of a letter equal one dot; those between letters, three dots; space between words, five dots. Note that a slight increase between two parts of a letter will make it sound like two letters.

are only too glad to be patient and considerate if you tell them that you are a newcomer. But the surest way to incur their scorn is to try to impress them with your "lightening speed," and then to request them to send more slowly when they come back at you at the same speed.

Stress your copying ability; never stress your sending ability. It should be obvious that if you try to send faster than you can receive, your ear will not recognize any mistakes which your hand may make.

Using the Key Figure 4 shows the proper position of the hand, fingers and wrist when manipulating a telegraph or radio key. The forearm should rest naturally on the desk. It is preferable that the key be placed far enough back from the edge of the table (about 18 inches) that the elbow can rest on the table. Otherwise, pressure of the table edge on the arm will tend to hinder the circulation of the blood and weaken the ulnar nerve at a point where it is close to the surface, which in turn will tend to increase fatigue considerably.

The knob of the key is grasped lightly with the thumb along the edge; the index and third fingers rest on the top towards the front or far edge. The hand moves with a free up and down motion, the wrist acting as a fulcrum. The power must come entirely from the arm muscles. The third and index fingers will bend slightly during the sending but not because of deliberate effort to manipulate the finger muscles. Keep your finger muscles just tight enough to act as a "cushion" for the

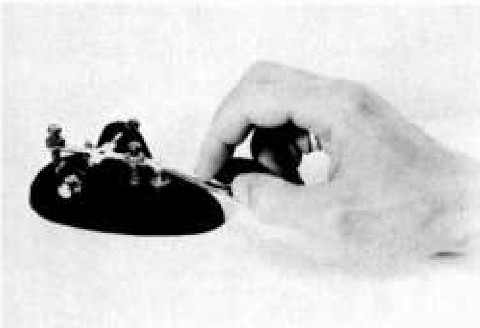


Figure 4.
PROPER POSITION OF THE FINGERS FOR OPERATING A TELEGRAPH KEY.
The fingers hold the knob and act as a cushion. The hand rests lightly on the key. The muscles of the forearm provide the power, the wrist acting as the fulcrum. The power should not come from the fingers, but rather from the forearm muscles.

arm motion and let the slight movement of the fingers take care of itself. The key's spring is adjusted to the individual wrist and should be neither too stiff nor too loose. Use a moderately stiff tension at first and gradually lighten it as you become more proficient. The separation between the contacts must be the proper amount for the desired speed, being somewhat under 1/16 inch for slow speeds and slightly closer together (about 1/32 inch) for faster speeds. Avoid extremes in either direction.

Do not allow the muscles of arm, wrist, or fingers to become tense. Send with a full, free arm movement. Avoid like the plague any finger motion other than the slight cushioning effect mentioned above.

Stick to the regular hand key for learning code. No other key is satisfactory for this purpose. Not until you have thoroughly mastered both sending and receiving at the maximum speed in which you are interested should you tackle any form of automatic or semi-automatic key such as the Vibroplex ("bug") or an electronic key.

Difficulties Should you experience difficulty in increasing your code speed after you have once memorized the characters, there is no reason to become discouraged. It is more difficult for some people to learn code than for others, but there is no justification for the contention sometimes made that "some people just can't learn the code." It is not a matter of intelligence; so don't feel ashamed if you seem to experience a little more than the usual difficulty in learning code. Your reaction time may be a little slower or your coordination not so good. If this is the case, remember *you can still learn the code.* You may never learn to send and receive at 40 w.p.m., but you can learn sufficient speed for all non-commercial purposes and even for most commercial purposes if you have patience, and refuse to be discouraged by the fact that others seem to pick it up more rapidly.

When the sending operator is sending just a bit too fast for you (the best speed for practice), you will occasionally miss a signal or a small group of them. When you do, leave a blank space; do not spend time futilely trying to recall it; dismiss it, and center attention on the next letter; otherwise you'll miss more. Do not ask the sender any questions until the transmission is finished.

To prevent guessing and get equal practice on the less common letters, depart occasionally from plain language material and use a jumble of letters in which the usually less commonly used letters predominate.

As mentioned before, many students put a greater space after the dash in the letter *B* than between other parts of the same letter so it sounds like *TS. C, F, Q, V, X, Y* and *Z* often give similar trouble. Make a list of words or arbitrary combinations in which these letters predominate and practice them, both sending and receiving until they no longer give you trouble. Stop everything else and stick at them. So long as they give you trouble you are not ready for anything else.

Follow the same procedure with letters which you may tend to confuse such as *F* and *L,* which are often confused by beginners. Keep at it until you *always* get them right without having to stop *even an instant* to think about it.

If you do not instantly recognize the sound of any character, you have not learned it; go back and practice your alphabet further. You should never have to omit writing down every signal you hear except when the transmission is too fast for you.

Write down what you hear, not what you

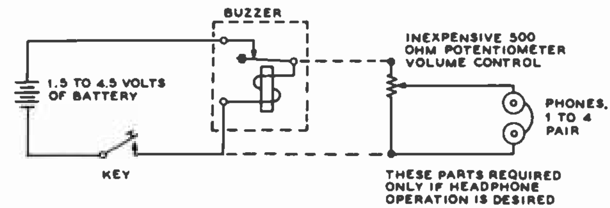


Figure 5.

THE SIMPLEST CODE PRACTICE SET CONSISTS OF A KEY AND A BUZZER.

The buzzer is adjusted to give a steady, high-pitched "whine." If desired, the phones may be omitted, in which case the buzzer should be mounted firmly on a sounding board. Crystal, magnetic, or dynamic earphones may be used. Additional sets of phones should be connected in parallel, not in series.

think it should be. It is surprising how often the word which you guess will be wrong.

Copying Behind All good operators copy several words behind, that is, while one word is being received, they are writing down or typing, say, the fourth or fifth previous word. At first this is very difficult, but after sufficient practice it will be found actually to be *easitr* than copying close up. It also results in more accurate copy and enables the receiving operator to capitalize and punctuate copy as he goes along. It is not recommended that the beginner attempt to do this until he can send and receive accurately and with ease at a speed of at least 12 words a minute.

It requires a considerable amount of training to dissociate the action of the subconscious mind from the direction of the conscious mind. It may help some in obtaining this training to write down two columns of short words. Spell the first word in the first column out loud while writing down the first word in the second column. At first this will be a bit awkward, but you will rapidly gain facility with practice. Do the same with all the words, and then reverse columns.

Next try speaking aloud the words in the one column while writing those in the other column; then reverse columns.

After the foregoing can be done easily, try sending with your key the words in one column while spelling those in the other. It won't be easy at first, but it is well worth keeping after if you intend to develop any real code proficiency. Do *not* attempt to catch up. There is a natural tendency to close up the gap, and you must train yourself to overcome this.

Next have your code companion send you a word either from a list or from straight text; do not write it down yet. Now have him send the next word; *after* receiving this second word, write down the first word. After receiving the third word, write the second word; and so on. Never mind how slowly you must go, even if it is only two or three words per minute. *Stay behind.*

It will probably take quite a number of practice sessions before you can do this with any facility. After it is relatively easy, then try staying two words behind; keep this up until it is easy. Then try three words, four words, and five words. The more you practice keeping received material in mind, the easier

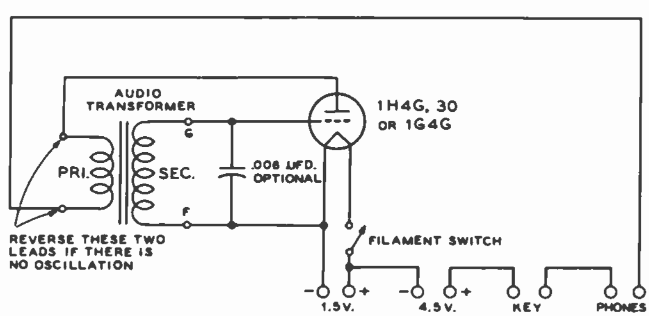


Figure 6.

SIMPLE VACUUM-TUBE CODE PRACTICE OSCILLATOR.

Power is furnished by a dry cell and a 4½-volt "C" battery. If the 0.006-μfd. capacitor is omitted, a higher pitched note will result. The note may have too low a pitch even without the capacitor unless the smallest, least expensive audio transformer available is used. The earphones must be of the magnetic or dynamic type since the plate current of the oscillator must flow through the phones.

it will be to stay behind. It will be found easier at first to copy material with which one is fairly familiar, then gradually switch to less familiar material.

Automatic Code Machines The two practice sets which are described in this chapter are of most value when you have someone with whom to practice. Automatic code machines are not recommended to anyone who can possibly obtain a companion with whom to practice, someone who is also interested in learning the code. If you are unable to enlist a code partner and have to practice by yourself, the best way to get receiving practice is by the use of a tape machine (automatic code sending machine) with several practice tapes. Or you can use a set of phonograph code practice records. The records are of use only if you have a phonograph whose turntable speed is readily adjustable. The tape matchine can be rented by the month for a reasonable fee.

Once you can copy about 10 w.p.m. you can also get receiving practice by listening to slow sending stations on your receiver. Many amateur stations send slowly particularly when working far distant stations. When receiving conditions are particularly poor many commercial stations also send slowly, sometimes repeating every word. Until you can copy around 10 w.p.m. your receiver isn't much use, and either another operator or a machine or records are necessary for getting receiving practice after you have once memorized the code.

Code Practice Sets If you don't feel too foolish doing it, you can secure a measure of code practice with the help of a partner by sending "dit-dah" messages to each other while riding to work, eating lunch, etc. It is better, however, to use a buzzer or code practice oscillator in conjunction with a regular telegraph key.

As a good key may be considered an investment it is wise to make a well-made key your first purchase. Regardless of what type code practice set you use, you will need a key, and later on you will need one to key your transmitter. If you get a good key to begin with, you won't have to buy another one later.

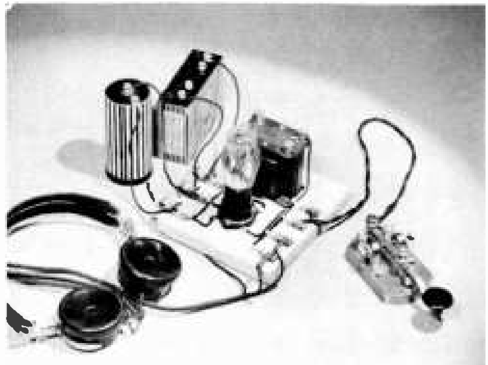


Figure 7.
THE CIRCUIT OF FIGURE 6 IS USED IN THIS BATTERY-OPERATED CODE PRACTICE OSCILLATOR.

A tube and audio transformer essentially comprise the oscillator. Fahnestock clips screwed to the base-board are used to make connection to the batteries, key, and phones.

The key should be rugged and have fairly heavy contacts. Not only will the key stand up better, but such a key will contribute to the "heavy" type of sending so desirable for radio work. Morse (telegraph) operators use a "light" style of sending and can send somewhat faster when using this light touch. But, in radio work static and interference are often present, and a slightly heavier dot is desirable. If you use a husky key, you will find yourself automatically sending in this manner.

To generate a tone simulating a code signal as heard on a receiver, either a mechanical buzzer or an audio oscillator may be used. Figure 5 shows a simple code-practice set using a buzzer which may be used directly simply by mounting the buzzer on a sounding board, or the buzzer may be used to feed from one to four pairs of conventional high-impedance phones.

An example of the audio-oscillator type of code-practice set is illustrated in figures 6 and 7. Any type of battery-filament tube may be used in this circuit to make up a satisfactory "howler" for code-practice work. The circuit is shown in figure 6.

CHAPTER TWO

Direct Current Circuits

All naturally occurring matter (excluding artificially produced radioactive substances) is made up of 92 fundamental constituents called *elements*. These elements can exist either in the free state such as iron, oxygen, carbon, copper, tungsten, and aluminum, or in chemical unions commonly called *compounds*. The smallest unit which still retains all the original characteristics of an element is the *atom*.

Combinations of atoms, or subdivisions of compounds, result in another fundamental unit, the *molecule*. The molecule is the smallest unit of any compound. All reactive elements when in the gaseous state also exist in the molecular form, made up of two or more atoms. The nonreactive or noble gaseous elements helium, neon, argon, krypton, xenon, and radon are the only gaseous elements that ever exist in a stable monatomic state at ordinary temperatures.

The Atom An atom is an extremely small unit of matter — there are literally billions of them making up so small a piece of material as a speck of dust. But to understand the basic theory of electricity and hence of radio, we must go further and divide the atom into its main components, a positively charged nucleus and a cloud of negatively charged particles that surround the nucleus. These particles, swirling around the nucleus in elliptical orbits at an incredible rate of speed, are called orbital electrons.

It is upon the behavior of these electrons, when freed from the atom, that depends the study of electricity and radio, as well as allied sciences. Actually it is possible to subdivide the nucleus of the atom into a dozen or so different particles, but this further subdivision can be left to quantum mechanics and atomic physics. As far as the study of electronics is concerned it is only necessary for the reader to think of the normal atom as being composed of a nucleus having a net positive charge that is exactly neutralized by the one or more orbital electrons surrounding it.

The atoms of different elements differ in respect to the charge on the positive nucleus and in the number of electrons revolving around this charge. They range all the way from hydrogen, having a net charge of one on the nucleus and one orbital electron, to uranium with a net charge of 92 and 92 orbital electrons. The number of orbital electrons is called the *atomic number* of the element.

From the above it must not be thought that the electrons revolve in a haphazard manner around the nucleus. Rather, the electrons in an element having a large atomic number are grouped into "shells" having a definite number of electrons. The only atoms in which these shells are completely filled are those of the inert or noble gases mentioned before; all other elements have one or more uncompleted shells of electrons. If the uncompleted shell is nearly empty, the element is *metallic* in character, being most metallic when there is only one electron in the outer shell. If the incomplete shell lacks only one or two electrons, the element is usually *non-metallic*. Elements with a shell about half completed will exhibit both non-metallic and metallic character; carbon, silicon, germanium, and arsenic are examples of this type of element.

In metallic elements these outer-shell electrons are rather loosely held. Consequently, there is a continuous helter-skelter movement of these electrons and a continual shifting from one atom to another. The electrons which move about in a substance are called *free electrons,* and it is the ability of these electrons to drift from atom to atom which makes possible the *electric current.*

If the free electrons are numerous and loosely held, the element is a good conductor. On the other hand, if there are few free electrons, as is the case when the electrons in an outer shell are tightly held, the element is a poor conductor. If there are virtually no free electrons, the element is a good insulator.

2-1 Fundamental Electrical Units and Relationships

Electromotive Force: Potential Difference The free electrons in a conductor move constantly about and change their position in a haphazard manner. To produce a drift of electrons or *electric current* along a wire it is necessary that there be a difference in pressure or *potential* between the two ends of the wire. This *potential difference* can be produced by connecting a source of *electrical potential* to the ends of the wire.

As will be explained later, there is an excess of electrons at the negative terminal of a battery and a deficiency of electrons at the positive terminal, due to chemical action. When the battery is connected to the wire, the deficient atoms at the positive terminal attract free electrons from the wire in order to become neutral. The attracting of electrons continues through the wire, and finally the excess electrons at the negative terminal of the battery are attracted by the positively charged atoms at the end of the wire. Other sources of electrical potential (in addition to a battery) are: an electrical generator (dynamo), a thermocouple, an electrostatic generator (static machine), a photoelectric cell, and a crystal or piezoelectric generator.

Thus it is seen that a potential difference is the result of a difference in the number of electrons between the two (or more) points in question. The force or pressure due to a potential difference is termed the *electromotive force,* usually abbreviated *e.m.f.* or *E.M.F.* It is expressed in units called *volts.*

It should be noted that for there to be a potential difference between two bodies or points it is not necessary that one have a positive charge and the other a negative charge. If two bodies each have a negative charge, but one more negative than the other, the one with the lesser negative charge will act as though it were positively charged *with respect to the other body.* It is the *algebraic* potential difference that determines the force with which electrons are attracted or repulsed, the potential of the earth being taken as the zero reference point.

The Electric Current The flow of electrons along a conductor due to the application of an electromotive force constitutes an electric current. This drift is in addition to the irregular movements of the electrons. However, it must not be thought that each free electron travels from one end of the circuit to the other. On the contrary, each free electron travels only a short distance before colliding with an atom; this collision generally knocking off one or more electrons from the atom, which in turn move a short distance and collide with other atoms, knocking off other electrons. Thus, in the general drift of electrons along a wire carrying an electric current, each electron travels only a short distance and the excess of electrons at one end and the deficiency at the other are balanced by the source of the e.m.f. When this source is removed the state of normalcy returns; there is still the rapid interchange of free electrons between atoms, but there is no general trend or "net movement" in either one direction or the other.

Ampere and Coulomb There are two units of measurement associated with current, and they are often confused. The *rate of flow* of electricity is stated in *amperes.* The unit of *quantity* is the *coulomb.* A coulomb is equal to 6.28×10^{18} electrons, and when this quantity of electrons flows by a given point in every second, a current of one ampere is said to be flowing. An ampere is equal to one coulomb per second; a coulomb is, conversely, equal to one ampere-second. Thus we see that *coulomb* indicates *amount,* and *ampere* indicates *rate of flow* of electric current.

Many textbooks speak of current flow as being from the positive terminal of the e.m.f. source through the conductor to the negative terminal. Nevertheless, it has long been an established fact that the current flow in a metallic conductor is the *electronic* flow from the negative terminal of the source of voltage through the conductor to the positive terminal. This is easily seen from a study of the foregoing explanation of the subject. The only exceptions to the electronic direction of flow occur in gaseous and electrolytic conductors where the flow of positive ions toward the cathode or negative electrode constitutes a positive flow in the opposite direction to the electronic flow. (An ion is an atom, molecule, or particle which either lacks one or more electrons, or else has an excess of one or more electrons.)

In radio work the terms "electron flow" and "current" are becoming accepted as being synonymous, but the older terminology is still accepted in the electrical (industrial) field. Because of the confusion this sometimes causes, it is often safer to refer to the direction of electron flow rather than to the direction of the "current." Since electron flow consists actually of a passage of *negative* charges, current flow and *algebraic* electron flow do pass in the same direction.

Resistance The flow of current in a material depends upon the ease with which electrons can be detached from the atoms of the material and upon its molecular structure. In other words, the easier it is to detach electrons from the atoms the more free electrons there will be to contribute to the flow of current, and the fewer collisions that occur between free electrons and atoms the greater will be the total electron flow.

The opposition to a steady electron flow is called the *resistance* of a material, and is one of its physical properties.

The unit of resistance is the *ohm.* Every substance has a *specific resistance,* usually expressed as *ohms per mil-foot,* which is determined by the material's molecular structure and temperature. A mil-foot is a piece of material one circular mil in area and one foot long. Another measure of resistivity frequently used is expressed in the units *microhms per centimeter cube.* The resistance of a uniform

TABLE OF RESISTIVITY

Material	Resistivity in Ohms per Circular Mil-Foot	Temp. Coeff. of resistance per °C at 20° C.
Aluminum	17	0.0049
Brass	45	0.003 to 0.007
Cadmium	46	0.0038
Chromium	16	0.00
Copper	10.4	0.0039
Iron	59	0.006
Silver	9.8	0.004
Zinc	36	0.0035
Nichrome	650	0.0002
Constantan	295	0.00001
Manganin	290	0.00001
Monel	255	0.0019

length of a given substance is directly proportional to its length and specific resistance, and inversely proportional to its cross-sectional area. A wire with a certain resistance for a given length will have twice as much resistance if the length of the wire is doubled. For a given length, doubling the cross-sectional area of the wire will *halve* the resistance, while doubling the *diameter* will reduce the resistance to *one fourth.* This is true since the cross-sectional area of a wire varies as the square of the diameter. The relationship between the resistance and the linear dimensions of a conductor may be expressed by the following equation:

$$R = \frac{r\, l}{A}$$

Where

R = resistance in ohms
r = resistivity in *ohms per mil-foot*
l = length of conductor in feet
A = cross-sectional area in circular mils

The resistance also depends upon temperature, increasing with increases in temperature for most substances (including most metals), due to increased electron acceleration and hence a greater number of impacts between electrons and atoms. However, in the case of some substances such as carbon and glass the temperature coefficient is negative, which means that the resistance decreases as the temperature increases. This is also true of electrolytes. The temperature may be raised by the external application of heat, or by the flow of the current itself. In the latter case, the temperature is raised by the heat generated when the electrons and atoms collide.

Conductors and Insulators In the molecular structure of many materials such as glass, porcelain, and mica all elec-

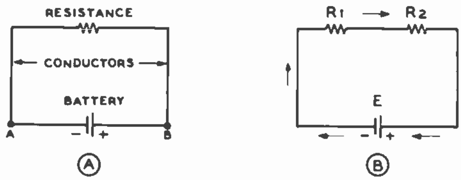


Figure 1.
SIMPLE SERIES CIRCUITS.
At (A) the battery is in series with a single resistor. At (B) the battery is in series with two resistors, the resistors themselves being in series. The arrows indicate the direction of electron flow.

trons are tightly held within their orbits and there are comparatively few free electrons. This type of substance will conduct an electric current only with great difficulty and is known as an *insulator*. An insulator is said to have a high electrical *resistance*.

On the other hand, materials that have a large number of free electrons are known as *conductors*. Most metals, those elements which have only one or two electrons in their outer shell, are good conductors. Silver, copper, and aluminum, in that order, are the best of the common metals used as conductors and are said to have the greatest *conductivity*, or lowest resistance to the flow of an electric current.

Fundamental Electrical Units These units are the *volt*, the *ampere*, and the *ohm*. They were mentioned in the preceding paragraphs, but were not completely defined in terms of fixed, known quantities.

The fundamental unit of *current*, or *rate of flow* of electricity is the ampere. A current of one ampere will deposit silver from a specified solution of silver nitrate at a rate of 1.118 milligrams per second.

The international standard for the ohm is the resistance offered by a uniform column of mercury at 0° C., 14.4521 grams in mass, of constant cross-sectional area and 106.300 centimeters in length. The expression *megohm* (1,000,000 ohms) is also sometimes used when speaking of very large values of resistance.

A volt is the e.m.f. that will produce a current of one ampere through a resistance of one ohm. The standard of electromotive force is the Weston cell which at 20° C. has a potential of 1.0183 volts across its terminals. This cell is used only for reference purposes in a bridge circuit, since only an infinitesimal amount of current may be drawn from it without disturbing its characteristics.

Ohm's Law The relationship between the electromotive force (voltage), the flow of current (amperes), and the resistance which impedes the flow of current (ohms), is very clearly expressed in a simple but highly valuable law known as *Ohm's law*. This law states that *the current in amperes is equal to the voltage in volts divided by the resistance in ohms.* Expressed as an equation:

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

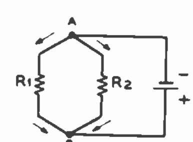
If the voltage (E) and resistance (R) are known, the current (I) can be readily found. If the voltage and current are known, and the resistance is unknown, the resistance (R) is equal to $\frac{E}{I}$. When the voltage is the unknown quantity, it can be found by multiplying $I \times R$. These three equations are all secured from the original by simple transposition. The expressions are here repeated for quick reference:

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

where I is the current in amperes,
R is the resistance in ohms,
E is the electromotive force in volts.

Applications of Ohm's Law All electrical circuits fall into one of three classes: series circuits, parallel circuits, and series-parallel circuits. A series circuit is one in which the current flows in a single continuous path and is of the same value at every

Figure 2.
SIMPLE PARALLEL CIRCUIT.



The two resistors R_1 and R_2 are said to be in parallel since the flow of current is offered two parallel paths. An electron leaving point A will pass either through R_1 or R_2, but not through both, to reach the positive terminal of the battery. If a large number of electrons are considered, the greater number will pass through whichever of the two resistors has the lower resistance.

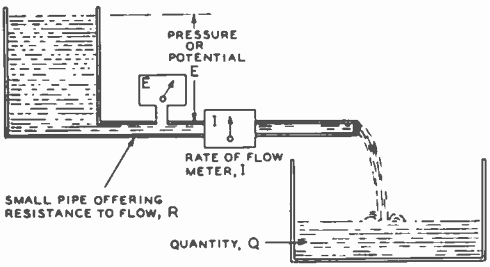


Figure 3.
WATER ANALOGY OF CURRENT AND POTENTIAL.
The instrument E indicates the water pressure, which is equivalent to potential or voltage. The instrument I indicates the rate-of-flow, which is equivalent to current. The quantity of water which has passed through the system is equivalent to the quantity of electrons which has passed through a circuit, usually expressed in coulombs.

point in the circuit. In a parallel circuit there are two or more current paths between two points in the circuit, as shown in figure 2. Here the current divides at A, part going through R_1 and part through R_2, and combines at B to return to the battery. Figure 4 shows a series-parallel circuit. There are two paths between points A and B as in the parallel circuit, and in addition there are two resistances in series in each branch of the parallel combination. Two other examples of series-parallel arrangements appear in figure 5. The way in which the current splits to flow through the parallel branches is shown by the arrows.

In every circuit, each of the parts has some resistance: the batteries or generator, the connecting conductors, and the apparatus itself. Thus, if each part has some resistance, no matter how little, and a current is flowing through it, there will be a voltage drop across it. In other words, there will be a potential difference between the two ends of the circuit element in question. This drop in voltage is equal to the product of the current and the resistance, hence it is called the *IR* drop.

The source of voltage has an *internal* resistance, and when connected into a circuit so that current flows, there will be an IR drop in the source just as in every other part of the circuit. Thus, if the terminal voltage of the source could be measured in a way that would cause no current to flow, it would be found to be more than the voltage measured when a current flows by the amount of the IR drop in the source. The voltage measured with no current flowing is termed the *no load* voltage; that measured with current flowing is the *load* voltage. It is apparent that a voltage source having a low internal resistance is most desirable, in order that the internal IR drop will be as small as possible, thereby making the load voltage more nearly equal to the no load voltage.

Resistances in Series The current flowing in a series circuit is equal to the voltage impressed divided by the *total* resistance across which the voltage is impressed. Since the same current flows through every part of the circuit, it is merely necessary to add all the individual resistances to obtain the total resistance. Expressed as a formula:

$$R_{total} = R_1 + R_2 + R_3 + \ldots + R_N .$$

Of course, if the resistances happened to be all of the same value, the total resistance would be the resistance of one multiplied by the number in the circuit.

Resistances in Parallel Consider two resistors, one of 100 ohms and one of 10 ohms, connected in parallel as in figure 2, with a voltage of 10 volts applied across the combination. The same voltage is across each resistor, so the current through each can be easily calculated.

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

$$E = 10 \text{ volts}$$
$$R = 100 \text{ ohms}$$

$$I = \frac{10}{100} = 0.1 \text{ ampere}$$

$$E = 10 \text{ volts}$$
$$R = 10 \text{ ohms}$$

$$I = \frac{10}{10} = 1.0 \text{ ampere}$$

Until it divides at A, the entire current of 1.1 amperes is flowing through the conductor from the battery, and again from B through the conductor to the battery. Since this is more current than flows through the smaller resistor it is evident that the resistance of the parallel combination must be less than 10 ohms, the resistance of the smaller resistor. We can find this value by applying Ohm's law.

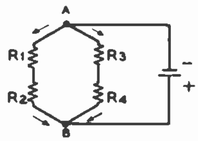


Figure 4. SERIES-PARALLEL CIRCUIT.

In this type of circuit the resistors are arranged in series groups, and these seriesed groups are then placed in parallel.

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

$$E = 10 \text{ volts}$$
$$I = 1.1 \text{ amperes}$$

$$R = \frac{10}{1.1} = 9.09 \text{ ohms}$$

The resistance of the parallel combination is 9.09 ohms.

Mathematically, we can derive a simple formula for finding the effective resistance of two resistors connected in parallel.
This formula is:

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

where R is the unknown resistance,
R_1 is the resistance of the first resistor,
R_2 is the resistance of the second resistor.

If the effective value required is known, and it is desired to connect one unknown resistor in parallel with one of known value, the following transposition of the above formula will simplify the problem of obtaining the unknown value:

$$R_2 = \frac{R_1 \times R}{R_1 - R}$$

where R is the effective value required,
R_1 is the known resistor,
R_2 is the value of the unknown resistance necessary to give R when in parallel with R_1.

The resultant value of placing a number of unlike resistors in parallel is equal to the reciprocal of the sum of the reciprocals of the various resistors. This can be expressed as:

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

The effective value of placing any number of unlike resistors in parallel can be determined from the above formula. However, it is commonly used only when there are three or more resistors under consideration, since the simplified formula given before is more convenient when only two resistors are being used.

When two or more resistors of the same value are placed in parallel, the effective resistance of the paralleled resistors is equal to the value of one of the resistors divided by the number of resistors in parallel.

The effective value of resistance of two or more resistors connected in parallel is *always* less than the value of the lowest resistance in the combination. It is well to bear this simple rule in mind, as it will assist greatly in approximating the value of paralleled resistors.

Resistors in Series-Parallel To find the total resistance of several resistors connected in series-parallel, it is usually easiest to apply either the formula for series resistors or the parallel resistor formula first, in order to reduce the original arrangement to a simpler one. For instance, in figure 4 the series resistors should be added in each branch, then there will be but two resistors in parallel to be calculated. Similarly in figure 6, although here there will be three parallel resistors after adding the series resistors in each branch. In figure 5 the paralleled resistors should be reduced to the equivalent series value, and then the series resistance values can be added.

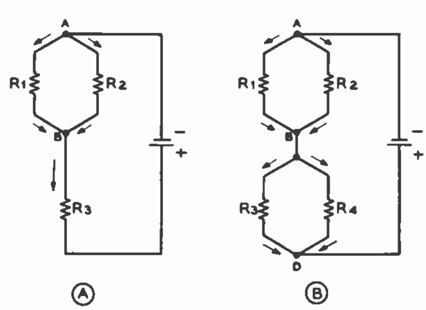


Figure 5. OTHER COMMON SERIES-PARALLEL CIRCUITS.

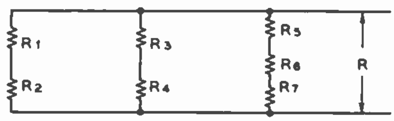


Figure 6.
ANOTHER TYPE OF SERIES-PARALLEL CIRCUIT.

Resistances in series-parallel can be solved by combining the series and parallel formulas into one similar to the following (refer to figure 6):

$$R = \frac{1}{\frac{1}{R_1 + R_2} + \frac{1}{R_3 + R_4} + \frac{1}{R_5 + R_6 + R_7}}$$

Voltage Dividers A voltage divider is exactly what its name implies: a resistor or a series of resistors connected across a source of voltage from which various lesser values of voltage may be obtained by connection to various points along the resistor.

A voltage divider serves a most useful purpose in a radio receiver, transmitter or amplifier, because it offers a simple means of obtaining plate, screen, and bias voltages of different values from a common power supply source. It may also be used to obtain very low voltages of the order of .01 to .001 volt with a high degree of accuracy, even though a means of measuring such voltages is lacking. The procedure for making these measurements can best be given in the following example.

Assume that an accurately calibrated voltmeter reading from 0 to 150 volts is available, and that the source of voltage is exactly 100 volts. This 100 volts is then impressed through a resistance of exactly 1,000 ohms. It will, then, be found that the voltage along various points on the resistor, with respect to the grounded end, is exactly proportional to the resistance at that point. From Ohm's law, the current would be 0.1 ampere; this current remains unchanged since the original value of resistance (1,000 ohms) and the voltage source (100 volts) are unchanged. Thus, at a 500-ohm point on the resistor (half its entire resistance), the voltage will likewise be halved or reduced to 50 volts.

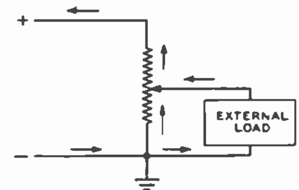


Figure 7.
SIMPLE VOLTAGE DIVIDER CIRCUIT.
The arrows indicate the manner in which the current flow divides between the voltage divider itself and the external load circuit.

The equation (E = I × R) gives the proof: E = 500 × 0.1 = 50. At the point of 250 ohms on the resistor, the voltage will be one-fourth the total value, or 25 volts (E = 250 × 0.1 = 25). Continuing with this process, a point can be found where the resistance measures exactly 1 ohm and where the voltage equals 0.1 volt. It is, therefore, obvious that if the original source of voltage and the resistance can be measured, it is a simple matter to predetermine the voltage at any point along the resistor, provided that the current remains constant, and provided that no current is taken from the tap-on point unless this current is taken into consideration.

Figuring Voltage Dividers Proper design of a voltage divider for any type of radio equipment is a relatively simple matter. The first consideration is the amount of "bleeder current" to be drawn. In addition, it is also necessary that the desired voltage and the exact current at each tap on the voltage divider be known.

Figure 7 illustrates the flow of current in a simple voltage divider and load circuit. The light arrows indicate the flow of bleeder current, while the heavy arrows indicate the flow of the load current. The design of a combined bleeder resistor and voltage divider, such as is commonly used in radio equipment, is illustrated in the following example.

A power supply delivers 300 volts and is conservatively rated to supply all needed current for the receiver and still allow a bleeder current of 10 milliamperes. The following voltages are wanted: 75 volts at 2 milliamperes for the detector tube, 100 volts at 5 milliamperes for the screens of the tubes, and 250 volts at 20 milliamperes for the plates of

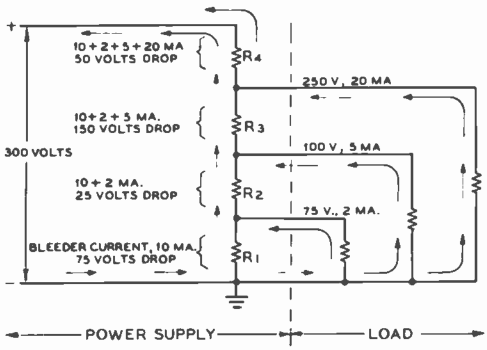


Figure 8.
MORE COMPLEX VOLTAGE DIVIDER.
The method for computing the values of the resistors is discussed in the accompanying text.

the tubes. The required voltage drop across R_1 is 75 volts, across R_2 25 volts, across R_3 150 volts, and across R_4 it is 50 volts. These values are shown in the diagram of figure 8. The respective current values are also indicated. Apply Ohm's law:

$$R_1 = \frac{E}{I} = \frac{75}{.01} = 7{,}500 \text{ ohms.}$$

$$R_2 = \frac{E}{I} = \frac{25}{.012} = 2{,}083 \text{ ohms.}$$

$$R_3 = \frac{E}{I} = \frac{150}{.017} = 8{,}823 \text{ ohms.}$$

$$R_4 = \frac{E}{I} = \frac{50}{.037} = 1{,}351 \text{ ohms.}$$

$$R_{Total} = 7{,}500 + 2{,}083 + 8{,}823 + 1{,}351 = 19{,}757 \text{ ohms.}$$

A 20,000-ohm resistor with three sliding taps will be of the approximately correct size, and would ordinarily be used because of the difficulty in securing four separate resistors of the exact odd values indicated, and because no adjustment would be possible to compensate for any slight error in estimating the probable currents through the various taps.

When the sliders on the resistor once are set to the proper point, as in the above example, the voltages will remain constant at the values shown as long as the current remains a constant value.

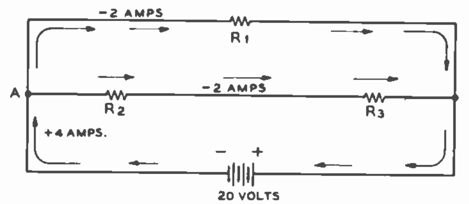


Figure 9.
ILLUSTRATING KIRCHHOFF'S FIRST LAW.
The current flowing toward point "A" is equal to the current flowing away from point "A".

Disadvantages of Voltage Dividers One of the serious disadvantages of the voltage divider becomes evident when the current drawn from one of the taps changes. It is obvious that the voltage drops are interdependent and, in turn, the individual drops are in proportion to the current which flows through the respective sections of the divider resistor. The only remedy lies in providing a heavy steady bleeder current in order to make the individual currents so small a part of the total current that any change in current will result in only a slight change in voltage. This can seldom be realized in practice because of the excessive values of bleeder current which would be required.

Kirchhoff's Laws Ohm's law is all that is necessary to calculate the values in simple circuits, such as the preceding examples; but in more complex problems, involving several loops or more than one voltage in the same closed circuit, the use of *Kirchhoff's laws* will greatly simplify the calculations. These laws are merely rules for applying Ohm's law.

Kirchhoff's first law is concerned with net current to a point in a circuit and states that:

At any point in a circuit the current flowing toward the point is equal to the current flowing away from the point.

Stated in another way: if currents flowing to the point are considered positive, and those flowing from the point are considered negative, the sum of all currents flowing toward and away from the point — taking signs into account — is equal to zero. Such a sum is known as an *algebraic sum;* so the law can be stated in another way, as: *The algebraic sum of all currents entering and leaving a point is zero.*

Figure 9 illustrates this first law. Since the effective resistance of the network of resistors is 5 ohms, it can be seen that 4 amperes flow toward point A, and 2 amperes flow away through the two 5-ohm resistors in series, while the remaining 2 amperes flow away through the 10-ohm resistor. Thus, there are 4 amperes flowing to point A and 4 amperes flowing away from the point. If R is the effective resistance of the network (5 ohms), $R_1 = 10$ ohms, $R_2 = 5$ ohms, $R_3 = 5$ ohms, and $E = 20$ volts, we can set up the following equation:

$$\frac{E}{R} - \frac{E}{R_1} - \frac{E}{R_2 + R_3} = 0$$

$$\frac{20}{5} - \frac{20}{10} - \frac{20}{5 + 5} = 0$$

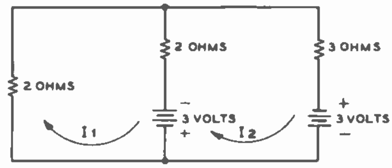
$$4 - 2 - 2 = 0$$

Kirchhoff's second law is concerned with net voltage drop around a closed loop in a circuit and states that:

In any closed path or loop in a circuit the sum of the IR drops must equal the sum of the applied e.m.f.'s.

The second law also may conveniently be stated in terms of an algebraic sum as: *The algebraic sum of all voltage drops around a closed path or loop in a circuit is zero.* The applied e.m.f.'s (voltages) are considered positive, while IR drops taken in the direction of current flow (including the internal drop of the sources of voltage) are considered negative.

Figure 10 shows an example of the application of Kirchhoff's laws to a comparatively simple circuit consisting of three resistors and two batteries. First assume an arbitrary direction of current flow in each closed loop of the circuit. Next draw an arrow to indicate the direction of current flow assumed so that you will not forget. Then equate the sum of all IR drops plus battery drops around each loop to zero. You will need one equation for each unknown to be determined. Then solve the equations for the unknown currents in the general manner indicated in figure 10. If the answer comes out positive the direction of current flow you originally assumed was correct. If the answer comes out negative, the current flow is in the opposite direction to the arrow which was drawn originally. This is illustrated in the example of figure 10 where the direction of flow of I_1 is opposite to the direction assumed in the sketch.



1. SET VOLTAGE DROPS AROUND EACH LOOP EQUAL TO ZERO.

   $I_1 2_{(OHMS)} + 2(I_1 - I_2) + 3 = 0$ (FIRST LOOP)

   $-6 + 2(I_2 - I_1) + 3I_2 = 0$ (SECOND LOOP)

2. SIMPLIFY

   $2I_1 + 2I_1 - 2I_2 + 3 = 0$ $\qquad$ $2I_2 - 2I_1 + 3I_2 - 6 = 0$

   $\frac{4I_1 + 3}{2} = I_2$ $\qquad$ $5I_2 - 2I_1 - 6 = 0$

   $\frac{2I_1 + 6}{5} = I_2$

3. EQUATE

   $$\frac{4I_1 + 3}{2} = \frac{2I_1 + 6}{5}$$

4. SIMPLIFY

   $20I_1 + 15 = 4I_1 + 12$

   $I_1 = -\frac{3}{16}$ AMPERE

5. RE-SUBSTITUTE

   $$I_2 = \frac{-\frac{12}{16} + 3}{2} = \frac{2\frac{1}{4}}{2} = 1\frac{1}{8} \text{ AMPERE}$$

Figure 10.
ILLUSTRATING KIRCHHOFF'S SECOND LAW.
The voltage drop around any closed loop in a network is equal to zero.

Power in Resistive Circuits In order to cause electrons to flow through a conductor, constituting a current flow, it is necessary to apply an electromotive force (voltage) across the circuit. Less power is expended in creating a small current flow through a given resistance than in creating a large one; so it is necessary to have a unit of power as a reference.

The unit of electrical power is the *watt,* which is the rate of energy consumption when an e.m.f. of 1 volt forces a current of 1 ampere through a circuit. The power in a resistive circuit is equal to the product of the voltage applied across, and the current flowing in, a given circuit. Hence: P (watts) = E (volts) × I (amperes).

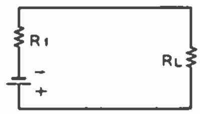


Figure 11. MATCHING OF RESISTANCES.

To deliver the greatest amount of power to the load, the load resistance R_L should be equal to the internal resistance of the battery R_I.

Since it is often convenient to express power in terms of the resistance of the circuit and the current flowing through it, a substitution of IR for E (E = IR) in the above formula gives: $P = IR \times I$ or $P = I^2R$. In terms of voltage and resistance, $P = E^2/R$. Here, $I = E/R$ and when this is substituted for I the original formula becomes $P = E \times E/R$, or $P = E^2/R$. To repeat these three expressions:

$$P = EI,\ P = I^2R,\ \text{and}\ P = E^2/R,$$

where P is the power in watts,
E is the electromotive force in volts, and
I is the current in amperes.

To apply the above equations to a typical problem: The voltage drop across a cathode resistor in a power amplifier stage is 50 volts; the plate current flowing through the resistor is 150 milliamperes. The number of watts the resistor will be required to dissipate is found from the formula: $P = EI$, or $50 \times .150 = 7.5$ watts (.150 amperes is equal to 150 milliamperes). From the foregoing it is seen that a 7.5-watt resistor will safely carry the required current, yet a 10- or 20-watt resistor would ordinarily be used to provide a safety factor.

In another problem, the conditions being similar to those above, but with the resistance and current being the *known* factors, the solution is obtained as follows: $P = I^2R = .0225 \times 333.33 = 7.5$. If only the voltage and resistance are known, $P = E^2/R = 2500/333.33 = 7.5$ watts. It is seen that all three equations give the same results; the selection of the particular equation depends only upon the known factors.

Power, Energy, and Work It is important to remember that power (expressed in watts, horsepower, etc.). represents the *rate* of energy consumption or the *rate* of doing work. But when we pay our electric bill to the power company we have purchased a specific *amount* of *energy* or *work* expressed in the common units of *kilowatt-hours*. Thus *rate* of energy consumption (watts or kilowatts) multiplied by *time* (seconds, minutes or hours) gives us total energy or work. Other units of energy are the watt-second, BTU, calorie, erg, and joule.

Heating Effect Heat is generated when a source of voltage causes a current to flow through a resistor (or, for that matter, through any conductor). As explained earlier, this is due to the fact that heat is given off when free electrons collide with the atoms of the material. More heat is generated in high resistance materials than in those of low resistance, since the free electrons must strike the atoms harder to knock off other electrons. As the heating effect is a function of the current flowing and the resistance of the circuit, the power expended in heat is given by the second formula: $P = I^2R$.

2-2 Electrostatics — Capacitors

Electrical energy can be stored in an electrostatic field. A device capable of storing energy in such a field is called *capacitor* (in earlier usage the term *condenser* was frequently used but the IRE standards call for the use of capacitor instead of condenser) and is said to have a certain *capacitance*. The *energy* stored in an electrostatic field is expressed in *joules* (watt seconds) and is equal to $CE^2/2$, where C is the capacitance in *farads* (a unit of capacitance to be discussed) and E is the potential in volts. The *charge* is equal to CE, the charge being expressed in coulombs.

Capacitance and Capacitors Two metallic plates separated from each other by a thin layer of insulating material (called a *dielectric*, in this case), become a *capacitor*. When a source of d-c potential is momentarily applied across these plates, they may be said to become charged. If the same two plates are then joined together momentarily by means of a wire, the capacitor will *discharge*.

When the potential was first applied, electrons immediately flowed from one plate to the other through the battery or such source of d-c potential as was applied to the capacitor plates. However, the circuit from plate to plate in the capacitor was *incomplete* (the two plates being separated by an insulator) and thus the electron flow ceased, meanwhile establishing a shortage of electrons on one plate and a surplus of electrons on the other.

Remember that when a deficiency of electrons exists at one end of a conductor, there is always a tendency for the electrons to move about in such a manner as to re-establish a state of balance. In the case of the capacitor herein discussed, the surplus quantity of electrons on one of the capacitor plates cannot move to the other plate because the circuit has been broken; that is, the battery or d-c potential was removed. This leaves the capacitor in a *charged* condition; the capacitor plate with the electron *deficiency* is *positively* charged, the other plate being *negative.*

In this condition, a considerable stress exists in the insulating material (dielectric) which separates the two capacitor plates, due to the mutual attraction of two unlike potentials on the plates. This stress is known as *electrostatic* energy, as contrasted with *electromagnetic* energy in the case of an inductor. This charge can also be called *potential energy* because it is capable of performing work when the charge is released through an external circuit.

In case it is difficult for the reader to understand why the charge is proportional to the voltage but the energy is proportional to the voltage squared, the following analogy may make things clear.

The charge represents a definite amount of electricity, a given number of electrons. The potential energy possessed by these electrons depends not only upon their number, but also upon their potential or voltage.

Compare the electrons to water, and two capacitors to standpipes, a 1 μfd. capacitor to a standpipe having a cross section of 1 square inch and a 2 μfd. capacitor to a standpipe having a cross section of 2 square inches. The charge will represent a given volume of water, as the "charge" simply indicates a certain number of electrons. Suppose the water is equal to 5 gallons.

Now the potential energy, or capacity for doing work, of the 5 gallons of water will be twice as great when confined to the 1 sq. in. standpipe as when confined to the 2 sq. in. standpipe. Yet the volume of water, or "charge" is the same in either case.

Likewise a 1 μfd. capacitor charged to 1000 volts possesses twice as much potential energy as does a 2 μfd. capacitor charged to 500 volts, though the *charge* (expressed in *coulombs:* $Q = CE$) is the same in either case.

The Unit of Capacitance: The Farad If the external circuit of the two capacitor plates is completed by joining the terminals together with a piece of wire, the electrons will rush immediately from one plate to the other through the external circuit and establish a state of equilibrium. This latter phenomenon explains the *discharge* of a capacitor. The amount of stored energy in a charged capacitor is dependent upon the charging potential, as well as a factor which takes into account the *size* of the plates, *dielectric thickness, nature* of the dielectric, and the *number* of plates. This factor, which is determined by the foregoing, is called the *capacitance* of a capacitor and is expressed in *farads.*

The farad is such a large unit of capacitance that it is rarely used in radio calculations, and the following more practical units have, therefore, been chosen:

1 *microfarad* = 1/1,000,000 *of a farad, or* .000001 *farad, or* 10^{-6} *farads.*

1 *micro-microfarad* = 1/1,000,000 *of a microfarad, or* .000001 *microfarad, or* 10^{-6} *microfarads.* μμf— pf

1 *micro-microfarad* = *one-millionth of one-millionth of a farad, or* 10^{-12} *farads.*

If the capacitance is to be expressed in *microfarads* in the equation given for *energy storage,* the factor C would then have to be divided by 1,000,000, thus:

$$\text{Stored energy in joules} = \frac{C \times E^2}{2 \times 1{,}000{,}000}$$

This storage of energy in a capacitor is one of its very important properties, particularly in those capacitors which are used in power supply filter circuits.

Dielectric Materials Although any substance which has the characteristics of a good insulator may be used as a dielectric material, commercially manufactured capacitors make use of dielectric materials which have been selected because their characteristics are particularly suited to the job at hand. Air

TABLE OF DIELECTRIC MATERIALS			
MATERIAL	DIELECTRIC CONSTANT 10 MC.	POWER FACTOR 10 MC.	SOFTENING POINT FAHRENHEIT
ANILINE-FORMALDEHYDE RESIN	3.4	0.004	260°
CASTOR OIL	4.67		
CELLULOSE ACETATE	3.7	0.04	180°
GLASS, WINDOW	6–8	POOR	2000°
GLASS, PYREX	4.5	0.02	
METHYL-METHACRYLATE—LUCITE	2.8	0.007	160°
MICA	5.4	0.0003	
MYCALEX, MYKROY	7.0	0.002	650°
PHENOL-FORMALDEHYDE, LOW-LOSS YELLOW	5.0	0.015	270°
PHENOL-FORMALDEHYDE BLACK BAKELITE	5.5	0.03	350°
PORCELAIN	7.0	0.005	2800°
POLYETHYLENE	2.25	0.0003	220°
POLYSTYRENE	2.55	0.0002	175°
QUARTZ, FUSED	4.2	0.0002	2600°
RUBBER, HARD-EBONITE	2.8	0.007	150°
STEATITE	6.1	0.003	2700°
SULFUR	3.8	0.003	236°
TITANIUM DIOXIDE	100-175	0.0006	2700°
TRANSFORMER OIL	2.2	0.003	
UREA-FORMALDEHYDE	5.0	0.05	260°
VINYL RESINS	4.0	0.02	200°
WOOD, MAPLE	4.4	POOR	

FIGURE 12.

is a very good dielectric material, but an air-spaced capacitor does not have a high capacitance since the dielectric constant of air is only slightly greater than one. A group of other commonly used dielectric materials is listed in figure 12.

Dielectric Constant The capacitance of a capacitor is determined by the thickness and nature of the dielectric material between plates. Certain materials offer a greater capacitance than others, depending upon their physical makeup and chemical constitution. This property is expressed by a constant K, called the dielectric constant.

Dielectric Breakdown If the charge becomes too great for a given thickness of a certain dielectric, the capacitor will break down, i.e., the dielectric will puncture. It is for this reason that capacitors are rated in the manner of the amount of voltage they will safely withstand as well as the capacitance in microfarads. This rating is commonly expressed as the *d-c working voltage.*

Calculation of Capacitance The capacitance of two parallel plates is given with good accuracy by the following formula:

$$C = 0.2248 \times K \times \frac{A}{t},$$

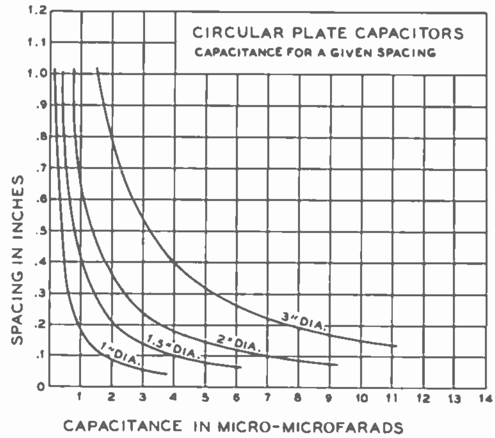


Figure 13.
Through the use of this chart it is possible to determine the required plate diameter (with the necessary spacing established by peak voltage considerations) for a circular-plate neutralizing capacitor. The capacitance given is for a dielectric of air and the spacing given is between adjacent faces of the two plates.

where C = capacitance in micro-microfarads,
K = dielectric constant of spacing material,
A = area of dielectric in square inches,
t = thickness of dielectric in inches.

This formula indicates that the capacitance is directly proportional to the area of the plates and inversely proportional to the thickness of the dielectric (spacing between the plates). This simply means that when the area of the plate is doubled, the spacing between plates remaining constant, the capacitance will be doubled. Also, if the area of the plates remains constant, and the plate spacing is doubled, the capacitance will be reduced to half.

The above equation also shows that capacitance is directly proportional to the dielectric constant of the spacing material. An air-spaced capacitor that has a capacitance of 100 $\mu\mu$fd. in air would have a capacitance of 467 $\mu\mu$fd. when immersed in castor oil, because the dielectric constant of castor oil is 4.67 times as great as the dielectric constant of air.

Where the area of the plates is definitely set, and when it is desired to know the spacing needed to secure a required capacitance,

$$t = \frac{A \times 0.2248 \times K}{C}$$

where all units are expressed just as in the preceding formula. This formula is not confined to capacitors having only square or rectangular plates, but also applies when the plates are circular in shape. The only change will be the calculation of the *area* of such circular plates; this area can be computed by squaring the *radius* of the plate, then multiplying by 3.1416, or "pi." Expressed as an equation:

$$A = 3.1416 \times r^2,$$

where r = radius in inches.

The capacitance of a multi-plate capacitor can be calculated by taking the capacitance of one section and multiplying this by the number of dielectric spaces. In such cases, however, the formula gives no consideration to the effects of edge capacitance; so the capacitance as calculated will not be entirely accurate. These additional capacitances will be but a small part of the effective total capacitance, particularly when the plates are reasonably large and thin, and the final result will, therefore, be within practical limits of accuracy.

Capacitors in Parallel and in Series Equations for calculating capacitances of capacitors in *parallel* connections are the same as those for resistors in *series:*

$$C = C_1 + C_2, \text{ etc.}$$

Capacitors in *series* connection are calculated in the same manner as are resistors in *parallel* connection.

The formulas are repeated: (1) For two or more capacitors of *unequal* capacitance in series:

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

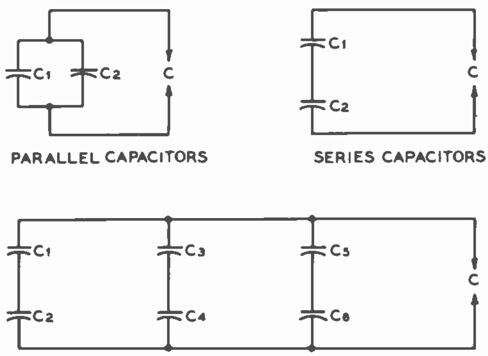
$$\text{or } \frac{1}{C} = \frac{1}{C_1} + \frac{1}{C_2} + \frac{1}{C_3}$$

(2) *Two* capacitors of *unequal* capacitance in series:

$$C = \frac{C_1 \times C_2}{C_1 + C_2}$$

(3) Three capacitors of *equal* capacitance in series:

$$C = \frac{C_1}{3}$$ where C_1 is the common capacitance.



PARALLEL CAPACITORS SERIES CAPACITORS

CAPACITORS IN SERIES-PARALLEL

Figure 14.
CAPACITORS IN SERIES, PARALLEL, AND SERIES-PARALLEL.

(4) Three or more capacitors of *equal* capacitance in series.

$$C = \frac{\text{Value of common capacitance}}{\text{Number of capacitors in series}}$$

(5) Six capacitors in series parallel:

$$C = \frac{1}{\frac{1}{C_1} + \frac{1}{C_2}} + \frac{1}{\frac{1}{C_3} + \frac{1}{C_4}} + \frac{1}{\frac{1}{C_5} + \frac{1}{C_6}}$$

Capacitors in A-C and D-C Circuits When a capacitor is connected into a direct-current circuit, it will block the d.c., or stop the flow of current. Beyond the initial movement of electrons during the period when the capacitor is being charged, there will be no flow of current because the circuit is effectively broken by the dielectric of the capacitor.

Strictly speaking, a very small current may actually flow because the dielectric of the capacitor may not be a perfect insulator. This minute current flow is the leakage current previously referred to and is dependent upon the internal d-c resistance of the capacitor. This leakage current is usually quite noticeable in most types of electrolytic capacitors.

When an alternating current is applied to a capacitor, the capacitor will charge and discharge a certain number of times per second in accordance with the frequency of the alternating voltage. The electron flow in the charge and discharge of a capacitor when an a-c potential is applied constitutes an alternating current, in effect. It is for this reason that a capacitor will pass an alternating current yet offer practically infinite opposition to a direct current. These two properties are repeatedly in evidence in a radio circuit.

Voltage Rating of Capacitors in Series Any good paper dielectric filter capacitor has such a high internal resistance (indicating a good dielectric) that the exact resistance will vary considerably from capacitor to capacitor even though they are made by the same manufacturer and are of the same rating. Thus, when 1000 volts d.c. is connected across two 1-μfd. 500-volt capacitors in series, the chances are that the voltage will divide unevenly and one capacitor will receive more than 500 volts and the other less than 500 volts.

Voltage Equalizing Resistors By connecting a half-megohm 1-watt carbon resistor across each capacitor, the voltage will be equalized because the resistors act as a voltage divider, and the internal resistances of the capacitors are so much higher (many megohms) that they have but little effect in disturbing the voltage divider balance.

Carbon resistors of the inexpensive type are not particularly accurate (not being designed for precision service); therefore it is advisable to check several on an accurate ohmmeter to find two that are as close as possible in resistance. The exact resistance is unimportant, just so it is the same for the two resistors used.

Capacitors in Series on A.C. When two capacitors are connected in series, *alternating* voltage pays no heed to the relatively high internal resistance of each capacitor, but divides across the capacitors in inverse proportion to the *capacitance*. Because, in addition to the d.c. across a capacitor in a filter or audio amplifier circuit there is usually an a-c or a-f voltage component, it is inadvisable to series-connect capacitors of unequal capacitance even if dividers are provided to keep the d.c. within the ratings of the individual capacitors.

For instance, if a 500-volt 1-μfd. capacitor is used in series with a 4-μfd. 500-volt capacitor across a 250-volt a-c supply, the 1-μfd. capacitor will have 200 volts a.c. across it and the 4-μfd. capacitor only 50 volts. An equalizing divider to do any good in this case would have to be of very low resistance because of the comparatively low impedance of the capacitors *to a.c.* Such a divider would draw excessive current and be impracticable.

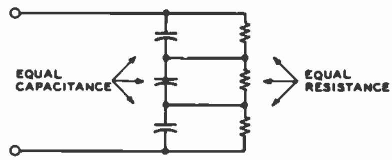


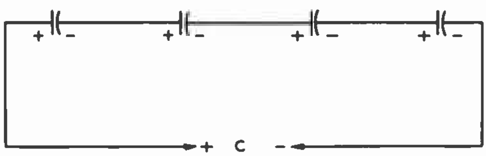
Figure 15.
SHOWING THE USE OF VOLTAGE EQUALIZING RESISTORS ACROSS CAPACITORS CONNECTED IN SERIES.

The safest rule to follow is to use only capacitors of the same capacitance and voltage rating and to install matched high resistance proportioning resistors across the various capacitors to equalize the d-c voltage drop across each capacitor. This holds regardless of how many capacitors are series-connected.

Electrolytic Capacitors Electrolytic capacitors use a very thin film of oxide as the dielectric, and are polarized; that is, they have a positive and a negative terminal which must be properly connected in a circuit; otherwise, the oxide will break down and the capacitor will overheat. The unit then will no longer be of service. When electrolytic capacitors are connected in series, the positive terminal is always connected to the positive lead of the power supply; the negative terminal of the capacitor connects to the positive terminal of the *next* capacitor in the series combination. The method of connection for electrolytic capacitors in series is shown in figure 16.

Similar electrolytic capacitors, of the same capacitance and made by the same manufacturer, have more nearly uniform internal resistance, though it still will vary considerably. However, the variation is not nearly as great as encountered in paper capacitors, and the lowest d-c voltage is across the weakest (leakiest) electrolytic capacitor of a series group.

As an electrolytic capacitor begins to show signs of breaking down from excessive volt-



POLARIZED CAPACITORS (ELECTROLYTIC) IN SERIES

Figure 16.

age, the leakage current goes up, which tends to heat the capacitor and aggravate the condition. However, when used in series with one or more others, the lower resistance (higher leakage current) tends to put less d-c voltage on the weakening capacitor and more on the remaining ones. Thus, the capacitor with the *lowest* leakage current, usually the *best* capacitor, has the highest voltage across it. For this reason, dividing resistors are not essential across series-connected electrolytic capacitors.

2-3 Magnetism and Electromagnetism

The common bar or horseshoe magnet is familiar to most people. The magnetic field which surrounds it causes the magnet to attract other magnetic materials, such as iron nails or tacks. Exactly the same kind of magnetic field is set up around any conductor carrying a current, but the field exists only while the current is flowing.

Magnetic Fields Before a potential, or voltage, is applied to a conductor there is no external field, because there is no general movement of the electrons in one direction. However, the electrons do progressively move along the conductor when an e.m.f. is applied, the direction of motion depending upon the polarity of the e.m.f. Since each electron has an electric field about it, the flow of electrons causes these fields to build up into a resultant external field which acts in a plane at right angles to the direction in which the current is flowing. This field is known as the *magnetic field*.

The magnetic field around a current-carrying conductor is illustrated in figure 17. The direction of this magnetic field depends entirely upon the direction of electron drift or current flow in the conductor. When the flow is toward the observer, the field about the conductor is clockwise; when the flow is away from the observer, the field is counter-clockwise. This is easily remembered if the left hand is clenched, with the thumb outstretched and pointing in the direction of electron flow. The fingers then indicate the direction of the magnetic field around the conductor.

Each electron adds its field to the total external magnetic field, so that the greater the number of electrons moving along the conductor, the stronger will be the resulting field.

One of the fundamental laws of magnetism is that *like poles repel one another and unlike poles attract one another.* This is true of current-carrying conductors as well as of permanent magnets. Thus, if two conductors are placed side by side and the current in each is flowing in the same direction, the magnetic fields will also be in the same direction and will combine to form a larger and stronger field. If the current flow in adjacent conductors is in opposite directions, the magnetic fields oppose each other and tend to cancel.

The magnetic field around a conductor may be considerably increased in strength by winding the wire into a coil. The field around each wire then combines with those of the adjacent turns to form a total field through the coil which is concentrated along the axis of the coil and behaves externally in a way similar to the field of a bar magnet.

If the left hand is held so that the thumb is outstretched and parallel to the axis of a coil, with the fingers curled to indicate the direction of electron flow around the turns of the coil, the thumb then points in the direction of the north pole of the magnetic field.

The Magnetic Circuit In the magnetic circuit, the units which correspond to current, voltage, and resistance in the electrical circuit are flux, magnetomotive force, and reluctance.

Flux; Flux Density As a current is made up of a drift of electrons, so is a magnetic field made up of lines of force, and the total number of lines of force in a given magnetic circuit is termed the *flux.* The flux depends upon the material, cross section, and length of the magnetic circuit, and it varies directly as the current flowing in the circuit. The unit of flux is the *maxwell,* and the symbol is the Greek letter ϕ (phi).

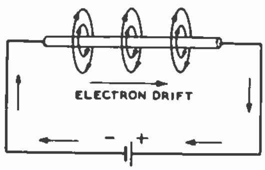


Figure 17.
LEFT-HAND RULE.
Showing the direction of the magnetic lines of force produced around a conductor carrying an electric current.

Flux density is the number of lines of force per unit area. It is expressed in *gauss* if the unit of area is the square centimeter (1 gauss = 1 line of force per square centimeter), or in *lines per square inch.* The symbol for flux density is B if it is expressed in gausses, or B if expressed in lines per square inch.

Magnetomotive Force The force which produces a flux in a magnetic circuit is called *magnetomotive force.* It is abbreviated m.m.f. and is designated by the letter F. The unit of magnetomotive force is the *gilbert,* which is equivalent to $1.26 \times \mathrm{NI}$, where N is the number of turns and I is the current flowing in the circuit in amperes.

The m.m.f. necessary to produce a given flux density is stated in gilberts per centimeter (*oersteds*) (H), or in ampere-turns per inch (H).

Reluctance Magnetic reluctance corresponds to electrical resistance, and is the property of a material that opposes the creation of a magnetic flux in the material. It is expressed in *rels,* and the symbol is the letter R. A material has a reluctance of 1 rel when an m.m.f. of 1 ampere-turn (NI) generates a flux of 1 line of force in it. Combinations of reluctances are treated the same as resistances in finding the total effective reluctance. The *specific reluctance* of any substance is its reluctance per unit volume.

Except for iron and its alloys, most common materials have a specific reluctance very nearly the same as that of a vacuum, which, for all practical purposes, may be considered the same as the specific reluctance of air.

Ohm's Law for Magnetic Circuits The relations between flux, magnetomotive force, and reluctance are exactly the same as the relations between current, voltage, and resistance in the electrical circuit. These can be stated as follows:

$$\phi = \frac{F}{R} \qquad R = \frac{F}{\phi} \qquad F = \phi R$$

where ϕ = flux, F = m.m.f., and R = reluctance.

Permeability Permeability expresses the ease with which a magnetic field may be set up in a material as compared with the effort required in the case of air. Iron, for example, has a permeability of around 2000 times that of air, which means that a given amount of magnetizing effect produced in an iron core by a current flowing through a coil of wire will produce 2000 times the *flux density* that the same magnetizing effect would produce in air. It may be expressed by the ratio B/H or B/H. In other words,

$$\mu = \frac{B}{H} \text{ or } \mu = \frac{\mathrm{B}}{\mathrm{H}}$$

where μ is the permeability, B is the flux density in gausses, B is the flux density in lines per square inch, H is the m.m.f. in gilberts per centimeter (oersteds), and H is the m.m.f. in ampere-turns per inch. These relations may also be stated as follows:

$$H = \frac{B}{\mu} \text{ or } \mathrm{H} = \frac{\mathrm{B}}{\mu}, \text{ and } B = H\mu \text{ or } \mathrm{B} = \mathrm{H}\mu$$

It can be seen from the foregoing that permeability is inversely proportional to the specific reluctance of a material.

Saturation Permeability is similar to *electric conductivity.* There is, however, one important difference: the permeability of magnetic materials is not independent of the magnetic current (flux) flowing through it, although electrical conductivity is substantially independent of the electric current in a wire. When the flux density of a magnetic conductor has been increased to the *saturation point,* a further increase in the magnetizing force will not produce a corresponding increase in flux density.

Calculations To simplify magnetic circuit calculations, a magnetization curve may be drawn for a given unit of material. Such a curve is termed a B-H curve, and is arrived at by experiment. B-H curves for most common magnetic materials are available in many reference books, so none will be given here.

Residual Magnetism; Retentivity The magnetism remaining in a material after the magnetizing force is removed is called *residual magnetism*. *Retentivity* is the property which causes a magnetic material to have residual magnetism after having been magnetized.

Hysteresis; Coercive Force *Hysteresis* is the characteristic of a magnetic system which causes a loss of power due to the fact that a negative magnetizing force must be applied to reduce the residual magnetism to zero. This negative force is termed *coercive force*. By "negative" magnetizing force is meant one which is of the opposite polarity with respect to the original magnetizing force. Hysteresis loss is apparent in transformers and chokes by the heating of the core.

Inductance If the switch shown in figure 17 is opened and closed, a pulsating direct current will be produced. When it is first closed, the current does not instantaneously rise to its maximum value, but builds up to it. While it is building up, the magnetic field is expanding around the conductor. Of course, this happens in a small fraction of a second. If the switch is then opened, the current stops and the magnetic field contracts quickly. This expanding and contracting field will induce a current in any other conductor that is part of a continuous circuit which it cuts. Such a field can be obtained in the way just mentioned by means of a vibrator interruptor, or by applying a.c. to the circuit in place of the battery. Varying the resistance of the circuit will also produce the same effect. This inducing of a current in a conductor due to a varying current in another conductor not in actual contact is called *electromagnetic induction*.

Self-induction If an alternating current flows through a coil the varying magnetic field around each turn cuts itself and the adjacent turn and *induces a voltage in the coil of opposite polarity to the applied e.m.f.* The amount of induced voltage depends upon the number of turns in the coil, the current flowing in the coil, and the number of lines of force threading the coil. The voltage so induced is known as a *counter-e.m.f.* or *back-e.m.f.*, and the effect is termed *self-induction*. When the applied voltage is building up, the counter-e.m.f. opposes the rise; when the applied voltage is decreasing, the counter-e.m.f. is of the same polarity and tends to maintain the current. Thus, it can be seen that self-induction tends to prevent any change in the current in the circuit.

The storage of energy in a magnetic field is expressed in *joules* and is equal to $(LI^2)/2$. (A joule is equal to 1 watt-second. L is defined immediately following.)

The Unit of Inductance; The Henry Inductance is usually denoted by the letter L, and is expressed in *henrys*. A coil has an inductance of 1 henry when a voltage of 1 volt is induced by a current change of 1 ampere per second. The henry, while commonly used in audio frequency circuits, is too large for reference to inductance coils such as those used in radio frequency circuits; *millihenry* or *microhenry* is more commonly used, in the following manner:

1 *henry* = 1,000 *millihenrys*, or 10^3 *millihenrys.*

1 *millihenry* = 1/1,000 *of a henry*, .001 *henry, or* 10^{-3} *henry.*

1 *microhenry* = 1/1,000,000 *of a henry, or* .000001 *henry, or* 10^{-6} *henry.*

1 *microhenry* = 1/1,000 *of a millihenry*, .001 *or* 10^{-3} *millihenrys.*

1,000 *microhenrys* = 1 *millihenry.*

Mutual Induction When one coil is near another, a varying current in one will produce a varying magnetic field which cuts the turns of the other coil, inducing a current in it. This induced current is also varying, and will therefore induce another current in the first coil. This reaction between two coupled circuits is called *mutual induction*, and can be calculated and expressed in henrys. The symbol for mutual inductance is M. Two circuits thus joined are said to be *inductively coupled*.

The magnitude of the mutual inductance

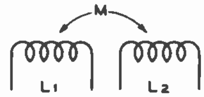


Figure 18.
MUTUAL INDUCTANCE.
The quantity M represents the mutual inductance between the two coils L_1 and L_2.

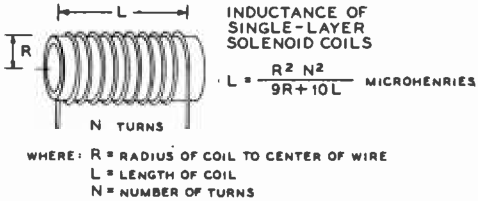


Figure 19.
FORMULA FOR CALCULATING INDUCTANCE.
Through the use of the equation and the sketch shown above the inductance of single-layer solenoid coils can be calculated with an accuracy of about one per cent for the types of coils normally used in the h-f and v-h-f range.

depends upon the shape and size of the two circuits, their positions and distances apart, and the permeability of the medium. The extent to which two inductors are coupled is expressed by a relation known as *coefficient of coupling*. This is the ratio of the mutual inductance actually present to the maximum possible value.

The formula for mutual inductance is $L = L_1 + L_2 + 2M$ when the coils are poled so that their fields add. When they are poled so that their fields buck, then $L = L_1 + L_2 - 2M$.

If a 3 henry coil and a 4 henry coil are placed so that there is no coupling between them, then the combined inductance of the two in series will be 7 henrys. But if the coils are placed in inductive relation to each other, the inductance of the two in series will be either greater or less than 7 henrys, depending upon whether the polarity is such that the mutual inductance aids the self-inductance or bucks the self-inductance. If the total inductance of the two coils when coupled measured either 6 or 8 henrys, then the mutual inductance would be (from the formula) ½ henry.

Inductors in Parallel Inductors in parallel are combined exactly as are resistors in parallel, provided that they are far enough apart so that the mutual inductance is entirely negligible.

Inductors in Series Inductors in series are additive, just as are resistors in series, again provided that no mutual inductance exists. In this case, the total inductance L is:

$$L = L_1 + L_2 + \ldots\ldots\ldots\ldots\text{etc.}$$

Where mutual inductance does exist:

$$L = L_1 + L_2 + 2M,$$

where M is the mutual inductance.

This latter expression assumes that the coils are connected in such a way that all flux linkages are in the same direction, i.e., additive. If this is not the case and the mutual linkages *subtract* from the self-linkages, the following formula holds:

$$L = L_1 + L_2 - 2M,$$

where M is the mutual inductance.

Core Material Ordinary magnetic cores cannot be used for radio frequencies because the *eddy current and hysteresis losses* in the core material become enormous as the frequency is increased. The principal use for conventional magnetic cores is in the audio-frequency range below approximately 15,000 cycles, whereas at very low frequencies (50 to 60 cycles) their use is mandatory if an appreciable value of inductance is desired.

An air core inductor of only 1 henry inductance would be quite large in size, yet values as high as 500 henrys are commonly available in small iron core chokes. The inductance of a coil with a magnetic core will vary with the amount of current (both a.c. and d.c.) which passes through the coil. For this reason, iron core chokes that are used in power supplies have a certain inductance rating at a *predetermined value of d.c.*

The permeability of air does not change with flux density; so the inductance of iron core coils often is made less dependent upon flux density by making part of the magnetic path air, instead of utilizing a closed loop of iron. This incorporation of an *air gap* is necessary in many applications of iron core coils, particularly where the coil carries a considerable d-c component. Because the permeability of air is so much lower than that of iron, the air gap need comprise only a small fraction of the magnetic circuit in order to provide a substantial proportion of the total reluctance.

Iron-core inductors may be used at radio frequencies if the iron is in a very finely divided form, as in the case of the powdered iron cores used in some types of r-f coils and i-f transformers. These cores are made of extremely small particles of iron. The particles are treated with an insulating material so that each particle will be insulated from the others, and the treated powder is molded with a binder into cores. Eddy current losses are greatly reduced, with the result that these special iron cores are entirely practical in circuits which operate up to 100 Mc. in frequency.

Time Constant—RC and RL Circuits When switch S in figure 20 is placed in position 1, a voltmeter across capacitor C will indicate the manner in which the capacitor will charge through the resistor R from battery B. If relatively large values are used for R and C, and if a v-t voltmeter which draws negligible current is used to measure the voltage *e*, the rate of charge of the capacitor may actually be plotted with the aid of a stop watch.

It will be found that the voltage *e* will begin to rise rapidly from zero the instant the switch is closed. Then, as the capacitor begins to charge, the rate of change of voltage across the capacitor will be found to decrease, the charging taking place more and more slowly as the capacitor voltage *e* approaches the battery voltage E. Actually, it will be found that in any given interval a constant percentage of the remaining difference between *e* and E will be delivered to the capacitor as an increase in voltage. A voltage which changes in this manner is said to increase *logarithmically*, or is said to follow an *exponential* curve.

A mathematical analysis of the charging of a capacitor in this manner would show that the relationship between the battery voltage E and the voltage across the capacitor *e* could be expressed in the following manner:

$$e = E\,(1 - \epsilon^{-t/RC})$$

where *e*, E, R, and C have the values discussed above, ϵ = 2.716 (the base of Naperian or natural logarithms), and t represents the time which has elapsed since the closing of the switch. With t expressed in seconds, R and C may be expressed in farads and ohms, or R and C may be expressed in microfarads and megohms. The product RC is called the *time constant* of the circuit, and is expressed in seconds. As an example, if R is one megohm and C is one microfarad, the time constant RC will be equal to the product of the two, or one second.

When the elapsed time t is equal to the time constant of the RC network under consideration, the exponent of ϵ becomes −1. Now ϵ^{-1} is equal to $1/\epsilon$, or 1/2.716, which is 0.368. The quantity (1-0.368) then is equal to 0.632. Expressed as a percentage, the above means that the voltage across the capacitor will have increased to 63.2 per cent of the battery voltage in an interval equal to the time constant or RC product of the circuit. Then, during the next period equal to the time constant of the RC combination, the voltage

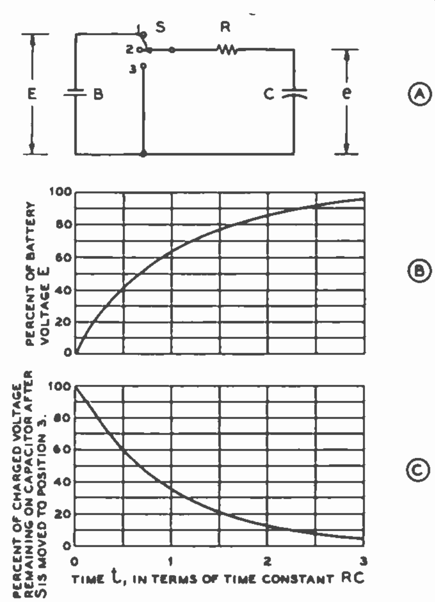


Figure 20.
TIME CONSTANT OF AN R-C CIRCUIT.
Shown at (A) is the circuit upon which is based the curves of (B) and (C). (B) shows the rate at which capacitor C will charge from the instant at which switch S is placed in position 1. (C) shows the discharge curve of capacitor C from the instant at which switch S is placed in position 3.

across the capacitor will have risen to 63.2 per cent of the remaining difference in voltage, or 86.5 per cent of the applied voltage E.

In the case of a series combination of a resistor and an inductor, as shown in figure 21, the current through the combination follows a very similar law to that given above for the voltage appearing across the capacitor in an RC series circuit. The equation for the current through the combination is:

$$i = \frac{E}{R}\left(1 - \epsilon^{-t\frac{R}{L}}\right)$$

where i represents the current at any instant through the series circuit, E represents the applied voltage, and R represents the total resistance of the resistor and the d-c resistance of the inductor in series. Thus the time constant of the RL circuit is L/R, with R expressed in ohms and L expressed in henrys.

When the switch in figure 20 is moved to position 3 after the capacitor has been charged, the capacitor voltage will drop in the manner shown in figure 20C. In this case the voltage

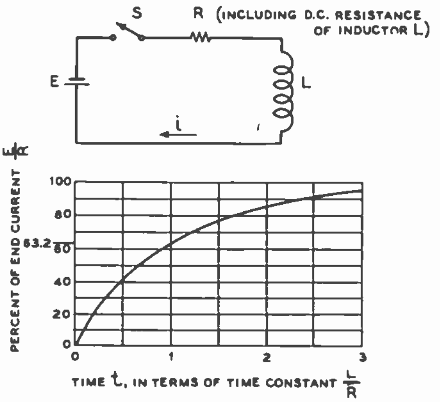


Figure 21.
TIME CONSTANT OF AN R-L CIRCUIT.
Note that the time constant for the increase in current through an R-L circuit is identical to the rate of increase in voltage across the capacitor in an R-C circuit.

across the capacitor will decrease to 36.8 per cent of the initial voltage (will make 63.2 per cent of the total drop) in a period of time equal to the time constant of the RC circuit.

CHAPTER THREE

Alternating Current Circuits

The previous chapter has been devoted to a discussion of circuits and circuit elements upon which is impressed a current consisting of a flow of electrons in one direction. This type of unidirectional current flow is called direct current, abbreviated *d.c.* Equally as important in radio and communications work, and more important in power practice, is a type of current flow whose direction of electron flow reverses periodically. The reversal of flow may take place at a low rate, as in the case of power systems, or it may take place millions of times per second in the case of communications frequencies. This type of current flow is called *alternating current,* abbreviated *a.c.*

Frequency of an Alternating Current An alternating current is one whose amplitude of current flow periodically rises from zero to a maximum in one direction, decreases to zero and changes its direction, rises to maximum in the opposite direction, and decreases to zero again. This complete process, starting from zero, passing through two maximums in opposite directions, and returning to zero again, is called a *cycle.* The number of times per second that a current passes through the complete cycle is called the *frequency* of the current in cycles per second. One and one quarter cycles of an alternating current wave are illustrated diagrammatically in figure 1.

Frequency Spectrum At present the usable frequency range for alternating electrical currents extends over the literally enormous frequency range from about 15 cycles per second to perhaps 30,000,000,000 cycles per second. It is obviously impracticable to use a frequency designation in c.p.s. for such enormously high frequencies, so three common units which are multiples of one cycle per second have been established.

These units are:

(1) the kilocycle (abbr., kc.), 1000 c.p.s.

(2) the Megacycle (abbr., Mc.), 1,000,000 c.p.s. or 1000 kc.

(3) the kilo-Megacycle (abbr., kMc.), 1,000,000,000 c.p.s. or 1000 Mc.

With easily handled units such as these we can classify the entire usable frequency range into frequency bands.

The frequencies falling between about 15 and 20,000 c.p.s. are called *audio* frequencies, abbreviated *a.f.,* since these frequencies are audible to the human ear when converted from electrical to acoustical signals by a loudspeaker or headphone. Frequencies in the vicinity of 60 c.p.s. also are called *power* frequencies, since they are commonly used to distribute electrical power to the consumer.

The frequencies falling between 10,000 c.p.s. (10 kc.) and 30,000,000,000 c.p.s. (30 kMc.) are commonly called *radio* frequencies, abbreviated *r.f.,* since they are commonly used

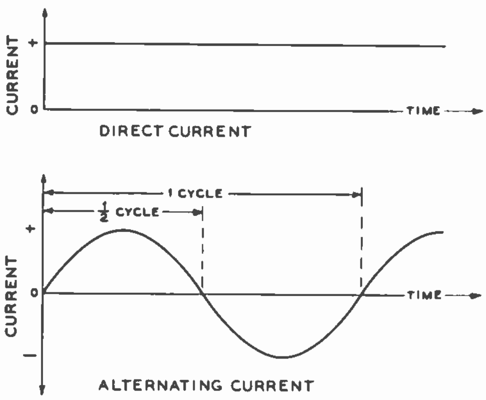


Figure 1.
ALTERNATING CURRENT AND DIRECT CURRENT.
Graphical comparison between unidirectional (direct) current and alternating current as plotted against time.

in radio communication and allied arts. The radio-frequency spectrum has been classified into seven frequency bands, each one of which is ten times as high in frequency as the one just below it in the spectrum (except for the v-l-f band at the bottom end of the spectrum). The present spectrum, with classifications, is given below.

Frequency	*Classification*	*Abbrev.*
10 to 30 kc.	Very-low frequencies	v.l.f.
30 to 300 kc.	Low frequencies	l.f.
300 to 3000 kc.	Medium frequencies	m.f.
3 to 30 Mc.	High frequencies	h.f.
30 to 300 Mc.	Very-high frequencies	v.h.f.
300 to 3000 Mc.	Ultra-high frequencies	u.h.f.
3 to 30 kMc.	Super-high frequencies	s.h.f.

Generation of Alternating Current Faraday discovered that if a conductor which forms part of a closed circuit is moved through a magnetic field so as to cut across the lines of force, a current will flow in the conductor. He also discovered that, if a conductor in a second closed circuit is brought near the first conductor and the current in the first one is varied, a current will flow in the second conductor. This effect is known as *induction*, and the currents so generated are *induced currents*. In the latter case it is the lines of force which are moving and cutting the second conductor, due to the varying current strength in the first conductor.

A current is induced in a conductor if there is a relative motion between the conductor and a magnetic field, its direction of flow depending upon the direction of the relative motion between the conductor and the field, and its strength depends upon the intensity of the field, the rate of cutting lines of force, and the number of turns in the conductor.

Alternators A machine that generates an alternating current is called an *alternator* or *a-c* generator. Such a machine in its basic form is shown in figure 2. It consists of two permanent magnets, M, the opposite poles of which face each other and are machined so that they have a common radius. Between these two poles, north (N) and south (S), a substantially constant magnetic field exists. If a conductor in the form of C is suspended so that it can be freely rotated between the two poles, and if the opposite ends of conductor C are brought to collector rings, there will be a flow of alternating current when conductor C is rotated. This current will flow out through the collector rings R and brushes B to the external circuit, X-Y.

The field intensity between the two pole pieces is substantially constant over the entire area of the pole face. However, when the conductor is moving parallel to the lines of force at the top or bottom of the pole faces, no lines are being cut. As the conductor moves on across the pole face it cuts more and more lines of force for each unit distance of travel, until it is cutting the maximum number of lines when opposite the center of the pole. Therefore, zero current is induced in the conductor at the instant it is midway between the two poles, and maximum current is induced when it is opposite the center of the

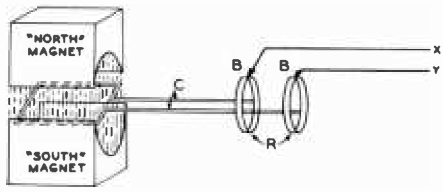


Figure 2.
THE ALTERNATOR.
Semi-schematic representation of the simplest form of the alternator.

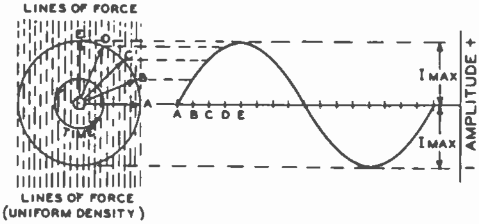


Figure 3.
OUTPUT OF THE ALTERNATOR.
Graph showing sine-wave output current of the alternator of figure 2.

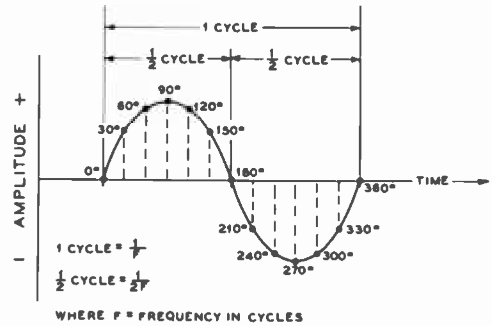


Figure 4.
THE SINE WAVE.
Illustrating one cycle of a sine wave. One complete cycle of alternation is broken up into 360 degrees. Then one-half cycle is 180 degrees, one-quarter cycle is 90 degrees, and so on down to the smallest division of the wave. A cosine wave has a shape identical to a sine wave but is shifted 90 degrees in phase —in other words the wave begins at full amplitude, the 90-degree point comes at zero amplitude, the 180-degree point comes at full amplitude in the opposite direction of current flow, and so forth.

pole face. After the conductor has rotated through 180° it can be seen that its position with respect to the pole pieces will be exactly opposite to that when it started. Hence, the second 180° of rotation will produce an alternation of current in the opposite direction to that of the first alternation.

The current does *not* increase directly as the angle of rotation, but rather as the *sine* of the angle; hence, such a current has the mathematical form of a *sine wave*. Although most electrical machinery does not produce a strictly pure sine curve, the departures are usually so slight that the assumption can be regarded as fact for most practical purposes. All that has been said in the foregoing paragraphs concerning alternating current also is applicable to alternating voltage.

The rotating arrow to the left in figure 3 represents a conductor rotating in a constant magnetic field of uniform density. The arrow also can be taken as a *vector* representing the strength of the magnetic field. This means that the length of the arrow is determined by the strength of the field (number of lines of force), which is constant. Now if the arrow is rotating at a constant rate (that is, with constant *angular velocity*), then the voltage developed across the conductor will be proportional to the rate at which it is cutting lines of force, which rate is proportional to the vertical distance between the tip of the arrow and the horizontal base line.

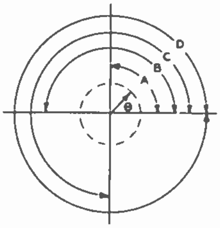
If EO is taken as unity or a voltage of 1, then the voltage (vertical distance from tip of arrow to the horizontal base line) at point C for instance may be determined simply by referring to a table of sines and looking up the sine of the angle which the arrow makes with the horizontal.

When the arrow has traveled from A to point E, it has traveled 90 degrees or one quarter cycle. The other three quadrants are not shown because their complementary or mirror relationship to the first quadrant is obvious.

It is important to note that time units are represented by *degrees* or quadrants. The fact that AB, BC, CD, and DE are equal chords (forming equal quadrants) simply means that the arrow (conductor or vector) is traveling at a constant speed, because these points on the radius represent the passage of equal units of time.

The whole picture can be represented in another way, and its derivation from the foregoing is shown in figure 3. The time base is represented by a straight line rather than by angular rotation. Points A, B, C, etc., represent the same units of time as before. When the voltage corresponding to each point is projected to the corresponding time unit, the familiar *sine curve* is the result.

Radian Notation From figure 1 we see that the value of an a-c wave varies continuously. But it is often of importance to know the amplitude of the wave in terms of the total amplitude at any instant



WHERE:
θ (THETA) = PHASE ANGLE = $2\pi ft$
A = $\frac{\pi}{2}$ RADIANS OR 90°
B = π RADIANS OR 180°
C = $\frac{3\pi}{2}$ RADIANS OR 270°
D = 2π RADIANS OR 360°
1 RADIAN = 57.324 DEGREES

Figure 5.
ILLUSTRATING RADIAN NOTATION.
The radian is a unit of phase angle, equal to 57.324 degrees. It is commonly used in mathematical relationships involving phase angles since such relationships are simplified when radian notation is used.

or at any time within the cycle. To be able to establish the instant in question we must be able to divide the cycle into parts. We could divide the cycle into eighths, hundredths, or any other ratio that suited our fancy. However, it is much more convenient mathematically to divide the cycle either into *electrical degrees* (360° represent one cycle) or into *radians.* A radian is an arc of a circle equal to the radius of the circle; hence there are 2π radians per cycle — or per circle (since there are π diameters per circumference, there are 2π radii).

Both radian notation and electrical degree notation are used in discussions of alternating current circuits. However, trigonometric tables are much more readily available in terms of degrees than radians, so the following simple conversions are useful.

$$2\pi \text{ radians} = 1 \text{ cycle} = 360°$$
$$\pi \text{ radians} = \tfrac{1}{2} \text{ cycle} = 180°$$
$$\frac{\pi}{2} \text{ radians} = \tfrac{1}{4} \text{ cycle} = 90°$$
$$\frac{\pi}{3} \text{ radians} = 1/6 \text{ cycle} = 60°$$
$$\frac{\pi}{4} \text{ radians} = \tfrac{1}{8} \text{ cycle} = 45°$$
$$1 \text{ radian} = \frac{1}{2\pi} \text{ cycle} = 57.3°$$

When the conductor in the simple alternator of figure 2 has made one complete revolution it has generated one cycle and has rotated through 2π radians. The expression $2\pi f$ then represents the number of radians in one cycle multiplied by the number of cycles per second (the frequency) of the alternating voltage or current. The expression then represents the number of radians per second through which the conductor has rotated. Hence $2\pi f$ represents the angular velocity of the rotating conductor, or of the rotating vector which represents any alternating current or voltage, expressed in radians per second.

In technical literature the expression $2\pi f$ is often replaced by ω, the lower-case Greek letter omega. Velocity multiplied by time gives the distance travelled, so $2\pi ft$ (or ωt) represents the angular distance through which the rotating conductor or the rotating vector has travelled since the reference time $t=0$. In the case of a sine wave the reference time $t=0$ represents that instant when the voltage or the current, whichever is under discussion, also is equal to zero.

Instantaneous Value of Voltage or Current The instantaneous voltage or current is proportional to the sine of the angle through which the rotating vector has travelled since reference time $t=0$. Hence, when the peak value of the a-c wave amplitude (either voltage or current amplitude) is known, and the angle through which the rotating vector has travelled is established, the amplitude of the wave at this instant can be determined through use of the following expression:

$$e = E_{max} \sin 2\pi ft,$$

where e = the instantaneous voltage
E = maximum crest value of voltage,
f = frequency in cycles per second, and
t = period of time which has elapsed since $t=0$ expressed as a fraction of one second.

The instantaneous current can be found from the same expression by substituting i for e and I_{max} for E_{max}.

It is often easier to visualize the process of determining the instantaneous amplitude by ignoring the frequency and considering only one cycle of the a-c wave. In this case, for a sine wave, the expression becomes:

$$e = E_{max} \sin \theta$$

where θ represents the angle through which the vector has rotated since time (and amplitude) were zero. As examples:

when $\theta = 30°$
sin $\theta = 0.5$
so $e = 0.5\ E_{max}$

when $\theta = 60°$
sin $\theta = 0.866$
so $e = 0.866\ E_{max}$

when $\theta = 90°$
sin $\theta = 1.0$
so $e = E_{max}$

when $\theta = 1$ radian
sin $\theta = 0.8415$
so $e = 0.8415\ E_{max}$

Effective Value of an Alternating Current The instantaneous value of an alternating current or voltage varies continuously throughout the cycle. So some value of an a-c wave must be chosen to establish a relationship between the effectiveness of an a-c and a d-c voltage or current. The heating value of an alternating current has been chosen to establish the reference between the *effective* values of a-c and d-c. Thus an alternating current will have an effective value of 1 ampere when it produces the same heat in a resistor as does 1 ampere of direct current.

This effective value is derived by taking the instantaneous values of current over a cycle of alternating current, squaring these values, taking an average of the squares, and then taking the square root of the average. By this procedure, the effective value becomes known as the *root mean square* or r.m.s. value. This is the value that is read on a-c voltmeters and a-c ammeters. The r.m.s. value is 70.7 (for sine waves only) per cent of the peak or maximum instantaneous value and is expressed as follows:

$$E_{eff} \text{ or } E_{r.m.s.} = 0.707 \times E_{max} \text{ or}$$

$$I_{eff} \text{ or } I_{r.m.s.} = 0.707 \times I_{max}.$$

The following relations are extremely useful in radio and power work:

$$E_{r.m.s.} = 0.707 \times E_{max}, \text{ and}$$

$$E_{max} = 1.414 \times E_{r.m.s.}$$

Rectified Alternating Current or Pulsating Direct Current If an alternating current is passed through a rectifier, it emerges in the form of a current of *varying amplitude* which flows in *one* direction only. Such a current is known as *rectified a.c.* or *pulsating d.c.* A typical wave form of a pulsating direct current as would be obtained from the output of a full-wave rectifier is shown in figure 6.

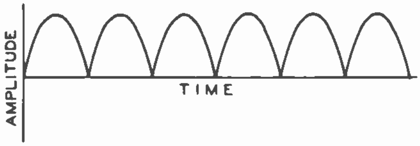


Figure 6.
FULL-WAVE RECTIFIED SINE WAVE.
Waveform obtained at the output of a full-wave rectifier being fed with a sine wave and having 100 per cent rectification efficiency. Each pulse has the same shape as one-half cycle of a sine wave. This type of current is known as pulsating direct current.

Measuring instruments designed for d.c. operation will not read the peak or instantaneous maximum value of the pulsating d-c output from the rectifier; they will read only the *average value*. This can be explained by assuming that it could be possible to cut off some of the peaks of the waves, using the cut-off portions to fill in the spaces that are open, thereby obtaining an *average* d-c value. A milliammeter and voltmeter connected to the adjoining circuit, or across the output of the rectifier, will read this average value. It is related to *peak* value by the following expression:

$$E_{avg} = 0.636 \times E_{max}$$

It is thus seen that the average value is 63.6 per cent of the peak value.

Relationship Between Peak, R.M.S. or Effective, and Average Values To summarize the three most significant values of an a-c sine wave: the peak value is equal to 1.41 times the r.m.s. or effective, and the r.m.s. value is equal to 0.707 times the peak value; the average value of a full-wave rectified a-c wave is 0.636 times the peak value, and the average value of a rectified wave is equal to 0.9 times the r.m.s. value.

R.M.S. = 0.707 × Peak
Average = 0.636 × Peak

Average = 0.9 × R.M.S.
R.M.S. = 1.11 × Average

Peak = 1.414 × R.M.S.
Peak = 1.57 × Average

Applying Ohm's Law to Alternating Current Ohm's law applies equally to direct or alternating current, *provided* the circuits under consideration are purely resistive, that is, circuits which have neither inductance (coils) nor capacitance (capacitors). Problems which involve tube filaments, drop resistors, electric lamps, heaters or similar resistive devices can be solved from Ohm's law, regardless of whether the current is direct or alternating. When a capacitor or coil is made a part of the circuit, a property common to either, called *reactance*, must be taken into consideration. Ohm's law still applies to a-c circuits containing reactance, but additional considerations are involved; these will be discussed in a later paragraph.

Inductive Reactance As was stated in Chapter Two, when a changing current flows through an inductor a back- or counter-electromotive force is developed; this force opposes any change in the initial current. This property of an inductor causes it to offer opposition or *impedance* to a change in current. The measure of impedance offered by an inductor to an alternating current of a given frequency is known as its *inductive reactance*. This is expressed as X_L:

$$X_L = 2\pi fL,$$

where X_L = inductive reactance expressed in ohms.

$\pi = 3.1416$ $(2\pi = 6.283)$,
f = frequency in cycles,
L = inductance in henrys.

Inductive Reactance at R.F. It is very often necessary to compute inductive reactance at radio frequencies. The same formula may be used, but to make it less cumbersome the inductance is expressed in *millihenrys* and the frequency in *kilocycles*. For higher frequencies and smaller values of inductance, frequency is expressed in *megacycles* and inductance in *microhenrys*. The basic equation need not be changed, since the multiplying factors for inductance and frequency appear in numerator and denominator, and hence are cancelled out. However, it is not possible in the same equation to express L in millihenrys and f in cycles without conversion factors.

Capacitive Reactance It has been explained that inductive reactance is the measure of the ability of an inductor to offer impedance to the flow of an alternating current. Capacitors have a similar property although in this case the opposition is to any change in the voltage across the capacitor. This property is called *capacitive reactance* and is expressed as follows:

$$X_c = \frac{1}{2\pi fC},$$

where X_c = capacitive reactance in ohms,
π = 3.1416,
f = frequency in cycles,
C = capacitance in farads.

Capacitive Reactance at R. F. Here again, as in the case of inductive reactance, the units of capacitance and frequency can be converted into smaller units for practical problems encountered in radio work. The equation may be written:

$$X_c = \frac{1{,}000{,}000}{2\pi fC},$$

where f = frequency in megacycles,
C = capacitance in micro-microfarads.

In the audio range it is often convenient to express frequency (f) in *cycles* and capacitance (C) in *microfarads*, in which event the same formula applies.

Phase When an alternating current flows through a purely resistive circuit, it will be found that the current will go through maximum and minimum in perfect step with the voltage. In this case the current is said to be in step or *in phase* with the voltage. For this reason, Ohm's law will apply equally well for a.c. or d.c. where pure resistances are concerned, provided that the same values of the wave (either peak or r.m.s.) for both voltage and current are used in the calculations.

However, in calculations involving alternating currents the voltage and current are not necessarily in phase. The current through the circuit may lag behind the voltage, in which case the current is said to have *lagging* phase; lagging phase is caused by inductive reactance.

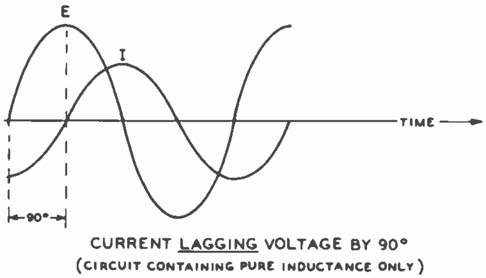


Figure 7.
LAGGING PHASE ANGLE.
Showing the manner in which the current lags the voltage in an a-c circuit containing pure inductance only. The lag is equal to one-quarter cycle or 90 degrees.

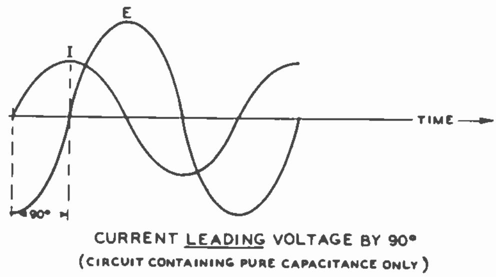


Figure 8.
LEADING PHASE ANGLE.
Showing the manner in which the current leads the voltage in an a-c circuit containing pure capacitance only. The lead is equal to one-quarter cycle or 90 degrees.

Or the current may reach its maximum value ahead of the voltage, in which case the current is said to have *leading* phase; a leading phase angle is caused by capacitive reactance.

In an electrical circuit containing reactance only, the current will either lead or lag the voltage by 90°. If the circuit contains inductive reactance only, the current will lag the voltage by 90°. If only capacitive reactance is in the circuit, the current will lead the voltage by 90°.

Reactances in Combination Inductive and capacitive reactance have exactly opposite effects on the phase relation between current and voltage in a circuit. Hence when they are used in combination their effects tend to neutralize. The combined effect of a capacitive and an inductive reactance is often called the *net reactance* of a circuit. The net reactance is found by subtracting the capacitive reactance from the inductive reactance, $X = X_L - X_C$.

The result of such a combination of pure reactances may be either positive, in which case the positive reactance is greater so that the net reactance is inductive, or it may be negative in which case the capacitive reactance is greater so that the net reactance is capacitive, or zero in which case the circuit is said to be *resonant*. The condition of resonance will be discussed in a later section. Note that inductive reactance is always taken as being positive while capacitive reactance is always taken as being negative.

Impedance; Circuits Containing Reactance and Resistance Pure reactances introduce a phase angle of 90° between voltage and current; pure resistance introduces no phase shift between voltage and current. Hence we cannot add a reactance and a resistance directly. When a reactance and a resistance are used in combination the resulting phase angle of current flow with respect to the impressed voltage lies somewhere between plus or minus 90° and 0° depending upon the relative magnitudes of the reactance and the resistance.

The term *impedance* is a general term which can be applied to any electrical entity which impedes the flow of current. Hence the term may be used to designate a resistance, a pure reactance, or a complex combination of both reactance and resistance. The designation for impedance is Z. An impedance must be defined in such a manner that both its magnitude and its phase angle are established. The designation may be accomplished in either of two ways — one of which is convertible into the other by simple mathematical operations.

The first method of designating an impedance is actually to specify both the resistive and the reactive component in the form $R + jX$. In this form R represents the resistive component in ohms and X represents the reactive component. The "j" merely means that the X component is reactive and thus cannot be added directly to the R component. Plus jX means that the reactance is positive or inductive, while if minus jX were given it

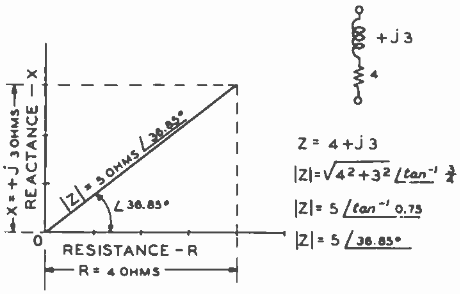


Figure 9.
THE IMPEDANCE TRIANGLE.
Showing the graphical construction of a triangle for obtaining the net (scalar) impedance resulting from the connection of a resistance and a reactance in series. Shown also alongside is the alternative mathematical procedure for obtaining the values associated with the triangle.

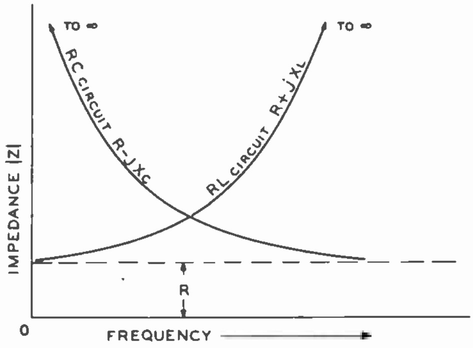


Figure 10.
IMPEDANCE AGAINST FREQUENCY FOR R-L AND R-C CIRCUITS.
The impedance of an R-C circuit approaches infinity as the frequency approaches zero (d.c.), while the impedance of a series R-L circuit approaches infinity as the frequency approaches infinity. The impedance of an R-C circuit approaches the impedance of the series resistor as the frequency approaches infinity, while the impedance of a series R-L circuit approaches the impedance of the resistor as the frequency approaches zero (d.c.)

would mean that the reactive component was negative or capacitive.

The second method of representing an impedance is to specify its absolute magnitude and the phase angle of current with respect to voltage, in the form $Z \angle \theta$. Figure 9 shows graphically the relationship between the two common ways of representing an impedance.

The construction of figure 9 is called an impedance diagram. Through the use of such a diagram we can add graphically a resistance and a reactance to obtain a value for the resulting impedance in the scalar form. With zero at the origin, resistances are plotted to the right, positive values of reactance (inductive) in the upward direction, and negative values of reactance (capacitive) in the downward direction.

Note that the resistance and reactance are drawn as the two sides of a right triangle, with the hypotenuse representing the resulting impedance. Hence it is possible to determine mathematically the value of a resultant impedance through the familiar right-triangle relationship — the square of the hypotenuse is equal to the sum of the squares of the other two sides:

$$Z^2 = R^2 + X^2$$

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

Note also that the angle θ included between R and Z can be determined from any of the following trigonometric relationships:

$$\sin \theta = \frac{X}{|Z|}$$

$$\cos \theta = \frac{R}{|Z|}$$

$$\tan \theta = \frac{X}{R}$$

One common problem is that of determining the scalar magnitude of the impedance, $|Z|$, and the phase angle θ, when resistance and reactance are known; hence, of converting from the $Z = R + jX$ to the $|Z| \angle \theta$ form. In this case we use two of the expressions just given:

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

$$\tan \theta = \frac{X}{R}, \left(\text{or } \theta = \tan^{-1} \frac{X}{R}\right)$$

The inverse problem, that of converting from the $|Z| \angle \theta$ to the $R + jX$ form is done with the following relationships, both of which are obtainable by simple division from the trigonometric expressions just given for determining the angle θ:

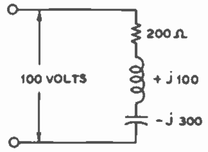


Figure 11.
SERIES R-L-C CIRCUIT.

$$R = |Z| \quad \cos \theta$$
$$jX = |Z| \ j \ \sin \theta$$

By simple addition these two expressions may be combined to give the relationship between the two most common methods of indicating an impedance:

$$R + jX = |Z| \ (\cos \theta + j \sin \theta)$$

In the case of impedance, resistance, or reactance, the unit of measurement is the ohm; hence the ohm may be thought of as a unit of *opposition to current flow*, without reference to the relative phase angle between the applied voltage and the current which flows.

Further, since both capacitive and inductive reactance are functions of frequency, impedance will vary with frequency. Figure 10 shows the manner in which |Z| will vary with frequency in an RL series circuit and in an RC series circuit.

Series RLC Circuits In a series circuit containing R, L, and C, the impedance is determined as discussed before except that the reactive component in the expressions becomes the net reactance — the difference between X_L and X_C. Hence $(X_L - X_C)$ may be substituted for X in the equations. Thus:

$$|Z| = \sqrt{R^2 + (X_L - X_C)^2}$$
$$\theta = \tan^{-1} \frac{(X_L - X_C)}{R}$$

A series RLC circuit thus may present an impedance which is capacitively reactive if the net reactance is capacitive, inductively reactive if the net reactance is inductive, or resistive if the capacitive and inductive reactances are equal.

Addition of Complex Quantities The addition of complex quantities (for example, impedances in series) is

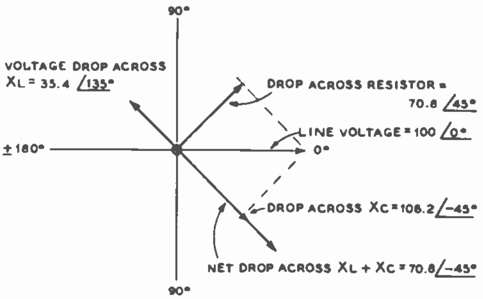


Figure 12.
Graphical construction of the voltage drops associated with the series R-L-C circuit of figure 11.

quite simple if the quantities are in the rectangular form. If they are in the polar form they only can be added graphically, unless they are converted to the rectangular form by the relationships previously given. As an example of the addition of complex quantities in the rectangular form, the equation for the addition of impedances is:

$$(R_1 + jX_1) + (R_2 + jX_2) = (R_1 + R_2) + j(X_1 + X_2)$$

For example if we wish to add the impedances (10 + j50) and (20 − j30) we obtain:

$$\begin{aligned}(10 + j50) + (20 - j30) &= (10 + 20) + j(50 + (-30)) \\ &= 30 + j(50-30) \\ &= 30 + j20\end{aligned}$$

Multiplication and Division of Complex Quantities It is often necessary in solving certain types of circuits to multiply or divide two complex quantities. It is a much simpler mathematical operation to multiply or divide complex quantities if they are expressed in the polar form. Hence if they are given in the rectangular form they should be converted to the polar form before multiplication or division is begun. Then the multiplication is accomplished by multiplying the |Z| terms together and *adding* algebraically the $\angle\theta$ terms, as:

$$(|Z_1| \angle\theta_1)(|Z_2| \angle\theta_2) = |Z_1| \, |Z_2| \angle\theta_1 + \theta_2$$

For example, suppose that the two impedances |20| ∠43° and |32| ∠−23° are to be multiplied. Then:

$$(|20| \angle 43°)\ (|32| \angle -23°) = |20 \cdot 32| \angle 43° + (-23°)$$
$$= 640 \angle 20°$$

Division is accomplished by dividing the denominator into the numerator, and *subtracting* the angle of the denominator from that of the numerator, as:

$$\frac{|Z_1| \angle \theta_1}{|Z_2| \angle \theta_2} = \frac{|Z_1|}{|Z_2|} \angle \theta_1 - \theta_2$$

For example, suppose that an impedance of $|50| \angle 67°$ is to be divided by an impedance of $|10| \angle 45°$. Then:

$$\frac{|50| \angle 67°}{|10| \angle 45°} = \frac{|50|}{|10|} \angle 67° - 45°$$
$$= |5| \angle 22°$$

Ohm's Law for Complex Quantities The simple form of Ohm's Law used for d-c circuits may be stated in a more general form for application to a-c circuits involving either complex quantities or simple resistive elements. The form is:

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

in which, in the general case, I, E, and Z are complex (vector) quantities. In the simple case where the impedance is a pure resistance with an a-c voltage applied, the equation simplifies to the familiar I = E/R. In any case the applied voltage may be expressed either as peak, r.m.s., or average; the resulting current always will be in the same type of units as used to define the voltage.

In the more general case vector algebra must be used to solve the equation. And, since either division or multiplication is involved, the complex quantities should be expressed in the polar form. As an example, take the case of the series circuit shown in figure 11, with 100 volts applied. The impedance of the series circuit can best be obtained first in the rectangular form, as:

$$200 + j(100-300) = 200 - j200$$

Now, to obtain the current we must convert this impedance to the polar form.

$$|Z| = \sqrt{200^2 + (-200)^2}$$
$$= \sqrt{40{,}000 + 40{,}000}$$
$$= \sqrt{80{,}000}$$
$$= 282\ \Omega$$

$$\theta = \tan^{-1}\frac{X}{R} = \tan^{-1}\frac{-200}{200} = \tan^{-1} -1$$
$$= -45°.$$

Therefore $Z = 282 \angle -45°$

Note that in a series circuit the resulting impedance takes the sign of the largest reactance in the series combination.

Where a slide-rule is being used to make the computations, the impedance may be found without any addition or subtraction operations by finding the angle θ first, and then using the trigonometric equation below for obtaining the impedance. Thus:

$$\theta = \tan^{-1}\frac{X}{R} = \tan^{-1}\frac{-200}{200} = \tan^{-1} -1$$
$$= -45°$$

Then $|Z| = \dfrac{R}{\cos\theta}$ $\qquad \cos -45° = 0.707$

$$|Z| = \frac{200}{0.707} = 282\ \Omega$$

Since the applied voltage will be the reference for the currents and voltages within the circuit, we may define it as having a zero phase angle: $E = 100 \angle 0°$.
Then

$$I = \frac{100 \angle 0°}{282 \angle -45°} = 0.354 \angle 0° - (-45°)$$
$$= 0.354 \angle 45° \text{ amperes.}$$

This same current must flow through all three elements of the circuit, since they are in series and the current through one must already have passed through the other two. Hence the voltage drop across the resistor (whose phase angle of course is 0°) is:

$$E = I\,R$$
$$E = (0.354 \angle 45°)\ (200 \angle 0°)$$
$$= 70.8 \angle 45° \text{ volts}$$

The voltage drop across the inductive reactance is:

$$E = I\ X_L$$

$$E = (0.354\ \angle 45^\circ)\ (100\ \angle 90^\circ)$$
$$= 35.4\ \angle 135^\circ \text{ volts}$$

Similarly, the voltage drop across the capacitive reactance is:

$$E = I\ X_C$$

$$E = (0.354\ \angle 45^\circ)\ (300\ \angle -90^\circ)$$
$$= 106.2\ \angle -45^\circ$$

Note that the voltage drop across the capacitive reactance is greater than the supply voltage. This condition often occurs in a series RLC circuit, and is explained by the fact that the drop across the capacitive reactance is cancelled to a lesser or greater extent by the drop across the inductive reactance. It is often desirable in a problem such as the above to check the validity of the answer by adding vectorially the voltage drops across the components of the series circuit to make sure that they add up to the supply voltage—or to use the terminology of Kirchhoff's Second Law, to make sure that the voltage drops across all elements of the circuit, including the source taken as negative, is equal to zero.

In the general case of the addition of a number of voltage vectors in series it is best to resolve the voltages into their in-phase and out-of-phase components with respect to the supply voltage. Then these components may be added directly. Hence:

$$\begin{aligned} E_R &= 70.8\ \angle 45^\circ \\ &= 70.8\ (\cos 45^\circ + j \sin 45^\circ) \\ &= 70.8\ (0.707 + j0.707) \\ &= 50 + j50 \end{aligned}$$

$$\begin{aligned} E_L &= 35.4\ \angle 135^\circ \\ &= 35.4\ (\cos 135^\circ + j \sin 135^\circ) \\ &= 35.4\ (-0.707 + j0.707) \\ &= -25 + j25 \end{aligned}$$

$$\begin{aligned} E_C &= 106.2\ \angle 45^\circ \\ &= 106.2\ (\cos -45^\circ + j \sin -45^\circ) \\ &= 106.2\ (0.707 - j0.707) \\ &= 75 - j75 \end{aligned}$$

$$\begin{aligned} E_R + E_L + E_C &= (50 + j50) + (-25 + j25) \\ &\quad + (75 - j75) \\ &= (50 - 25 + 75) + j(50 + 25 - 75) \\ &= 100 + j0 \\ &= 100\ \angle 0^\circ, \text{ which is equal to the} \end{aligned}$$

supply voltage.

Checking by Construction on the Complex Plane It is frequently desirable to check computations involving complex quantities by constructing vectors representing the quantities on the complex plane. Figure 12 shows such a construction for the quantities of the problem just completed. Note that the answer to the problem may be checked by constructing a parallelogram with the voltage drop across the resistor as one side and the net voltage drop across the capacitor plus the inductor (these may be added agrebraically as they are 180° out of phase) as the adjacent side. The vector sum of these two voltages, which is represented by the diagonal of the parallelogram, is equal to the supply voltage of 100 volts at zero phase angle.

Resistance and Reactance in Parallel In a series circuit, such as just discussed, the current through all the elements which go to make up the series circuit is the same. But the voltage drops across each of the components are, in general, different from one another. Conversely, in a parallel RLC or RX circuit the voltage is, obviously, the same across each of the elements. But the currents through each of the elements are usually different.

There are many ways of solving a problem involving paralleled resistance and reactance; several of these ways will be described. In general, it may be said that the impedance of a number of elements in parallel is solved using the same relations as are used for solving resistors in parallel, except that complex quantities are employed. The basic relation is:

$$\frac{1}{Z_{tot}} = \frac{1}{Z_1} + \frac{1}{Z_2} + \frac{1}{Z_3} + \ldots\ldots,$$

or when only two impedances are involved:

$$Z_{tot} = \frac{Z_1 Z_2}{Z_1 + Z_2}$$

As an example using the two-impedance relation, take the simple case, illustrated in figure 13, of a resistance of 6 ohms in parallel with a capacitive reactance of 4 ohms. To simplify the first step in the computation it is best to put the impedances in the polar form for the numerator, since multiplication is involved, and in the rectangular form for the addition in the denominator.

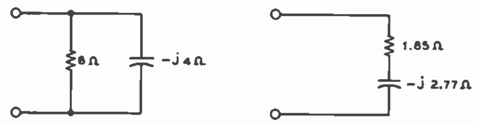


Figure 13.
THE EQUIVALENT SERIES CIRCUIT.
Showing a parallel R-C circuit and the equivalent series R-C circuit which represents the same net impedance as the parallel circuit.

$$Z_{tot} = \frac{(6\angle 0°)\ \ (4\angle -90°)}{6 - j4}$$

$$= \frac{24\angle -90°}{6 - j4}$$

Then the denominator is changed to the polar form for the division operation:

$$\theta = \tan^{-1}\frac{-4}{6} = \tan^{-1} - 0.667 = -33.7°$$

$$|Z| = \frac{6}{\cos -33.7°} = \frac{6}{0.832} = 7.21 \text{ ohms}$$

$$6 - j4 = 7.21\ \angle -33.7°$$

Then:

$$Z_{tot} = \frac{24\ \angle -90°}{7.21\ \angle -33.7°} = 3.33\ \angle -56.3°$$

$$= 3.33\ (\cos -56.3° + j \sin -56.3°)$$

$$= 3.33\,[0.5548 + j\ (-0.832)]$$

$$= 1.85 - j\ 2.77$$

Equivalent Series Circuit Through the series of operations in the previous paragraph we have converted a circuit composed of two impedances in parallel into an *equivalent series circuit* composed of impedances in series. An equivalent series circuit is one which, as far as the terminals are concerned, acts identically to the original parallel circuit; the current through the circuit and the power dissipation of the resistive elements are the same for a given voltage at the specified frequency.

We can check the equivalent series circuit with respect to the original circuit of figure 15 by assuming that one volt a.c. (at the frequency where the capacitive reactance in the parallel circuit is 4 ohms) is applied to the terminals of both.

In the parallel circuit the current through the resistor will be 1/6 ampere (0.166 a.) while the current through the capacitor will be j ¼ ampere (+ j 0.25 a.). The total current will be the sum of these two currents, or 0.166 + j 0.25 a. Adding these vectorially we obtain:

$$|I| = \sqrt{0.166^2 + 0.25^2} = \sqrt{0.09} = 0.3 \text{ a.}$$

The dissipation in the resistor will be $1^2/6 = 0.166$ watts.

In the case of the equivalent series circuit the current will be:

$$|I| = \frac{E}{|Z|} = \frac{1}{3.33} = 0.3 \text{ a.}$$

And the dissipation in the resistor will be:

$$W = I^2R = 0.3^2 \times 1.85$$
$$= 0.9 \times 1.85$$
$$= 0.166 \text{ watts}$$

So we see that the equivalent series circuit checks exactly with the original parallel circuit.

Parallel RLC Circuits In solving a more complicated circuit made up of more than two impedances in parallel we may elect to use either of two methods of solution. These methods are called the *admittance* method and the *assumed-voltage* method. However, the two methods are equivalent since both use the sum-of-reciprocals equation:

$$\frac{1}{Z_{tot}} = \frac{1}{Z_1} + \frac{1}{Z_2} + \frac{1}{Z_3} \ldots\ldots$$

In the admittance method we use the relation $Y = 1/Z$, where $Y = G + jB$; Y is called the *admittance,* defined above, G is the *conductance* or R/Z^2 and B is the *susceptance* or $-X/Z^2$. Then $Y_{tot} = 1/Z_{tot} = Y_1 + Y_2 + Y_3 \ldots.$ In the assumed-voltage method we multiply both sides of the equation above by E, the assumed voltage, and add the currents, as:

$$\frac{E}{Z_{tot}} = \frac{E}{Z_1} + \frac{E}{Z_2} + \frac{E}{Z_3} \ldots = I_{Z1} + I_{Z2} + I_{Z3} \ldots$$

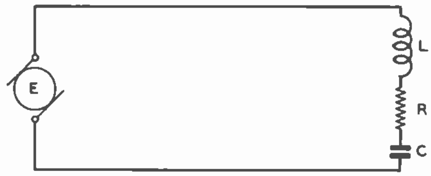


Figure 14.
SERIES RESONANT CIRCUIT.

Then the impedance of the parallel combination may be determined from the relation:

$$Z_{tot} = E/I_{Ztot}$$

Resonant Circuits

A series circuit such as shown in figure 14 is said to be in *resonance* when the applied frequency is such that the capacitive reactance is exactly balanced by the inductive reactance. At this frequency the two reactances will cancel in their effects, and the impedance of the circuit will be at a minimum so that maximum current will flow. In fact, as shown in figure 15, the net impedance of a series circuit at resonance is equal to the resistance which remains in the circuit after the reactances have been cancelled.

Resonant Frequency Some resistance is always present in a circuit because it is possessed in some degree by both the inductor and the capacitor. If the frequency of the alternator E is varied from nearly zero to some high frequency, there will be one particular frequency at which the inductive reactance and capacitive reactance will be equal. This is known as the *resonant frequency,* and in a series circuit it is the frequency at which the circuit current will be a maximum. Such series resonant circuits are chiefly used when it is desirable to allow a certain frequency to pass through the circuit (low impedance to this frequency), while at the same time the circuit is made to offer considerable opposition to currents of other frequencies.

If the values of inductance and capacitance both are fixed, there will be only one resonant frequency.

If both the inductance and capacitance are made variable, the circuit may then be changed or *tuned,* so that a number of combinations of inductance and capacitance can resonate at the same frequency. This can be more easily understood when one considers that inductive reactance and capacitive reactance travel in opposite directions as the frequency is changed. For example, if the frequency were to remain constant and the values of inductance and capacitance were then changed, the following combinations would have equal *reactance:*

Frequency is constant at 60 cycles.

L is expressed in henrys.

C is expressed in microfarads (.000001 farad.)

L	X_L	C	X_C
.265	100	26.5	100
2.65	1,000	2.65	1,000
26.5	10,000	.265	10,000
265.00	100,000	.0265	100,000
2,650.00	1,000,000	.00265	1,000,000

Frequency of Resonance From the formula for resonance,

$$2\pi fL = \frac{1}{2\pi fC},$$ the resonant frequency

can readily be solved. In order to isolate f on one side of the equation, merely multiply both sides by $2\pi f$, thus giving:

$$4\pi^2 f^2 L = \frac{1}{C}.$$

Divided by the quantity $4\pi^2 L$, the result is:

$$f^2 = \frac{1}{4\pi^2 LC}.$$

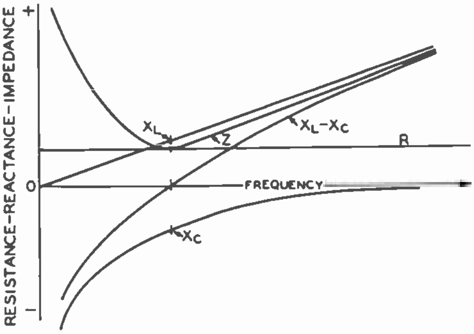


Figure 15.
IMPEDANCE OF A SERIES-RESONANT CIRCUIT.
Showing the variation in reactance of the separate elements and in the net impedance of a series resonant circuit (such as figure 14) with changing frequency. The short vertical lines are drawn at the point of resonance ($X_L - X_C = 0$) in the series circuit.

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

Then, by taking the square root of both sides:

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

where f = frequency in cycles,
L = inductance in henrys,
C = capacitance in farads.

It is more convenient to express L and C in smaller units, especially in making radio-frequency calculations; f can also be expressed in megacycles or kilocycles. A very useful group of such formulas is:

$$f^2 = \frac{25{,}330}{LC} \text{ or } L = \frac{25{,}330}{f^2C} \text{ or } C = \frac{25{,}330}{f^2L}$$

where f = frequency in megacycles,
L = inductance in microhenrys,
C = capacitance in micromicrofarads.

Impedance of Series Resonant Circuits The impedance across the terminals of a series resonant circuit (figure 14) is:

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

where Z = impedance in ohms,
r = resistance in ohms,
X_C = capacitive reactance in ohms,
X_L = inductive reactance in ohms.

From this equation, it can be seen that the impedance is equal to the vector sum of the circuit resistance and the *difference* between the two reactances. Since at the resonant frequency X_L equals X_C, the difference between them (figure 15) is obviously zero, so that at resonance the impedance is simply equal to the resistance of the circuit; therefore, because the resistance of most normal radio-frequency circuits is of a very low order, the impedance is also low.

At frequencies higher and lower than the resonant frequency, the difference between the reactances will be a definite quantity and will add with the resistance to make the impedance higher and higher as the circuit is tuned off the resonant frequency.

If X_C should be greater than X_L, then the term $(X_L - X_C)$ will give a negative number. However, this is nothing to worry about because when the difference is squared the product is always positive. This means that the smaller reactance is subtracted from the larger, regardless of whether it be capacitive or inductive, and the difference squared.

Current and Voltage in Series Resonant Circuits Formulas for calculating currents and voltages in a series resonant circuit are similar to those of Ohm's law.

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

The complete equations:

$$I = \frac{E}{\sqrt{r^2 + (X_L - X_C)^2}}$$

$$E = I\sqrt{r^2 + (X_L - X_C)^2}$$

Inspection of the above formulas will show the following to apply to series resonant circuits: When the impedance is low, the current will be high; conversely, when the impedance is high, the current will be low.

Since it is known that the impedance will be very low at the resonant frequency, it follows that the current will be a maximum at this point. If a graph is plotted of the current against the frequency either side of resonance, the resultant curve becomes what is known as a *resonance curve*. Such a curve is shown in figure 16, the frequency being plotted against *current* in the series resonant circuit.

Several factors will have an effect on the shape of this resonance curve, of which resistance and L-to-C ratio are the important considerations. The curves B and C in figure 16 show the effect of adding increasing values of resistance to the circuit. It will be seen that the peaks become less and less prominent as the resistance is increased; thus, it can be said that the *selectivity* of the circuit is thereby *decreased*. Selectivity in this case can be defined as the ability of a circuit to discriminate against frequencies adjacent to the resonant frequency.

Voltage Across Coil and Capacitor in Series Circuit Because the a-c or r-f voltage across a coil and capacitor is proportional to the reactance (for a given current), the actual voltages across the coil and across the capacitor may be many times greater than the *terminal* voltage of the circuit. Furthermore, since the individual reactances can be very high, the voltage across the capacitor, for example, may be high enough to cause flashover, even though the applied voltage is of a value considerably below that at which the capacitor is rated.

Circuit Q—Sharpness of Resonance An extremely important property of a capacitor or an inductor is its factor-of-merit, more generally called its Q. It is this factor, Q, which primarily determines the sharpness of resonance of a tuned circuit. This factor can be expressed as the ratio of the reactance to the resistance, as follows:

$$Q = \frac{2\pi fL}{R},$$

where R = total resistance.

Skin Effect The actual resistance in a wire or an inductor can be far greater than the d-c value when the coil is used in a radio-frequency circuit; this is because the current does not travel through the entire cross-section of the conductor, but has a tendency to travel closer and closer to the surface of the wire as the frequency is increased. This is known as the *skin effect.*

The actual current-carrying portion of the wire is decreased, as a result of the skin effect, so that the ratio of a-c to d-c resistance of the wire, called the *resistance ratio,* is increased. The resistance ratio of wires to be used at frequencies below about 500 kc. may be materially reduced through the use of *litz* wire. Litz wire, of the type commonly used to wind the coils of 455-kc. i-f transformers, may consist of 3 to 10 strands of insulated wire, about No. 40 in size, with the individual strands connected together only at the ends of the coils.

Variation of Q with Frequency Examination of the equation for determining Q might give rise to the thought that even though the resistance of an inductor increases with frequency, the inductive reactance does likewise, so that the Q might be a constant. Actually, however, it works out in practice that the Q of an inductor will reach a relatively broad maximum at some particular frequency. Hence, coils normally are designed in such a manner that the peak in their curve of Q with frequency will occur at the normal operating frequency of the coil in the circuit for which it is designed.

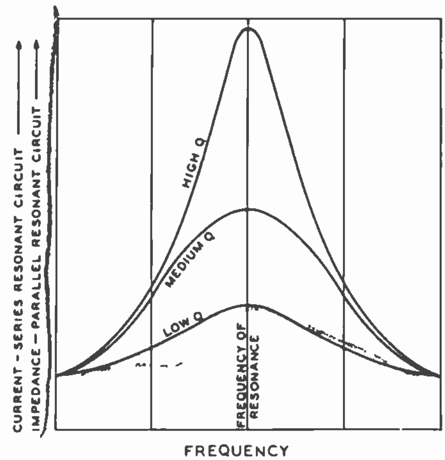


Figure 16.
RESONANCE CURVE.
Showing the increase in impedance at resonance for a parallel-resonant circuit, and similarly, the increase in current at resonance for a series-resonant circuit. The sharpness of resonance is determined by the Q of the circuit, as illustrated by a comparison between A, B, and C.

The Q of a capacitor ordinarily is much higher than that of the best coil. Therefore, it usually is the merit of the coil that limits the overall Q of the circuit.

At audio frequencies the core losses in an iron-core inductor greatly reduce the Q from the value that would be obtained simply by dividing the reactance by the resistance. Obviously the core losses also represent circuit resistance, just as much so as though the loss occurred in the wire itself.

Parallel Resonance In radio circuits, parallel resonance (more correctly termed *anti-resonance*) is more frequently encountered than series resonance; in fact, it is the basic foundation of receiver and transmitter circuit operation. A circuit is shown in figure 17.

The "Tank" Circuit In this circuit, as contrasted with a circuit for series resonance, L (inductance) and C (capaci-

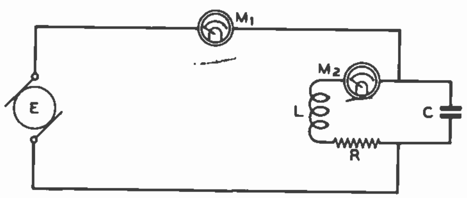


Figure 17.
PARALLEL-RESONANT CIRCUIT.
The inductance L and capacitance C comprise the reactive elements of the parallel-resonant (anti-resonant) tank circuit, and the resistance R indicates the sum of the r-f resistance of the coil and capacitor, plus the resistance coupled into the circuit from the external load. In most cases the tuning capacitor has much lower r-f resistance than the coil and can therefore be ignored in comparison with the coil resistance and the coupled-in resistance. The instrument M_1 indicates the "line current" which keeps the circuit in a state of oscillation — this current is the same as the fundamental component of the plate current of a Class C amplifier which might be feeding the tank circuit. The instrument M_2 indicates the "tank current" which is equal to the line current multiplied by the operating Q of the tank circuit.

tance) are connected in *parallel*, yet the *combination* can be considered to be in series with the remainder of the circuit. This combination of L and C, in conjunction with R, the resistance which is principally included in L, is sometimes called a *tank* circuit because it effectively functions as a storage tank when incorporated in vacuum tube circuits.

Contrasted with series resonance, there are two kinds of current which must be considered in a parallel resonant circuit: (1) the line current, as read on the indicating meter M_1, (2) the circulating current which flows within the parallel L-C-R portion of the circuit. See figure 17.

At the resonant frequency, the line current (as read on the meter M_1) will drop to a very low value although the circulating current in the L-C circuit may be quite large. It is interesting to note that the parallel resonant circuit acts in a distinctly opposite manner to that of a series resonant circuit, in which the current is at a maximum and the impedance is minimum at resonance. It is for this reason that in a parallel resonant circuit the principal consideration is one of impedance rather than current. It is also significant that the *impedance* curve for *parallel* circuits is very nearly identical to that of the *current* curve for *series* resonance. The impedance at resonance is expressed as:

$$Z = \frac{(2\pi fL)^2}{R},$$

where Z = impedance in ohms,
L = inductance in henrys,
f = frequency in cycles,
R = resistance in ohms.

Or, impedance can be expressed as a function of Q as:

$$Z = 2\pi fLQ,$$

showing that the impedance of a circuit is directly proportional to its effective Q at resonance.

The curves illustrated in figure 16 can be applied to parallel resonance. Reference to the curve will show that the effect of adding resistance to the circuit will result in both a broadening out and a lowering of the peak of the curve. Since the voltage of the circuit is directly proportional to the impedance, and since it is this voltage that is applied to the grid of the vacuum tube in a detector or amplifier circuit, the impedance curve must have a sharp peak in order for the circuit to be *selective*. If the curve is broad-topped in shape, both the desired signal and the interfering signals at close proximity to resonance will give nearly equal voltages on the grid of the tube, and the circuit will then be *non-selective;* i.e., it will tune broadly.

Effect of L/C Ratio in Parallel Circuits In order that the highest possible voltage can be developed across a parallel resonant circuit, the impedance of this circuit must be very high. The impedance will be greater with conventional coils of limited Q when the ratio of inductance-to-capacitance is great, that is, when L is large as compared with C. When the resistance of the circuit is very low, X_L will equal X_C at maximum impedance. There are innumerable ratios of L and C that will have *equal* reactance, at a given resonant frequency, exactly as is the case in a series resonant circuit.

In practice, where a certain value of inductance is tuned by a variable capacitance over a fairly wide range in frequency, the L/C ratio will be small at the lowest frequency and large at the high-frequency end.

The circuit, therefore, will have unequal gain and selectivity at the two ends of the band of frequencies which is being tuned. Increasing the Q of the circuit (lowering the resistance) will obviously increase *both* the selectivity and gain.

Circulating Tank Current at Resonance The Q of a circuit has a definite bearing on the circulating tank current at resonance. This tank current is very nearly the value of the line current multiplied by the effective circuit Q. For example: an r-f line current of 0.050 amperes, with a circuit Q of 100, will give a circulating tank current of approximately 5 amperes. From this it can be seen that both the inductor and the connecting wires in a circuit with a high Q must be of very low resistance, particularly in the case of high power transmitters, if heat losses are to be held to a minimum.

Because the voltage across the tank at resonance is determined by the Q, it is possible to develop very high peak voltages across a high Q tank with but little line current.

Effect of Coupling on Impedance If a parallel resonant circuit is coupled to another circuit, such as an antenna output circuit, the impedance and the effective Q of the parallel circuit is decreased as the coupling becomes closer. The effect of closer (tighter) coupling is the same as though an actual resistance were added in series with the parallel tank circuit. The resistance thus coupled into the tank circuit can be considered as being *reflected* from the output or load circuit to the driver circuit.

Tank Circuit Flywheel Effect When the plate circuit of a Class B or Class C operated tube (defined in Chapter Five) is connected to a parallel resonant circuit tuned to the same frequency as the exciting voltage for the amplifier, the plate current serves to maintain this L/C circuit in a state of oscillation.

The plate current is supplied in short pulses which do not begin to resemble a sine wave, even though the grid may be excited by a sine-wave voltage. These spurts of plate current are converted into a sine wave in the plate tank circuit by virtue of the "Q" or "flywheel effect" of the tank.

If a tank did not have some resistance losses, it would, when given a "kick" with a single pulse, continue to oscillate indefinitely. With a moderate amount of resistance or "friction" in the circuit the tank will still have inertia, and continue to oscillate with decreasing amplitude for a time after being given a "kick." With such a circuit, almost pure sine-wave voltage will be developed across the tank circuit even though power is supplied to the tank in short pulses or spurts, so long as the spurts are evenly spaced with respect to time and have a frequency that is the same as the resonant frequency of the tank.

Another way to visualize the action of the tank is to recall that a resonant tank with moderate Q will discriminate strongly against harmonics of the resonant frequency. The distorted plate current pulse in a Class C amplifier contains not only the fundamental frequency (that of the grid excitation voltage) but also higher harmonics. As the tank offers low impedance to the harmonics and high impedance to the fundamental (being resonant to the latter), only the fundamental — a sine-wave voltage — appears across the tank circuit in substantial magnitude.

Loaded and Unloaded Q Confusion sometimes exists as to the relationship between the unloaded and the loaded Q of the tank circuit in the plate of an r-f power amplifier. In the normal case the loaded Q of the tank circuit is determined by such factors as the operating conditions of the amplifier, bandwidth of the signal to be emitted, permissible level of harmonic radiation, and such factors. The normal value of *loaded* Q for an r-f amplifier used for communications service is from perhaps 6 to 20. The *unloaded* Q of the tank circuit determines the efficiency of the output circuit and is determined by the losses in the tank coil, its leads and plugs and jacks if any, and by the losses in the tank capacitor which ordinarily are very low. The unloaded Q of a good quality large diameter tank coil in the high-frequency range may be as high as 500 to 800, and values greater than 300 are quite common.

Tank Circuit Efficiency Since the unloaded Q of a tank circuit is determined by the minimum losses in the tank, while the loaded Q is determined by useful loading of the tank circuit from the external

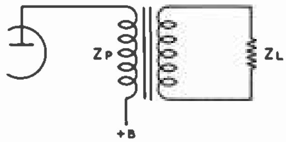


Figure 18.
IMPEDANCE-MATCHING TRANSFORMER.
The reflected impedance Z_p varies directly in proportion to the secondary load Z_L, and directly in proportion to the square of the primary-to-secondary turns ratio.

load in addition to the internal losses in the tank circuit, the relationship between the two Q values determines the operating efficiency of the tank circuit. Expressed in the form of an equation, the loaded efficiency of a tank circuit is:

$$\text{Tank efficiency} = \left(1 - \frac{Q_u}{Q_l}\right) \times 100$$

where Q_u = unloaded Q of the tank circuit
Q_l = loaded Q of the tank circuit

As an example, if the unloaded Q of the tank circuit for a class C r-f power amplifier is 400, and the external load is coupled to the tank circuit by an amount such that the loaded Q is 20, the tank circuit efficiency will be: eff. = (1 − 400/20) × 100, or (1 − 0.05) × 100, or 95 per cent. Hence 5 per cent of the power output of the class C amplifier will be lost as heat in the tank circuit and the remaining 95 per cent will be delivered to the load.

Transformers

When two coils are placed in such inductive relation to each other that the lines of force from one cut across the turns of the other and induce a current in so doing, the combination can be called a *transformer*. The name is derived from the fact that energy is transformed from one winding to another. The inductance in which the original flux is produced is called the *primary;* the inductance which *receives* the induced current is called the *secondary*. In a radio receiver power transformer, for example, the coil through which the 110-volt a.c. passes is the *primary,* and the coil from which a higher or lower voltage than the a-c line potential is obtained is the *secondary.*

Transformers can have either air or magnetic cores, depending upon whether they are to be operated at radio or audio frequencies. The reader should thoroughly impress upon his mind the fact that current can be transferred from one circuit to another *only* if the primary current is changing or alternating. From this it can be seen that a power transformer cannot possibly function as such when the primary is supplied with non-pulsating d.c.

A power transformer usually has a magnetic core which consists of laminations of iron, built up into a square or rectangular form, with a center opening or window. The secondary windings may be several in number, each perhaps delivering a different voltage. The secondary voltages will be proportional to the number of turns and to the primary voltage.

Types of Transformers Transformers are used in alternating-current circuits to transfer power at one voltage and impedance to another circuit at another voltage and impedance. There are three main classifications of transformers: those made for use in power-frequency circuits, those made for audio-frequency applications, and those made for radio frequencies. Power-frequency transformers are discussed in Chapter 25, *Power Supplies;* design and application data on power transformers is given in this chapter. The application of audio-frequency transformers is given in Chapter 5, particularly in the section devoted to *Audio Frequency Power Amplifiers.* Radio frequency transformers are also discussed in Chapter 5 in the section devoted to *Tuned R-F Voltage Amplifiers.*

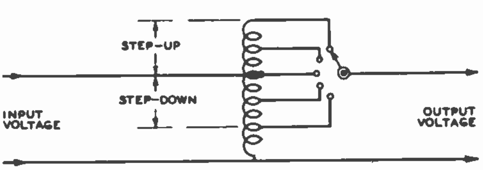


Figure 19.
THE AUTO-TRANSFORMER.
Schematic diagram of an auto-transformer showing the method of connecting it to the line and to the load. When only a small amount of step up or step down is required, the auto-transformer may be much smaller physically than would be a transformer with a separate secondary winding. Continuously variable auto-transformers (Variac's and Powerstat's) are widely used commercially.

The Auto Transformer The type of transformer in figure 19, when wound with heavy wire over an iron core, is a common device in primary power circuits for the purpose of increasing or decreasing the line voltage. In effect, it is merely a continuous winding with taps taken at various points along the winding, the input voltage being applied to the bottom and also to one tap on the winding. If the output is taken from this same tap, the voltage ratio will be 1-to-1; i.e., the input voltage will be the same as the output voltage. On the other hand, if the output tap is moved down toward the common terminal, there will be a step-down in the turns ratio with a consequent step-down in voltage.

The opposite holds true if the output terminal is moved upward from the middle input terminal; there will be a voltage step-up in this case. The initial setting of the middle input tap is chosen so that the number of turns will have sufficient reactance to keep the no-load primary current at a reasonably low value.

Electric Filters

There are many applications where it is desirable to pass a d-c component without passing a superimposed a-c component, or to pass all frequencies above or below a certain frequency while rejecting or attenuating all others, or to pass only a certain band or bands of frequencies while attenuating all others.

All of these things can be done by suitable combinations of inductance, capacitance and resistance. However, as whole books have been devoted to nothing but electric filters, it can be appreciated that it is possible only to touch upon them superficially in a general coverage book.

A filter acts by virtue of its property of offering very high impedance to the undesired frequencies, while offering but little impedance

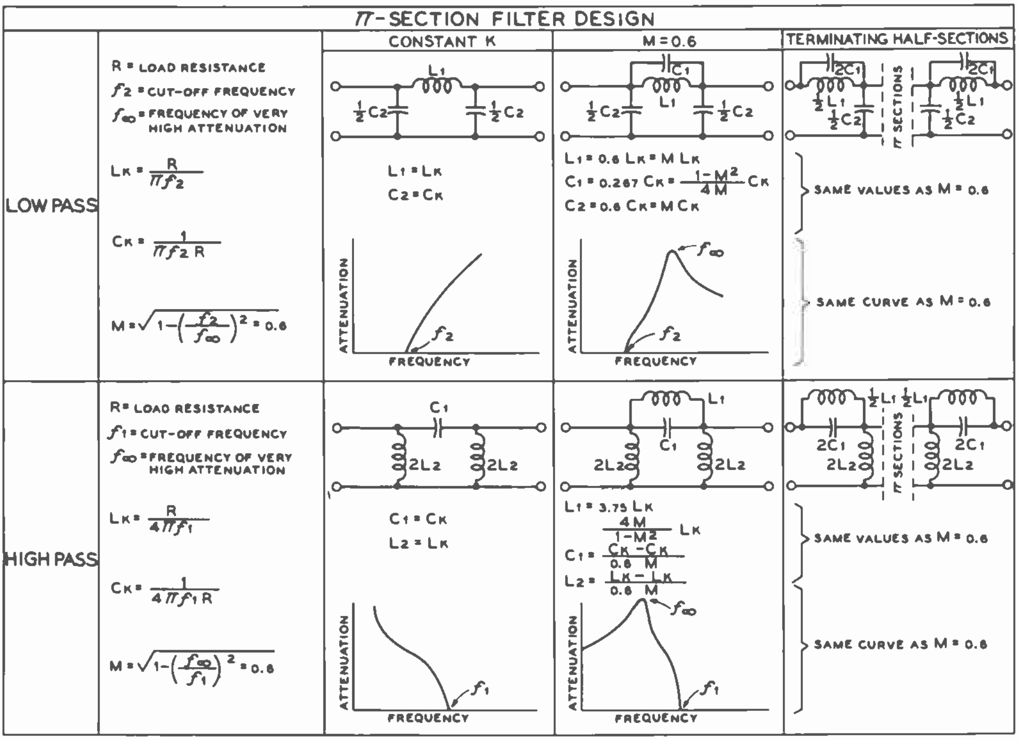


Figure 20.

Through the use of the curves and equations which accompany the diagrams in the illustration above it is possible to determine the correct values of inductance and capacitance for the usual types of pi-section filters.

to the desired frequencies. This will also apply to d.c. with a superimposed a-c component, as d.c. can be considered as an alternating current of zero frequency so far as filter discussion goes.

Types of Filters Sometimes a shunt or series element of an L-C filter is resonated with a reactance of opposite sign. When this is done, the section is known as an *M-derived* section. If the complementary reactance is added to a series arm, the section is said to be *shunt derived;* if added to the shunt arm, *series derived.*

A derived filter has sharper cut off than a regular constant K filter, but has less attenuation than the constant K section at frequencies far removed from cut off. The effect of resonating the series inductance of a π section filter to form an M-derived filter is shown in figure 20. The "notch" frequency is determined by the resonant frequency of the filter element which is tuned. The closer the resonant frequency is made to cut off, the sharper will be the cut off attenuation, but the less will be the attenuation at several times the cut off frequency.

The amount of attenuation obtained at the "notch" when a derived section is used is determined by the effective Q of the resonant arm.

Oftentimes constant-K sections and derived sections are cascaded to obtain the combined characteristic of sharp cut off and good remote-frequency attenuation. Such a filter is known as a *composite* filter.

All filters have some *insertion loss.* This is the attenuation (substantially uniform) provided to frequencies within the pass band. The insertion loss varies with the kind of filter, the Q of capacitors and inductors used, and the type termination employed.

Electric Filter Design Electric wave filters have long been used in some amateur stations in the audio channel to reduce the transmission of unwanted high frequencies and hence to reduce the bandwidth occupied by a radiophone signal. The effectiveness of a properly designed and properly used filter circuit in reducing QRM and sideband splatter should not be underestimated.

The chart of figure 20 gives design data and procedure on the pi-section type of filter. M-derived sections with an M of 0.6 will be found to be most satisfactory as the input section (or half-section) of the usual filter since the input impedance of such a section is most constant over the pass band of the filter section.

Simple filters may use either L, T, or π sections. Since the π section is the more commonly used type figure 20 gives design data and characteristics for this type of filter.

CHAPTER FOUR

Vacuum Tube Principles

In the previous chapters we have seen the manner in which an electric current flows through a metallic conductor as a result of an electron drift. This drift, which takes place when there is a difference in potential between the ends of the metallic conductor, is in addition to the normal random electron motion between the molecules of the conductor.

An electric current can be caused to flow through other media than a metallic conductor. One such medium is an ionized solution, such as the sulfuric acid electrolyte in a storage battery; this type of current flow is called electrolytic conduction. Further, it was shown at about the turn of the century that an electric current can be carried by a stream of free electrons in an evacuated chamber. The flow of a current in such a manner is said to take place by electronic conduction. The study of electron tubes (also called vacuum tubes, or valves) is actually the study of the control and use of electronic currents within an evacuated or partially evacuated chamber.

Since the current flow in an electron tube takes place in an evacuated chamber, there must be located within the enclosure both a source of electrons and a collector for the electrons which have been emitted. The electron source is called the *cathode*, and the electron collector is usually called the *anode*. Some external source of energy must be applied to the cathode in order to impart sufficient velocity to the electrons within the cathode material to enable them to overcome the surface forces and thus escape into the surrounding medium. In the usual types of electron tubes the cathode energy is applied in the form of heat; electron emission from a heated cathode is called thermionic emission. In another common type of electron tube, the photoelectric cell, energy in the form of light is applied to the cathode to cause photoelectric emission.

Thermionic Emission Emission of electrons from the cathode of a thermionic electron tube takes place when the cathode of the tube is heated to a temperature sufficiently high that the free electrons in the emitter have sufficient velocity to overcome the restraining forces at the surface of the material. These surface forces vary greatly with different materials. Hence different types of cathodes must be raised to various temperatures to obtain adequate quantities of electron emission. The several types of emitters found in common types of transmitting and receiving tubes will be described in the following paragraphs.

Cathode Types The emitters or cathodes as used in present-day thermionic electron tubes may be classified into two groups: the directly-heated or filament type and the indirectly-heated or heater-cathode type. Directly-heated emitters may be further subdivided into three important groups, all of which are commonly used in modern

Figure 1.
CONVENTIONAL ELECTRON-TUBE TYPES.

vacuum tubes. These classifications are: the pure-tungsten filament, the thoriated-tungsten filament, and the oxide-coated filament.

The Pure Tungsten Filament Pure tungsten wire was used as the filament in nearly all (platinum was used occasionally) the earlier transmitting and receiving tubes. However, the thermionic efficiency of tungsten wire as an emitter (the number of milliamperes emission per watt of filament heating power) is quite low, the filaments become fragile after use, their life is rather short, and they are susceptible to burnout at any time. Pure tungsten filaments must be run at bright white heat (about 2500° Kelvin). For these reasons, tungsten filaments have been replaced in all applications where another type of filament could be used. They are, however, still universally employed in large water-cooled tubes and in certain large, high-power air-cooled triodes where another filament type would be unsuitable. Tungsten filaments are the most satisfactory for high-power, high-voltage tubes where the emitter is subjected to positive ion bombardment due to the residual gas content of the tubes. Tungsten is not adversely affected by such bombardment.

The Thoriated-Tungsten Filament In the course of experiments made upon tungsten emitters, it was found that filaments made from tungsten having a small amount of thoria (thorium oxide) as an impurity had much greater emission than those made from the pure metal. Subsequent development has resulted in the highly efficient carburized thoriated-tungsten filament as used in virtually all medium-power transmitting tubes today.

Thoriated-tungsten emitters consist of a tungsten wire containing from 1% to 2% thoria. The activation process varies between different manufacturers of vacuum tubes, but it is essentially as follows: (1) the tube is outgassed; (2) the filament is burned for a short period at about 2800° Kelvin to clean the surface and reduce some of the thoria within the filament to metallic thorium; (3) the filament is burned for a longer period at about 2100° Kelvin to form a layer of thorium on the surface of the tungsten; (4) the temperature is reduced to about 1600° Kelvin and some pure hydrocarbon gas is admitted to form a layer of tungsten carbide on the surface of the tungsten. This layer of tungsten carbide reduces the rate of thorium evaporation from the surface at the normal operating temperature of the filament and thus increases the operating life of the vacuum tube. Thorium evaporation from the surface is a natural consequence of the operation of the thoriated-tungsten filament. The carburized layer on the tungsten wire plays another role in acting as a reducing agent to produce new thorium from the thoria to replace that lost by evaporation. This new thorium continually diffuses to the surface during the normal operation of the filament. The last process, (5), in the activation of a thoriated tungsten filament consists of re-evacuating the envelope and then burning or ageing the new filament for a considerable period of time at the normal operating temperature of approximately 1900° K.

One thing to remember about any type of filament, particularly the thoriated type, is that the emitter deteriorates practically as fast when "standing by" (no plate current) as it does with any normal amount of emission load. Also, a thoriated filament may be either temporarily or permanently damaged by a heavy overload which may strip the surface layer of thorium from the filament.

Reactivating Thoriated-Tungsten Filaments Thoriated-tungsten filaments (and *only* thoriated-tungsten filaments) which have lost emission as a result of insufficient filament voltage, a severe temporary overload, a less severe extended overload, or even normal operation may quite frequently be reactivated to their original characteristics by a process similar to that of the original activation. However, only filaments which have not approached too close to the end of their useful life may be successfully reactivated. The filament found in certain makes of tubes may be reactivated three or four times before it will cease to operate as a thoriated emitter.

The actual process of reactivation is relatively simple. The tube which has gone flat is placed in a socket to which only the two filament wires have been connected. The filament is then "flashed" for about 20 to 40 seconds at about 1½ times normal rated voltage. The filament will become extremely bright during this time and, if there is still some thoria left in the tungsten and if the tube didn't originally fail as a result of an air leak, some of this thoria will be reduced to metallic thorium. The filament is then burned at 15 to 25 per cent overvoltage for from 30 minutes to 3 to 4 hours to bring this new thorium to the surface.

The tube should then be tested to see if it shows signs of renewed life. If it does, but is still weak, the burning process should be continued at about 10 to 15 per cent overvoltage for a few more hours. This should bring it back almost to normal. If the tube checks still very low after the first attempt at reactivation, the complete process can be repeated as a last effort.

As has been mentioned above, thoriated-tungsten filaments are operated at about 1900° K or at a bright yellow heat. A burnout at normal filament voltage is almost an unheard of occurrence. The ratings placed upon tubes

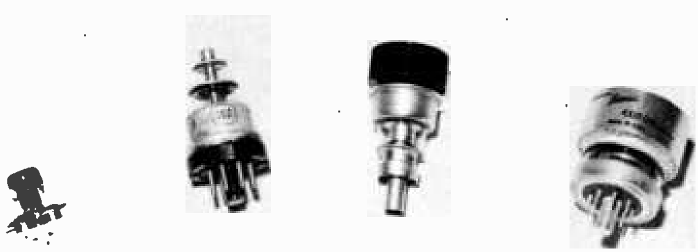


Figure 2.
V-H-F and U-H-F TUBE TYPES.

The tube to the left in this photograph is a 955 "acorn" triode. The 6F4 acorn triode is very similar in appearance to the 955 but has two leads brought out each for the grid and for the plate connection. The second tube is a 446A "lighthouse" triode. The 2C40, 2C43, and 2C44 are more recent examples of the same type tube and are essentially the same in external appearance. The third tube from the left is a 2C39 "oilcan" tube. This tube type is essentially the inverse of the lighthouse variety since the cathode and heater connections come out the small end and the plate is the large finned radiator on the large end. The use of the finned plate radiator makes the oilcan tube capable of approximately 10 times as much plate dissipation as the lighthouse type. The tube to the right is the 4X150A beam tetrode. This tube, a comparatively recent release, is capable of somewhat greater power output than any of the other tube types shown, and is rated for full output at 500 Mc. and at reduced output at frequencies greater than 1000 Mc.

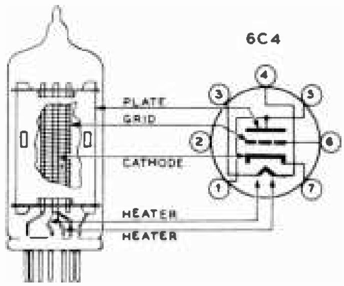


Figure 3.
CUT-AWAY DRAWING OF A 6C4 TRIODE.

by the manufacturers are figured for a life expectancy of 1000 hours. Certain types may give much longer life than this but the average transmitting tube will give from 1000 to 5000 hours of useful life.

The Oxide-Coated Filament The most *efficient* of all modern filaments is the oxide-coated type which consists of a mixture of barium and strontium oxides coated upon a wire or strip usually of a nickel alloy. This type of filament operates at a dull-red to orange-red temperature (1050° to 1170° K) at which temperature it will emit large quantities of electrons. The oxide-coated filament is somewhat more efficient than the thoriated-tungsten type in small sizes and it is considerably less expensive to manufacture. For this reason all receiving tubes and quite a number of the low-powered transmitting tubes use the oxide-coated filament. Another advantage of the oxide-coated emitter is its extremely long life — the average tube can be expected to run from 3000 to 5000 hours, and when loaded very lightly, tubes of this type have been known to give 50,000 hours of life before their characteristics changed to any great extent.

Oxide filaments are unsatisfactory for use at high *continuous* plate voltages because: (1) their activity is seriously impaired by the high temperature necessary to de-gas the high-voltage tubes and, (2) the positive ion bombardment which takes place even in the best evacuated high-voltage tube causes destruction of the oxide layer on the surface of the filament.

Oxide-coated emitters have been found capable of emitting an enormously large current pulse with a high applied voltage for a very short period of time without damage. This characteristic has proved to be of great value in radar work. For example, the relatively small cathode in a microwave magnetron may be called upon to deliver 25 to 50 *amperes* at an applied voltage of perhaps 25,000 volts for a period in the order of one microsecond. After this large current pulse has been passed, plate voltage normally will be removed for 1000 microseconds or more so that the cathode surface may be restored in time for the next pulse of current. If the cathode were to be subjected to a continuous current drain of this magnitude, it would be destroyed in an exceedingly short period of time.

The activation of oxide-coated filaments also varies with tube manufacturers but consists essentially in heating the wire which has been coated with a mixture of barium and strontium carbonates to a temperature of about 1500° Kelvin for a time and then applying a potential of 100 to 200 volts through a protective resistor to limit the emission current. This process reduces the carbonates to oxides thermally, cleans the filament surface of foreign materials, and activates the cathode surface.

Reactivation of oxide-coated filaments is not possible since there is always more than sufficient reduction of the oxides and diffusion of the metals to the surface of the filament to meet the emission needs of the cathode.

The Heater Cathode The heater type cathode was developed as a result of the requirement for a type of emitter which could be operated from alternating current and yet would not introduce a-c ripple modulation even when used in low-level stages. It consists essentially of a small nickel-alloy cylinder with a coating of strontium and barium oxides on its surface similar to the coating used on the oxide-coated filament. Inside the cylinder is an insulated heater element consisting usually of a double spiral of tungsten wire. The heater may operate on any voltage from 2 to 117 volts, although 6.3 is by far the most common value. The heater is operated at quite a high temperature so that the cathode itself usually may be brought to operating temperature in a matter of 15 to 30 seconds. Heat coupling between the heater and the cathode is mainly by radiation, although there is some thermal conduction through the insulating coating on the heater wire, as this coating is also in contact with the cathode thimble.

Indirectly heated cathodes are employed in all a-c operated tubes which are designed to operate at a low level either for r-f or a-f use. However, some receiver power tubes use heater cathodes (6L6, 6V6, 6F6, and 6K6-GT) as do some of the low-power transmitter tubes (802, 807, 815, 3E29, 2E26, 5763, etc.). Heater cathodes are employed almost exclusively when a number of tubes are to be operated in series as in an a-c–d-c receiver. A heater cathode is often called a uni-potential cathode because there is no voltage drop along its length as there is in the directly-heated or filament cathode.

The Emission Equation The emission of electrons from a heated cathode is quite similar to the evaporation of molecules from the surface of a liquid. The molecules which leave the surface are those having sufficient kinetic (heat) energy to overcome the forces at the surface of the liquid. As the temperature of the liquid is raised, the average velocity of the molecules is increased, and a greater number of molecules will acquire sufficient energy to be evaporated. The evaporation of electrons from the surface of a thermionic emitter is similarly a function of average electron velocity, and hence is a function of the temperature of the emitter.

Electron emission per unit area of emitting surface is a function of the temperature T in degrees Kelvin, the work function of the emitting surface b (which is a measure of the surface forces of the material and hence of the energy required of the electron before it may escape), and of the constant A which also varies with the emitting surface. The relationship between emission current in amperes per square centimeter, I, and the above quantities can be expressed as:

$$I = AT^2\epsilon^{-b/T} \qquad (1)$$

Secondary Emission The bombarding of most metals and a few insulators by electrons will result in the emission of other electrons by a process called *secondary emission.* The secondary electrons are literally knocked from the surface layers of the bombarded material by the primary electrons which strike the material. The number of secondary electrons emitted per primary electron varies from a very small percentage to as high as 5 to 10 secondary electrons per primary.

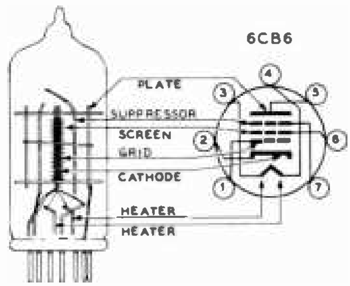


Figure 4.
CUT-AWAY DRAWING OF A 6CB6 PENTODE.

The phenomena of secondary emission is undesirable for most thermionic electron tubes. However, the process is used to advantage in certain types of electron tubes such as the image orthicon (TV camera tube) and the electron-multiplier type of photo-electric cell. In types of electron tubes which make use of secondary emission, such as the type 931 photo cell, the secondary-electron-emitting surfaces are specially treated to provide a high ratio of secondary to primary electrons. Thus a high degree of current amplification in the electron-multiplier section of the tube is obtained.

The Space Charge Effect As a cathode is heated so that it begins to emit, those electrons which have been discharged into the surrounding space form a negatively charged cloud in the immediate vicinity of the cathode. This cloud of electrons around the cathode is called the *space charge.* The electrons comprising the charge are continuously changing, since those electrons making up the original charge fall back into the cathode and are replaced by others emitted by it.

The Diode If a cathode capable of being heated either indirectly or directly is placed in an evacuated envelope along with a plate, such a two-element vacuum tube is called a diode. The diode is the simplest of all vacuum tubes and is the fundamental type from which all the others are derived; hence, the diode and its characteristics will be discussed first.

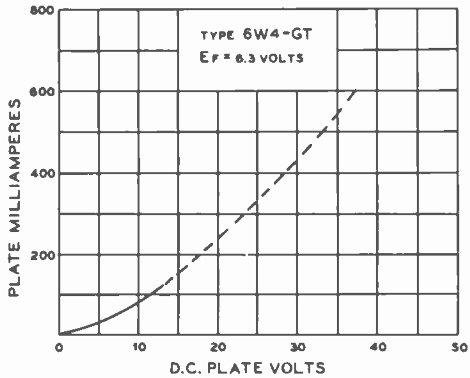


Figure 5.
AVERAGE PLATE CHARACTERISTICS OF A POWER DIODE.

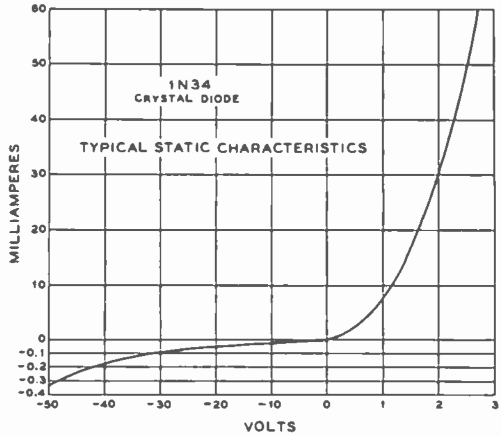


Figure 6.
TYPICAL CHARACTERISTICS OF A CRYSTAL DIODE.

Characteristics of the Diode When the cathode within a diode is heated, it will be found that a few of the electrons leaving the cathode will leave with sufficient velocity to reach the plate. If the plate is electrically connected back to the cathode, the electrons which have had sufficient velocity to arrive at the plate will flow back to the cathode through the external circuit. This small amount of initial plate current is an effect found in all two-element vacuum tubes.

If a battery or other source of d-c voltage is placed in the external circuit between the plate and cathode so that it places a positive potential on the plate, the flow of current from the cathode to plate will be increased. This is due to the strong attraction offered by the positively charged plate for any negatively charged particles.

Space-Charge-Limited Current At moderate values of plate voltage the current flow from cathode to anode is limited by the space charge of electrons around the cathode. Increased values of plate voltage will tend to neutralize a greater portion of the cathode space charge and hence will cause a greater current to flow.

Under these conditions, with plate current limited by the cathode space charge, the plate current is not linear with plate voltage. In fact it may be stated in general that the plate-current flow in electron tubes does not obey Ohm's Law. Rather, plate current increases as the three-halves power of the plate voltage. The relationship between plate voltage, E, and plate current, I, can be expressed as:

$$I = K\,E^{3/2} \qquad (2)$$

where K is a constant determined by the geometry of the element structure within the electron tube.

Plate Current Saturation As plate voltage is raised to the potential where the cathode space charge is neutralized, all the electrons that the cathode is capable of emitting are being attracted to the plate. The electron tube is said then to have reached *saturation* plate current. Further increase in plate voltage will cause only a relatively small increase in plate current. The initial point of plate current saturation is sometimes called the point of Maximum Space-Charge-Limited Emission (MSCLE).

The degree of flattening in the plate-voltage plate-current curve after the MSCLE point will vary with different types of cathodes. This effect is shown in figure 7. The flattening is quite sharp with a pure tungsten emitter. With thoriated tungsten the flattening is smoothed somewhat, while with an oxide-coated cathode the flattening is quite gradual. The gradual saturation in emission with an oxide-coated emitter is generally considered to result from a lowering of the surface work function by the field at the cathode resulting from the plate potential.

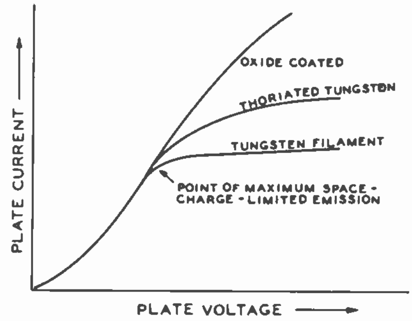


Figure 7.
MAXIMUM SPACE-CHARGE-LIMITED EMISSION FOR DIFFERENT TYPES OF EMITTERS.

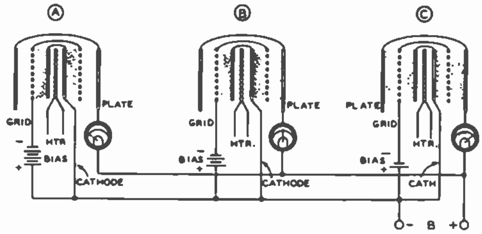


Figure 8.
ACTION OF THE GRID IN A TRIODE.
(A) shows the triode tube with cutoff bias on the grid. Note that all the electrons emitted by the cathode remain inside the grid mesh. (B) shows the same tube with an intermediate value of bias on the grid. Note the medium value of plate current and the fact that there is a reserve of electrons remaining within the grid mesh. (C) shows the operation with a relatively small amount of bias which with certain tube types will allow substantially all the electrons emitted by the cathode to reach the plate. Emission is said to be saturated in this case. In a majority of tube types a high value of positive grid voltage is required before plate-current saturation takes place.

Another effect encountered with oxide-coated cathodes is a relatively enormous short-term emission capability. The cathodes used in radar magnetrons are capable of emission current density up to 100 amperes per square centimeter for a period of the order of one microsecond. After the emission of a large current pulse the cathode must be permitted to rest for several hundred microseconds before it is capable of emitting the next pulse.

Electron Energy Dissipation The current flowing in the plate-cathode space of a conducting electron tube represents the energy required to accelerate electrons from the zero potential of the cathode space charge to the potential of the anode. Then, when these accelerated electrons strike the anode, the energy associated with their velocity is immediately released to the anode structure. In normal electron tubes this energy release appears as heating of the plate or anode structure.

The Triode If an element consisting of a mesh or spiral of wire is inserted concentric with the plate and between the plate and the cathode, such an element will be able to control by electrostatic action the cathode-to-plate current of the tube. The new element is called a grid, and a vacuum tube containing a cathode, grid, and plate is commonly called a triode.

If this new element through which the electrons must pass in their course from cathode to plate is made negative with respect to the cathode, the negative charge on this grid will effectively repel the negatively charged electrons (like charges repel; unlike charges attract) back into the space charge surrounding the cathode. Hence, the number of electrons which are able to pass through the grid mesh and reach the plate will be reduced, and the plate current will be reduced accordingly. As a matter of fact, if the charge on the grid is made sufficiently negative, all the electrons leaving the cathode will be repelled back to it and the plate current will be reduced to zero. Any d-c voltage placed upon a grid is called a *bias* (especially so when speaking of a control grid). The smallest negative voltage which will cause cutoff of plate current at a particular plate voltage is called the value of *cutoff bias.*

The amount of plate current in a triode is a result of the net field at the cathode from interaction between the field caused by the grid bias and that caused by the plate voltage. Hence, both grid bias and plate voltage affect the plate current. In all normal tubes a small change in grid bias has a considerably greater effect than a similar change in plate voltage. The ratio between the change in grid bias and the change in plate current which will cause the same small change in plate current is

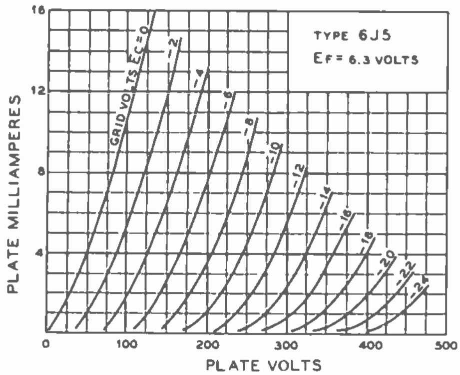


Figure 9.

NEGATIVE-GRID CHARACTERISTICS OF A TYPICAL TRIODE.

Average plate characteristics of this type are most commonly used in determining the Class A operating characteristics of a triode amplifier stage.

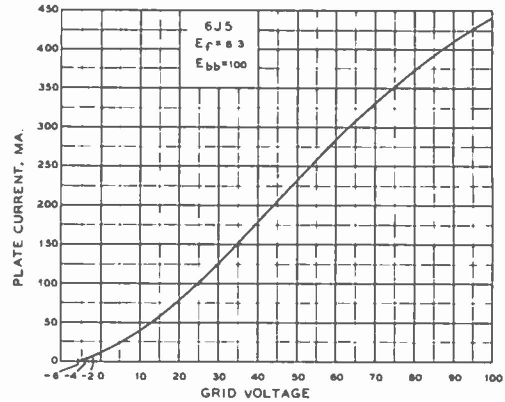


Figure 10.

POSITIVE-GRID CHARACTERISTICS OF A TYPICAL TRIODE.

Plate characteristics of this type are most commonly used in determining the pulse-signal operating characteristics of a triode amplifier stage. Note the large emission capability of the oxide-coated heater cathode in tubes of the general type of the 6J5.

called the *amplification factor* or μ of the electron tube. Expressed as an equation:

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

with i_p constant (Δ represents a small change).

The μ can be determined experimentally by making a small change in grid bias, thus slightly changing the plate current. The plate current is then returned to the original value by making a change in the plate voltage. The ratio of the change in plate voltage to the change in grid voltage is the μ of the tube under the operating conditions chosen for the test.

Current Flow in a Triode In a diode it was shown that the electrostatic field at the cathode was proportional to the plate potential, E_p, and that the total cathode current was proportional to the three-halves power of the plate voltage. Similarly, in a triode it can be shown that the field at the cathode space charge is proportional to the equivalent voltage ($E_g + E_p/\mu$), where the amplification factor, μ, actually represents the relative efficacy of grid potential and plate potential in producing a field at the cathode.

It would then be expected that the cathode current in a triode would be proportional to the three-halves power of ($E_g + E_p/\mu$). This has been proved to be true. The cathode current of a triode can be represented with fair accuracy by the expression:

$$\text{Cathode current} = K\left(E_g + \frac{E_p}{\mu}\right)^{3/2} \qquad (3)$$

where K is a constant determined by element geometry within the triode.

Plate Resistance The plate resistance of a vacuum tube is the ratio of a change in plate voltage to the change in plate current which the change in plate voltage produces. To be accurate, the changes should be very small with respect to the operating values. Expressed as an equation:

$$R_p = \frac{\Delta E_p}{\Delta I_p} \qquad E_g = \text{constant} \quad \Delta = \text{small change}$$

The plate resistance can also be determined by the experiment mentioned above. By noting the change in plate current as it occurs when the plate voltage is changed (grid voltage held constant), and by dividing the latter by the former, the plate resistance can be determined. Plate resistance is expressed in ohms.

Transconductance The mutual conductance, also referred to as *trans-*

conductance, is the ratio of a change in the plate current to the change in grid voltage which brought about the plate current change, the plate voltage being held constant. Expressed as an equation:

$$G_m = \frac{\Delta I_p}{\Delta E_g} \qquad E_p = \text{constant} \quad \Delta = \text{small change}$$

The transconductance is also numerically equal to the amplification factor divided by the plate resistance. $G_m = \mu / R_p$.

Transconductance is most commonly expressed in microreciprocal-ohms or *micromhos.* However, since transconductance expresses change in plate current as a function of a change in grid voltage, a tube is often said to have a transconductance of so many milliamperes-per-volt. If the transconductance in milliamperes-per-volt is multiplied by 1000 it will then be expressed in micromhos. Thus the transconductance of a 6A3 could be called either 5.25 ma./volt or 5250 micromhos.

Tetrode or Screen-Grid Tube In the preceding chapter it was mentioned that two conductors separated by a dielectric form a *capacitor,* or that there is *capacitance* between them. Since the electrodes in a vacuum tube are conductors and they are separated by a dielectric, vacuum, there is capacitance between them. Although the interelectrode capacitances are so small as to be of little consequence in audio-frequency work, they are large enough to be of considerable importance when triodes are operated at radio frequencies.

The quest for a simple and easily usable method of eliminating the effects of the grid-to-plate capacitance of the triode led to the development of the *screen-grid* tube or *tetrode.* When another grid is added between the grid and plate of a vacuum tube the tube is called a tetrode, and because the new grid is called a *screen,* as a result of its screening or shielding action, the tube is often called a screen-grid tube. The interposed screen grid acts as an electrostatic shield between the grid and plate, with the consequence that the grid-to-plate capacitance is reduced. Although the screen grid is maintained at a positive voltage with respect to the cathode of the tube, it is maintained at ground potential with respect to r.f. by means of a by-pass capacitor of very low reactance at the frequency of operation.

In addition to the shielding effect, the screen grid serves another very useful purpose. Since the screen is maintained at a positive potential, it serves to increase or accelerate the flow of electrons to the plate. There being large openings in the screen mesh, most of the electrons pass through it and on to the plate. Due also to the screen, the plate current is largely independent of plate voltage, thus making for high amplification. When the screen voltage is held at a constant value, it is possible to make large changes in plate voltage without appreciably affecting the plate current.

When the electrons from the cathode approach the plate with sufficient velocity, they dislodge electrons upon striking the plate. This effect of *bombarding* the plate with high velocity electrons, with the consequent dislodgement of other electrons from the plate, gives rise to the condition of secondary emission which has been discussed in a previous paragraph. This effect can cause no particular difficulty in a triode because the secondary electrons so emitted are eventually attracted back to the plate. In the screen-grid tube, however, the screen is close to the plate and is maintained at a positive potential. Thus, the screen will attract these electrons which have been knocked from the plate, particularly when the plate voltage falls to a lower value than the screen voltage, with the result that the plate current is lowered and the amplification is decreased.

The Pentode The undesirable effects of secondary emission from the plate can be greatly reduced if another element is added between the screen and plate. This additional element is called a *suppressor,* and tubes in which it is used are called *pentodes.* The suppressor grid is sometimes connected to cathode within the tube, sometimes it is brought out to a connecting pin on the tube base, but in any case it is established negative with respect to the minimum plate voltage. The secondary electrons that would travel to the screen if there were no suppressor are diverted back to the plate. The plate current is, therefore, not reduced and the amplification possibilities are increased.

Pentodes for audio applications are designed so that the suppressor increases the limits to which the plate voltage may swing; therefore the consequent power output and gain can be very great. Pentodes for radio-frequency service function in such a manner that the suppressor allows high voltage gain, at the same time permitting fairly high gain at low plate voltage. This holds true even if the plate voltage is the same or slightly lower than the screen voltage.

Beam Power Tubes A beam power tube makes use of another method for suppressing secondary emission. In this tube there are four electrodes: a cathode, a grid, a screen, and a plate, so spaced and placed that secondary emission from the plate is suppressed without actual power. Because of the manner in which the electrodes are spaced, the electrons which travel to the plate are slowed down when the plate voltage is low, almost to zero velocity in a certain region between screen and plate. For this reason the electrons form a stationary cloud, a *space charge*. The effect of this space charge is to repel secondary electrons emitted from the plate and thus cause them to return to the plate. In this way, secondary emission is suppressed.

Another feature of the beam power tube is the low current drawn by the screen. The screen and the grid are spiral wires wound so that each turn in the screen is shaded from the cathode by a grid turn. This alignment of the screen and the grid causes the electrons to travel in sheets between the turns of the screen so that very few of them strike the screen itself. This forming of the electron stream into sheets or beams increases the charge density in the screen-plate region and assists in the formation of the space charge in this region.

Because of the effective suppressor action provided by the space charge, and because of the low current drawn by the screen, the beam power tube has the advantages of high power output, high power sensitivity, and high efficiency. The 6L6 is such a beam power tube, designed for use in the power amplifier stages of receivers and speech amplifiers or modulators. Larger tubes employing the beam-power principle are being made by various manufacturers for use in the radio-frequency stages of transmitters. These tubes feature extremely high power sensitivity (a very small amount of driving power is required for a large output), good plate efficiency, and low grid-to-plate capacitance.

Grid-Screen Mu Factor The grid-screen mu factor (μ_{sg}) is analogous to the amplification factor in a triode, except that the screen of a pentode or tetrode is substituted for the plate of a triode. μ_{sg} denotes the ratio of a change in grid voltage to a change in screen voltage, each of which will produce the same change in screen current. Expressed as an equation:

$$\mu_{sg} = \frac{\Delta E_{sg}}{\Delta E_g} \qquad I_{sg} = \text{constant} \quad \Delta = \text{small change}$$

The grid-screen mu factor is important in determining the operating bias of a tetrode or pentode tube. The relationship between control-grid potential and screen potential determines the plate current of the tube as well as the screen current since the plate current is essentially independent of the plate voltage in tubes of this type. In other words, when the tube is operated at cutoff bias as determined by the screen voltage and the grid-screen mu factor (determined in the same way as with a triode, by dividing the operating voltage by the mu factor) the plate current will be substantially at cutoff as will be the screen current. The grid-screen mu factor is numerically equal to the amplification factor of the same tetrode or pentode tube when triode connected.

Current Flow in Tetrodes and Pentodes Equation (3) is the expression for total cathode current in a triode tube. The expression for the total cathode current of tetrode and pentode tubes is the same, except that the screen-grid voltage and the grid-screen μ factor are used in place of the plate voltage and μ of the triode. The equation is:

$$\text{Cathode current} = K\left(E_g + \frac{E_{sg}}{\mu_{sg}}\right)^{3/2} \quad (4)$$

Cathode current, of course, is the sum of the screen and plate current, plus control grid current in the event that the control grid is positive with respect to the cathode. It will be noted that total *cathode* current is independent of plate voltage in a tetrode or pentode. Also, in the usual tetrode or pentode the *plate* current also is substantially independent of plate voltage over the usual operating range—which means simply that the effective plate resistance of such tubes is relatively high. However, when the plate voltage falls below the normal operating range the plate current falls sharply while the screen current rises to such a value that the total cathode current remains substantially constant. Hence, the screen grid in a tetrode or pentode will almost invariably be damaged by excessive dissipation if the plate voltage is removed while the screen voltage is still being applied from a low-impedance source.

The Effect of Grid Current Equations (3) and (4) have shown how the total cathode current in triodes, tetrodes, and pentodes is a function of the potentials applied to the various electrodes. Hence, if only one electrode is positive with respect to the cathode (such as would be the case in a triode acting as a class A amplifier) all the cathode current goes to the plate. But when both screen and plate are positive in a tetrode or pentode, the cathode current divides between the two elements. Hence the screen current is taken from the total cathode current, while the balance goes to the plate. Further, if the control grid in a tetrode or pentode is operated at a positive potential the total cathode current is divided between all three elements which have a positive potential. Hence, in a tube which is receiving a large excitation voltage, it may be said that the control grid robs electrons from the output electrode during the period that the grid is positive. Hence it always is necessary to limit the peak positive excursion of the control grid.

Coefficients of Tetrodes and Pentodes In general it may be stated that the amplification factor of tetrode and pentode tubes is a coefficient which is not of much use to the designer. In fact the amplification factor is seldom given on the design data sheets of such tubes. Its value is usually very high, due to the relatively high plate resistance of such tubes, but bears little relationship to the stage gain which actually will be obtained with such tubes.

On the other hand, the grid-plate transconductance is the most important coefficient of pentode and tetrode tubes. Gain per stage can be computed directly when the G_m is known. The grid-plate transconductance of a tetrode or pentode tube can be calculated through use of the expression:

$$G_m = \frac{\Delta I_p}{\Delta E_g},$$

with E_{sg} and E_p constant.

The plate resistance of such tubes is of less importance than in the case of triodes, though it is often of value in determining the amount of damping a tube will exert upon the impedance in its plate circuit. Plate resistance is calculated from:

$$R_p = \frac{\Delta E_p}{\Delta I_p},$$

with E_G and E_{sg} constant.

Mixer and Converter Tubes The superheterodyne receiver will be described in detail in Chapter Six, but suffice to say here that such a receiver always includes at least one stage for changing the frequency of the incoming signal to the fixed frequency of the main intermediate amplifier in the receiver. The frequency changing process is accomplished by selecting the beat-note difference frequency between a locally generated oscillation and the incoming signal frequency. If the oscillator signal is supplied by a separate tube, the frequency changing tube is called a *mixer*. Alternatively, the oscillation may be generated by additional elements within the frequency changer tube. In this case the frequency changer is commonly called a *converter* tube.

Conversion Conductance The conversion conductance (G_c) is a coefficient of interest in the case of mixer or converter tubes, or of conventional triodes, tetrodes, or pentodes operating as frequency changers. The conversion conductance is the ratio of a change in the signal-grid voltage at the input frequency to a change in the output current at the converted frequency. Hence G_c in a mixer is essentially the same as transconductance in amplifier with the exception that the input signal and the output current are on different frequencies. The value of G_c in conventional mixer tubes is from 300 to 1000 micromhos. The value of G_c in an amplifier tube operated as a mixer is approximately 0.3 times the G_m of the tube operated as an amplifier. The voltage gain of a mixer stage is equal to G_cZ_L where Z_L is the impedance of the plate load into which the mixer tube operates.

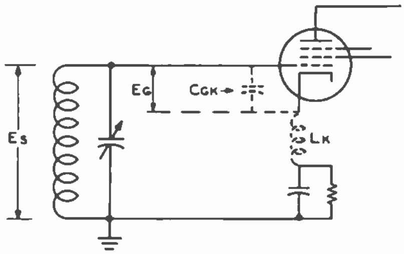


Figure 11.

SHOWING THE EFFECT OF CATHODE LEAD INDUCTANCE.

The degenerative action of cathode lead inductance tends to reduce the effective grid-to-cathode voltage with respect to the voltage available across the input tuned circuit. Cathode lead inductance also introduces undesirable coupling between the input and the output circuits.

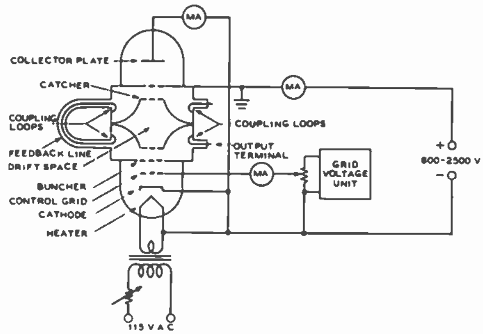


Figure 12.

TWO-CAVITY KLYSTRON OSCILLATOR.

A conventional two-cavity klystron is shown with a feedback loop connected between the two cavities so that the tube may be used as an oscillator.

Electron Tubes at Very High Frequencies As the frequency of operation of the usual type of electron tube is increased above about 20 Mc., certain assumptions which are valid for operation at lower frequencies must be re-examined. First, we find that lead inductances from the socket connections to the actual elements within the envelope no longer are negligible. Second, we find that electron transit time no longer may be ignored; an appreciable fraction of a cycle of input signal may be required for an electron to leave the cathode space charge, pass through the grid wires, and travel through the space between grid and plate.

Effects of Lead Inductance The effect of lead inductance is two-fold. First, as shown in figure 11, the combination of grid-lead inductance, grid-cathode capacitance, and cathode lead inductance tends to reduce the effective grid-cathode signal voltage for a constant voltage at the tube terminals as the frequency is increased. Second, cathode lead inductance tends to introduce undesired coupling between the various elements within the tube.

Tubes especially designed for v-h-f and u-h-f use have had their lead inductances minimized. The usual procedures for reducing lead inductance are: (1) using heavy lead conductors or several leads in parallel (examples are the 6SH7 and 6AK5), (2) scaling down the tube in all dimensions to reduce both lead inductances and interelectrode capacitances (examples are the 6AK5, 6F4, and other acorn and miniature tubes), and (3) the use of very low inductance extensions of the elements themselves as external connections (examples are lighthouse tubes such as the 2C40, oilcan tubes such as the 2C29, and many types of v-h-f transmitting tubes).

Effects of Transit Time When an electron tube is operated at a frequency high enough so that electron transit time between cathode and plate is an appreciable fraction of a cycle at the input frequency, several undesirable effects take place. First, the grid takes power from the input signal even though the grid is negative at all times. This comes about since the grid will have changed its potential during the time required for an electron to pass from cathode to plate. Due to interaction and a resulting phase difference between the field associated with the grid and that associated with a moving electron, the grid presents a resistance to an input signal in addition to its normal "cold" capacitance. Further, as a result of this action, plate current no longer is in phase with grid voltage.

An amplifier stage operating at a frequency high enough so that transit time is appreciable:

(a) Is difficult to excite as a result of grid loss from the equivalent input grid resistance,

(b) Is capable of less output since transconductance is reduced and plate current is not in phase with grid voltage.

The effects of transit time increase with the square of the operating frequency — hence they increase rapidly as frequency is increased above the value where they become just appreciable. These effects may be reduced by scaling down tube dimensions — a procedure which also reduces lead inductance. Further, transit-time effects may be reduced by the obvious procedure of increasing electrode potentials so that electron velocity will be increased. However, due to the law of electron motion in an electric field, transit time is increased only as the square root of the ratio of operating potential increase; hence this expedient is of limited value due to other limitations upon operating voltages of small electron tubes.

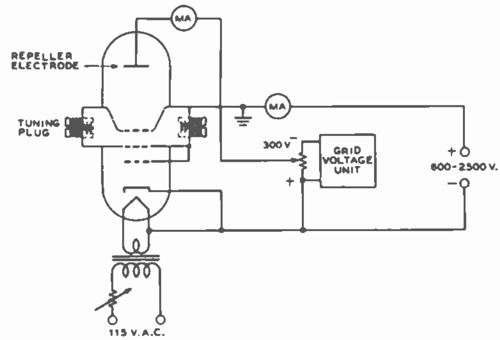


Figure 13.

REFLEX KLYSTRON OSCILLATOR.

A conventional reflex klystron oscillator of the type commonly used as a local oscillator in superheterodyne receivers operating above about 2000 Mc. is shown above. Frequency modulation of the output frequency of the oscillator, or a-f-c operation in a receiver, may be obtained by varying the negative voltage on the repeller electrode.

Special Microwave Electron Tubes Due primarily to the limitation imposed by transit time, conventional negative-grid electron tubes are capable of affording worthwhile amplification and power output only to a definite upper frequency. This upper frequency limit varies from perhaps 100 Mc. for conventional tube types to about 4000 Mc. for specialized types such as the lighthouse tube. Above the limiting frequency the conventional negative-grid tube no longer is practicable and recourse must be taken to totally different types of electron tubes in which electron transit time is not a limitation to operation. Three of the most important of such microwave tube types are the ***klystron***, the ***magnetron***, and the ***travelling-wave tube.***

The Power Klystron The klystron is a type of electron tube in which electron transit time is used to advantage. Such tubes comprise, as shown in figure 12, a cathode, a focussing electrode, a resonator connected to a pair of grids which afford *velocity modulation* of the electron beam (called the "buncher"), a *drift space*, and another resonator connected to a pair of grids (called the "catcher"). A *collector* for the expended electrons may be included at the end of the tube, or the catcher may also perform the function of electron collection.

The tube operates in the following manner: The cathode emits a stream of electrons which is focussed into a beam by the focussing electrode. The stream passes through the buncher where it is acted upon by any field existing between the two grids of the buncher cavity. When the potential between the two grids is zero, the stream passes through without change in velocity. But when the potential between the two grids of the buncher is increasingly positive in the direction of electron motion, the velocity of the electrons in the beam is increased. Conversely, when the field becomes increasingly negative in the direction of the beam (corresponding to the other half cycle of the exciting voltage from that which produced electron acceleration) the velocity of the electrons in the beam is decreased.

When the velocity-modulated electron beam reaches the drift space, where there is no field, those electrons which have been sped up on one half-cycle overtake those immediately ahead which were slowed down on the other half-cycle. In this way, the beam electrons become bunched together. As the bunched groups pass through the two grids of the catcher cavity, they impart pulses of energy to these grids. The catcher grid-space is charged to different voltage levels by the passing electron bunches, and a corresponding oscillating field is set up in the catcher cavity. The catcher is designed to resonate at the frequency of the velocity-modulated beam, or at a harmonic of this frequency.

In the klystron amplifier, energy delivered

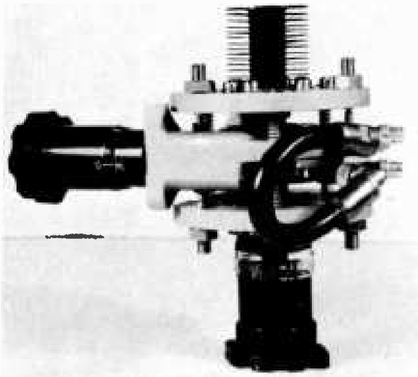


Figure 14.
TUNABLE KLYSTRON OSCILLATOR.
Showing a two-cavity klystron oscillator tube with its tuning mechanism (Sperry).

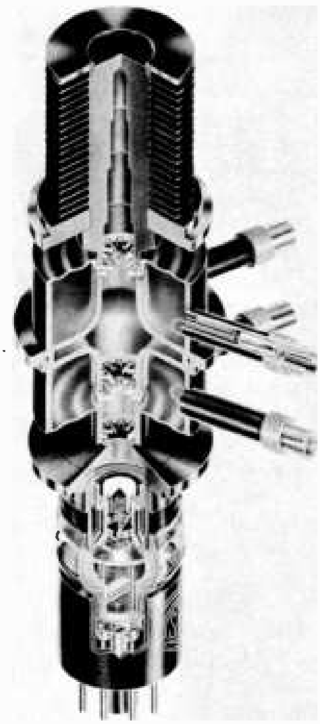


Figure 15.
CUT-AWAY VIEW OF A TUNABLE KLYSTRON.
The tuning mechanism has been removed in this cut-away view of a tunable klystron tube. The flexible diaphragms which serve as the top enclosure of the top cavity and the bottom enclosure of the bottom cavity allow the volumes of the two cavities to be varied by the tuning mechanism.

by the buncher to the catcher grids is greater than that applied to the buncher cavity by the input signal. In the klystron oscillator (figure 12), a feedback loop connects the two cavities. Coupling to either buncher or catcher is provided by small loops which enter the cavities by way of concentric lines, as shown in figure 12.

The klystron is an electron-coupled device. When used as an oscillator, its output voltage is rich in harmonics. Klystron oscillators of various types afford power outputs ranging from less than 1 watt to many thousand watts. Operating efficiency varies between 5 and 30 per cent. Frequency may be shifted to some extent by varying the beam voltage. Tuning is carried on mechanically in some klystrons by altering (by means of knob settings) the shape of the resonant cavity.

The Reflex Klystron The two-cavity klystron as described in the preceding paragraphs is primarily used as a transmitting device since quite reasonable amounts of power are made available in its output circuit. However, for applications where a much smaller amount of power is required—power levels in the milliwatt range—for low-power transmitters, receiver local oscillators, etc., another type of klystron having only a single cavity is more frequently used.

The theory of operation of the single-cavity klystron is essentially the same as the multicavity type with the exception that the velocity-modulated electron beam, after having left the "buncher" cavity is reflected back into the area of the buncher again by a repeller electrode as illustrated in figure 13. The potentials on the various electrodes are adjusted to the value such that proper bunching of the electron beam will take place just as a particular portion of the velocity-modulated beam reenters the area of the resonant cavity. Since this type of klystron has only one circuit it

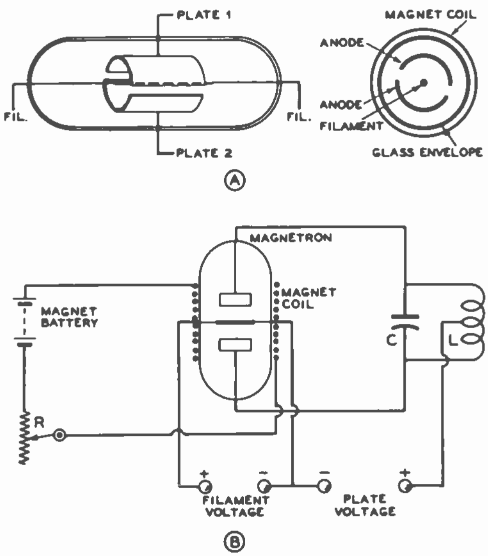


Figure 16.
SIMPLE MAGNETRON OSCILLATOR.
An external tank circuit is used with this type of magnetron oscillator for operation in the lower u-h-f range.

can be used only as an oscillator and not as an amplifier. Effective modulation of the frequency of a single-cavity klystron for FM work can be obtained by modulating the repeller electrode voltage.

The Magnetron The magnetron is an s-h-f oscillator tube normally employed where very high values of peak power or moderate amounts of average power are required in the range from perhaps 700 Mc. to 30,000 Mc. Special magnetrons were developed for wartime use in radar equipments which had peak power capabilities of several million watts (megawatts) output at frequencies in the vicinity of 3000 Mc. The normal duty cycle of operation of these radar equipments was approximately 1/10 of one per cent (the tube operated about 1/1000 of the time and rested for the balance of the operating period) so that the average power output of these magnetrons was in the vicinity of 1000 watts.

In its simplest form the magnetron tube is a filament-type diode with two half-cylindrical plates or anodes situated coaxially with respect to the filament. The construction is illustrated in figure 16A. The anodes of the magnetron are connected to a resonant circuit as illustrated on figure 16B. The tube is surrounded by an electromagnet coil which, in turn, is connected to a low-voltage d-c energizing source through a rheostat R for controlling the strength of the magnetic field. The field coil is oriented so that the lines of magnetic force it sets up are parallel to the axis of the electrodes.

Under the influence of the strong magnetic field, electrons leaving the filament are deflected from their normal paths and move in circular orbits within the anode cylinder. This effect results in a negative resistance which sustains oscillations. The oscillation frequency is very nearly the value determined by L and C, in figure 16B. In other magnetron circuits, the frequency may be governed by the electron rotation, no external tuned circuits being employed. Wavelengths of less than 1 centimeter have been produced with such circuits.

More complex magnetron tubes employ no external tuned circuit, but utilize instead one or more resonant cavities which are integral with the anode structure. Figure 17 shows a magnetron of this type having a multi-cellular anode of eight cavities. It will be noted, also, that alternate cavities (which would operate at the same polarity when the tube is oscillating) are strapped together. Strapping was found to improve the efficiency and stability of high-power radar magnetrons. In most radar applications of magnetron oscillators a powerful permanent magnet of controlled characteristics is employed to supply the magnetic field rather than to use an electromagnet.

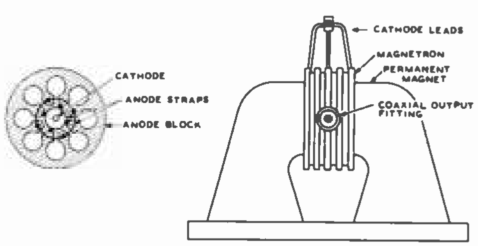


Figure 17.
MODERN MULTI-CAVITY MAGNETRON.
Illustrated is an external-anode strapped magnetron of the type commonly used in radar equipment for the 10-cm. range. A permanent magnet of the general type used with such a magnetron is shown in the right-hand portion of the drawing, with the magnetron in place between the pole pieces of the magnet.

The Cathode-Ray Tube The cathode-ray tube is a special type of electron tube which permits the visual observation of electrical signals. It may be incorporated into an oscilloscope for use as a test instrument or it may be the display device for a radar equipment or a television receiver.

A cathode-ray tube always includes an electron gun for producing a stream of electrons, a grid for controlling the intensity of the electron beam, and a luminescent screen for converting the impinging electron beam into visible light. Such a tube always operates in conjunction with either a built-in or an external means for focussing the electron stream into a narrow beam, and a means for deflecting the electron beam in accordance with an electrical signal.

The main electrical difference between types of cathode-ray tubes lies in the means employed for focussing and deflecting the electron beam. The beam may be focussed and/or deflected either electrostatically or magnetically, since a stream of electrons can be acted upon either by an electrostatic or a magnetic field. In an electrostatic field the electron beam tends to be deflected toward the positive termination of the field. In a magnetic field the stream tends to be deflected at right angles to the field. Further, an electron beam tends to be deflected so that it is normal (perpendicular) to the equipotential lines of an electrostatic field—and it tends to be deflected so that it is parallel to the lines of force in a magnetic field.

Large cathode-ray tubes used as kinescopes in television receivers usually are both focused and deflected magnetically. On the other hand, the medium-size c-r tubes used in oscilloscopes and small television receivers usually are both focussed and deflected electrostatically. But c-r tubes for special applications may be focussed magnetically and deflected electrostatically or vice versa.

There are advantages and disadvantages to both types of focussing and deflection. However, it may be stated that electrostatic deflection is much better than magnetic deflection when high-frequency waves are to be displayed on the screen; hence the almost universal use of this type of deflection for oscillographic work. But when a tube is operated at a high value of accelerating potential so as to obtain a bright display on the face of the tube for television or radar work, the use of magnetic deflection becomes desirable since it is relatively much easier to deflect a high-velocity electron beam magnetically than electrostatically. However, an ion trap is required with magnetic deflection since the heavy negative ions emitted by the cathode are not materially deflected by the magnetic field and hence would burn an "ion spot" in the center of the luminescent screen. With electrostatic deflection the heavy ions are deflected equally as well as the electrons in the beam so that an ion spot is not formed.

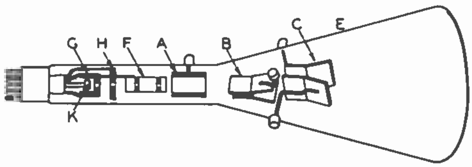


Figure 18.

TYPICAL CATHODE-RAY TUBE.

The tube illustrated is of the electrostatic-deflection electrostatic-focus type usually used in oscilloscopes. Most current oscilloscope tubes have all the electrodes brought out through the base of the tube. Television tubes, which almost always are of the magnetic-focus magnetic-deflection type, do not include the focussing electrode or the deflecting plates, and the high-voltage accelerating anode usually is connected to a conducting coating on the inside of the glass envelope.

The construction of a typical electrostatic-focus, electrostatic-deflection cathode-ray tube is illustrated in the pictorial diagram of figure 18. The indirectly heated cathode K releases free electrons when heated by the enclosed filament. The cathode is surrounded by a cylinder G, which has a small hole in its front for the passage of the electron stream. Although this element is not a wire mesh as is the usual grid, it is known by the same name because its action is similar: it controls the electron stream when its negative potential is varied.

Next in order is found the first accelerating anode, H, which resembles another disk or cylinder with a small hole in its center. This electrode is run at a high or moderately high positive voltage, to accelerate the electrons towards the far end of the tube.

The focussing electrode, F, is a sleeve which usually contains two small disks, each with a small hole.

After leaving the focussing electrode, the electrons pass through another accelerating anode, A, which is operated at a high positive potential. In some tubes this electrode is operated at a higher potential than the first accelerating electrode, H, while in other tubes both accelerating electrodes are operated at the same potential.

The electrodes which have been described up to this point constitute the "electron gun," which produces the free electrons and focusses them into a slender, concentrated, rapidly-traveling stream for projecting onto the viewing screen.

To make the tube useful, means must be provided for deflecting the electron beam along two axes at right angles to each other. The more common tubes employ electrostatic deflection plates, one pair to exert a force on the beam in the vertical plane and one pair to exert a force in the horizontal plane. These plates are designated as B and C in figure 18.

Standard oscilloscope practice with small cathode-ray tubes calls for connecting one of the B plates and one of the C plates together and to the high voltage accelerating anode. With the newer three-inch tubes and with five-inch tubes and larger all four deflecting plates are commonly used for deflection. The *positive* high voltage is grounded, instead of the negative as is common practice in amplifiers, etc., in order to permit operation of the deflecting plates at a d-c potential at or near ground.

In the average electrostatic-deflection c-r tube the spot will be fairly well centered if all four deflection plates are returned to the potential of the second anode (ground). However, for accurate centering and to permit moving the entire trace either horizontally or vertically to permit display of a particular waveform, horizontal and vertical centering controls usually are provided on the front of the oscilloscope.

After the spot is once centered, it is necessary only to apply a positive or negative voltage (with respect to ground) to one of the ungrounded or "free" deflector plates in order to move the spot. If the voltage is positive with respect to ground, the beam will be attracted toward that deflector plate, while if negative the beam and spot will be repulsed. The amount of deflection is directly proportional to the voltage (with respect to ground) that is applied to the free electrode.

With the larger-screen higher-voltage tubes it becomes necessary to place deflecting voltage on both horizontal and both vertical plates. This is done for two reasons: First, the amount of deflection voltage required by the high-voltage tubes is so great that a transmitting tube operating from a high voltage supply would be required to attain this voltage without distortion. By using push-pull deflection with two tubes feeding the deflection plates, the necessary plate supply voltage for the deflection amplifier is halved. Second, a certain amount of de-focussing of the electron stream is always present on the extreme excursions in deflection voltage when this voltage is applied only to one deflecting plate. When the deflecting voltage is fed in push-pull to both deflecting plates in each plane, there is no defocussing because the *average* voltage acting on the electron stream is zero, even though the *net* voltage (which causes the deflection) acting on the stream is twice that on either plate.

The fact that the beam is deflected by a magnetic field is important even in an oscilloscope which employs a tube using electrostatic deflection, because it means that precautions must be taken to protect the tube from the transformer fields and sometimes even the earth's magnetic field. This normally is done by incorporating a magnetic shield around the tube and by placing any transformers as far from the tube as possible, oriented to the position which produces minimum effect upon the electron stream.

Screen Materials — "Phosphors" At least five types of luminescent screen materials are commonly available on the various types of c-r tubes commercially available. These screen materials are called *phosphors;* each of the five phosphors is best suited to a particular type of applica-cation. The P-1 phosphor, which has a green fluorescence with medium persistence, is almost invariably used for oscilloscope tubes for visual observation. The P-4 phosphor, with white fluorescence and medium persistence, is used on television viewing tubes ("Kinescopes"). The P-5 and P-11 phosphors, with blue fluorescence and very short persistence, are used primarily in oscilloscopes where photographic recording of the trace is to be obtained. The P-7 phosphor, which has a blue flash and a long-persistence greenish-yellow persistence, is used primarily for radar displays where retention of the image for several seconds after the initial signal display is required.

Gas Tubes The space charge of electrons in the vicinity of the cathode in a diode causes the plate-to-cathode voltage drop to be a function of the current being carried between the cathode and the plate. This voltage drop can be rather high when large currents are being passed, causing a considerable amount of energy loss which shows up as plate dissipation. However, this negative space charge can be neutralized by the presence of the proper density of positive ions in the space between the cathode and anode. The positive ions may be obtained by the introduction of the proper amount of gas or a small amount of mercury into the envelope of the tube. When the voltage drop across the tube reaches the ionization potential of the gas or mercury vapor, the gas molecules will become ionized to form positive ions. The positive ions then tend to neutralize the space charge in the vicinity of the cathode. The voltage drop across the tube then remains constant at the ionization potential of the gas up to a current drain equal to the maximum emission capability of the cathode. The voltage drop varies between 10 and 20 volts, depending upon the particular gas employed, up to the maximum current rating of the tube.

Mercury-vapor tubes, although very widely used, have the disadvantage that they must be operated within a specified temperature range (25° to 70° C.) in order that the mercury vapor pressure within the tube shall be within the proper range. If the temperature is too low, the drop across the tube becomes too high causing immediate overheating and possible damage to the elements. If the temperature is too high, the vapor pressure is too high, and the voltage at which the tube will "flash back" is lowered to the point where destruction of the tube may take place. Since the ambient temperature range specified above is within the normal room temperature range, no trouble will be encountered under normal operating conditions. However, by the substitution of xenon gas for mercury it is possible to produce a rectifier with characteristics comparable to those of the mercury-vapor tube except that the tube is capable of operating over the range from approximately −70° to 90° C. The 3B25 rectifier is an example of this type of tube.

Thyratron Tubes If a grid is inserted between the cathode and plate of a mercury-vapor gaseous-conduction rectifier, a negative potential placed upon the added element will increase the plate-to-cathode voltage drop required before the tube will ionize or "fire." The potential upon the control grid will have no effect on the plate-to-cathode drop after the tube has ionized. However, the grid voltage may be adjusted to such a value that conduction will take place only over the desired portion of the cycle of the a-c voltage being impressed upon the plate of the rectifier.

CHAPTER FIVE

Vacuum Tube Amplifiers

The ability of the control grid of a vacuum tube to control large amounts of plate power with a small amount of grid energy allows the vacuum tube to be used as an amplifier. It is this ability of vacuum tubes to amplify an extremely small amount of energy up to almost any level without change in anything except amplitude which makes the vacuum tube such an extremely valuable adjunct to modern electronics and communication.

Symbols for Vacuum-Tube Parameters As an assistance in simplifying and shortening expressions involving vacuum-tube parameters, the following symbols will be used throughout this book:

Tube Constants

μ—amplification factor
R_p—plate resistance
G_m—transconductance
μ_{sg}—grid-screen mu factor
G_c—conversion transconductance (mixer tube)

Interelectrode Capacitances

C_{gk}—grid-cathode capacitance
C_{gp}—grid-plate capacitance
C_{pk}—plate-cathode capacitance
C_{in}—Input capacitance (tetrode or pentode)
C_{out}—output capacitance (tetrode or pentode)

Electrode Potentials

E_{bb}—d-c plate supply voltage (a positive quantity)
E_{cc}—d-c grid supply voltage (a negative quantity)
E_{gm}—peak grid excitation voltage (½ total peak-to-peak grid swing)
E_{pm}—peak plate voltage (½ total peak-to-peak plate swing)
e_p—instantaneous plate potential
e_g—instantaneous grid potential
e_{pmin}—minimum instantaneous plate voltage
e_{gmp}—maximum positive instantaneous grid voltage
E_p—static plate voltage
E_g—static grid voltage
e_{co}—cut off bias

Electrode Currents

I_b—average plate current
I_c—average grid current
I_{pm}—peak fundamental plate current
i_{pmax}—maximum instantaneous plate current
i_{gmax}—maximum instantaneous grid current
I_p—static plate current
I_g—static grid current

Other Symbols

P_i—plate power input
P_o—plate power output
P_p—plate dissipation
P_d—grid driving power (grid plus bias losses)
P_g—grid dissipation
N_p—plate efficiency (expressed as a decimal)
θ_p—one-half angle of plate current flow
θ_g—one-half angle of grid current flow
R_L—load resistance
Z_L—load impedance

Vacuum-Tube Constants The relationships between certain of the electrode potentials and currents within a vacuum tube are reasonably constant under specified conditions of operation. These relationships are called *vacuum-tube constants* and are listed in the data published by the manufacturers of vacuum tubes. The defining equations for the basic vacuum-tube constants are given in Chapter Four.

Interelectrode Capacitances The values of interelectrode capacitance published in vacuum-tube tables are the static values measured, in the case of triode for example, as shown in figure 1. The static capacitances are simply as shown in the drawing, but when a tube is operating as amplifier there is another consideration known as *Miller Effect* which causes the dynamic input capacitance to be different from the static value. The output capacitance of an amplifier is essentially the same as the static value such as is given in the published tube tables. The grid-to-plate capacitance is also the same as the published static value, but since the C_{gp} acts as a small capacitance coupling energy back from the plate circuit to the grid circuit, the dynamic input capacitance is equal to the static value plus an amount (frequently much greater in the case of a triode) determined by the gain of the stage, the plate load impedance, and the C_{gp} feedback capacitance. The total value for an audio amplifier stage can be expressed in the following equation:

$$C_{gk\ (dynamic)} = C_{gk\ (static)} + (A + 1)\ C_{gp}$$

where C_{gk} is the grid-to-cathode capacitance, C_{gp} is the grid-to-plate capacitance, and A is the stage gain. This expression assumes that the vacuum tube is operating into a resistive load such as would be the case with an audio stage working into a resistance plate load in the middle audio range.

The more complete expression for the input admittance (vector sum of capacitance and resistance) of an amplifier operating into any type of plate load is as follows:

$$\text{Input capacitance} = C_{gk} + (1 + A \cos \theta)\ C_{gp}$$

$$\text{Input resistance} = - \frac{\left(\dfrac{1}{\omega C_{gp}}\right)}{A \sin \theta}$$

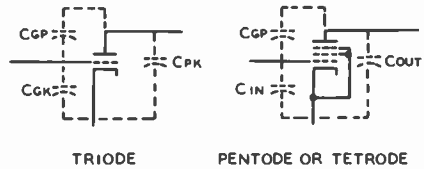


Figure 1.
Static Interelectrode Capacitances Within a Triode, Pentode, or Tetrode.

Where: C_{gk} = grid-to-cathode capacitance
C_{gp} = grid-to-plate capacitance
A = voltage amplification of the tube alone
θ = phase angle of the plate load impedance, positive for inductive loads, negative for capacitive

It can be seen from the above that if the plate load impedance of the stage is capacitive or inductive, there will be a resistive component in the input admittance of the stage. The resistive component of the input admittance will be positive (tending to load the circuit feeding the grid) if the load impedance of the plate is capacitive, or it will be negative (tending to make the stage oscillate) if the load impedance of the plate is inductive.

Neutralization of Interelectrode Capacitance Neutralization of the effects of interelectrode capacitance is employed most frequently in the case of radio frequency power amplifiers. Before the introduction of the tetrode and pentode tube, triodes were employed as neutralized Class A amplifiers in receivers. This practice has been largely superseded in the present state of the art through the use of tetrode and pentode tubes in which the C_{gp} or feedback capacitance has been reduced to such a low value that neutralization of its effects is not necessary to prevent oscillation and instability.

5-1 Classes and Types of Vacuum-Tube Amplifiers

Vacuum-tube amplifiers are grouped into various classes and sub-classes according to the type of work they are intended to perform. The difference between the various classes is determined primarily by the value of average grid bias employed and the maximum value of the exciting signal to be impressed upon the grid.

Class A Amplifier A Class A amplifier is an amplifier biased and supplied with excitation of such amplitude that plate current flows continuously (360° of the exciting voltage waveshape) and grid current does not flow at any time. Such an amplifier is normally operated in the center of the grid-voltage plate-current transfer characteristic and gives an output waveshape which is a substantial replica of the input waveshape.

Class A_1 Amplifier This is another term applied to the Class A amplifier in which grid current does not flow over any portion of the input wave cycle.

Class A_2 Amplifier This is a Class A amplifier operated under such conditions that the grid is driven positive over a portion of the input voltage cycle, but plate current still flows over the entire cycle.

Class AB_1 Amplifier This is an amplifier operated under such conditions of grid bias and exciting voltage that plate current flows for more than one-half the input voltage cycle but for less than the complete cycle. In other words the operating angle of plate current flow is appreciably greater than 180° but less than 360°. The suffix 1 indicates that grid current does not flow over any portion of the input cycle.

Class AB_2 Amplifier A Class AB_2 amplifier is operated under essentially the same conditions of grid bias as the Class AB_1 amplifier mentioned above, but the exciting voltage is of such amplitude that grid current flows over an appreciable portion of the input wave cycle.

Class B Amplifier A Class B amplifier is biased substantially to cutoff of plate current (without exciting voltage) so that plate current flows essentially over one-half the input voltage cycle. The operating angle of plate current flow is essentially 180°. The Class B amplifier is almost always excited to such an extent that grid current flows.

Class C Amplifier A Class C amplifier is biased to a value greater than the value required for plate current cutoff and is excited with a signal of such amplitude that grid current flows over an appreciable period of the input voltage waveshape. The angle of plate current flow in a Class C amplifier is appreciably less than 180°, or in other words, plate current flows appreciably less than one-half the time. Actually, the conventional operating conditions for a Class C amplifier are such that plate current flows for 120° to 150° of the exciting voltage waveshape.

Types of Amplifiers There are three general types of amplifier circuits in use. These types are classified on the basis of the *return* for the input and output circuits. Conventional amplifiers are called *cathode return* amplifiers since the cathode is effectively grounded and acts as the common return for both the input and output circuits. The second type is known as a plate return amplifier or *cathode follower* since the plate circuit is effectively at ground for the input and output signal voltages and the output voltage or power is taken between cathode and plate. The third type is called a *grid-return* or *grounded-grid* amplifier since the grid is effectively at ground potential for input and output signals and output is taken between grid and plate.

Chapter Organization

As an assistance to the reader, the balance of this chapter has been subdivided into six main sections: Audio-Frequency Voltage Amplifiers, Audio-Frequency Power Amplifiers, Radio-Frequency Voltage Amplifiers, Radio-Frequency Power Amplifiers, Feedback Amplifiers, Video-Frequency Amplifiers.

AUDIO-FREQUENCY VOLTAGE AMPLIFIERS

Audio-frequency voltage amplifiers are most frequently employed in two general applications: First, to build up the voltage output of the detector or demodulator in a receiver to a level sufficient to excite the grid of the audio-frequency power amplifier which drives the loudspeaker; and second, to increase the voltage output of a microphone or other type of pickup to a level sufficient to excite the grid of an audio-frequency power amplifier.

5-2 Resistance-Capacitance Coupled Audio-Frequency Amplifiers

Present practice in the design of audio-frequency voltage amplifiers is almost exclusively to use resistance-capacitance coupling between the low-level stages. Both triodes and pentodes are used; triode amplifier stages will be discussed first.

R-C Coupled Triode Stages Figure 2 illustrates the standard circuit for a resistance-capacitance coupled amplifier stage utilizing a triode tube with cathode bias.

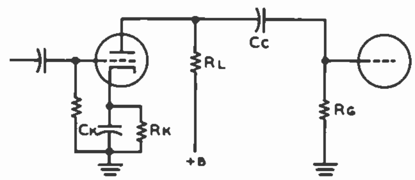
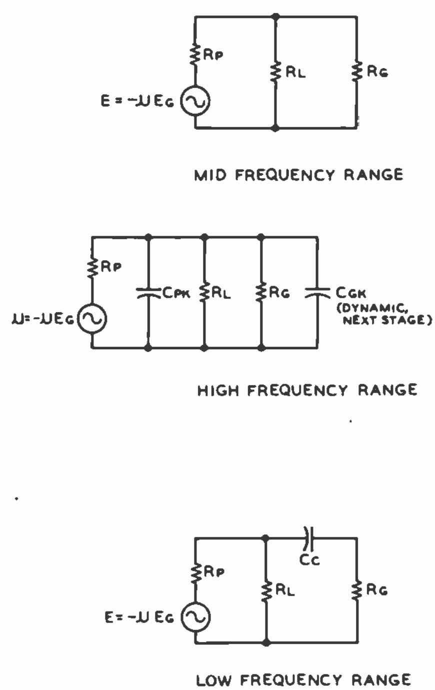


Figure 2.

Standard circuit for resistance-capacitance coupled triode amplifier stage. Values for components can be determined with the aid of Table I.

In conventional audio-frequency amplifier design such stages are used at medium voltage levels (from 0.01 to 5 volts peak on the grid of the tube) and use medium-μ triodes such as the 6J5 or high-μ triodes such as the 6SF5 or 6SL7-GT. Normal voltage gain for a single stage of this type is from 10 to 70, depending upon the tube chosen and its operating con-



$$A = \frac{\mu R_L R_G}{R_P(R_L+R_G)+R_L R_G}$$

$$\frac{A_{\text{HIGH FREQ.}}}{A_{\text{MID FREQ.}}} = \frac{1}{\sqrt{1+(R_{EQ}/X_S)^2}}$$

$$R_{EQ} = \frac{R_L}{1+\frac{R_L}{R_G}+\frac{R_L}{R_P}}$$

$$X_S = \frac{1}{2\pi F\,(C_{PK}+C_{GK\,(\text{DYNAMIC})})}$$

$$\frac{A_{\text{LOW FREQ.}}}{A_{\text{MID FREQ.}}} = \frac{1}{\sqrt{1+(X_C/R)^2}}$$

$$X_C = \frac{1}{2\pi F C_C}$$

$$R = R_G + \frac{R_L R_P}{R_L+R_P}$$

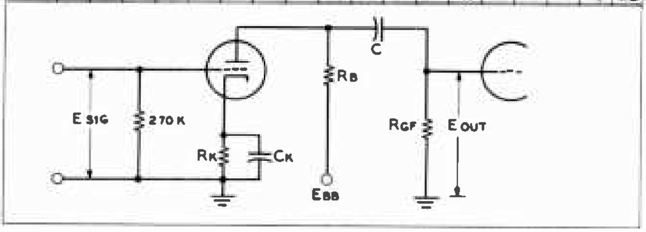
Figure 3.

Equivalent circuits and gain equations for a triode R-C coupled amplifier stage. In using these equations, be sure to select the values of mu and R_p which are proper for the static current and voltages with which the tube will operate. These values may be obtained from curves published in the RCA Tube Handbook HB-3.

TABLE I

RESISTANCE-CAPACITANCE COUPLED TRIODE AUDIO VOLTAGE AMPLIFIER

	6J5, 6J5-G, 7A4, 6F8-G, 6SN7-GT, 7N7 (1 TRI. UNIT)						6SQ7, 7B6, 75, 2A6, 6B6-G (TRIODE UNIT)						1LE3, 1E4-G					
E_{BB}	250 VOLTS						250 VOLTS						90 VOLTS					
R_B	47 K		100 K		270 K		100 K		270 K		470 K		47 K		100 K		270 K	
R_{GF}	0.1	0.27	0.1	0.47	0.27	0.47	0.27	0.47	0.47	1.0	0.47	1.0	0.1	0.27	0.10	0.47	0.27	0.47
R_K	1500	2200	2700	3900	6800	8200	1800	1800	3300	3900	3900	4700	0.70	0.64	0.45	0.38	0.199	0.187
I_B	2.79	2.4	1.49	1.31	0.61	0.56	0.73	0.73	.395	.385	.288	.281						
E_C	-4.18	-5.28	-4.03	-5.11	-4.15	-4.74	-1.31	-1.31	-1.30	-1.42	-1.12	-1.25	-1.8	-2.1	-1.5	-2.0	-1.5	-1.7
E_B	119	137	101	119	85	94	177	177	143.5	151.5	114.5	124.5	57.1	60	45	52	36.2	39.8
E_{SIG}	1.0	1.0	1.0	1.0	1.0	1.0	0.1	0.1	0.1	0.1	0.1	0.1	0.5	0.5	0.5	0.5	0.5	0.5
E_{OUT}	14.8	15	15.2	16.2	15.9	16.2	4.37	4.78	5.92	6.15	6.24	6.75	3.94	4.2	4.32	4.76	5.0	5.2
GAIN	14.8	15	15.2	16.2	15.9	16.2	43.7	47.8	59.2	61.3	62.4	67.5	7.9	8.4	8.65	9.5	10.0	10.4
% DIST.	1.4	1.4	1.8	1.3	1.6	1.3	0.8	0.7	0.8	0.7	0.8	0.7	1.7	1.4	1.7	1.3	2.4	2.2
E_{SIG}	2.7	3.5	2.55	3.3	2.6	3.05	0.55	0.55	0.53	0.61	0.40	0.53	1.27	1.48	1.06	1.41	1.06	1.2
E_{OUT}	39.9	52.5	38.4	53.0	42	49.4	23.9	26.0	31.2	37	25	38	10	12.4	9.15	13.4	10.6	12.5
GAIN	14.7	15.0	15	18.1	15.9	16.2	43.5	47.4	59.0	60.6	62.4	61.5	7.88	8.4	8.65	9.5	10	10.4
% DIST.	4.1	4.9	4.9	4.8	4.7	4.5	4.5	4.0	4.0	4.5	3.3	3.8	4.7	5.0	4.7	5.0	5.0	5.0
	6C4, 12AU7						6AQ6, 6AQ7-GT, 6AT6, 6Q7, 6Q7-GT, 6SZ7, 6T7G, 6T8						7F8					
E_{BB}	250 VOLTS						250 VOLTS						250 VOLTS					
R_B	47 K		100 K		270 K		100 K		270 K		470 K		47 K		100 K		270 K	
R_{GF}	0.1	0.27	0.1	0.47	0.27	0.47	0.27	0.47	0.47	1.0	0.47	1.0	0.1	0.27	0.1	0.47	0.27	0.47
R_K	1000	1000	1500	1800	4700	6800	1800	2200	3900	3900	4700	5600	390	470	820	1000	2200	2200
I_B	3.2	3.2	1.78	1.72	.684	0.63	.917	0.83	0.44	0.44	.312	0.29	3.0	2.86	1.58	1.50	0.66	0.66
E_C	-3.2	-3.2	-2.87	-3.10	-3.21	-4.26	-1.65	-1.83	-1.72	-1.72	-1.47	-1.62	-1.17	-1.34	-1.29	-1.50	-1.45	-1.45
E_B	150.5	150.5	72	76	65	80	158	167	131	131	103	114	109	115	92	100	72	72
E_{SIG}	1.0	1.0	1.0	1.0	1.0	1.0	0.1	0.1	0.1	0.1	0.1	0.1	0.1	0.1	0.1	0.1	0.1	0.1
E_{OUT}	13.5	14.1	13.8	14.3	13.4	13.2	4.0	4.1	5.0	5.25	5.25	5.55	3.38	3.82	3.56	3.65	3.40	3.60
GAIN	13.5	14.1	13.6	14.3	13.4	13.2	40	41	50	52.5	52.5	55.5	33.8	38.2	35.6	36.5	34.0	36.0
% DIST.	3.3	3.1	3.6	2.8	2.5	2.0	0.6	0.5	0.5	0.4	0.5	0.4	1.1	0.9	1.0	0.7	0.8	0.7
E_{SIG}	1.70	1.70	1.34	1.70	1.80	2.52	0.87	1.03	0.97	0.97	0.77	0.90	0.4	0.55	0.50	0.70	0.60	0.60
E_{OUT}	23.0	24.0	18.5	24.5	24.1	33.1	33.6	41.5	46.6	48.6	38.6	48.5	13.5	21.0	17.6	25.5	20.4	21.6
GAIN	13.5	14.1	13.8	14.3	13.4	13.1	38.6	40.2	48	50.4	50.4	54	33.8	38.2	35.6	36.4	34	36
% DIST.	4.9	4.6	5.0	5.0	4.9	5.0	4.0	4.8	4.8	3.6	3.9	3.7	4.0	4.6	4.6	4.9	4.5	4.2



ditions. Triode tubes are normally used in the last voltage amplifier stage of an R-C amplifier since their harmonic distortion with large output voltage (25 to 75 volts) is less than with a pentode tube.

Voltage Gain per Stage The voltage gain per stage of a resistance-capacitance coupled triode amplifier can be calculated with the aid of the equivalent circuits and expressions for the mid-frequency, high-frequency, and low-frequency range given in figure 3.

A triode R-C coupled amplifier stage is normally operated with values of cathode resistor and plate load resistor such that the actual voltage on the tube is approximately one-half the d-c plate supply voltage. To assist the designer of such stages, data on operating conditions for commonly used tubes is published in the RCA Tube Handbook HB-3. It is assumed, in the case of the gain equations of figure 3, that the cathode by-pass capacitor, C_k, has a reactance that is low with respect to the cathode resistor at the lowest frequency to be passed by the amplifier stage.

R-C Coupled Pentode Stages Figure 4 illustrates the standard circuit for a resistance-capacitance coupled pentode amplifier stage. Cathode bias is used and the screen voltage is supplied through a dropping resistor from the plate voltage supply. In conventional audio-frequency amplifier design such stages are normally used at low voltage levels (from 0.00001 to 0.1 volts peak on the grid of the tube) and use moderate-G_m pentodes such as the 6SJ7. Normal voltage gain for a stage of this type is from 60 to 250, depending upon the tube chosen and its operating conditions. Pentode tubes are ordinarily used in the first stage of an R-C amplifier where the

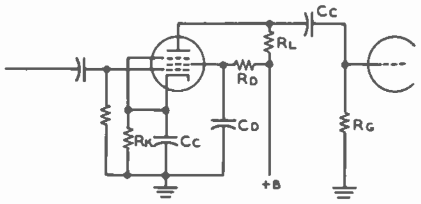


Figure 4.

Standard circuit for resistance-capacitance coupled pentode amplifier stage. Values of circuit constants may be determined with the aid of Table II.

high gain which they afford is of greatest advantage and where only a small voltage output is required from the stage.

The voltage gain per stage of a resistance-capacitance coupled pentode amplifier can be calculated with the aid of the equivalent circuits and expressions for the mid-frequency, high-frequency, and low-frequency range given in figure 5.

To assist the designer of such stages, data on operating conditions for commonly used types of tubes is published in the RCA Tube Handbook HB-3. It is assumed, in the case of the gain equations of figure 5, that the cathode by-pass capacitor, C_k, has a reactance that is low with respect to the cathode resistor at the lowest frequency to be passed by the stage. It is additionally assumed that the reactance of the screen by-pass capacitor, C_d, is low with respect to the screen dropping resistor, R_d, at the lowest frequency to be passed by the amplifier stage.

Cascade Voltage Amplifier Stages When voltage amplifier stages are operated in such a manner that the output voltage of the first is fed to the grid of the second, and so forth, such stages are said to be *cascaded*. The total voltage gain of cascaded amplifier stages is obtained by taking the product of the voltage gains of each of the successive stages.

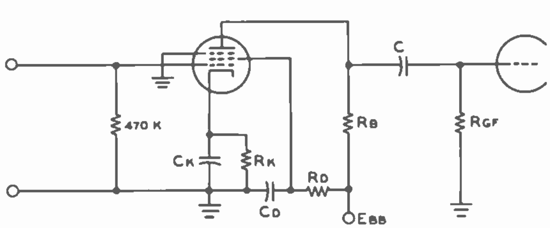
Sometimes the voltage gain of an amplifier stage is rated in decibels. Voltage gain is converted into decibels gain through the use of the following expression: db = 20 $\log_{10}$ A, where A is the voltage gain of the stage. The total gain of cascaded voltage amplifier stages can be obtained by *adding* the number of decibels gain in each of the cascaded stages. This subject is covered in Chapter Twenty-seven, *Reference Data.*

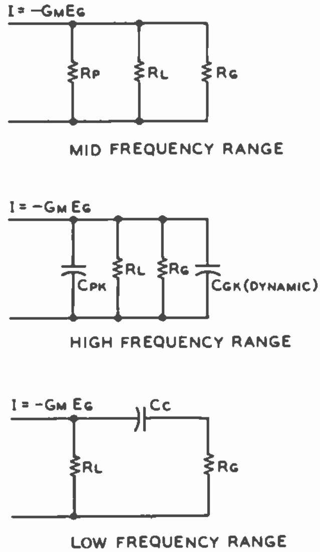
TABLE II
RESISTANCE-CAPACITANCE COUPLED PENTODE AUDIO VOLTAGE AMPLIFIER

	6SJ7								
E_{BB}	300 VOLTS								
R_B	100 K			250 K			500 K		
R_{GF}	0.1	.25	0.5	.25	0.5	1.0	0.5	1.0	2.0
R_D	.35	.37	.47	.89	1.1	1.2	2.0	2.2	2.5
R_K	500	530	590	850	860	910	1300	1400	1530
C_K	11.8	10.9	9.9	8.5	7.4	6.9	6.0	5.8	5.2
C_D	0.1	.09	.09	.07	.06	.06	.06	.05	.04
C	.019	.016	.007	.011	.004	.003	.004	.002	.002
E_O	72	96	101	79	88	98	64	79	89
GAIN	67	98	104	139	167	185	200	236	263

	6J7, 6J7-G, 6W7-G								
E_{BB}	300 VOLTS								
R_B	100 K			250 K			500 K		
R_{GF}	0.1	.25	0.5	0.25	0.5	1.0	0.5	1.0	2.0
R_D	.44	0.5	.53	1.2	1.2	1.45	2.45	2.9	2.95
R_K	500	450	600	1100	1200	1300	1700	2200	2300
C_K	8.5	8.3	8.0	5.5	5.4	5.8	4.2	4.1	4.0
C_D	.07	.07	.06	.04	.04	.05	.04	.04	.04
C	.02	.01	.006	.008	.005	.005	.005	.003	.003
E_O	55	81	96	81	104	110	75	97	100
GAIN	61	82	94	104	140	185	161	350	240

	6AU6, 6SH7								
E_{BB}	300 VOLTS								
R_B	100 K			220 K			470 K		
R_{GF}	0.1	0.22	0.47	0.22	0.47	1.0	0.47	1.0	2.2
R_D	0.2	0.24	0.26	0.42	0.5	0.55	1.0	1.1	1.2
R_K	500	600	700	1000	1000	1100	1800	1900	2100
C_K	18.0	16.4	15.3	12.4	12.0	11.0	8.0	7.6	7.3
C_D	.13	.11	.11	.1	.098	.09	.075	.065	.08
C	.019	.011	.006	.009	.007	.003	.0045	.0028	.0018
E_O	76	103	129	92	108	122	94	105	122
GAIN	109	145	168	164	230	282	248	318	371





$$A = G_M R_{EQ}$$

$$R_{EQ} = \frac{R_L}{1 + \frac{R_L}{R_G} + \frac{R_L}{R_P}}$$

$$\frac{A_{\text{HIGH FREQ.}}}{A_{\text{MID FREQ.}}} = \frac{1}{\sqrt{1 + (R_{EQ}/X_S)^2}}$$

$$R_{EQ} = \frac{R_L}{1 + \frac{R_L}{R_G} + \frac{R_L}{R_P}}$$

$$X_S = \frac{1}{2\pi f\,(C_{PK} + C_{GK}\ \text{(DYNAMIC)})}$$

$$\frac{A_{\text{LOW FREQ.}}}{A_{\text{MID FREQ.}}} = \frac{1}{\sqrt{1 + (X_C/R)^2}}$$

$$X_C = \frac{1}{2\pi f\,C_C}$$

$$R = R_G + \frac{R_L\,R_P}{R_L + R_P}$$

Figure 5.

Equivalent circuits and gain equations for a pentode R-C coupled amplifier stage. In using these equations be sure to select the values of G_m and R_p which are proper for the static currents and voltages with which the tube will operate. These values may be obtained from curves published in the RCA Tube Handbook HB-3.

5-3 Other Interstage Coupling Methods

Figure 6 illustrates, in addition to resistance-capacitance interstage coupling, seven additional methods in which coupling between two successive stages of an audio-frequency amplifier may be accomplished. Although resistance-capacitance coupling is most commonly used, there are certain circuit conditions wherein coupling methods other than resistance capacitance are more effective.

Transformer Coupling Transformer coupling, as illustrated in figure 6B, is seldom used at the present time between two successive single-ended stages of an audio amplifier. There are several reasons why resistance coupling is favored over transformer coupling between two successive single-ended stages. These are: (1) a transformer having frequency characteristics comparable with a properly designed R-C stage is very expensive; (2) transformers, unless they are very well shielded, will pick up inductive hum from nearby power and filament transformers; (3) the phase characteristics of step-up interstage transformers are very poor, making very difficult the inclusion of a transformer of this type within a feedback loop; and (4) transformers are heavy.

However, there is one circuit application where a step-up interstage transformer is of considerable assistance to the designer; this is the case where it is desired to obtain a large amount of voltage to excite the grid of a cathode follower or of a high-power Class A amplifier from a tube operating at a moderate plate voltage. Under these conditions it is possible to obtain a peak voltage on the secondary of the transformer of a value somewhat greater than the d-c plate supply voltage of the tube supplying the primary of the transformer.

Push-Pull Transformer Interstage Coupling Push-pull transformer coupling between two stages is illustrated in figure 6C. This interstage coupling arrangement is fairly commonly used. The system is particularly effective when it is desired, as in the system just described, to obtain a fairly high voltage to excite the grids of a high-power audio stage. The arrangement is also very good when it is desired to apply feedback to the grids of the push-pull stage by applying the feedback voltage to the low-potential sides of the two push-pull secondaries.

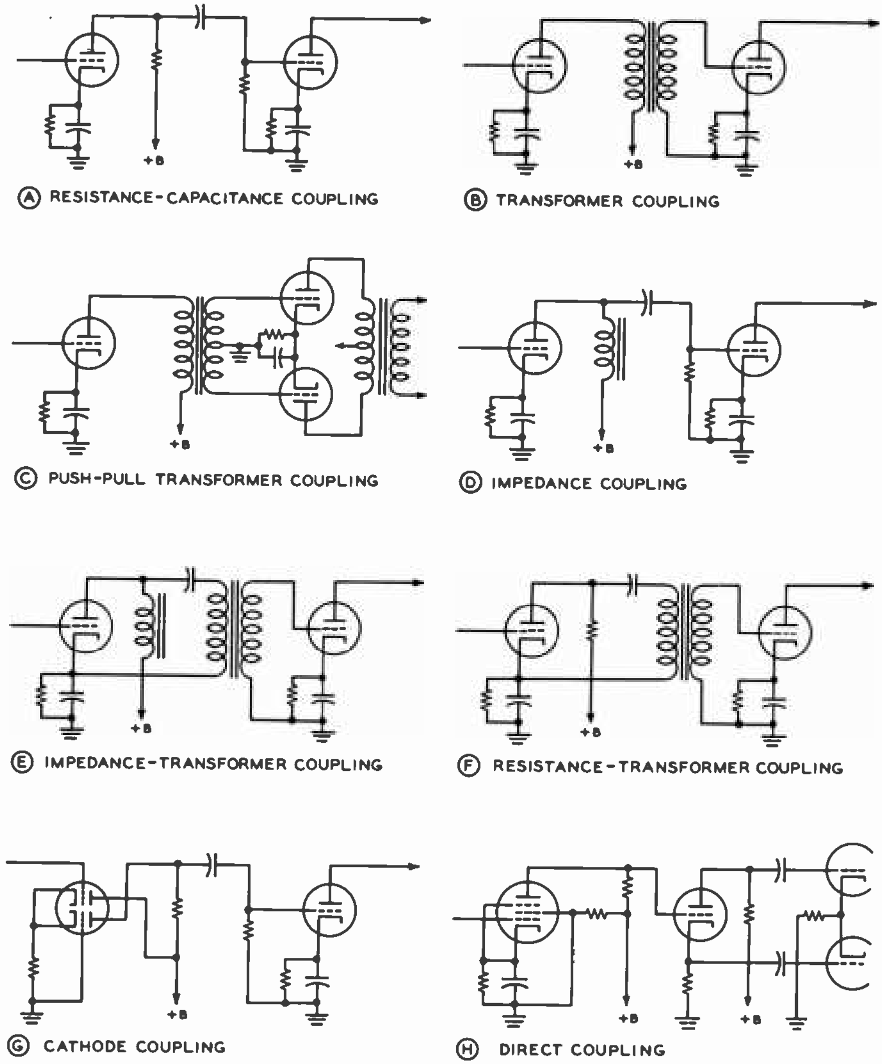
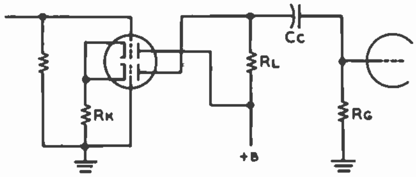


Figure 6.
Interstage coupling methods for audio-frequency voltage amplifiers.

Impedance Coupling Impedance coupling between two stages is shown in figure 6D. This circuit arrangement is seldom used, but it offers one strong advantage over R-C interstage coupling. This advantage is the fact that, since the operating voltage on the tube with the impedance in the plate circuit is the plate supply voltage, it is possible to obtain approximately twice the peak voltage output that it is possible to obtain with R-C coupling. This is because, as has been mentioned before, the d-c plate voltage on an R-C stage is approximately one-half the plate supply voltage.

Impedance-Transformer and Resistance-Transformer Coupling These two circuit arrangements, illustrated in figures 6E and 6F, are employed when it is desired to use transformer



$$G_M' = -G_M \frac{G}{2G+1} \qquad G = R_K G_M \left(1 + \frac{1}{\mu}\right)$$

R_K = CATHODE RESISTOR

$$R_P' = R_P \frac{2G+1}{G+1} \qquad G_M = G_M \text{ OF EACH TUBE}$$

μ = μ OF EACH TUBE

$$\mu' = -\mu \frac{G}{G+1} \qquad R_P = R_P \text{ OF EACH TUBE}$$

EQUIVALENT FACTORS INDICATED ABOVE BY (') ARE THOSE OBTAINED BY USING AN AMPLIFIER WITH A PAIR OF SIMILAR TUBE TYPES IN CIRCUIT SHOWN ABOVE.

Figure 7.

Equivalent factors for a pair of similar triodes operating as a cathode-coupled audio-frequency voltage amplifier.

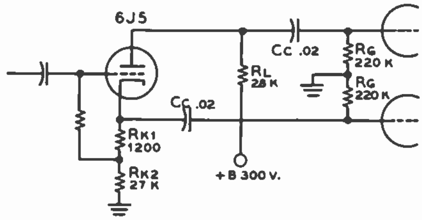
coupling for the reasons cited above, but where it is desired that the d-c plate current of the amplifier stage be isolated from the primary of the coupling transformer. With most types of high-permeability wide-response transformers it is necessary that there be no direct-current flow through the windings of the transformer. The impedance-transformer arrangement of figure 6E will give a higher voltage output from the stage but is not often used since the plate coupling impedance (choke) must have very high inductance and very low distributed capacitance in order not to restrict the range of the transformer which it and its associated tube feed. The resistance-transformer arrangement of figure 6F is ordinarily quite satisfactory where it is desired to feed a transformer from a voltage amplifier stage with no d-c in the transformer primary.

Cathode Coupling The cathode coupling arrangement of figure 6G has been widely used only comparatively recently. One outstanding characteristic of such a circuit is that there is no phase reversal between the grid and the plate circuit. All other common types of interstage coupling are accompanied by a 180° phase reversal between the grid circuit and the plate circuit of the tube.

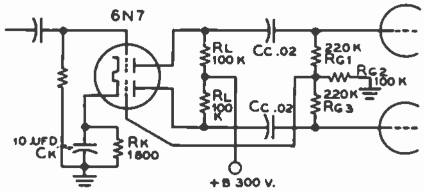
Figure 7 gives the expressions for determining the appropriate factors for an equivalent triode obtained through the use of a pair of similar triodes connected in the cathode-coupled circuit shown. With these equivalent triode factors it is possible to use the expressions shown in figure 3 to determine the gain of the stage at different frequencies. The input capacitance of such a stage is less than that of one of the triodes, the effective grid-to-plate capacitance is very much less (it is so much less that such a stage may be used as an r-f amplifier without neutralization), and the output capacitance is approximately equal to the grid-to-plate capacitance of one of the triode sections. This circuit is particularly effective with tubes such as the 6J6, 6N7, and 6SN7-GT which have two similar triodes in one envelope. An appropriate value of cathode resistor to use for such a stage is the value which would be used for the cathode resistor of a conventional amplifier using *one* of the same type tubes with the values of plate voltage and load resistance to be used for the cathode-coupled stage.

Inspection of the equations in figure 7 shows that as the cathode resistor is made smaller, to approach zero, the G_m approaches zero, the plate resistance approaches the R_p of one tube, and the mu approaches zero. As the cathode resistor is made very large the G_m approaches one half that of a single tube of the same type, the plate resistance approaches twice that of one tube, and the mu approaches the same value as one tube. But since the G_m of each tube decreases as the cathode resistor is made larger (since the plate current will decrease on each tube) the optimum value of cathode resistor will be found to be in the vicinity of the value mentioned in the previous paragraph.

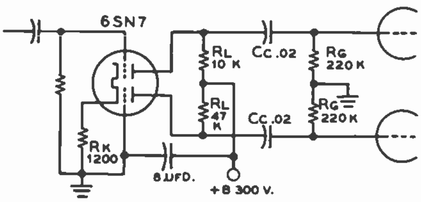
Direct Coupling Direct coupling between successive amplifier stages (plate of first stage connected directly to the grid of the succeeding stage) is complicated by the fact that the grid of an amplifier stage must be operated at an average negative potential with respect to the cathode of that stage. However, if the cathode of the second amplifier stage can be operated at a potential more positive than the plate of the preceding stage by the amount of the grid bias on the second amplifier stage, this direct connection between the plate of one stage and the grid of the succeeding stage can be used. Figure 6H illustrates an application of this principle in the coupling of a pentode amplifier stage to the grid of a "hot-cathode" phase inverter. In this arrangement the values of cathode, screen,



Ⓐ "HOT CATHODE" PHASE INVERTER



Ⓑ "FLOATING PARAPHRASE" PHASE INVERTER



Ⓒ CATHODE COUPLED PHASE INVERTER

Figure 8.

Three popular phase-inverter circuits with recommended values for circuit components.

and plate resistor in the pentode stage are chosen such that the plate of the pentode is at approximately 0.3 times the plate supply potential. The succeeding phase-inverter stage then operates with conventional values of cathode and plate resistor (same value of resistance) in its normal manner. This type of phase inverter is described in more detail in the section to follow.

5-4 Phase Inverters

It is necessary in order to excite the grids of a push-pull stage that voltages equal in amplitude and opposite in polarity be applied to the two grids. These voltages may be obtained through the use of a push-pull input transformer such as is shown in figure 6C. It is possible also, without the attendant bulk and expense of a push-pull input transformer, to obtain voltages of the proper polarity and phase through the use of a so-called *phase-inverter* stage. There are a large number of phase inversion circuits which have been developed and applied but the three shown in figure 8 have been found over a period of time to be the most satisfactory from the point of view of the number of components required and from the standpoint of the accuracy with which the two out-of-phase voltages are held to the same amplitude with changes in supply voltage and changes in tubes.

"Hot-Cathode" Phase Inverter Figure 8A illustrates the hot-cathode type of phase inverter. This type of phase inverter is the simplest of the three types since it requires only one tube and a minimum of circuit components. It is particularly simple when directly coupled from the plate of a pentode amplifier stage as shown in figure 6H. The circuit does, however, possess the following two disadvantages: (1) the cathode of the tube must run at a potential of approximately 0.3 times the plate supply voltage above the heater when a grounded common heater winding is used for this tube as well as the other heater-cathode tubes in a receiver or amplifier; (2) the circuit actually has a loss in voltage from its input to either of the output grids — about 0.9 times the input voltage will be applied to each of these grids. This does represent a voltage gain of about 1.8 in *total* voltage output with respect to input (grid-to-grid output voltage) but it is still small with respect to the other two phase inverter circuits shown.

Recommended component values for use with a 6J5 tube in this circuit are shown in figure 8A. If it is desired to use another tube in this circuit, appropriate values for the operation of that tube as a conventional amplifier can be obtained from manufacturer's tube data. The value of R_L obtained should be divided by two, and this new value of resistance placed in the circuit as R_L. The value of R_k from Table I should then be used as R_{k1} in this circuit, and then the total of R_{k1} and R_{k2} should be made to equal R_L.

"Floating Paraphrase" Phase Inverter An alternate type of phase inverter sometimes called the "floating paraphrase" is illustrated in figure 8B. This circuit is quite often used with a 6N7 tube, and appropriate values for the 6N7 tube in this application are shown. The circuit shown with the values given will give a voltage gain of approximately 21 from the input grid to each of the grids of the succeeding stage. It is capable of approximately 70 volts peak output to each grid.

The circuit inherently has a small unbalance in output voltage. This unbalance can be eliminated, if it is required for some special application, by making the resistor Rg_1 a few per cent lower in resistance value than Rg_2.

Cathode-Coupled Phase Inverter The circuit shown in figure 8C gives approximately one-half the voltage gain from the input grid to either of the grids of the succeeding stage that would be obtained from a single tube of the same type operating as a conventional R-C amplifier stage. Thus, with a 6SN7-GT tube as shown (two 6J5's in one envelope) the voltage gain from the input grid to either of the output grids will be approximately 7—the gain is, of course, 14 from the input to both output grids. The phase characteristics are such that the circuit is commonly used in deriving push-pull deflection voltage for a cathode-ray tube from a single-ended input signal. However, the frequency response of such a circuit is not good enough for broad-band television work.

AUDIO FREQUENCY POWER AMPLIFIERS

5-5 Single-Ended Triode Amplifiers

Figure 9 illustrates five circuits for the operation of Class A triode amplifier stages. Since the cathode current of a triode Class A_1 (no grid current) amplifier stage is constant with and without excitation, it is common practice to operate the tube with cathode bias. Recommended operating conditions in regard to plate voltage, grid bias, and load impedance for conventional triode amplifier stages are given in the RCA Tube Manual, HB-3.

It is possible, under certain conditions to operate single-ended triode amplifier stages (and pentode and tetrode stages as well) with grid excitation of sufficient amplitude that grid current is taken by the tube on peaks. This type of operation is called Class A_2 and is characterized by increased plate-circuit efficiency over straight Class A amplification without grid current. The normal Class A_1 amplifier power stage will operate with a plate circuit efficiency of from 20 per cent to perhaps 35 per cent. Through the use of Class A_2 operation it is possible to increase this plate circuit efficiency to approximately 38 to 45 per cent. However, such operation requires careful choice of the value of plate load impedance, a grid bias supply with good regulation (since the tube draws grid current on peaks although the plate current does not change with signal), and a driver tube with moderate power capability to excite the grid of the Class A_2 tube.

Figures 9D and 9E illustrate two methods of connection for such stages. Tubes such as the 845, 849, and 304TL (and also the 813 beam tetrode with appropriate screen supply) are suitable for such a stage. In each case the grid bias is approximately the same as would be used for a Class A_1 amplifier using the same tube, and as mentioned before, fixed bias must be used along with an audio driver of good regulation — preferably a triode stage with a 1:1 or step-down driver transformer. In each case it will be found that the correct value of plate load impedance will be increased about 40 per cent over the value recommended by the tube manufacturer for Class A_1 operation of the tube.

Both Class A_1 and Class A_2 power output, load impedance, and second harmonic distortion can be calculated with the aid of the published average plate characteristic curves of the tube under consideration and the three equations given in figure 10. It is simply necessary to draw a trial load line on the tube characteristics, take the values of voltage and current from the points of intersection between the load line and the tube curves, and substitute these values in the equations given. A sample load line for a type 2A3 tube is included in figure 10.

The correct values for R_L in figure 9 can be obtained from the RCA Receiving Tube Hand-

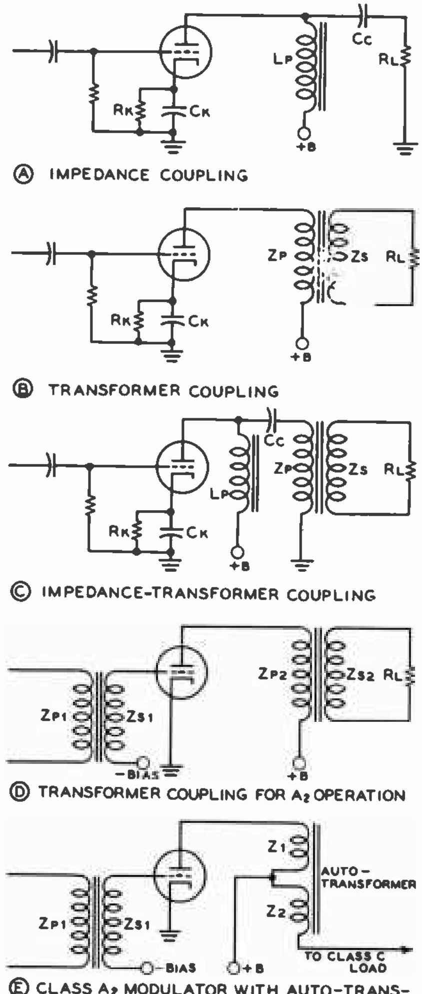
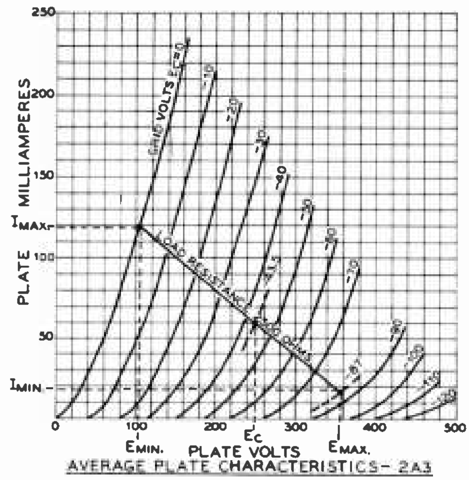


Figure 9.

Output coupling arrangements for single-ended Class A triode audio-frequency power amplifiers.



$\mu = 4.2$ $R_p = 800$ OHMS
PLATE DISSIPATION = 15 WATTS

LOAD RESISTANCE

$$R_L = \frac{E_{MAX.} - E_{MIN.}}{I_{MAX.} - I_{MIN.}} \text{ OHMS}$$

POWER OUTPUT

$$P_O = \frac{(I_{MAX.} - I_{MIN.})\ (E_{MAX.} - E_{MIN.})}{8} \text{ WATTS}$$

SECOND HARMONIC DISTORTION

$$D_2 = \frac{\frac{(I_{MAX.} + I_{MIN.})}{2} - I_O}{I_{MAX.} - I_{MIN.}} \times 100 \text{ PERCENT}$$

Figure 10.

Formulas for determining the operating conditions for a Class A triode single-ended audio-frequency power output stage. A typical load line has been drawn on the average plate characteristics of a type 2A3 tube to illustrate the procedure.

book, or they may be obtained by dividing the value of grid bias listed for the tube by the operating plate current of the tube. The value of C_k should be such that at the lowest audio frequency it is desired to pass with the stage the reactance of C_k is somewhat less than the resistance of R_k. Similarly, when an inductor is used for a plate feed choke as shown in figures 9A and 9C, the reactance of L_p should be somewhat greater than the correct value of load resistance for the tube at the lowest audio frequency to be passed. The coupling capacitor C_c, where used, should be low with respect to R_L at the lowest frequency to be passed.

Where a transformer is employed as shown in figure 9 to couple the energy from the plate of the output tube to the load circuit, the turns ratio of the transformer, between primary and secondary, should be equal to the square root of the impedance ratio. Thus, for example, if the recommended plate load impedance of the tube is 10,000 ohms and the load impedance to be fed is 500 ohms, the impedance ratio is 20, and the turns ratio between primary and secondary should be equal to the square root of 20 or 4.47.

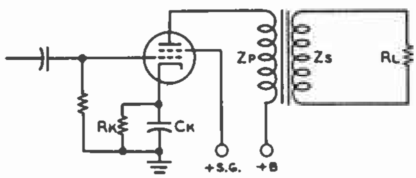


Figure 11.
Normal single-ended pentode or beam tetrode audio-frequency power output stage.

5-6 Single-Ended Tetrode or Pentode Power Audio Amplifier Stages

Figure 11 illustrates the conventional circuit for a single-ended tetrode or pentode amplifier stage. Tubes of this type have largely replaced triodes in the output audio stages of receivers and low-power amplifiers due to the higher power efficiency of such stages and the greater plate circuit efficiency with which they operate. As an example, a type 45 tube operating at a plate voltage of 250 volts requires a peak grid swing of 50 volts and operates at a plate circuit efficiency of about 20 per cent, while a type 6V6 tetrode operating at the same plate and screen voltage as the type 45 requires a peak grid swing of only 12.5 volts and delivers about 35 per cent of its total input power in the form of useful output.

Tetrode and pentode tubes do, however, introduce a considerably greater amount of harmonic distortion into their output, and their plate circuit impedance (which acts in a receiver to damp loudspeaker overshoot and ringing, and acts in a driver stage to give good regulation) is many times higher than that of an equivalent triode. The application of negative feedback to an amplifier employing tetrode and pentode tubes acts both to reduce distortion and to reduce the effective plate circuit impedance of the stage. The application of negative feedback to such amplifiers is discussed in Section 5-15 of this chapter.

5-7 Push-Pull Class A and Class AB Audio Amplifier Stages

A number of advantages are obtained through the use of the push-pull connection of two or four tubes in an audio-frequency power amplifier. Two conventional circuits for the use of triode and tetrode tubes in the push-pull connection are shown in figure 12. The two main advantages of the push-pull circuit arrangement are: (1) the magnetizing effect of the plate currents of the output tubes is cancelled in the windings of the output transformer; (2) even harmonics of the input signal (second and fourth harmonics primarily) generated in the push-pull stage are cancelled when the tubes are balanced.

The cancellation of even harmonics generated in the stage allows the tubes to be operated Class AB—in other words the tubes may be operated with bias and input signals of such amplitude that the plate current of alternate tubes may be cut off during a portion of the input voltage cycle. If a tube were operated in such a manner in a single-ended amplifier the second harmonic amplitude generated would be prohibitively high.

Push-pull Class AB operation allows a plate circuit efficiency of from 45 to 60 per cent to be obtained in an amplifier stage depending upon whether or not the exciting voltage is

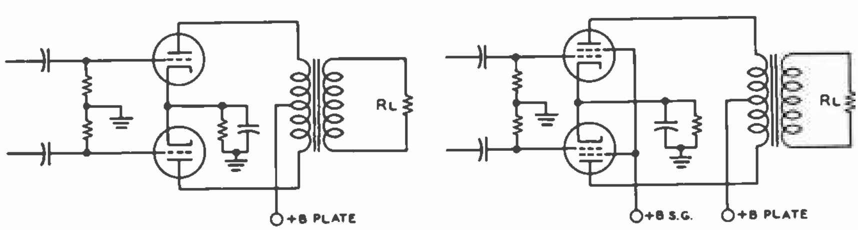


Figure 12.

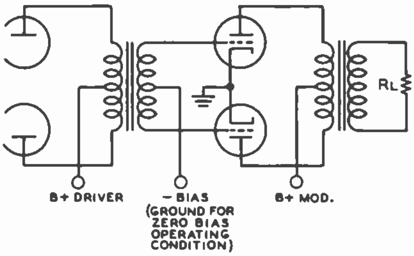


Figure 13.
Class B audio-frequency power amplifier.

of such amplitude that grid current is drawn by the tubes. If grid current is taken on input voltage peaks the amplifier is said to be operating Class AB_2 and the plate circuit efficiency can be as high as the upper value just mentioned. If grid current is not taken by the stage it is said to be operating Class AB_1 and the plate circuit efficiency will be toward the lower end of the range just quoted. In all Class AB amplifiers the plate current will increase from 40 to 150 per cent over the no-signal value when full signal is applied.

5-8 Class B Audio-Frequency Power Amplifiers

The Class B audio-frequency power amplifier operates at a higher plate-circuit efficiency than any of the previously described types of audio power amplifiers. Full-signal plate-circuit efficiencies of 60 to 70 per cent are readily obtainable with the tube types at present available for this type of work. Since the plate circuit efficiency is higher, smaller tubes of lower plate dissipation may be used in a Class B power amplifier of a given power output than can be used in any other conventional type of audio amplifier. An additional factor in favor of the Class B audio amplifier is the fact that the power input to the stage is relatively low under no-signal conditions. It is for these reasons that this type of amplifier has largely superseded other types in the generation of audio-frequency power levels from perhaps 100 watts on up to levels of approximately 150,000 watts as required for large short-wave broadcast stations.

There are attendant disadvantageous features to the operation of a power amplifier of this type; but all these disadvantages can be overcome by proper design of the circuits associated with the power amplifier stage. These disadvantages are: (1) The Class B audio amplifier requires driving power in its grid circuit; this disadvantage can be overcome by the use of an oversize power stage preceding the Class B stage with a step-down transformer between the driver stage and the Class-B grids. Degenerative feedback is sometimes employed to reduce the plate impedance of the driver stage and thus to improve the voltage regulation under the varying load presented by the Class B grids. (2) The Class B stage requires a constant value of average grid bias to be supplied in spite of the fact that the grid current on the stage is zero over most of the cycle but rises to values as high as one-third of the peak plate current on the peak of the exciting voltage cycle. Special regulated bias supplies have been used for this application, or B batteries can be used. However, a number of tubes especially designed for Class B audio amplifiers have been developed which require zero average grid bias for their operation. The 811A, 838, 805, 203Z, 809, HY-5514, and TZ-40 are examples of this type of tube. All these so-called "zero-bias" tubes have rated operating conditions up to moderate plate voltages wherein they can be operated without grid bias. As the plate voltage is increased to their maximum ratings, however, a small amount of grid bias, such as could be obtained from several 4½-volt C batteries, is required.

And (3), a Class B audio-frequency power amplifier or modulator requires a source of plate supply voltage having reasonably good regulation. This requirement led to the development of the "swinging choke." The swinging choke is essentially a conventional filter choke in which the core air gap has been reduced. This reduction in the air gap allows the choke to have a much greater value of inductance with low current values such as are encountered with no signal or small signal being applied to the Class B stage. With a higher value of current such as would be taken by a Class B stage with full signal applied the inductance of the choke drops to a much lower value. With a swinging choke of this type, having adequate current rating, as the input inductor in the filter system for a rectifier power supply, the regulation will be improved to a point which is normally adequate for a power supply for a Class B amplifier or modulator stage. Swinging-choke power supplies

TABLE III

CLASS AB_2 AND CLASS B AUDIO-FREQUENCY POWER AMPLIFIERS

TUBES (2)	PLATE VOLTAGE	GRID BIAS	FILAMENT VOLTAGE	PEAK A.F. GRID TO GRID	ZERO SIGNAL PLATE CURRENT	MAX. SIGNAL PLATE CURRENT	LOAD RESISTANCE	MAX. SIGNAL DRIVING POWER	AVERAGE SINE-WAVE POWER OUTPUT
2E30	180 EP 180 ESG	-22.5	6.0	75	18	100	2500	0.23	7.4
	250 EP 250 ESG	-30	6.0	87	40	120	3800	0.2	17
6V6 (AB_1)	285 EP 285 ESG	-19	6.3	38	70	92	8000	NO GRID CUR.	14
6F6 (AB_2)	350 EP 250 ESG	340 Ω RESISTOR	6.3	94	54	77	10000	0.2	19
6L6 (AB_1)	360 EP 270 ESG	250 Ω RESISTOR	6.3	57	66	100	9000	NO GRID CUR.	24.5
6L6 (AB_2)	360 EP 225 ESG	-18	6.3	52	78	142	6000	0.140	31
6L6 (AB_2)	360 EP 270 ESG	-22.5	6.3	72	88	205	3800	0.270	47
2E26 (PAIR)	400 EP 125 ESG	-15	6.3 OR 12.6	60	20	150	6200	0.36	42
815 (SINGLE)	500 EP 125 ESG	-15	6.3 OR 12.6	60	22	150	8000	0.36	54
807	500 EP 300 ESG	-25	6.3	78	100	240	4240	0.2	75
	600 EP 300 ESG	30	6.3	78	60	200	6400	0.1	80
	750 EP 300 ESG	-32	6.3	92	60	240	6950	0.2	120
HY-69	600 EP 300 ESG	-35	6.0	100	50	200	6500	0.3	60
809	700	0	6.3	160	70	250	6200	3.4	120
	1000	-9	6.3	155	40	200	11600	2.7	145
811	1000	0	6.3	160	30	260	6600	4.0	165
	1250	0	6.3	150	48	240	12000	3.4	210
	1500	-9	6.3	150	20	200	17800	3.0	220
35T	1000	-6	5.0	240	67	240	7900	7.0	140
	1500	-25	5.0	250	45	200	16200	5.0	200
	2000	-40	5.0	255	34	167	27500	4.0	235
203A	1000	-35	10.0	310	26	320	6900	10.0	200
	1250	-45	10.0	330	26	320	9000	11.0	260
211	1000	-77	10.0	380	20	320	6900	7.5	200
	1250	-100	10.0	410	20	320	9000	8.0	260
838	1000	0	10.0	200	106	320	6900	7.0	200
	1250	0	10.0	200	148	320	9000	7.5	260
5514	750	0	7.5	93	46	200	8400	3.0	105
	1250	0	7.5	118	64	300	10000	4.5	270
	1500	-4.5	7.5	146	50	350	10500	6.5	400
8005	1000	-55	10.0	290	40	320	8000	4.0	250
	1250	-70	10.0	310	40	310	10000	4.0	300
828	EP=1700 ESG=750 ESUP=60	-120	10.0	240	50	248	16200	NO GRID CUR.	300
	EP=2000 ESG 750 ESUP.=60	-120	10.0	240	50	270	18500	NO GRID CUR.	385
75TL	1500	-105	5.0	450	67	285	11000	6.0	280
	2000	-160	5.0	534	50	250	18000	5.0	350
100TH	2000	-35	5.0	310	60	280	15000	7.0	380
	3000	-65	5.0	335	40	215	31000	5.0	450
4-125A (AB_1)	EP=2000 ESG=600	-94	5.0	188	50	240	13400	NO GRID CUR.	230
	EP=2500 ESG=600	-96	5.0	192	50	232	20300	NO GRID CUR.	330
4-125A (AB_2)	EP=2000 ESG=350	-45	5.0	210	72	300	13600	3.1	350
	EP=3000 ESG=350	-51	5.0	196	55	260	27700	2.5	520
8003	1350	-100	10.0	480	40	490	6000	10.5	460
813	EP=2250 ESG=750	-90	10.0	230	45	315	18500	0.1	515
	EP=2500 ESG=750	-95	10.0	235	35	360	17000	0.35	650
810	2000	-50	10.0	345	60	420	11000	10.0	590
	2250	-60	10.0	380	70	450	11600	13.0	725
4-250A (AB_1)	EP=2000 ESG=500	-88	5.0	176	110	405	9170	NO GRID CUR.	460
	EP=3000 ESG=500	-93	5.0	186	120	417	15000	NO GRID CUR.	750
4-250A (AB_2)	EP=2000 ESG=300	-48	5.0	198	120	510	8000	2.3	650
	EP=3000 ESG=300	-53	5.0	198	125	473	16000	1.9	1040
304TL (AB_1)	2000	-160	5.0 OR 10.0	320	200	546	5300	NO GRID CUR.	490
	3000	-260	5.0 OR 10.0	520	130	444	12000	NO GRID CUR.	730
304TL (B)	1500	-105	5.0 OR 10.0	500	270	1.14 AMP.	2750	30	1100
	2000	-180	5.0 OR 10.0	580	200	1.0 AMP.	4500	25	1400

are discussed in detail in the *Power Supply* chapter.

Recommended Operating Conditions Table III lists recommended operating conditions for a number of tube types that find frequent application as Class B a-f power amplifiers and modulators. Certain additional operating conditions are also given for other tube types as Class AB_1 and Class AB_2 power amplifiers and modulators.

It is often desirable to operate a pair of tubes as a Class B power amplifier at plate voltages somewhat different from the conditions listed in Table III or given as standard by the vacuum-tube manufacturers. The procedure given in the following paragraphs can be used to determine proper operating conditions for non-standard applications.

Calculation of Operating Conditions of Class B Power Amplifiers The following procedure can be used for the calculation of the operating conditions of Class B power amplifiers when they are to operate into a resistive load such as the type of load presented by a Class C power amplifier. This procedure will be found quite satisfactory for the application of vacuum tubes as Class B modulators when it is desired to operate the tubes under conditions which are not specified in the tube operating characteristics published by the tube manufacturer. The same procedure can be used with equal effectiveness for the calculation of the operating conditions of beam tetrodes as Class AB_2 amplifiers or modulators when the resting plate current on the tubes (no signal condition) is less than 25 or 30 per cent of the maximum-signal plate current.

First, with the average plate characteristics of the tube as published by the manufacturer before you, select a point on the $E_p = E_g$ (diode bend) line at about twice the plate current you expect the tubes to kick to under modulation. If beam tetrode tubes are concerned, select a point at about the same amount of plate current mentioned above, just to the right of the region where the I_b line takes a sharp curve downward. This will be the first trial point, and the plate voltage at the point chosen should be not more than about 20 per cent of the d-c voltage applied to the tubes if good plate-circuit efficiency is desired.

Second, note down the value of i_{pmax} and e_{pmin} at this point.

Third, subtract the value of e_{pmin} from the d-c plate voltage on the tubes.

Fourth, substitute the values obtained in the following equations:

$$P_o = \frac{i_{pmax}\,(E_{bb} - e_{pmin})}{2} = \text{Power output from 2 tubes}$$

$$R_L = 4\,\frac{(E_{bb} - e_{pmin})}{i_{pmax}}$$

$= \text{Plate-to-plate load for 2 tubes}$

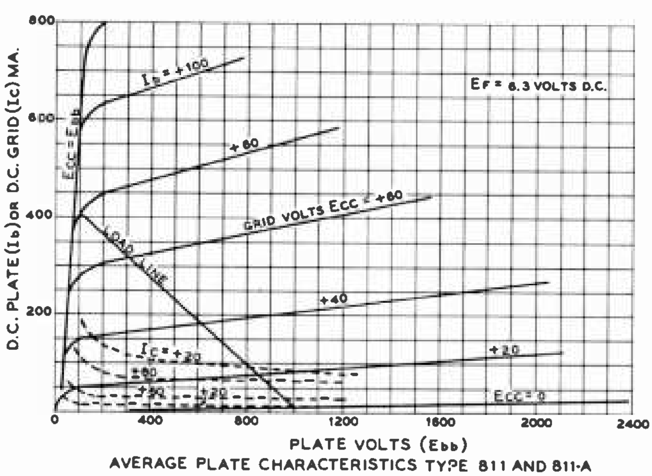


Figure 14.
Typical Class B a-f amplifier load line. The load line has been drawn on the average characteristics of a type 811 tube.

$$N_p = 78.5 \left(1 - \frac{e_{p\,min}}{E_{bb}}\right)$$

= Full signal plate efficiency

All the above equations are true for sine-wave operating conditions on the tubes concerned. However, if a speech clipper is being used in the speech amplifier, or if it is desired to calculate the operating conditions on the basis of the fact that the ratio of peak power to average power in a speech wave is approximately 4-to-1 as contrasted to the ratio of 2-to-1 in a sine wave — in other words, when non-sinusoidal waves such as plain speech or speech that has passed through a clipper are concerned, we are no longer concerned with average power output of the modulator as far as its capability of modulating a class-C amplifier is concerned; we are concerned with its *peak-power-output* capability.

Under these conditions we call upon other, more general relationships. The first of these is: It requires a *peak* power output *equal* to the Class-C stage input to modulate that input fully.

The second one is: The average power output required of the modulator is equal to the shape factor of the modulating wave multiplied by the input to the Class-C stage. The shape factor of unclipped speech is approximately 0.25. The shape factor of a sine wave is 0.5. The shape factor of a speech wave that has been passed through a clipper-filter arrangement is somewhere between 0.25 and 0.9 depending upon the amount of clipping that has taken place. With 15 or 20 db of clipping the shape factor may be as high as the figure of 0.9 mentioned above. This means that the audio power output of the modulator will be 90% of the input to the Class C stage. Thus with a kilowatt input we would be putting 900 watts of audio into the Class-C stage for 100 per cent modulation as contrasted to perhaps 250 watts for unclipped speech modulation of 100 per cent.

Sample Calculation for 811A Tubes Figure 14 shows a set of plate characteristics for a type 811A tube with a load line for Class B operation. Figure 15 lists a sample calculation for determining the proper operating conditions for obtaining approximately 185 watts output from a pair of the tubes with 1000 volts d-c plate potential.

SAMPLE CALCULATION

CONDITION: 2 TYPE 811 TUBES, E_{bb} = 1000

INPUT TO FINAL STAGE, 350 W.

PEAK POWER OUTPUT NEEDED = 350 + 6% = 370 W.

FINAL AMPLIFIER E_{bb} = 2000 V.

FINAL AMPLIFIER I_b = .175 A.

FINAL AMPLIFIER $Z_L = \frac{2000}{.175} = 11400\ \Omega$

EXAMPLE: CHOSE POINT ON 811 CHARACTERISTICS JUST TO RIGHT OF $E_{bb} = E_{cc}$.

I_p MAX. = .410 A. E_p MIN. = +100

I_g MAX. = .100 A. E_g MAX. = +80

PEAK P_o = .410 X (1000−100) = .410 X 900 = 369 W.

$R_L = 4 \times \frac{900}{.410} = 8800\ \Omega$

$N_p = 78.5 \left(1 - \frac{100}{1000}\right) = 78.5\,(.9) = 70.5\%$

W_o (AVERAGE WITH SINE WAVE) $= \frac{P_{o(PEAK)}}{2} = 184.5$ W

$W_{in} = \frac{184.5}{70.5} = 260$ W.

I_b (MAXIMUM WITH SINE WAVE) = 260 MA

W_g PEAK = .100 X 80 = 8 W.

DRIVING POWER $= \frac{W_{g\,PK}}{2} = 4$ W.

TRANSFORMER:

$\frac{Z_s}{Z_p} = \frac{11400}{8800} = 1.29$

TURNS RATIO $= \sqrt{\frac{Z_s}{Z_p}} = \sqrt{1.29} = 1.14$ STEP UP

Figure 15.

Typical calculation of operating conditions for a Class B a-f power amplifier using a pair of type 811 or 811A tubes. Plate characteristics and load line shown in figure 14.

Also shown in figure 15 is the method of determining the proper ratio for the modulation transformer to couple between the 811's or 811A's and the anticipated final amplifier which is to operate at 2000 plate volts and 175 ma. plate current.

Modulation Transformer Calculation The method illustrated in figure 15 can be used in general for the determination of the proper transformer ratio to couple between the modulator tube and the amplifier to be modulated. The procedure can be stated as follows: (1) Determine the proper plate-to-plate load impedance for the modulator tubes either by the use of the type of calculation shown in figure 15, by reference to Table III, or by referenec to the published characteristics on the tubes to be used. (2) Determine the load impedance which will be presented by the Class C amplifier stage to be modulated by dividing the operating plate voltage on that stage by the operating value of plate current in *amperes*. (3) Divide the Class C load impedance determined in (2) above by the plate-to-plate load impedance for the modulator tubes determined in (1) above. The ratio determined in this way is the secondary-to-primary *impedance* ratio. (4) Take the square root of this ratio to determine the secondary-to-primary *turns* ratio. If the turns ratio is greater than one the use of a step-up transformer is required. If the turns ratio as determined in this way is less than one a step-down transformer is called for.

If the procedure shown in figure 15 has been used to calculate the operating conditions for the modulator tubes, the transformer ratio calculation can be checked in the following manner: Divide the plate voltage on the modulated amplifier by the total voltage swing on the modulator tubes: $2\,(E_{bb} - e_{min})$. This ratio should be quite close numerically to the transformer turns ratio as previously determined. The reason for this condition is quite obvious; the ratio between the total primary voltage and the d-c plate supply voltage on the modulated stage is equal to the turns ratio of the transformer since a peak secondary voltage equal to the plate voltage on the modulated stage is required to modulate this stage 100 per cent.

Note Concerning Use of Clipper Speech Amplifier with Tetrode Modulator Tubes

When a clipper speech amplifier is used in conjunction with a Class B modulator stage, the plate current on that stage will kick to a higher value with modulation (due to the greater average power output and input) but the plate dissipation on the tubes will ordinarily be less than with sine-wave modulation. However, when tetrode tubes are used as modulators, the screen dissipation will be much greater than with sine-wave modulation. Care must be taken to insure that the screen dissipation rating on the modulator tubes is not exceeded under full modulation conditions with a clipper speech amplifier. The screen dissipation is equal to screen voltage times screen current.

5-9 Cathode-Follower Power Amplifiers

The cathode-follower amplifier was mentioned briefly at the beginning of this chapter under "Classes and Types of Amplifiers." The cathode-follower is essentially a power output

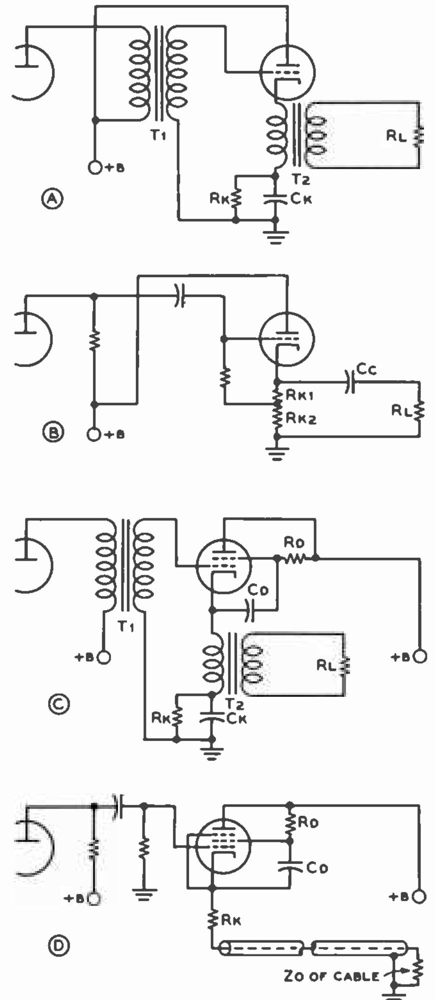


Figure 16.
Cathode-follower output circuits for audio or video amplifiers.

stage in which the exciting signal is applied between grid and ground, the plate is maintained at ground potential with respect to input and output signals, and the output signal is taken between cathode and ground. Figure 16 illustrates four types of cathode-follower power amplifiers in common usage and figure 17 shows the output impedance (R_o), and stage gain (A) of both triode and pentode (or tetrode) cathode-follower stages. It will be seen by inspection of the equations that the stage voltage gain is always less than one, that the output impedance of the stage is much less than the same stage operated as a conventional cathode-return amplifier. The output impedance for conventional tubes will be somewhere between 100 and 1000 ohms, depending primarily on the transconductance of the tube.

This reduction in gain and output impedance for the cathode-follower comes about since the stage operates as though it had 100 per cent degenerative feedback applied between its output and input circuit. Even though the voltage gain of the stage is reduced to a value less than one by the action of the degenerative feedback, the power gain of the stage if it is operating Class A is not reduced. Although more voltage is required to excite a cathode-follower amplifier than appears across the load circuit, since the cathode "follows" along with the grid, the relative grid-to-cathode voltage is essentially the same as in a conventional amplifier.

Although the cathode-follower gives no voltage gain, it is an effective power amplifier where it is desired to feed a low-impedance load, or where it is desired to feed a load of varying impedance with a signal having good regulation. This latter capability makes the cathode follower particularly effective as a driver for the grids of a Class B modulator stage.

The circuit of figure 16A is the type of amplifier, either single-ended or push-pull, which may be used as a driver for a Class B modulator or which may be used for other applications such as feeding a loudspeaker where unusually good damping of the speaker is desired. If the d-c resistance of the primary of the transformer T_2 is approximately the correct value for the cathode bias resistor for the amplifier tube, the components R_k and C_k need not be used. Figure 16B shows an arrangement which may be used to feed directly a value of load impedance which is equal to or higher than the cathode impedance of the amplifier tube. The value of C_c must be quite high, somewhat higher than would be used in a conventional circuit, if the frequency response of the circuit when operating into a low-impedance load is to be preserved.

Figures 16C and 16D show cathode-follower circuits for use with tetrode or pentode tubes. Figure 16C is a circuit similar to that shown in 16A and essentially the same comments apply in regard to the components R_k and C_k and the primary resistance of the transformer T_2. Notice also that the screen of the tube is maintained at the same signal potential as the cathode by means of coupling capacitor C_d. This capacitance should be large enough so that at the lowest frequency it is desired to pass through the stage its reactance will be low with respect to the dynamic screen-to-cathode resistance in parallel with R_d. T_2 in this stage as well as in the circuit of figure 16A should have the proper turns (or impedance) ratio to give the desired step-down or step-up from the cathode circuit to the load. Figure 16D is an arrangement frequently used in video systems for feeding a coaxial cable of relatively low impedance from a vacuum-tube amplifier. A pentode or tetrode tube with a cathode impedance as a cathode follower ($1/G_m$) approximately the same as the cable impedance should be chosen. The 6AG7 and 6AC7 have cathode impedances of the same order as the surge impedances of certain types of low-capacitance coaxial cable. An arrangement such as 16D is also usable for feeding coaxial cable with audio or r-f energy where it is desired to transmit the output signal over moderate distances. The resistor R_k is added to the circuit as shown if the cathode impedance of the tube used is lower than the characteristic impedance of the cable. If the output impedance of the stage is higher than the cable impedance a resistance of appropriate value is sometimes placed in parallel with the input end of the cable. The values of C_d and R_d should be chosen with the same considerations in mind as mentioned in the discussion of the circuit of figure 16C above.

The cathode follower may conveniently be used as a method of coupling r-f or i-f energy between two units separated a considerable distance. In such an application a coaxial cable

TRIODE: $\mu_{CF} = \frac{\mu}{\mu+1}$ $\qquad A = \frac{\mu R_L}{R_L(\mu+1)+R_P}$

$R_{O(CATHODE)} = \frac{R_P}{\mu+1}$ $\qquad R_L = \frac{(R_{K1}+R_{K2})R_L'}{R_{K1}+R_{K2}+R_L'}$

PENTODE: $R_{O(CATHODE)} = \frac{1}{G_M}$ $\qquad R_{EQ} = \frac{R_L}{1+R_L G_M}$

$A = G_M R_{EQ}$

Figure 17.

Equivalent factors for pentode (or tetrode) cathode-follower power amplifiers.

should be used to carry the r-f or i-f energy. One such application would be for carrying the output of a v-f-o to a transmitter located a considerable distance from the operating position. Another application would be where it is desired to feed a single-sideband demodulator, an FM adaptor, or another accessory with intermediate-frequency signal from a communications receiver. A tube such as a 6CB6 connected in a manner such as is shown in figure 16D would be adequate for the i-f amplifier coupler, while a 6L6 or a 6AG7 could be used in the output stage of a v-f-o as a cathode follower to feed the coaxial line which carries the v-f-o signal from the control unit to the transmitter proper.

TUNED R-F VOLTAGE AMPLIFIERS

Tuned r-f voltage amplifiers are used in receivers for the amplification of the incoming r-f signal and for the amplification of intermediate frequency signals after the incoming frequency has been converted to the intermediate frequency by the mixer stage. Signal frequency stages are normally called tuned r-f amplifiers and intermediate-frequency stages are called i-f amplifiers. Both tuned r-f and i-f amplifiers are operated Class A and normally operate at signal levels from a fraction of a microvolt to amplitudes as high as 10 to 50 volts at the plate of the last i-f stage in a receiver.

5-10 Grid-Circuit Considerations

Since the full amplification of a receiver follows the first tuned circuit, the operating conditions existing in that circuit and in its coupling to the antenna on one side and to the grid of the first amplifier stage on the other are of greatest importance in determining the signal-to-noise ratio of the receiver on weak signals.

First Tuned Circuit It is obviously of great importance that the highest ratio of signal-to-noise be impressed on the grid of the first r-f amplifier tube. Attaining the optimum ratio is a complex problem since noise will be generated in the antenna due to its equivalent radiation resistance (this noise is in addition to any noise of atmospheric origin) and in the first tuned circuit due to its equivalent coupled resistance at resonance. The noise voltage generated due to antenna radiation resistance and to equivalent tuned circuit resistance is similar to that generated in a resistor due to thermal agitation and is expressed by the following equation:

$$E_n^2 = 4kTR\Delta f$$

Where: E_n = r-m-s value of noise voltage over the interval Δf

k = Boltzman's constant = 1.374×10^{-23} joule per °K.

T = Absolute temperature °K.

R = Resistive component of impedance across which thermal noise is developed.

Δf = Frequency band across which voltage is measured.

In the above equation Δf is essentially the frequency band passed by the intermediate frequency amplifier of the receiver under consideration. This equation can be greatly simplified for the conditions normally encountered in communications work. If we assume the following conditions: T = 300° K or 27° C or 80.5° F, room temperature; Δf = 8000 cycles (the average pass band of a communications receiver or speech amplifier), the equation reduces to: $E_{r.m.s.} = 0.0115\sqrt{R}$ microvolts. Accordingly, the thermal-agitation voltage appearing in the center of half-wave antenna (assuming effective temperature to be 300° K) having a radiation resistance of 73 ohms is approximately 0.096 microvolts. Also, the thermal agitation voltage appearing across a 500,000-ohm grid resistor in the first stage of a speech amplifier is approximately 8 microvolts under the conditions cited above. Further, the voltage due to thermal agitation being impressed on the grid of the first r-f stage in a receiver by a first tuned circuit whose resonant resistance is 50,000 ohms is approximately 2.5 microvolts. Suffice to say, however, that the value of thermal agitation voltage appearing across the first tuned circuit when the antenna is properly coupled to this circuit will be very much less than this value.

It is common practice to match the impedance of the antenna transmission line to the input impedance of the grid of the first r-f amplifier stage in a receiver. This is the condition of antenna coupling which gives maximum gain in the receiver. However, when u-h-f tubes such as acorns and miniatures are used at frequencies somewhat less than their maximum capabilities, a significant improvement in *signal-to-noise* ratio can be attained by increasing the coupling between the antenna and first tuned circuit to a value greater than that which gives greatest signal amplitude out of the receiver. In other words, in the 10, 6, and 2 meter bands it is possible to attain somewhat improved signal-to-noise ratio by increasing antenna coupling to the point where the gain of the receiver is slightly reduced.

It is always possible, in addition, to obtain improved signal-to-noise ratio in a v-h-f receiver through the use of tubes which have improved input impedance characteristics at the frequency in question over conventional types.

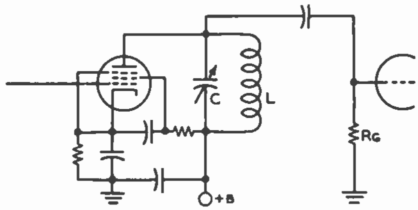
Noise Factor The limiting condition for sensitivity in any receiver is the thermal noise generated in the antenna and in the first tuned circuit. However, with proper coupling between the antenna and the grid of the tube, through the first tuned circuit, the noise contribution of the first tuned circuit can be made quite small. Unfortunately, though, the major noise contribution in a properly designed receiver is that of the first tube. The noise contribution due to electron flow and due to losses in the tube can be lumped into an equivalent value of resistance which, if placed in the grid circuit of a perfect tube having the same gain but no noise would give the same noise voltage output in the plate load. The equivalent noise resistance of tubes such as the 6SK7, 6SG7, etc., runs from 5000 to 10,000 ohms. Very high G_m tubes such as the 6AC7 and 6AK5 have equivalent noise resistances as low as 700 to 1500 ohms. The lower the value of equivalent noise resistance, the lower will be the noise output under a fixed set of conditions.

The equivalent noise resistance of a tube must not be confused with the actual input loading resistance of a tube. For highest signal-to-noise ratio in an amplifier the input loading resistance should be as high as possible so that the amount of voltage that can be developed from grid to ground by the antenna energy will be as high as possible, the equivalent noise resistance should be as low as possible so that the noise generated by this resistance will be lower than that attributable to the antenna and first tuned circuit, and the losses in the first tuned circuit should be as low as possible.

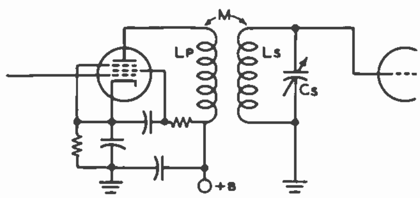
The absolute sensitivity of receivers has been designated in recent years in government and commercial work by an arbitrary dimensionless number known as "noise factor" or *N*. The noise factor is the ratio of noise output of a "perfect" receiver having a given amount of gain with a dummy antenna matched to its input, to the noise output of the receiver under measurement having the same amount of gain with the dummy antenna matched to its input. Although a perfect receiver is not a physically realizable thing, the noise factor of a receiver under measurement can be determined by calculation from the amount of additional noise (from a temperature-limited diode or other calibrated noise generator) required to increase the noise power output of a receiver by a predetermined amount.

Tube Input Loading As has been mentioned in a previous paragraph, greatest gain in a receiver is obtained when the antenna is matched, through the r-f coupling transformer, to the input resistance of the r-f tube. However, the higher the ratio of tube input resistance to equivalent noise resistance of the tube the higher will be the signal-to-noise ratio of the stage—and of course, the better will be the noise factor of the overall receiver. The input resistance of a tube is very high at frequencies in the broadcast band and gradually decreases as the frequency increases. Tube input resistance on conventional tube types begins to become an important factor at frequencies of about 25 Mc. and above. At frequencies above about 100 Mc. the use of conventional tube types becomes impracticable since the input resistance of the tube has become so much lower than the equivalent noise resistance that it is impossible to attain reasonable signal-to-noise ratio on any but very strong signals. Hence, special v-h-f tube types such as the 6AK5, 6AG5, and 6CB6 must be used.

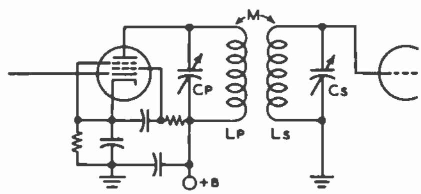
The lowering of the effective input resistance of a vacuum tube at higher frequencies is brought about by a number of factors. The



Ⓐ AMPLIFICATION AT RESONANCE (APPROX.) = $G_M \omega L Q$



Ⓑ AMPLIFICATION AT RESONANCE (APPROX.) = $G_M \omega M Q$



Ⓒ AMPLIFICATION AT RESONANCE (APPROX.) = $G_M K \dfrac{\omega\sqrt{L_P L_S}}{K^2 + \frac{1}{Q_P Q_S}}$

WHERE: 1. PRI. AND SEC. RESONANT AT SAME FREQUENCY
2. K IS COEFFICIENT OF COUPLING

IF PRI. AND SEC. Q ARE APPROXIMATELY THE SAME:
$\dfrac{\text{TOTAL BANDWIDTH}}{\text{CENTER FREQUENCY}} = 1.2\ K$

MAXIMUM AMPLITUDE OCCURS AT CRITICAL COUPLING — WHEN $K = \dfrac{1}{\sqrt{Q_P Q_S}}$

Figure 18.

Gain equations for pentode r-f amplifier stages operating into a tuned load.

first, and most obvious, is the fact that the dielectric loss in the internal insulators, and in the base and press of the tube increases with frequency. The second factor is due to the fact that a finite time is required for an electron to move from the space charge in the vicinity of the cathode, pass between the grid wires, and travel on to the plate. The fact that the electrostatic effect of the grid on the moving electron acts over an appreciable portion of a cycle at these high frequencies causes a current flow in the grid circuit which appears to the input circuit feeding the grid as a resistance. The decrease in input resistance of a tube due to electron transit time varies as the square of the frequency. The undesirable effects of transit time can be reduced in certain cases by the use of higher plate voltages. Transit time varies inversely as the square root of the applied plate voltage.

Cathode lead inductance is an additional cause of reduced input resistance at high frequencies. This effect has been reduced in certain tubes such as the 6SH7 and the 6AK5 by providing two cathode leads on the tube base. One cathode lead should be connected to the input circuit of the tube and the other lead should be connected to the by-pass capacitor for the plate return of the tube.

5-11 Plate-Circuit Considerations

Noise is generated in a vacuum tube by the fact that the current flow within the tube is not a smooth flow but rather is made up of the continuous arrival of particles (electrons) at a very high rate. This "shot effect" is a source of noise in the tube, but its effect is referred back to the grid circuit of the tube since it is included in the "equivalent noise resistance" discussed in the preceding paragraphs.

So, for the purpose of this section, it will be considered that the function of the plate load circuit of a tuned vacuum-tube amplifier is to deliver energy to the next stage with the greatest efficiency over the required band of frequencies. Three methods of interstage coupling for tuned r-f voltage amplifiers are shown in figure 18. In figure 18A omega (ω) is 2π times the resonant frequency of the circuit in the plate of the amplifier tube, and L and Q are the inductance and Q of the inductor L. In figure 18B the notation is the same and M is the mutual inductance between the primary coil and the secondary coil. In figure 18C the notation is again the same and k is the coefficient of coupling between the two tuned circuits. As the coefficient of coupling between the circuits is increased the bandwidth becomes greater but the response over the band becomes progressively more double-humped. The response over the band is the most flat when the Q's of primary and secondary are approximately the same and the value of each Q is equal to $1.75/k$.

Variable-Mu Tubes It is common practice to control the gain of a succession of r-f or i-f amplifier stages by varying the average bias on their control grids. However, as the bias is raised above the operating value on a conventional sharp-cutoff tube the tube becomes increasingly non-linear in operation as cutoff of plate current is approached. The effect of such non-linearity is to cause cross modulation between strong signals which appear on the grid of the tube. When a tube operating in such a manner is in one of the first stages of a receiver a number of signals are appearing on its grid simultaneously and cross modulation between them will take place. The result of this effect is to produce a large number of spurious signals in the output of the receiver—in most cases these signals will carry the modulation of both the carriers which have been cross modulated to produce the spurious signal.

The undesirable effect of cross modulation can be eliminated in most cases and greatly reduced in the balance through the use of a variable-mu tube in all stages which have a-v-c voltage or other large negative bias applied to their grids. The variable-mu tube has a characteristic which causes the cutoff of plate current to be gradual with an increase in grid bias, and the reduction in plate current is accompanied by a decrease in the effective amplification factor of the tube. Variable-mu tubes ordinarily have somewhat reduced G_m as compared to a sharp-cutoff tube of the same group. Hence the sharp-cutoff tube will perform best in stages to which a-v-c voltage is not applied.

RADIO-FREQUENCY POWER AMPLIFIERS

All modern transmitters in the medium-frequency range and an increasing percentage of those in the v-h-f and u-h-f ranges consist of a comparatively low-level source of radio-frequency energy which is multiplied in frequency and successively amplified to the desired power level. Microwave transmitters are still predominately of the self-excited oscillator type, but when it is possible to use r-f amplifiers in s-h-f transmitters the flexibility of their application will be increased. The following portion of this chapter will be devoted, however, to the method of operation and calculation of operating characteristics of r-f power amplifiers for operation in the range of approximately 3.5 to 500 Mc.

5-12 Class C R-F Power Amplifiers

The majority of r-f power amplifiers fall into the Class C category since such stages can be made to give the best plate circuit efficiency of any present type of vacuum-tube amplifier. Hence, the cost of tubes for such a stage and the cost of the power to supply that stage is least for any given power output. Nevertheless, the Class C amplifier gives less power gain than either a Class A or Class B amplifier under similar conditions since the grid of a Class C stage must be driven highly positive over the portion of the cycle of the exciting wave when the plate voltage on the amplifier is low, and must be at a large negative potential over a large portion of the cycle so that no plate current will flow except when plate voltage is very low. This, in fact, is the fundamental reason why the plate circuit efficiency of a Class C amplifier stage can be made high—plate current is cut off at all times except when the plate-to-cathode voltage drop across the tube is at its lowest value. Class C amplifiers almost invariably operate into a tuned tank circuit as a load, and as a result are used as amplifiers of a single frequency or of a comparatively narrow band of frequencies.

Figure 19 shows the relationships between the various voltages and currents over one cycle of the exciting grid voltage for a Class C amplifier stage. The notation given in figure 19 and in the discussion to follow is the same as given at the first of this chapter under "Symbols for Vacuum-Tube Parameters."

The various manufacturers of vacuum tubes

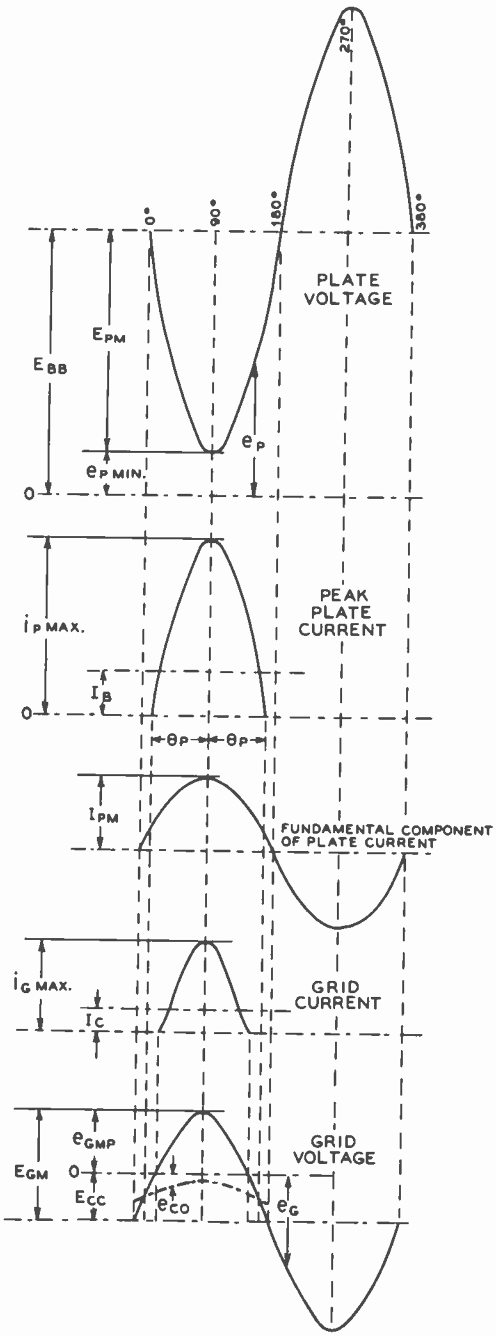


Figure 19.
Instantaneous electrode and tank circuit voltages and currents for a Class C r-f power amplifier.

publish booklets listing in adequate detail alternative Class C operating conditions for the tubes which they manufacture. In addition, operating condition sheets for any particular type of vacuum tube are available for the asking from the different vacuum-tube manufacturers. It is, nevertheless, often desirable to determine optimum operating conditions for a tube under a particular set of circumstances. To assist in such calculations the following paragraphs are devoted to a method of calculating Class C operating conditions which is moderately simple and yet sufficiently accurate for all practical purposes.

Calculation of Class C Amplifier Operating Characteristics* Although Class C operating conditions can be determined with the aid of the more conventional grid voltage-plate current operating curves, the calculation is considerably simplified if the alternative "constant-current curve" of the tube in question is used. This is true since the operating line of a Class C amplifier is a straight line on a set of constant-current curves. A set of constant-current curves on the 250TH tube with a sample load line drawn thereon is shown in figure 22.

In calculating and predicting the operation of a vacuum tube as a Class C radio-frequency amplifier, the considerations which determine the operating conditions are plate efficiency, power output required, maximum allowable plate and grid dissipation, maximum allowable plate voltage and maximum allowable plate current. The values chosen for these factors will depend both upon the demands of a particular application and upon the tube chosen.

The plate and grid currents of a Class C amplifier tube are periodic pulses, the durations of which are always less than 180 degrees. For this reason the average grid current, average plate current, power output, driving power, etc., cannot be directly calculated but must be determined by a Fourier analysis from points selected at proper intervals along the line of operation as plotted upon the constant-current characteristics. This may be done either analytically or graphically. While the Fourier analysis has the advantage of accuracy, it also has the disadvantage of being tedious and involved.

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* Adapted from a procedure given in the Mar.-April 1945, Eimac News.

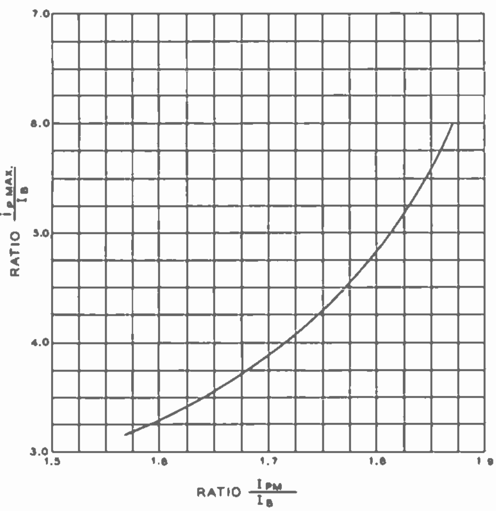


Figure 20.
Relationship between the peak value of the fundamental component of tube plate current, and average plate current; as compared to the ratio of the instantaneous peak value of tube plate current, and average plate current.

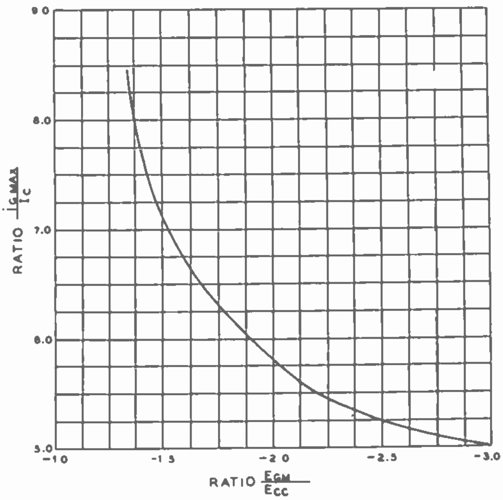


Figure 21.
Relationship between the ratio of the peak value of the fundamental component of the grid excitation voltage, and the average grid bias; as compared to the ratio between instantaneous peak grid current and average grid current.

The approximate analysis which follows has proved to be sufficiently accurate for most applications. This type of analysis also has the advantage of giving the desired information at the first trial. The system is direct in giving the desired information since the important factors, power output, plate efficiency, and plate voltage are arbitrarily selected at the beginning.

Method of Calculation The first step in the method to be described is to determine the power which must be delivered by the Class C amplifier. In making this determination it is well to remember that ordinarily from 5 to 10 per cent of the power delivered by the amplifier tube or tubes will be lost in well-designed tank and coupling circuits at frequencies below 20 Mc. Above 20 Mc. the tank and circuit losses are ordinarily somewhat above 10 per cent.

The plate power input necessary to produce the desired output is determined by the plate efficiency: $P_{in} = P_{out}/N_p$.

For most applications it is desirable to operate at the highest practicable efficiency. High-efficiency operation usually requires less expensive tubes and power supplies, and the amount of artificial cooling required is frequently less than for low-efficiency operation. On the other hand, high-efficiency operation usually requires more driving power and involves the use of higher plate voltages and higher peak tube voltages. The better types of triodes will ordinarily operate at a plate efficiency of 75 to 85 per cent at the highest rated plate voltage, and at a plate efficiency of 65 to 75 per cent at intermediate values of plate voltage.

The first determining factor in selecting a tube or tubes for a particular application is the amount of plate dissipation which will be required of the stage. The total plate dissipation rating for the tube or tubes to be used in the stage must be equal to or greater than that calculated from: $P_p = P_{in} - P_{out}$.

After selecting a tube or tubes to meet the power output and plate dissipation requirements it becomes necessary to determine from the tube characteristics whether the tube selected is capable of the desired operation and, if so, to determine the driving power, grid bias, and grid dissipation.

The complete procedure necessary to determine a set of Class C amplifier operating conditions is given in the following steps:*

1. Select the plate voltage, power output, and efficiency.

2. Determine plate input from: $P_{in} = P_{out}/N_p$.

3. Determine plate dissipation from: $P_p = P_{in} - P_{out}$. P_p must not exceed maximum rated plate dissipation for tube or tubes selected.

4. Determine average plate current from: $I_b = P_{in}/E_{bb}$.

5. Determine approximate i_{pmax} from:
   $i_{pmax} = 4.9\ I_b$ for $N_p = 0.85$
   $i_{pmax} = 4.5\ I_b$ for $N_p = 0.80$
   $i_{pmax} = 4.0\ I_b$ for $N_p = 0.75$
   $i_{pmax} = 3.5\ I_b$ for $N_p = 0.70$

6. Locate the point on constant-current characteristics where the constant plate current line corresponding to the approximate i_{pmax} determined in step 5 crosses the line of equal plate and grid voltages (diode line). Read e_{pmin} at this point. In a few cases the lines of constant plate current will inflect sharply upward before reaching the diode line. In these cases e_{pmin} should not be read at the diode line but at the point where the plate current line intersects a line drawn from the origin through these points of inflection.

7. Calculate E_{pm} from: $E_{pm} = E_{bb} - e_{pmin}$.

8. Calculate the ratio I_{pm}/I_b from:

$$\frac{I_{pm}}{I_b} = \frac{2\ N_p\ E_{bb}}{E_{pm}}$$

9. From the ratio of I_{pm}/I_b calculated in step 8 determine the ratio i_{pmax}/I_b from figure 20.

10. Calculate a new value for i_{pmax} from the ratio found in step 9.
    $i_{pmax} =$ (ratio from step 9) I_b

11. Read e_{gmp} and i_{gmax} from the constant-current characteristics for the values of e_{pmin} and i_{pmax} determined in steps 6 and 10.

12. Calculate the cosine of one-half the angle of plate current flow from:

$$\cos \theta_p = 2.32\left(\frac{I_{pm}}{I_b} - 1.57\right)$$

13. Calculate the grid bias voltage from:

$$E_{cc} = \frac{1}{1 - \cos\theta_p} \times \left[\cos\theta_p\left(\frac{E_{pm}}{\mu} - e_{gmp}\right) - \frac{E_{bb}}{\mu}\right]$$

14. Calculate the peak fundamental grid excitation voltage from:

$$E_{gm} = e_{gmp} - E_{cc}$$

15. Calculate the ratio E_{gm}/E_{cc} for the values of E_{cc} and E_{gm} found in steps 13 and 14.

16. Read i_{gmax}/I_c from figure 21 for the ratio E_{gm}/E_{cc} found in step 15.

17. Calculate the average grid current from the ratio found in step 16, and the value of i_{gmax} found in step 11:

$$I_c = \frac{i_{gmax}}{\text{Ratio from step 16}}$$

18. Calculate approximate grid driving power from:

$$P_d = 0.9\ E_{gm} I_c$$

19. Calculate grid dissipation from:

$$P_g = P_d + E_{cc} I_c$$

P_g must not exceed the maximum rated grid dissipation for the tube selected.

Sample Calculation A typical example of a Class C amplifier calculation is shown in the example below. Reference is made to figures 20, 21 and 22 in the calculation.

1. Desired power output—800 watts.
   Desired plate voltage—3500 volts.
   Desired plate efficiency—80 per cent ($N_p = 0.80$)

2. $P_{in} = 800/0.80 = 1000$ watts

3. $P_p = 1000 - 800 = 200$ watts
   Use 250TH; max. $P_p = 250$w.; $\mu = 37$.

4. $I_b = 1000/3500 \doteq 0.285$ ampere (285 ma.)
   Max. I_b for 250TH is 350 ma.

5. Approximate $i_{pmax} = 0.285 \times 4.5 = 1.28$ ampere

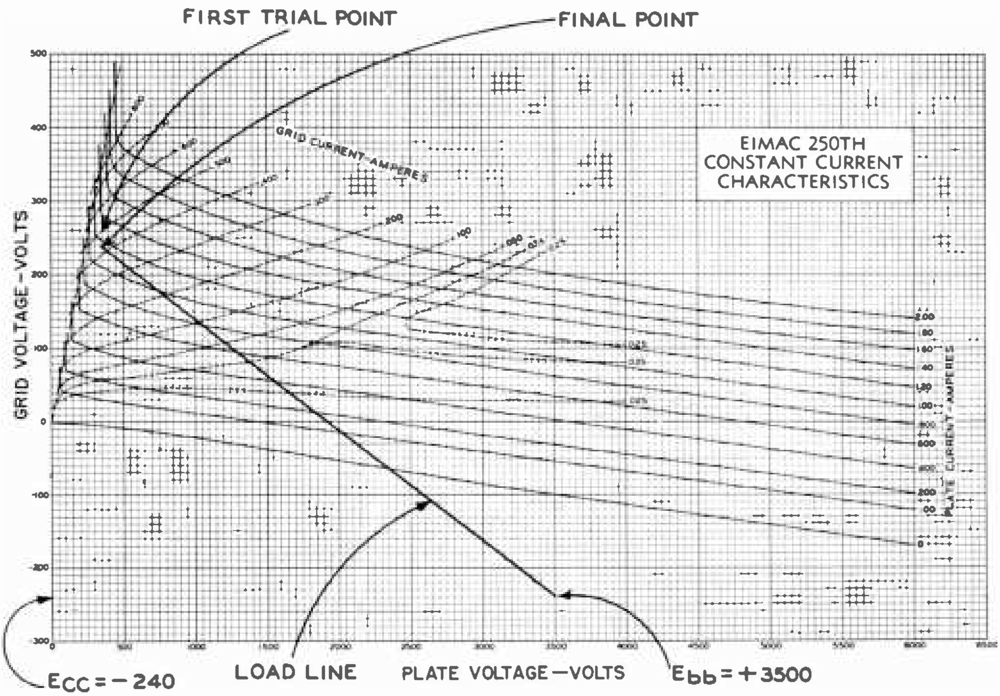


FIGURE 22

Active portion of the operating load line for an Eimac 250TH triode Class C r-f power amplifier, showing first trial point and the final operating point.

6. e_{pmin} = 260 volts (see figure 22, first trial point)

7. $E_{pm} = 3500 - 260 = 3240$ volts

8. $I_{pm}/I_b = 2 \times 0.80 \times 3500/3240 = 5600/3240 = 1.73$

9. $i_{pmax}/I_b = 4.1$ (from figure 20)

10. $i_{pmax} = 0.285 \times 4.1 = 1.17$

11. e_{gmp} = 240 volts
i_{gmax} = 0.430 amperes
(Both above from final point on figure 22)

12. $\cos \theta_p = 2.32\ (1.73 - 1.57) = 0.37$
($\theta_p = 68.3°$)

13. $$E_c = \frac{1}{1 - 0.37} \times \left[0.37\left(\frac{3240}{37} - 240\right) - \frac{3500}{37}\right]$$
$= -240$ volts

14. $E_{gm} = 240 - (-240) = 480$ volts grid swing

15. $E_{gm}/E_{cc} = 480/-240 = -2$

16. $i_{gmax}/I_c = 5.75$ (from figure 21)

17. $I_c = 0.430/5.75 = 0.075$ amp. (75 ma. grid current)

18. $P_d = 0.9 \times 480 \times 0.075 = 32.5$ watts driving power

19. $P_g = 32.5 - (-240 \times 0.75) = 14.5$ watts grid dissipation

Max. P_g for 250TH is 40 watts

The power output of any type of r-f amplifier is equal to:

$$I_{pm}E_{pm}/2 = P_o$$

I_{pm} can be determined, of course, from the ratio determined in step 8 above (in this type of calculation) by multiplying this ratio times I_b.

It is frequently of importance to know the value of load impedance into which a Class C amplifier operating under a certain set of conditions should operate. This is simply $R_L = E_{pm}/I_{pm}$. In the case of the operating conditions just determined for a 250TH amplifier stage the value of load impedance is:

$$R_L = \frac{E_{pm}}{I_{pm}} = \frac{3240}{.495} = 6600 \text{ ohms}$$

$$I_{pm} = \frac{I_{pm}}{I_b} \times I_b$$

Q of Amplifier Tank Circuit In order to obtain good plate tank circuit tuning and low radiation of harmonics from an amplifier it is necessary that the plate tank circuit have the correct Q. Charts giving compromise values of Q for Class C amplifiers are given in Chapter Seven, *Generation of R-F Energy*. However, the amount of inductance required for a specified tank circuit Q under specified operating conditions can be calculated from the following expression:

$$\omega L = \frac{R_L}{Q}$$

ω = 2 π × operating frequency
L = Tank inductance
R_L = Required tube load impedance
Q = Effective tank circuit Q

A tank circuit Q of 12 to 20 is recommended for all normal conditions. However, if a balanced push-pull amplifier is employed the tank receives two impulses per cycle and the circuit Q may be lowered somewhat from the above values.

Quick Method of Calculating Amplifier Plate Efficiency The plate circuit efficiency of a Class B or Class C r-f amplifier can be determined from the following facts. The plate circuit efficiency of such an amplifier is equal to the product of two factors, F_1, which is equal to the ratio of E_{pm} to E_{bb} ($F_1 = E_{pm}/E_{bb}$) and F_2, which is proportional to the one-half angle of plate current flow, θ_p. A graph of F_2 against both θ_p and cos θ_p is given in figure 23. Either θ_p or cos θ_p may be used to determine F_2. Cos θ_p may be determined either from the procedure previously given for making Class C

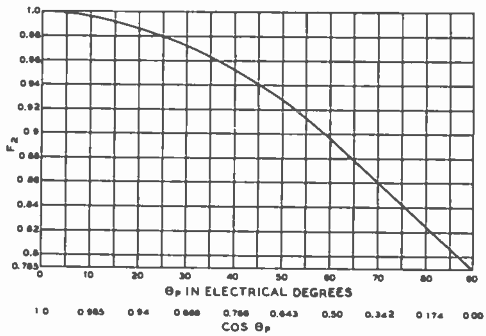


Figure 23.
Relationship between Factor F_2 and the half-angle of plate current flow in an amplifier with sine-wave input and output voltage, operating at a grid-bias voltage greater than cut-off.

amplifier computations or it may be determined from the following expression:

$$\cos \theta_p = -\frac{\mu E_{cc} + E_{bb}}{\mu E_{gm} - E_{pm}}$$

Example of Method It is desired to know the one-half angle of plate current flow and the plate circuit efficiency for an 812 tube operating under the following conditions which have been assumed from inspection of the data and curves given in the RCA Transmitting Tube Handbook HB-3:

1. E_{bb} = 1100 volts
   E_{cc} = −40 volts
   μ = 29
   E_{gm} = 120 volts
   E_{pm} = 1000 volts

2. $F_1 = E_{pm}/E_{bb} = 0.91$

3. $$\cos \theta_p = \frac{-29 \times 40 + 1100}{29 \times 120 - 1000} = \frac{60}{2480} = 0.025$$

4. F_2 = 0.79 (by reference to figure 23)

5. $N_p = F_1 \times F_2 = 0.91 \times 0.79 = 0.72$ (72 per cent efficiency)

F_1 could be called the plate-voltage-swing efficiency factor, and F_2 can be called the operating-angle efficiency factor or the maximum possible efficiency of any stage running with that value of half-angle of plate current flow.

N_p is, of course, only the ratio between power output and power input. If it is desired to determine the power input, exciting power, and grid current of the stage, these can be obtained through use of steps 7, 8, 9, and 10 of the previously given method for power input and output; and knowing that $i_{g\,max}$ is 0.095 ampere the grid circuit conditions can be determined through the use of steps 15, 16, 17, 18 and 19.

5-13 Class B Radio Frequency Power Amplifiers

Radio frequency power amplifiers operating under Class B conditions of grid bias and excitation voltage are used in two general types of applications in transmitters. The first general application is as a buffer amplifier stage where it is desired to obtain a high value of power amplification in a particular stage. A particular tube type operated with a given plate voltage will be capable of somewhat greater output for a certain amount of excitation power when operated as a Class B amplifier than when operated as a Class C amplifier. Calculation of the operating conditions for this type of Class B r-f amplifier can be carried out in a manner similar to that described in the previous paragraphs, except that the grid bias voltage is set on the tube before calculation at the value: $E_{cc} = - E_{bb}/\mu$. Since the grid bias is set at cutoff the one-half angle of plate current flow is 90°; hence cos θ_p is fixed at 0.00. The plate circuit efficiency for a Class B r-f amplifier operated in this manner can be determined in the following manner:

$$N_p = 78.5\left(\frac{E_{pm}}{E_{bb}}\right)$$

The second type of Class B r-f amplifier is the so-called "Class B linear amplifier" which is often used in transmitters for the amplification of a single-sideband signal or a conventional amplitude-modulated wave. Calculation of operating conditions is carried out in a manner similar to that previously described with the following exceptions: The first trial operating point is chosen on the basis of the 100 per cent positive modulation peak of the modulated exciting wave. The plate circuit and grid circuit peak voltages and currents can then be determined and the power input and output calculated. Then, with the exciting voltage reduced to one-half for the no-modulation condition of the exciting wave, and with the same value of load resistance reflected on the tube, the plate input and plate efficiency will drop to approximately one-half the values at the 100 per cent positive modulation peak and the power output of the stage will drop to one-fourth the peak-modulation value. On the negative modulation peak the input, efficiency, and output all drop to zero.

5-14 Special R-F Power Amplifier Circuits

The r-f power amplifier discussions of Sections 5-12 and 5-13 have been based on the assumption that a conventional grounded-cathode or cathode-return type of amplifier was in question. It is possible, however, as in the case of a-f and low-level r-f amplifiers to use circuits in which electrodes other than the cathode are returned to ground insofar as the signal potential is concerned. Both the plate-return or cathode-follower amplifier and the grid-return or grounded-grid amplifier are effective in certain circuit applications as tuned r-f power amplifiers.

Grounded-Grid R-F Power Amplifiers An undesirable aspect of the operation of cathode-return r-f power amplifiers using triode tubes is that such amplifiers must be neutralized. Principles and methods of neutralizing r-f power amplifiers are discussed in Chapter Seven, *Generation of R-F Energy*. As the frequency of operation of an amplifier is increased the stage becomes more and more difficult to neutralize due to inductance in the grid and plate leads of the tubes and in the leads to the neutralizing capacitors. In other words the bandwidth of neutralization decreases as the frequency is increased. In addition the very presence of the neutralizing capacitors adds additional undesirable capacitive loading to the grid and plate tank circuits of the tube or tubes. To look at the problem in another way, an amplifier that may be perfectly neutralized at a frequency of 30 Mc. may be completely out of neutralization at a frequency of 120 Mc. Hence, if there are circuits in both the grid and plate circuits which offer appreciable impedance at this high frequency it is quite possible that the stage may develop a "parasitic oscillation" in the vicinity of 120 Mc.

This condition of restricted-range neutralization of r-f power amplifiers can be greatly alleviated through the use of a cathode-return or grounded-grid r-f stage. The grounded-grid amplifier has the following advantages:

1. The output capacitance of a stage is reduced to approximately one-half the value which would be obtained if the same tube or tubes were operated as a conventional neutralized amplifier.

2. The tendency toward parasitic oscillations in such a stage is greatly reduced since the shielding effect of the control grid between the filament and the plate is effective over a broad range of frequencies.

3. The feedback capacitance within the stage is the plate-to-cathode capacitance which is ordinarily very much less than the grid-to-plate capacitance. Hence neutralization is ordinarily not required. If neutralization is required the neutralizing capacitors are very small in value and are cross connected between plates and cathodes in a push-pull stage, or between the opposite end of a split plate tank and the cathode in a single-ended stage.

The disadvantages of a grounded-grid amplifier are:

1. A large amount of excitation energy is required. However, only the normal amount of energy is lost in the grid circuit of the amplifier tube; all additional energy over this amount is delivered to the load circuit as useful output.

2. The cathode of a grounded-grid amplifier stage is "hot" to r.f. This means that the cathode must be fed through a suitable impedance from the filament supply, or the secondary of the filament transformer must be of the low-capacitance type and adequately insulated for the r-f voltage which will be present.

3. A grounded-grid r-f amplifier cannot be plate modulated 100 per cent unless the output of the exciting stage is modulated also. Approximately 70 per cent modulation of the exciter stage as the final stage is being modulated 100 per cent is recommended. However, the grounded-grid r-f amplifier is quite satisfactory as a Class B linear r-f amplifier for single sideband or conventional amplitude modulated waves or as an amplifier for a straight c-w or FM signal.

Figure 24 shows a simplified representation of a grounded-grid triode r-f power amplifier stage. The relationships between input and output power and the peak fundamental components of electrode voltages and currents are given below the drawing. The calculation of the complete operating conditions for a grounded-grid amplifier stage is somewhat more complex than that for a conventional amplifier because the input circuit of the tube is in series with the output circuit as far as the load is concerned. The primary result of this effect is, as stated before, that considerably more power is required from the driver stage. The normal power gain for a g-g stage is from 3 to 15 depending upon the grid circuit conditions chosen for the output stage. The higher the grid bias and grid swing required on the output stage, the higher will be the requirement from the driver.

Calculation of Operating Conditions of Grounded Grid R-F Amplifiers It is most convenient to determine the operating conditions for a Class B or Class C grounded-grid r-f power amplifier in a two-step process. The first step is to determine the plate-circuit and grid-circuit operating conditions of the tube as though it were to operate as a conventional cathode-return amplifier stage. The second step is then to add in the additional conditions imposed upon the operating conditions by the fact that the stage is to operate as a grounded-grid amplifier.

For the first step in the calculation the procedure given in Section 5-12 is quite satisfactory and will be used in the example to follow. Suppose we take for our example the case of a type 304TL tube operating at 2700 plate volts at a kilowatt input. Following through the procedure previously given:

1. Desired power output—850 watts
   Desired Plate voltage—2700 volts
   Desired plate efficiency—85 per cent ($N_p = 0.85$)

2. $P_{in} = 850/0.85 = 1000$ watts

3. $P_p = 1000 - 850 = 150$ watts
Type 304TL chosen; max. $P_p = 300$ watts, $\mu = 12$.

4. $I_b = 1000/2700 = 0.370$ ampere (370 ma.)

5. Approximate $i_{pmax} = 4.9 \times 0.370 = 1.81$ ampere

6. $e_{pmin} = 140$ volts (from 304TL constant-current curves)

7. $E_{pm} = 2700 - 140 = 2560$ volts

8. $I_{pm}/I_b = 2 \times 0.85 \times 2700/2560 = 1.79$

9. $i_{pmax}/I_b = 4.65$ (from figure 20)

10. $i_{pmax} = 4.65 \times 0.370 = 1.72$ amperes

11. $e_{gmp} = 140$ volts
$i_{gmax} = 0.480$ amperes

12. $\text{Cos}\ \theta_p = 2.32\ (1.79 - 1.57) = 0.51$
$\theta_p = 59°$

13. $$E_{cc} = \frac{1}{1 - 0.51} \times \left[0.51\left(\frac{2560}{12} - 140\right) - \frac{2700}{12}\right]$$
$$= -385 \text{ volts}$$

14. $E_{gm} = 140 - (-385) = 525$ volts

15. $E_{gm}/E_{cc} = -1.36$

16. $i_{gmax}/I_c =$ approx. 8.25 (extrapolated from figure 21)

17. $I_c = 0.480/8.25 = 0.058$ (58 ma. d-c grid current)

18. $P_d = 0.9 \times 525 \times 0.058 = 27.5$ watts

19. $P_g = 27.5 - (-385 \times 0.058) = 5.2$ watts
Max. P_g for 304TL is 50 watts

We can check the operating plate efficiency of the stage by the method described in Section 5-12 as follows:

$F_1 = E_{pm}/E_{bb} = 2560/2700 = 0.95$

F_2 for θ_p of 59° (from figure 23) $= 0.90$

$N_p = F_1 \times F_2 = 0.95 \times 0.90 =$ approx. 0.85 (85 per cent plate efficiency)

Now, to determine the operating conditions as a grounded-grid amplifier we must also know the peak value of the fundamental components of plate current. This is simply equal to $(I_{pm}/I_b)\ I_b$, or:

$I_{pm} = 1.79 \times 0.370 = 0.660$ amperes (from 4 and 8 above)

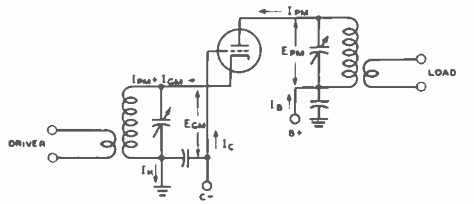
The total average power required of the driver (from figure 24) is equal to $E_{gm}I_{pm}/2$ (since the grid is grounded and the grid swing appears also as cathode swing) plus P_d which is 27.5 watts from 18 above. The total is:

$$\text{Total drive} = \frac{525 \times 0.660}{2} = 172.5 \text{ watts}$$

plus 27.5 watts or 200 watts

Therefore the total power output of the stage is equal to 850 watts (contributed by the 304TL) plus 172.5 watts (contributed by the driver) or 1022.5 watts. The cathode driving impedance of the 304TL (again referring to figure 24) is approximately:

$Z_k = 525/(0.660 + 0.116) =$ approximately 675 ohms.



POWER OUTPUT TO LOAD $= \dfrac{(E_{GM} + E_{PM})\, I_{PM}}{2}$ OR $\dfrac{E_{PM}\, I_{PM}}{2} + \dfrac{E_{GM}\, I_{PM}}{2}$

POWER DELIVERED BY OUTPUT TUBE $= \dfrac{E_{PM}\, I_{PM}}{2}$

POWER FROM DRIVER TO LOAD $= \dfrac{E_{GM}\, I_{PM}}{2}$

TOTAL POWER DELIVERED BY DRIVER $= \dfrac{E_{GM}\,(I_{PM} + I_{GM})}{2}$

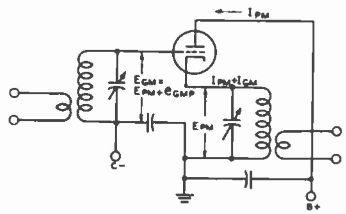
OR $\dfrac{E_{GM}\, I_{PM}}{2} + 0.9\, E_{GM}\, I_C$

POWER ABSORBED BY OUTPUT TUBE GRID AND BIAS SUPPLY

$= \dfrac{E_{GM}\, I_{GM}}{2}$ OR $0.9\, E_{GM}\, I_C$

Z_K = (APPROXIMATELY) $= \dfrac{E_{GM}}{I_{PM} + 1.8\, I_C}$

Figure 24.
GROUNDED-GRID CLASS B OR CLASS C AMPLIFIER.
The equations in the above figure give the relationships between the fundamental components of grid and plate potential and current, and the power input and power output of the stage. An expression for the approximate cathode impedance is given.



POWER OUTPUT TO LOAD $= \frac{E_{PM}(I_{PM}+I_{GM})}{2}$

POWER DELIVERED BY OUTPUT TUBE $= \frac{E_{PM} I_{PM}}{2}$

POWER FROM DRIVER TO LOAD $= \frac{E_{PM} I_{GM}}{2}$

TOTAL POWER FROM DRIVER $= \frac{E_{GM} I_{GM}}{2} = \frac{(E_{PM}+e_{GMP}) I_{GM}}{2}$

$= \text{APPROX } \frac{(E_{PM}+e_{GMP})\, 1.8\, I_C}{2}$

ASSUMING $I_{GM} \cong 1.8\, I_C$

POWER ABSORBED BY OUTPUT TUBE GRID AND BIAS SUPPLY

$= \text{APPROX } 0.9\,(E_{CC}+e_{GMP})\, I_C$

$Z_G = \frac{E_{GM}}{I_{GM}} = \text{APPROX } \frac{(E_{PM}+e_{GMP})}{1.8\, I_C}$

Figure 25.
CATHODE-FOLLOWER R-F POWER AMPLIFIER.
Showing the relationships between the tube potentials and currents and the input and output power of the stage. The approximate grid impedance also is given.

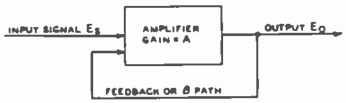
Plote-Return or Cathode-Follower R-F Power Amplifier Circuit diagram, electrode potentials and currents, and operating conditions for a cathode-follower r-f power amplifier are given in figure 25. This circuit can be used, in addition to the grounded-grid circuit just discussed, as an r-f amplifier with a triode tube and no additional neutralization circuit. However, the circuit will oscillate if the impedance from cathode to ground is allowed to become capacitive rather than inductive or resistive with respect to the operating frequency. The circuit is not recommended except for v-h-f or u-h-f work with coaxial lines as tuned circuits since the peak grid swing required on the r-f amplifier stage is approximately equal to the plate voltage on the amplifier tube if high-efficiency operation is desired. This means, of course, that the grid tank must be able to withstand slightly more peak voltage than the plate tank. Such a stage may not be plate modulated unless the driver stage is modulated the same percentage as the final amplifier. However, such a stage may be used as an amplifier or modulated waves (Class B linear) or as a c-w or FM amplifier.

The design of such an amplifier stage is essentially the same as the design of a grounded-grid amplifier stage as far as the first step is concerned. Then, for the second step the operating conditions given in figure 25 are applied to the data obtained in the first step. As an example, take the 304TL stage previously described. The total power required of the driver will be (from figure 25) approximately $(2700 \times 0.058 \times 1.8)/2$ or 141 watts. Of this 141 watts 27.5 watts (as before) will be lost as grid dissipation and bias loss and the balance of 113.5 watts will appear as output. The total output of the stage will then be approximately 963 watts.

Cathode Tank for G-G or C-F Power Amplifier The cathode tank circuit for either a grounded-grid or cathode-follower r-f power amplifier may be a conventional tank circuit if the filament transformer for the stage is of the low-capacitance high-voltage type. Conventional filament transformers, however, will not operate with the high values of r-f voltage present in such a circuit. If a conventional filament transformer is to be used the cathode tank coil may consist of two parallel heavy conductors (to carry the high filament current) by-passed at both the ground end and at the tube socket. The tuning capacitor is then placed between filament and ground. It is possible in certain cases to use two r-f chokes of special design to feed the filament current to the tubes, with a conventional tank circuit between filament and ground. Coaxial lines also may be used to serve both as cathode tank and filament feed to the tubes for v-h-f and u-h-f work.

5-15 Feedback Amplifiers

It is possible to modify the characteristics of an amplifier by feeding back a portion of the output to the input. All components, circuits and tubes included between the point where the feedback is taken off and the point where the feedback energy is inserted are said to be included within the feedback loop. An amplifier containing a feedback loop is said to be a feedback amplifier. One stage or any number of stages may be included within the feedback loop. However, the difficulty of obtaining proper operation of a feedback amplifier in-



$$\text{VOLTAGE AMPLIFICATION WITH FEEDBACK} = \frac{A}{1-A\beta}$$

A = GAIN IN ABSENCE OF FEEDBACK

β = FRACTION OF OUTPUT VOLTAGE FED BACK

β IS NEGATIVE FOR NEGATIVE FEEDBACK

$$\text{FEEDBACK IN DECIBELS} = 20 \log (1-A\beta)$$

$$= 20 \log \frac{\text{MID FREQ GAIN WITHOUT FEEDBACK}}{\text{MID FREQ GAIN WITH FEEDBACK}}$$

$$\text{DISTORTION WITH FEEDBACK} = \frac{\text{DISTORTION WITHOUT FEEDBACK}}{(1-A\beta)}$$

$$R_D = \frac{R_N}{1-A\beta\left(1+\frac{R_N}{R_L}\right)}$$

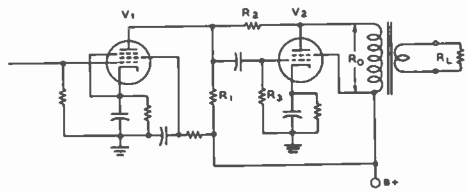
WHERE,

R_D = OUTPUT IMPEDANCE OF AMPLIFIER WITH FEEDBACK

R_N = OUTPUT IMPEDANCE OF AMPLIFIER WITHOUT FEEDBACK

R_L = LOAD IMPEDANCE INTO WHICH AMPLIFIER OPERATES

Figure 26.
FEEDBACK AMPLIFIER RELATIONSHIPS.



$$\text{DB FEEDBACK} = 20 \log \left[\frac{R_2 + R_A(G_{MV2} R_O)}{R_2}\right]$$

$$= 20 \log \left[\frac{R_2 + R_A(\text{VOLTAGE GAIN OF } V_2)}{R_2}\right]$$

$$\text{GAIN OF BOTH STAGES} = \left[G_{MV1}\left(\frac{R_B \times R_A}{R_B + R_A}\right)\right] \times (G_{MV2} R_O)$$

WHERE,

$$R_A = \frac{R_1 \times R_3}{R_1 + R_3}$$

$$R_B = \frac{R_2}{G_{MV2} R_O}$$

R_O = REFLECTED LOAD IMPEDANCE ON V_2

R_2 = FEEDBACK RESISTOR (USUALLY ABOUT 500 K)

$$\text{OUTPUT IMPEDANCE} = \frac{R_N R_2}{\left[R_2 + R_A(G_{MV2} R_O)\right] \times \left(1 + \frac{R_N}{R_O}\right)}$$

R_N = PLATE IMPEDANCE OF V_2

Figure 27.
SHUNT FEEDBACK CIRCUIT FOR PENTODES OR TETRODES.
This circuit requires only the addition of one resistor, R_2, to the normal circuit for such an application. The plate impedance and distortion introduced by the output stage are materially reduced.

creases with the bandwidth of the amplifier, and with the number of stages and circuit elements included within the feedback loop.

The gain and phase shift of any amplifier are functions of frequency. For any amplifier containing a feedback loop to be completely stable the gain of such an amplifier, as measured from the input back to the point where the feedback circuit connects to the input, must be less than one at the frequency where the feedback voltage is in phase with the input voltage of the amplifier. If the gain is equal to or more than one at the frequency where the fed back voltage is in phase with the input the amplifier will oscillate. This fact imposes a limitation upon the amount of feedback which may be employed in an amplifier which is to remain stable. If the reader is desirous of designing amplifiers in which a large amount of feedback is to be employed he is referred to a paper and a book on the subject by H. W. Bode.*

Feedback may be either negative or positive, and the feedback voltage may be proportional either to output voltage or output current. The most commonly used type of feedback with a-f or video amplifiers is negative feedback proportional to output voltage. Figure 26 gives the general operating conditions for feedback amplifiers. Note that the reduction in distortion is proportional to the reduction in gain of the amplifier and that the reduction in the output impedance of the amplifier is somewhat greater than the reduction in the gain by an amount which is a function of the ratio of the output impedance of the amplifier without feedback to the load impedance. The reduction in noise and hum in those stages included within the feedback loop is proportional to the reduction in gain. However, due to the reduction in gain of the output section of the amplifier somewhat increased gain is required of the stages preceding the stages included within the feedback loop. Therefore the noise and hum output of the entire amplifier may or may not be reduced dependent upon the relative contributions of the first part and the latter part of the amplifier to hum and noise. If most of the noise and hum is coming from the stages included within the feedback loop the undesired signals will be reduced in the output from the complete amplifier. It is most frequently true in conventional amplifiers

* H. W. Bode, "Relations Between Attenuation and Phase in Feedback Amplifiers," Bell System Technical Journal, July, 1940, pge. 421.
H. W. Bode, "Network Analysis and Feedback Amplifier Design," D. Van Nostrand Co., Inc., 250 Fourth Ave., New York 3, N. Y.

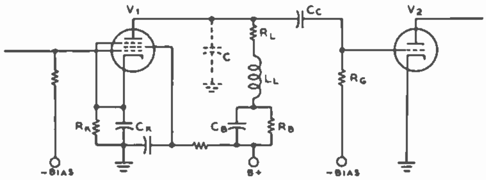
that hum and distortion come from the latter stages, hence these will be reduced by feedback, but thermal agitation and microphonic noise come from the first stage and will not be reduced but may be increased by feedback unless the feedback loop includes the first stage of the amplifier.

Figure 27 illustrates a very simple and effective application of negative voltage feedback to an output pentode or tetrode amplifier stage. The reduction in hum and distortion may amount to 15 to 20 db. The reduction in the effective plate impedance of the stage will be by a factor of 20 to 100 dependent upon the operating conditions. The circuit is commonly used in commercial equipment with tubes such as the 6SJ7 for V_1 and the 6V6 or 6L6 for V_2.

5-16 Video-Frequency Amplifiers

A video-frequency amplifier is one which has been designed to pass frequencies from the lower audio range (lower limit perhaps 50 cycles) to the middle r-f range (upper limit perhaps 4 to 6 megacycles). Such amplifiers, in addition to passing such an extremely wide frequency range, must be capable of amplifying this range with a minimum of amplitude, phase, and frequency distortion. Video amplifiers are commonly used in television, pulse communication, and radar work.

Tubes used in video amplifiers must have a high ratio of G_m to capacitance if a usable gain per stage is to be obtained. Commonly available tubes which have been designed for or are suitable for use in video amplifiers are: 6AU6, 6AG5, 6AK5, 6CB6, 6AC7, 6AG7, and 6K6-GT. Since, at the upper frequency limits of a video amplifier the input and output shunting capacitances of the amplifier tubes have rather low values of reactance, low values of coupling resistance along with peaking coils or other special interstage coupling impedances are usually used to flatten out the gain/frequency and hence the phase/frequency charac-



MID-FREQUENCY GAIN = $G_{M\,V1}\,R_L$

HIGH-FREQUENCY GAIN = $G_{M\,V1}\,Z_{\text{COUPLING NETWORK}}$

$C = C_{\text{OUT}\,V1} + C_{\text{IN}\,V2} + C_{\text{DISTRIBUTED}}$

FOR COMPROMISE HIGH FREQUENCY EQUALIZATION

$X_{LL} = 0.5\,X_C$ AT f_C

$R_L = X_C$ AT f_C

WHERE f_C = CUTOFF FREQUENCY OF AMPLIFIER

L_L = PEAKING INDUCTOR

FOR COMPROMISE LOW FREQUENCY EQUALIZATION

$R_B = R_K\,(G_{M\,V1}\,R_L)$

$R_B\,C_B = R_K\,C_K$

C_K = 25 TO 50 μFD IN PARALLEL WITH .001 MICA

C_B = CAPACITANCE FROM ABOVE WITH .001 MICA IN PARALLEL

Figure 28.
SIMPLE COMPENSATED VIDEO AMPLIFIER CIRCUIT.
Resistor R_L in conjunction with coil L_L serves to flatten the high-frequency response of the stage, while C_B and R_B serve to equalize the low-frequency response of this simple video amplifier stage.

teristic of the amplifier. Recommended operating conditions along with expressions for calculation of gain and circuit values are given in figure 28. Only a simple two-terminal interstage coupling network is shown in this figure.

The performance and gain-per-stage of a video amplifier can be improved by the use of increasingly complex two-terminal interstage coupling networks or through the use of four-terminal coupling networks or filters between successive stages. The reader is referred to Terman's *"Radio Engineer's Handbook"* for design data on such interstage coupling networks. A cathode follower is usually employed where it is desired to feed a video-frequency amplifier into a low-impedance load such as a coaxial cable. Cathode followers are discussed in Section 5-9.

CHAPTER SIX

Radio Receiver Fundamentals

A conventional reproducing device such as a loudspeaker or a pair of earphones is incapable of receiving directly the intelligence carried by the "carrier" wave of a radio transmitting station. It is necessary that an additional device, called a *radio receiver*, be placed between the receiving antenna and the loudspeaker or headphones.

Radio receivers vary widely in their complexity and basic design, depending upon the intended application and upon economic factors. A simple radio receiver for reception of radiotelephone signals can consist of an earphone, a silicon or germanium crystal as a carrier rectifier or *demodulator*, and a length of wire as an antenna. However, such a receiver is highly insensitive, and offers no significant discrimination between two signals in the same portion of the spectrum.

On the other hand, a dual-diversity receiver designed for single-sideband reception and employing double or triple detection might occupy several relay racks and would cost many thousands of dollars. However, conventional communications receivers are intermediate in complexity and performance between the two extremes. This chapter is devoted to the principles underlying the operation of such conventional communications receivers.

6-1 Detection or Demodulation

A detector or demodulator is a device for removing the modulation (demodulating) or detecting the intelligence carried by an incoming radio wave.

Radiotelephony Demodulation Figure 1 illustrates an elementary form of radiotelephony receiver employing a diode detector. Energy from a passing radio wave will induce a voltage in the antenna and cause a radio-frequency current to flow from antenna to ground through coil L_1. The alternating magnetic field set up around L_1 links with the turns of L_2 and causes an r-f current to flow through the parallel-tuned circuit, L_2-C_1. When variable capacitor C_1 is adjusted so that the tuned circuit is resonant at the frequency of the applied signal, the r-f voltage is maximum, as explained in Chapter 3. This r-f voltage is applied to the diode detector where it is rectified into a varying direct current and passed through the earphones. The variations in this current correspond to the voice modulation placed on the signal at the transmitter. As the earphone diaphragms vibrate back and forth in accordance with the pulsating current

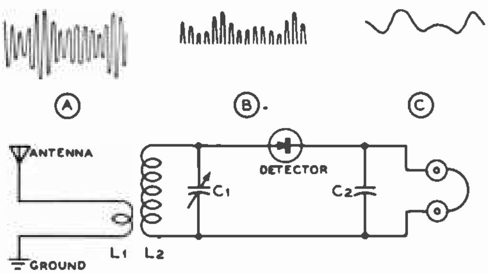


Figure 1.

ELEMENTARY FORM OF RECEIVER.

This is the basis of the "crystal set" type of receiver, although a vacuum diode may be used in place of the crystal diode. The tank circuit L_2-C_1 is tuned to the frequency it is desired to receive. The by-pass capacitor across the phones should have a low reactance to the carrier frequency being received, but a high reactance to the modulation on the received signal.

they audibly reproduce the modulation which was placed upon the carrier wave.

The operation of the detector circuit is shown graphically above the detector circuit in figure 1. The modulated carrier is shown at A, as it is applied to the antenna. B represents the same carrier, increased in amplitude, as it appears across the tuned circuit. In C the varying d-c output from the detector is seen.

Radiotelegraphy Reception Since a c-w telegraphy signal consists of an unmodulated carrier which is interrupted to form dots and dashes, it is apparent that such a signal would not be made audible by detection alone. While the keying is a form of modulation, it is composed of such low frequency components that the keying envelope itself is below the audible range for hand keying speeds. Some means must be provided whereby an audible tone is heard while the unmodulated carrier is being received, the tone stopping immediately when the carrier is interrupted.

The most simple means of accomplishing this is to feed a locally generated carrier of a slightly different frequency into the same detector, so that the incoming signal will *mix* with it to form an audible *beat note*. The difference frequency, or *heterodyne* as the beat note is known, will of course stop and start in accordance with the incoming c-w radiotelegraph signal, because the audible heterodyne can exist only when both the incoming and the locally generated carriers are present.

The Autodyne Detector The local signal which is used to beat with the desired c-w signal in the detector may be supplied by a separate low-power oscillator in the receiver itself, or the detector may be made to self-oscillate, and thus serve the dual purpose of detector and oscillator. A detector which self-oscillates to provide a beat note is known as an *autodyne* detector, and the process of obtaining feedback between the detector plate and grid is called *regeneration.*

An autodyne detector is most sensitive when it is barely oscillating, and for this reason a regeneration control is always included in the circuit to adjust the feedback to the proper amount. The regeneration control may be either a variable capacitor or a variable resistor, as shown in figure 2.

With the detector regenerative but not oscillating, it is also quite sensitive. When the circuit is adjusted to operate in this manner, modulated signals may be received with considerably greater strength than with a non-regenerative detector.

6-2 Superregenerative Receivers

At ultra-high frequencies, when it is desired to keep weight and cost at a minimum, a special form of the regenerative receiver known as the *superregenerator* is often used for radiotelephony reception. The superregenerator is essentially a regenerative receiver with a means provided to throw the detector rapidly in and out of oscillation. The frequency at which the detector is made to go in and out of oscillation varies with the frequency to be received, but is usually between 20,000 and 500,000 times a second. This superregenerative action considerably increases the sensitivity of the oscillating detector so that the usual "background hiss" is greatly amplified when no signal is being received. This hiss diminishes in proportion to the strength of the received signal, loud signals eliminating the hiss entirely.

Quench Methods There are two systems in common use for causing the detector to break in and out of oscillation rapidly. In one, a separate *interruption-fre-*

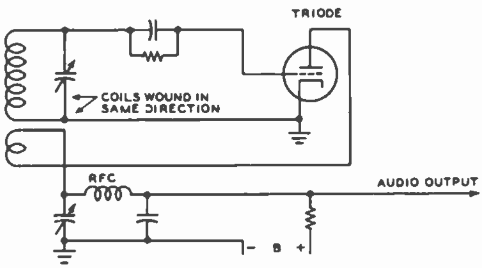
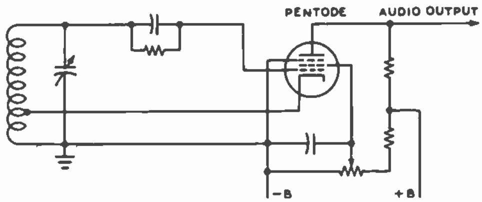
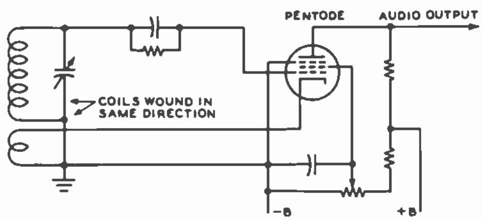


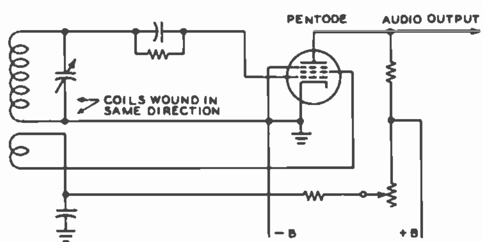
PLATE-TICKLER REGENERATION WITH "THROTTLE" CONDENSER REGENERATION CONTROL.



CATHODE-TAP REGENERATION WITH SCREEN VOLTAGE REGENERATION CONTROL.



CATHODE-COIL REGENERATION WITH SCREEN-VOLTAGE REGENERATION CONTROL.



SCREEN-GRID-TICKLER REGENERATION WITH SCREEN VOLTAGE REGENERATION CONTROL.

Figure 2.
REGENERATIVE DETECTOR CIRCUITS.
Regenerative detectors are seldom used at the present time due to their poor selectivity, but they do illustrate the simplest type of receiver which may be used either for radiophone or radiotelegraph reception.

quency oscillator is arranged so as to vary the voltage rapidly on one of the detector tube elements (usually the plate, sometimes the screen) at the high rate necessary. The interruption-frequency oscillator commonly uses a conventional tickler-feedback circuit with coils appropriate for its operating frequency.

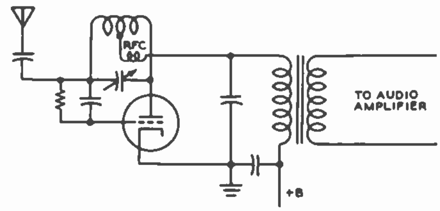


Figure 3.
SUPERREGENERATIVE DETECTOR CIRCUIT.
A self-quenched superregenerative detector such as illustrated above is capable of giving good sensitivity in the v-h-f range. However, the circuit has the disadvantage that its selectivity is relatively poor. Also, such a circuit should be preceded by an r-f stage to suppress the radiation of a signal by the oscillating detector.

The second, and simplest, type of superregenerative detector circuit is arranged so as to produce its own interruption frequency oscillation, without the aid of a separate tube. The detector tube damps (or "quenches") itself out of signal-frequency oscillation at a high rate by virtue of the use of a high value of grid leak and proper size plate-blocking and grid capacitors, in conjunction with an excess of feedback. In this type of "self-quenched" detector, the grid leak is quite often returned to the positive side of the power supply (through the coil) rather than to the cathode. A representative self-quenched superregenerative detector circuit is shown in figure 3.

Except where it is impossible to secure sufficient regenerative feedback to permit superregeneration, the self-quenching circuit is to be preferred; it is simpler, is self-adjusting as regards quenching amplitude, and can have good quenching wave form. To obtain as good results with a separately quenched superregenerator, very careful design is required. However, separately quenched circuits are useful when it is possible to make a certain tube oscillate on a very high frequency but is impossible to obtain enough regeneration for self-quenching action.

The optimum quenching frequency is a function of the signal frequency. As the operating frequency goes up, so does the optimum quenching frequency. When the quench frequency is too low, maximum sensitivity is not obtained. When it is too high, both sensitivity and selectivity suffer. In fact, the optimum quench frequency for an operating frequency below 15 Mc. is in the audible range. This makes the superregenerator impracticable for use on the lower frequencies.

The high background noise or hiss which is heard on a properly designed superregenerator when no signal is being received is not the quench frequency component "leaking through"; it is tube and tuned circuit fluctuation noise, indicating that the receiver is extremely sensitive.

A moderately strong signal will cause the background noise to disappear completely, because the superregenerator has an inherent and instantaneous automatic volume control characteristic. This same a-v-c characteristic makes the receiver comparatively insensitive to impulse noise such as ignition pulses—a highly desirable feature. This characteristic also results in appreciable distortion of a received radiotelephone signal, but not enough to affect the intelligibility.

The selectivity of a superregenerator is rather poor as compared to a superheterodyne, but is surprisingly good for so simple a receiver when figured on a percentage basis rather than absolute kc. bandwidth.

FM Reception A superregenerative receiver will receive frequency modulated signals with results comparing favorably with amplitude modulation if the frequency swing of the FM transmitter is sufficiently high. For such reception, the receiver is detuned slightly to either side of resonance.

Superregenerative receivers radiate a strong, broad, and rough signal. For this reason it is necessary in most applications to employ a radio frequency amplifier stage ahead of the detector, with thorough shielding throughout the receiver.

6-3 Superheterodyne Receivers

Because of its superiority and nearly universal use in all fields of radio reception, the theory of operation of the superheterodyne should be familiar to every radio student and experimenter. The following discussion concerns superheterodynes for amplitude-modulation reception. It is, however, applicable in part to receivers for frequency modulation. The points of difference between the two types of receivers, together with circuits required for FM reception, will be found in Chapter Nine.

Principle of Operation In the superheterodyne, the incoming signal is applied to a *mixer* consisting of a non-linear impedance such as a vacuum tube or a diode. The signal is mixed with a steady signal generated locally in an oscillator stage, with the result that a signal bearing all the modulation applied to the original signal *but of a frequency equal to the difference between the local oscillator and incoming signal frequencies* appears in the mixer output circuit. The output from the mixer stage is fed into a fixed-tuned *intermediate-frequency amplifier,* where it is amplified and detected in the usual manner, and passed on to the audio amplifier. Figure 4 shows a block diagram of the fundamental superheterodyne arrangement. The basic components are shown in heavy lines, the simplest superheterodyne consisting simply of these three units. However, a good communications receiver will comprise all of the elements shown, both heavy and dotted blocks.

Superheterodyne Advantages The advantages of superheterodyne reception are directly attributable to the use of the fixed-tuned intermediate-frequency (i-f) amplifier. Since all signals are converted to the intermediate frequency, this section of the receiver may be designed for optimum se-

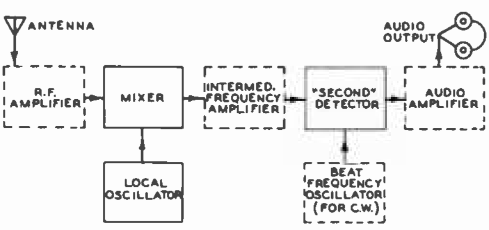


Figure 4.
ESSENTIAL UNITS OF A SUPERHETERODYNE RECEIVER.
The basic portions of the receiver are shown in solid blocks. Practicable receivers employ the dotted blocks in addition, and in addition usually include such additional circuits as a noise limiter, an a-v-c circuit, and a crystal filter in the i-f amplifier.

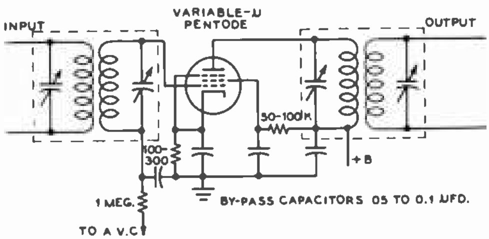


Figure 5.
TYPICAL I-F AMPLIFIER STAGE.

lectivity and high amplification. High amplification is easily obtained in the intermediate-frequency amplifier, since it operates at a relatively low frequency, where conventional pentode-type tubes give adequate voltage gain. A typical i-f amplifier is shown in figure 5.

From the diagram it may be seen that both the grid and plate circuits are tuned. The tuned circuits used for coupling between i-f stages are known as *i-f transformers.* These will be more fully discussed later in this chapter.

Choice of Intermediate Frequency The choice of a frequency for the i-f amplifier involves several considerations. One of these considerations concerns selectivity; the lower the intermediate frequency the greater the obtainable selectivity. On the other hand, a rather high intermediate frequency is desirable from the standpoint of *image* elimination, and also for the reception of signals from television and FM transmitters and modulated self-controlled oscillators, all of which occupy a rather wide band of frequencies, making a broad selectivity characteristic desirable. Images are a peculiarity common to all superheterodyne receivers, and for this reason they are given a detailed discussion later in this chapter.

While intermediate frequencies as low as 50 kc. are used where extreme selectivity is a requirement, and frequencies of 60 Mc. and above are used in some specialized forms of receivers, most present-day communications superheterodynes use intermediate frequencies around either 455 kc. or 1600 kc.

Home-type broadcast receivers almost always use an i-f in the vicinity of 455 kc., while auto receivers usually use a frequency of about 262 kc. The standard frequency for the i-f channel of FM receivers is 10.7 Mc. Television receivers usually use an i-f which covers the band between about 21.5 and 27 Mc., although a new band between 41 and 46 Mc. is coming into more common usage.

Arithmetical Selectivity Aside from allowing the use of fixed-tuned band-pass amplifier stages, the superheterodyne has an overwhelming advantage over the tuned radio frequency (t-r-f) type of receiver because of what is commonly known as *arithmetical selectivity.*

This can best be illustrated by considering two receivers, one of the t-r-f type and one of the superheterodyne type, both attempting to receive a desired signal at 10,000 kc. and eliminate a strong interfering signal at 10,010 kc. In the t-r-f receiver, separating these two signals in the tuning circuits is practically impossible, since they differ in frequency by only 0.1 per cent. However, in a superheterodyne with an intermediate frequency of, for example, 1000 kc., the desired signal will be converted to a frequency of 1000 kc. and the interfering signal will be converted to a frequency of 1010 kc., both signals appearing at the input of the i-f amplifier. In this case, the two signals may be separated much more readily, since they differ by 1 per cent, or 10 times as much as in the first case.

The Converter Stage The converter stage, or the *mixer,* of a superheterodyne receiver can be either one of two types: (1) it may use a single envelope *converter* tube, such as a 6K8, 6SA7, or 6BE6, or (2) it may use two tubes, or two sets of elements in the same envelope, in an oscillator-mixer arrangement. Figure 6 shows a group of circuits of both types to illustrate present practice with regard to types of converter stages.

Converter tube combinations such as shown in figures 6A and 6B are relatively simple and inexpensive, and they do an adequate job for most applications. With a converter tube such as the 6SB7-Y or the 6BA7 quite satisfactory performance may be obtained for the reception of relatively strong signals (as for example FM broadcast reception) up to frequencies in excess of 100 Mc. However, the equivalent input noise resistance of such tubes is of the order of 200,000 ohms, which is a rather high value indeed. So such tubes are

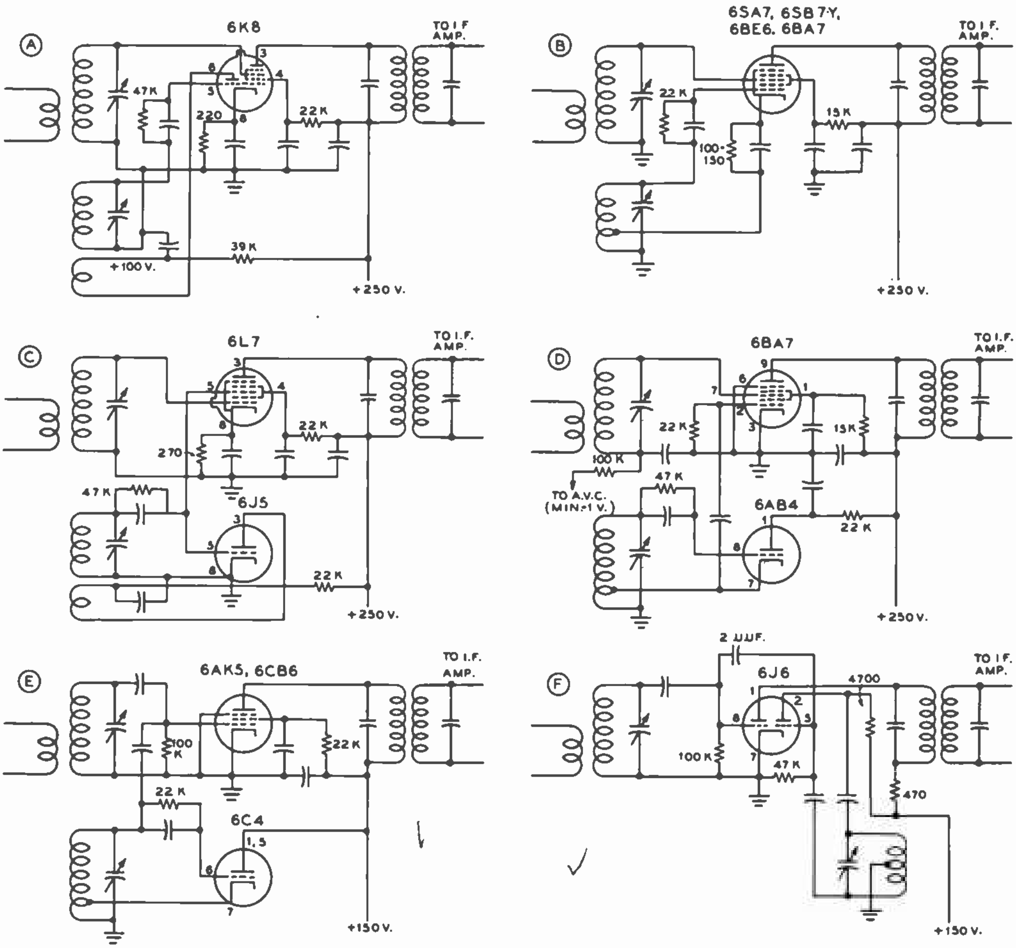


Figure 6.
TYPICAL FREQUENCY-CONVERTER (MIXER) STAGES.
The relative advantages of the different circuits are discussed in the text.

not suited for operation without an r-f stage in the high-frequency range if weak-signal reception is desired.

The 6L7 mixer circuit shown in figure 6C, and the 6BA7 circuit of figure 6D, also are characterized by an equivalent input noise resistance of several hundred thousand ohms, so that these also must be preceded by one or more r-f stages with a fairly high gain per stage if a low noise factor is desired of the complete receiver.

However, the circuit arrangements shown at figures 6E and 6F *are* capable of low-noise operation within themselves, so that these circuits may be fed directly from the antenna without an r-f stage and still provide a good noise factor to the complete receiver. Note that both these circuits use *control-grid injection* of both the incoming signal and the local-oscillator voltage. Hence, paradoxically, circuits such as these should be preceded by an r-f stage if local-oscillator radiation is to be held to any reasonable value of field intensity.

Diode Mixers As the frequency of operation of a superheterodyne receiver is increased above a few hundred megacycles the signal-to-noise ratio appearing in the plate circuit of the mixer tube when triodes or pentodes are employed drops to a prohibitively low value. At frequencies above the upper-frequency limit for conventional mixer stages, mixers of the diode type are most commonly employed. The diode may be either a vacuum-tube heater diode of a special u-h-f design such as the 9005, or it may be a crystal diode of the general type of the 1N21 through 1N28 series. Circuits and operating principles of diode mixers are discussed in more detail in Section 6-9 of this chapter.

6-4 Mixer Noise and Images

The effects of *mixer noise* and *images* are troubles common to all superheterodynes. Since both these effects can largely be obviated by the same remedy, they will be considered together.

Mixer Noise Mixer noise of the shot-effect type, which is evidenced by a hiss in the audio output of the receiver, is caused by small irregularities in the plate current in the mixer stage and will mask weak signals. Noise of an identical nature is generated in an amplifier stage, but due to the fact that the conductance in the mixer stage is considerably lower than in an amplifier stage using the same tube, the proportion of inherent noise present in a mixer usually is considerably greater than in an amplifier stage using a comparable tube.

Although this noise cannot be eliminated, its effects can be greatly minimized by placing sufficient signal-frequency amplification having a high signal-to-noise ratio ahead of the mixer. This remedy causes the signal output from the mixer to be large in proportion to the noise generated in the mixer stage. Increasing the gain *after* the mixer will be of no advantage in eliminating mixer noise difficulties; greater selectivity after the mixer will help to a certain extent, but cannot be carried too far, since this type of selectivity decreases the i-f band-pass and if carried too far will not pass the sidebands that are an essential part of a voice-modulated signal.

Triode Mixers A triode having a high transconductance is the *quietest* mixer tube, exhibiting somewhat less gain but a better signal-to-noise ratio than a comparable multi-grid mixer tube. However, below 30 Mc. it is possible to construct a receiver that will get down to the atmospheric noise level without resorting to a triode mixer. The additional difficulties experienced in avoiding "pulling," undesirable feedback, etc., when using a triode with control-grid injection tend to make multi-grid tubes the popular choice for this application on the lower frequencies.

On very high frequencies, where set noise rather than atmospheric noise limits the weak signal response, triode mixers are more widely used. A 6J6 miniature twin triode with grids in push-pull and plates in parallel makes an excellent mixer up to about 600 Mc.

Injection Voltage The amplitude of the injection voltage will affect the conversion transconductance of the mixer, and therefore should be made optimum if maximum signal-to-noise ratio is desired. If fixed bias is employed on the injection grid, the optimum injection voltage is quite critical. If cathode bias is used, the optimum voltage is not so critical; and if grid leak bias is employed, the optimum injection voltage is not at all critical just so it is adequate. Typical optimum injection voltages will run from 1 to 10 volts for control grid injection, and 45 volts or so for screen or suppressor grid injection.

Images There always are *two* signal frequencies which will combine with a given frequency to produce the same difference frequency. For example: assume a superheterodyne with its oscillator operating on a higher frequency than the signal, which is common practice in present superheterodynes, tuned to receive a signal at 14,100 kc. Assuming an i-f amplifier frequency of 450 kc., the mixer input circuit will be tuned to 14,100 kc., and the oscillator to 14,100 plus 450, or 14,550 kc. Now, a *strong* signal at the oscillator frequency plus the intermediate frequency (14,550 plus 450, or 15,000 kc.) will also give a difference frequency of 450 kc. in the mixer output and will be heard also. Note that the image is always *twice* the intermediate frequency away from the desired signal. Images cause "repeat points" on the tuning dial.

The only way that the image could be eliminated in this particular case would be

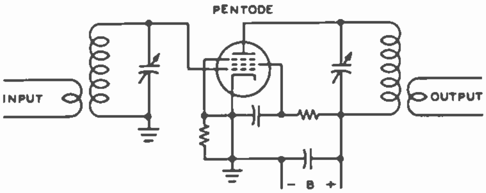


Figure 7.
TYPICAL PENTODE R-F AMPLIFIER STAGE.

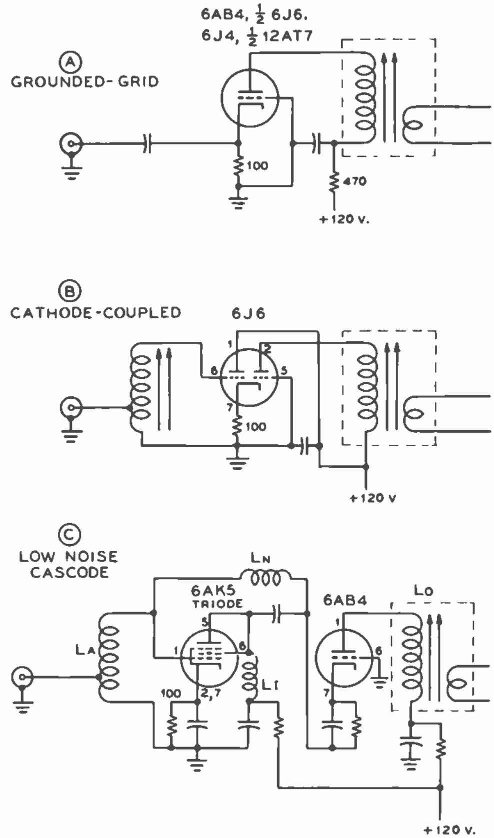


Figure 8.
TYPICAL TRIODE V-H-F R-F AMPLIFIER STAGES.
Triode r-f stages contribute the least amount of noise output for a given signal level, hence their frequent use in the v-h-f range.

to make the selectivity of the mixer input circuit, and any circuits preceding it, great enough so that the 15,000-kc. signal never reaches the mixer grid in sufficient amplitude to produce interference.

For any particular intermediate frequency, image interference troubles become increasingly greater as the frequency to which the signal-frequency portion of the receiver is tuned is increased. This is due to the fact that the percentage difference between the desired frequency and the image frequency decreases as the receiver is tuned to a higher frequency. The ratio of strength between a signal at the image frequency and a signal at the frequency to which the receiver is tuned producing equal output is known as the *image ratio.* The higher this ratio, the better the receiver in regard to image-interference troubles.

With but a single tuned circuit between the mixer grid and the antenna, and with 400-500 kc. i-f amplifiers, image ratios of 60 db and over are easily obtainable up to frequencies around 2000 kc. Above this frequency, greater selectivity in the mixer grid circuit through the use of additional tuned circuits between the mixer and the antenna is necessary if a good image ratio is to be maintained.

R-F Stages Since the necessary tuned circuits between the mixer and the antenna can be combined with tubes to form r-f amplifier stages, the reduction of the effects of mixer noise and the increasing of the image ratio can be accomplished in a single section of the receiver. When incorporated in the receiver, this section is known simply as an *r-f amplifier;* when it is a separate unit with a separate tuning control it is often known as a *preselector.* Either one or two stages are commonly used in the preselector or r-f amplifier. Some preselectors use regeneration to obtain still greater amplification and selectivity. An r-f amplifier or preselector embodying more than two stages rarely ever is employed since two stages will ordinarily give adequate gain to override mixer noise.

R-F Stages in the V-H-F Range Generally speaking, atmospheric noise in the frequency range above 30 Mc. is quite low—so low, in fact, that the noise generated within the receiver itself is greater than the noise received on the antenna. Hence it is of the greatest importance that internally generated noise be held to a minimum in a receiver. At frequencies much above 300 Mc. there is not too much that can be done at the present state of the art in the direction of

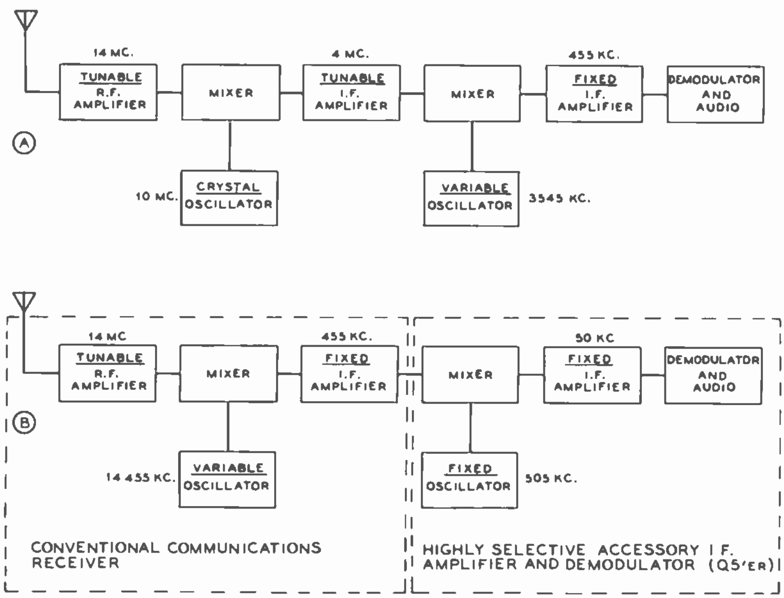


Figure 9.
TYPICAL DOUBLE-CONVERSION SUPERHETERODYNE RECEIVERS.
Illustrated at (A) is the basic circuit of a commercial double-conversion superheterodyne receiver. At (B) is illustrated the application of an accessory sharp i-f channel for obtaining improved selectivity from a conventional communications receiver through the use of the double-conversion principle.

reducing receiver noise below that generated in the converter stage. But in the v-h-f range, between 30 and 300 Mc., the receiver noise factor in a well designed unit is determined by the characteristics of the first r-f stage. These considerations are discussed in some detail in Section 5-10 of Chapter Five.

The usual v-h-f receiver, whether for communications or for FM or TV reception, uses a miniature pentode for the first r-f amplifier stage. The 6AK5 is the best of presently available types, with the 6CB6 and the 6AG5 closely approaching the 6AK5 in performance. But when gain in the first r-f stage is not so important, and the best noise factor must be obtained, the first r-f stage usually uses a triode.

Shown in figure 8 are three commonly used types of triode r-f stages for use in the v-h-f range. The circuit at (A) uses few components and gives a moderate amount of gain with very low noise. It is most satisfactory when the first r-f stage is to be fed directly from a low-impedance coaxial transmission line. Figure 8 (B) gives somewhat more gain than (A), but requires an input matching circuit. The effective gain of this circuit is somewhat reduced when it is being used to amplify a broad band of frequencies since the effective G_m of the cathode-coupled dual tube is somewhat less than half the G_m of either of the two tubes taken alone. Figure 8 (C) shows the recently popularized *cascode* circuit which, although somewhat complex, gives a value of stage gain about equal to that of a pentode while still retaining the low equivalent noise resistance of a triode stage.

Double Conversion As previously mentioned, the use of a higher intermediate frequency will also improve the image ratio, at the expense of i-f selectivity, by placing the desired signal and the image farther apart. To give both good image ratio at the higher frequencies and good selectivity in the i-f amplifier, a system known as *double conversion* is sometimes employed. In this system, the incoming signal is first converted to a rather high intermediate frequency, and then amplified and again converted, this time to a much lower frequency. The first intermediate frequency supplies the necessary wide separation between the image and the desired signal, while the second one supplies the bulk of the i-f selectivity.

The double-conversion system, as illustrated in figure 9, is receiving two general types of application at the present time. The first application is for the purpose of attaining extremely good stability in a communications receiver through the use of crystal control of the first oscillator. In such an arrangement, as used in several types of Collins receivers and in connection with the crystal-controlled converters described in Chapter 20, the first oscillator is crystal controlled and is followed by a tunable i-f amplifier which then is followed by a mixer stage and a fixed-tuned i-f amplifier on a much lower frequency. Through such a circuit arrangement the stability of the complete receiver is equal to the stability of the oscillator which feeds the second mixer, while the selectivity is determined by the bandwidth of the second, fixed i-f amplifier.

The second common application of the double-conversion principle is for the purpose of obtaining a very high degree of selectivity in the complete communications receiver. In this type of application, as illustrated in figure 9 (B), a conventional communications receiver is modified in such a manner that its normal i-f amplifier (which usually is in the 450 to 915 kc. range) instead of being fed to a demodulator and then to the audio system, is alternatively fed to a fixed-tune mixer stage and then into a much lower intermediate frequency amplifier before the signal is demodulated and fed to the audio system. The accessory i-f amplifier system (sometimes called a Q5'er) normally is operated on a frequency of 175 kc., 85 kc., or 50 kc.

6-5 Signal-Frequency Tuned Circuits

The signal-frequency tuned circuits in high-frequency superheterodynes and tuned radio frequency types of receivers consist of coils of either the solenoid or universal-wound types shunted by variable capacitors. It is in these tuned circuits that the causes of success or failure of a receiver often lie. The universal-wound type coils usually are used at frequencies below 2000 kc.; above this frequency the single-layer solenoid type of coil is more satisfactory.

Impedance and Q The two factors of greatest significance in determining the gain-per-stage and selectivity, respectively, of a tuned amplifier are tuned-circuit impedance and tuned-circuit Q. As explained in Chapter 3, Q is the ratio of reactance to resistance in the circuit. Since the resistance of modern capacitors is low at ordinary frequencies, the resistance usually can be considered to be concentrated in the coil. The resistance to be considered in making Q determinations is the r-f resistance, not the d-c resistance of the wire in the coil. The latter ordinarily is low enough that it may be neglected. The increase in r-f resistance over d-c resistance primarily is due to skin effect and is influenced by such factors as wire size and type, and the proximity of metallic objects or poor insulators, such as coil forms with high losses.

It may be seen from the curves shown in Chapter 3 that higher values of Q lead to better selectivity and increased r-f voltage across the tuned circuit. The increase in voltage is due to an increase in the circuit impedance with the higher values of Q.

Frequently it is possible to secure an increase in impedance in a resonant circuit, and consequently an increase in gain from an amplifier stage, by increasing the reactance through the use of larger coils and smaller tuning capacitors (higher L/C ratio).

Input Resistance Another factor which influences the operation of

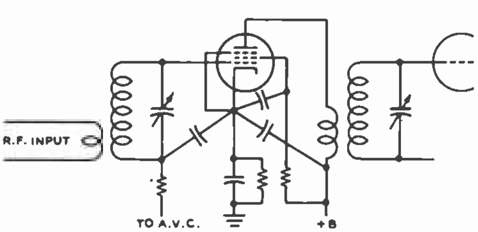


Figure 10. ILLUSTRATING "COMMON POINT" BY-PASSING.

To reduce the detrimental effects of cathode circuit inductance in v-h-f stages, all by-pass capacitors should be returned to the cathode terminal at the socket. Tubes with two cathode leads can give improved performance if the grid return is made to one cathode terminal while the plate and screen by-pass returns are made to the cathode terminal which is connected to the suppressor within the tube.

tuned circuits is the input resistance of the tubes placed across these circuits. At broadcast frequencies, the input resistance of most conventional r-f amplifier tubes is high enough so that it is not bothersome. But as the frequency is increased, the input resistance becomes lower and lower, until it ultimately reaches a value so low that no amplification can be obtained from the r-f stage.

The two contributing factors to the decrease in input resistance with increasing frequency are the transit time required by an electron traveling between the cathode and grid, and the inductance of the cathode lead common to both the plate and grid circuits. As the frequency becomes higher, the transit time can become an appreciable portion of the time required by an r-f cycle of the signal voltage, and current will actually flow into the grid. The result of this effect is similar to that which would be obtained by placing a resistance between the tube's grid and cathode.

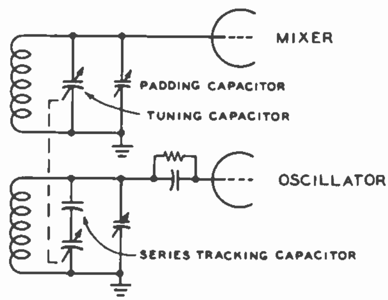


Figure 11.
SERIES TRACKING EMPLOYED IN THE H-F OSCILLATOR OF A SUPERHETERODYNE.
The series tracking capacitor permits the use of identical gangs in a ganged capacitor, since the tracking capacitor slows down the rate of frequency change in the oscillator so that a constant difference in frequency between the oscillator and the r-f stage (equal to the i-f amplifier frequency) may be maintained.

Superheterodyne Tracking Because the oscillator in a superheterodyne operates "offset" from the other front end circuits, it is necessary to make special provisions to allow the oscillator to track when similar tuning capacitor sections are ganged. The usual method of obtaining good tracking is to operate the oscillator on the high-frequency side of the mixer and use a series "tracking capacitor" to slow down the tuning rate of the oscillator. The oscillator tuning rate must be slower because it covers a smaller range than does the mixer when both are expressed as a percentage of frequency. At frequencies above 7000 kc. and with ordinary intermediate frequencies, the difference in percentage between the two tuning ranges is so small that it may be disregarded in receivers designed to cover only a small range, such as an amateur band.

A mixer and oscillator tuning arrangement in which a series tracking capacitor is provided is shown in figure 11. The value of the tracking capacitor varies considerably with different intermediate frequencies and tuning ranges, capacitances as low as .0001 μfd. being used at the lower tuning-range frequencies, and values up to .01μfd. being used at the higher frequencies.

Superheterodyne receivers designed to cover only a single frequency range, such as the standard broadcast band, sometimes obtain tracking between the oscillator and the r-f circuits by cutting the variable plates of the oscillator tuning section to a different shape from those used to tune the r-f stages.

Frequency Range Selection The frequency to which a receiver responds may be varied by changing the size of either the coils or the capacitors in the tuning circuits, or both. In short-wave receivers a combination of both methods is usually employed, the coils being changed from one band to another, and variable capacitors being used to tune the receiver across each band. In practical receivers, coils may be changed by one of two methods: a switch, controllable from the panel, may be used to switch coils of different sizes into the tuning circuits or, alternatively, coils of different sizes may be plugged manually into the receiver, the connection into the tuning circuits being made by suitable plugs on the coils. Where there are several "plug-in" coils for each band, they are sometimes arranged to a single mounting strip, allowing them all to be plugged in simultaneously.

Bandspread Tuning In receivers using large tuning capacitors to cover the short-wave spectrum with a minimum of coils, tuning is likely to be quite difficult, owing to the large frequency range covered by a small rotation of the variable capacitors. To alleviate this condition, some method of slowing down the tuning rate, or *bandspreading,* must be used.

Quantitatively, bandspread is usually designated as being inversely proportional to the range covered. Thus, a *large* amount of bandspread indicates that a *small* frequency range is covered by the bandspread control. Conversely, a *small* amount of bandspread is taken to mean that a *large* frequency range is covered by the bandspread dial.

Types of Bandspread Bandspreading systems are of two general types: electrical and mechanical. Mechanical systems are exemplified by high-ratio dials in which the tuning capacitors rotate much more slowly than the dial knob. In this system there is often a separate scale or pointer either connected or geared to the dial knob to facilitate accurate dial readings. However, there is a practical limit to the amount of mechanical bandspread which can be obtained in a dial and capacitor before the speed-reduction unit and capacitor bearings become prohibitively expensive. Hence, most receivers employ a combination of electrical and mechanical bandspread. In such a system, a moderate reduction in the tuning rate is obtained in the dial, and the rest of the reduction obtained by *electrical bandspreading.*

Stray Circuit Capacitance In this book and in other radio literature, mention is sometimes made of "stray" or *circuit capacitance.* This capacitance is in the usual sense defined as the capacitance remaining across a coil when all the tuning, bandspread, and padding capacitors across the circuit are at their minimum capacitance setting.

Circuit capacitance can be attributed to two general sources. One source is that due to the input and output capacitance of the tube when its cathode is heated. The input capacitance varies somewhat from the static or "cold" value when the tube is in actual operation. Such factors as plate load impedance, grid bias, and frequency will cause a change in input capacitance. However, in all except the extremely high-transconductance tubes, the published measured input capacitance is reasonably close to the effective value when the tube is used within its recommended frequency

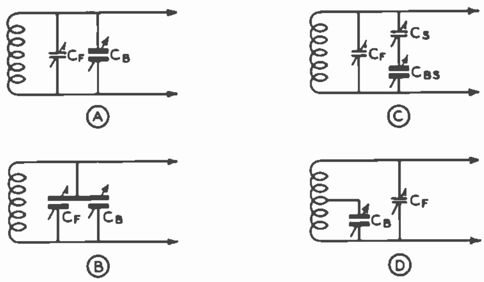


Figure 12.
BANDSPREAD CIRCUITS.
Parallel bandspread is illustrated at (A) and (B), series bandspread at (C), and tapped-coil bandspread at (D).

range. But in the high-transconductance types the effective capacitance will vary considerably from the published figures as operating conditions are changed.

The second source of circuit capacitance, and that which is more easily controllable, is that contributed by the minimum capacitance of the variable capacitors across the circuit and that due to capacitance between the wiring and ground. In well-designed high-frequency receivers, every effort is made to keep this portion of the circuit capacitance at a minimum since a large capacitance reduces the tuning range available with a given coil and prevents a good L/C ratio, and consequently a high-impedance tuned circuit, from being obtained.

A good percentage of stray circuit capacitance is due also to distributed capacitance of the coil and capacitance between wiring points and chassis.

Typical values of circuit capacitance may run from 10 to 75 μμfd. in high-frequency receivers, the first figure representing concentric-line receivers with acorn or miniature tubes and extremely small tuning capacitors, and the latter representing all-wave sets with bandswitching, large tuning capacitors, and conventional tubes.

6-6 I-F Tuned Circuits

I-f amplifiers usually employ bandpass circuits of some sort. A bandpass circuit is exactly what the name implies—a circuit for passing a band of frequencies. Bandpass arrangements can be designed for almost any degree of selectivity, the type used in any particular case depending upon the ultimate application of the amplifier.

I-F Transformers Intermediate frequency transformers ordinarily consist of two or more tuned circuits and some method of coupling the tuned circuits together. Some representative arrangements are shown in figure 13. The circuit shown at A is the conventional i-f transformer, with the coupling, M, between the tuned circuits being provided by inductive coupling from one coil to the other. As the coupling is increased, the selectivity curve becomes less peaked, and when a condition known as "critical coupling" is reached, the top of the curve begins to flatten out. When the coupling is increased still more, a dip occurs in the top of the curve.

The windings for this type of i-f transformer, as well as most others, nearly always consist of small, flat universal-wound pies mounted either on a piece of dowel to provide an air core or on powdered-iron for "iron core" i-f transformers. The iron-core transformers generally have somewhat more gain and better selectivity than equivalent air-core units.

The circuits shown at B and C are quite similar. Their only difference is the type of mutual coupling used, an inductance being used at B and a capacitance at C. The operation of both circuits is similar. Three resonant circuits are formed by the components. In B, for example, one resonant circuit is formed by L_1, C_1, C_2 and L_2 all in series. The frequency of this resonant circuit is just the same as that of a single one of the coils and capacitors, since the coils and capacitors are similar in both sides of the circuit, and the resonant frequency of the two capacitors and the two coils all in series is the same as that of a single coil and capacitor. The second resonant frequency of the complete circuit is determined by the characteristics of each half of the circuit containing the mutual coupling device. In B, this second frequency will be lower than the first, since the resonant frequency of L_1, C_1 and the inductance, M, or L_2, C_2 and M is lower than that of a single coil and capacitor, due to the inductance of M being added to the circuit.

The opposite effect takes place at C, where the common coupling impedance is a capacitor.

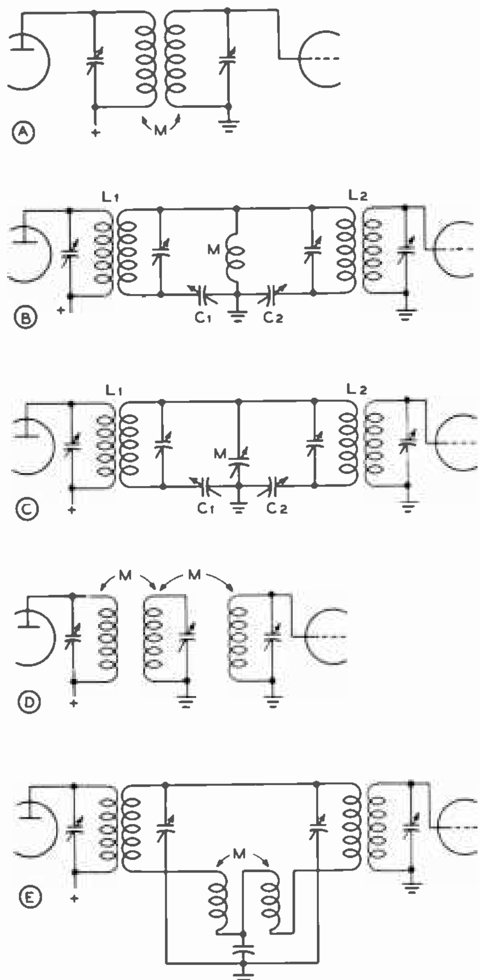


Figure 13.
I-F AMPLIFIER COUPLING ARRANGEMENTS.
The interstage coupling arrangements illustrated above give a better shape factor (more straight sided selectivity curve) than would the same number of tuned circuits coupled by means of tubes.

Thus, at C the second resonant frequency is higher than the first. In either case, however, the circuit has two resonant frequencies, resulting in a flat-topped selectivity curve. The width of the top of the curve is controlled by the reactance of the mutual coupling component. As this reactance is increased (inductance made greater, capacitance made smaller), the two resonant frequencies become further apart and the curve is broadened.

In the circuit of figure 13D, there is inductive coupling between the center coil and each of the outer coils. The result of this arrangement is that the center coil acts as a sharply tuned coupler between the other two. A signal somewhat off the resonant frequency of the transformer will not induce as much current in the center coil as will a signal of the correct frequency. When a smaller current is induced in the center coil, it in turn transfers a still smaller current to the output coil. The effective coupling between the outer coils increases as the resonant frequency is approached, and remains nearly constant over a small range and then decreases again as the resonant band is passed.

Another very satisfactory bandpass arrangement, which gives a very straight-sided, flat-topped curve, is the negative-mutual arrangement shown at E. Energy is transferred between the input and output circuits in this arrangement by both the negative-mutual coils, M, and the common capacitive reactance, C. The negative-mutual coils are interwound on the same form, and connected "backward."

Transformers usually are made tunable over a small range to permit accurate alignment in the circuit in which they are employed. This is accomplished either by means of a variable capacitor across a fixed inductance, or by means of a fixed capacitor across a variable inductance. The former usually employ either a mica-compression capacitor (designated "mica tuned"), or a small air dielectric variable capacitor (designated "air tuned"). Those which use a fixed capacitor usually employ a powdered iron core on a threaded rod to vary the inductance, and are known as "permeability tuned."

Shape Factor It is obvious that to pass modulation sidebands and to allow for slight drifting of the transmitter carrier frequency and the receiver local oscillator, the i-f amplifier must pass not a single frequency but a band of frequencies. The width of this pass band, usually 5 to 8 kc. at maximum width in a good communications receiver, is known as the "pass band," and is arbitrarily taken as the width between the two frequencies at which the response is attenuated 6 db, or is "6 db down." However, it is apparent that to discriminate against an interfering signal which is stronger than the desired signal, much more than 6 db attenuation is required. The attenuation arbitrarily taken to indicate adequate discrimination against an interfering signal is 60 db.

It is apparent that it is desirable to have the band width at 60 db down as narrow as possible, but it must be done without making the pass band (6 db down points) too narrow for satisfactory reception of the desired signal. The figure of merit used to show the ratio of bandwidth at 6 db down to that at 60 db down is designated *shape factor.* The ideal i-f curve, a rectangle, would have a shape factor of 1.0. The i-f shape factor in typical communications receivers runs from 3.0 to 5.5.

The most practicable method of obtaining a low shape factor for a given number of tuned circuits is to employ them in pairs, as in figure 13A, adjusted to *critical coupling* (the value at which two resonance points just begin to become apparent). If this gives too sharp a "nose" or pass band, then coils of lower Q should be employed, with the coupling maintained at the critical value. As the Q is lowered, closer coupling will be required for critical coupling.

Conversely if the pass band is too broad, coils of higher Q should be employed, the coupling being maintained at critical. If the pass band is made more narrow by using looser coupling instead of raising the Q and maintaining critical coupling, the shape factor will not be as good.

The *pass band* will not be much narrower for several pairs of identical, critically coupled tuned circuits than for a single pair. However, the *shape factor* will be greatly improved as each additional pair is added, up to about 5 pairs, beyond which the improvement for each additional pair is not significant. Commercially available communications receivers of good quality normally employ 3 or 4 double tuned transformers with coupling adjusted to critical or slightly less.

"Miller Effect" As mentioned previously, the dynamic input capacitance of a tube varies slightly with bias. As a-v-c voltage normally is applied to i-f tubes for radiotelephony reception, the effective grid-cathode capacitance varies as the signal strength varies, which produces the same effect as slight detuning of the i-f transformer. This effect is known as "Miller effect," and can be minimized to the extent that it is not troublesome either by using a fairly low L/C ratio in the

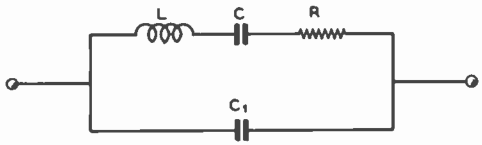


Figure 14.
ELECTRICAL EQUIVALENT OF QUARTZ FILTER CRYSTAL.
The crystal is equivalent to a very large value of inductance in series with small values of capacitance and resistance, with a larger though still small value of capacitance across the whole circuit (representing holder capacitance plus stray capacitances).

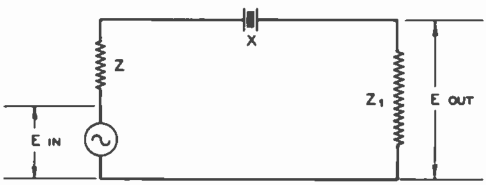


Figure 15.
EQUIVALENT OF CRYSTAL FILTER CIRCUIT.
For a given voltage out of the generator, the voltage developed across Z_1 depends upon the ratio of the impedance of X to the sum of the impedances of Z and Z_1. Because of the high Q of the crystal, its impedance changes rapidly with changes in frequency.

transformers or by incorporating a small amount of degenerative feedback, the latter being most easily accomplished by leaving part of the cathode resistor unbypassed for r.f.

Crystal Filters The pass band of an intermediate frequency amplifier may be made very narrow through the use of a piezoelectric filter crystal employed as a series resonant circuit in a bridge arrangement known as a *crystal filter*. The shape factor is quite poor, as would be expected when the selectivity is obtained from the equivalent of a single tuned circuit, but the very narrow pass band obtainable as a result of the extremely high Q of the crystal makes the crystal filter useful for c-w telegraphy reception. The pass band of a 455 kc. crystal filter may be made as narrow as 50 cycles, while the narrowest pass band that can be obtained with a 455 kc. tuned circuit of practicable dimensions is about 5 kc.

The electrical equivalent of a filter crystal is shown in figure 14. For a given frequency, L is very high, C very low, and R (assuming a good crystal of high Q) is very low. Capacitance C_1 represents the shunt capacitance of the electrodes, plus the crystal holder and wiring, and is many times the capacitance of C. This makes the crystal act as a parallel resonant circuit with a frequency only slightly higher than that of its frequency of series resonance. For crystal filter use it is the series resonant characteristic that we are primarily interested in.

The electrical equivalent of the basic crystal filter circuit is shown in figure 15. If the impedance of Z plus Z_1 is low compared to the impedance of the crystal X at resonance, then the current flowing through Z_1, and the voltage developed across it, will be almost in inverse proportion to the impedance of X, which has a very sharp resonance curve.

If the impedance of Z plus Z_1 is made *high* compared to the resonant impedance of X, then there will be no appreciable drop in voltage across Z_1 as the frequency departs from the resonant frequency of X until the point is reached where the impedance of X approaches that of Z plus Z_1. This has the effect of broadening out the curve of frequency versus voltage developed across Z_1, which is another way of saying that the selectivity of the crystal filter (but not the crystal proper) has been reduced.

In practicable filter circuits the impedances Z and Z_1 usually are represented by some form of tuned circuit, but the basic principle of operation is the same.

Practical Filters It is necessary to balance out the capacitance across the crystal holder (C_1, in figure 14) to prevent bypassing around the crystal undesired signals off the crystal resonant frequency. The balancing is done by a *phasing* circuit which takes out-of-phase voltage from a balanced input circuit and passes it to the output side of the crystal in proper phase to neutralize that passed through the holder capacitance. A representative practical filter arrangement is shown in figure 16. The phasing capacitor is indicated in the diagram by PC. The balanced input circuit may be obtained either through the use of a split-stator capacitor as shown, or by the use of a center-tapped input coil.

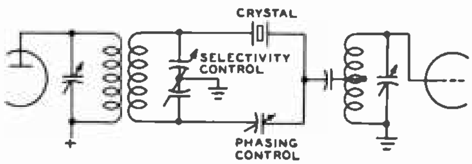


Figure 16.
TYPICAL CRYSTAL FILTER CIRCUIT.

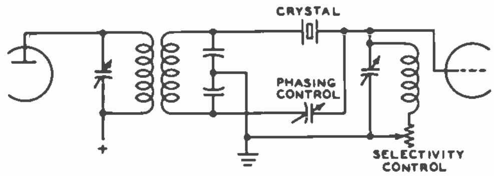


Figure 17.
VARIABLE SELECTIVITY CRYSTAL FILTER.
This circuit permits of a greater control of selectivity than does the circuit of figure 16, and does not require a split-stator variable capacitor.

Variable-Selectivity Filters In the circuit of figure 16, the selectivity is *minimum* with the crystal input circuit tuned to resonance, since at resonance the impedance of the tuned circuit is maximum. As the input circuit is detuned from resonance, however, the impedance decreases, and the selectivity becomes greater. In this circuit, the output from the crystal filter is tapped down on the i-f stage grid winding to provide a low value of series impedance in the output circuit. It will be recalled that for maximum selectivity, the total impedance in series with the crystal (both input and output circuits) must be low. If one is made low and the other is made variable, then the selectivity may be varied at will from sharp to broad.

The circuit shown in figure 17 also achieves variable selectivity by adding a variable impedance in series with the crystal circuit. In this case, the variable impedance is in series with the crystal output circuit. The impedance of the output tuned circuit is varied by varying the Q. As the Q is reduced (by adding resistance in series with the coil), the impedance decreases and the selectivity becomes greater. The input circuit impedance is made low by using a non-resonant secondary on the input transformer.

A variation of the circuit shown at figure 17 consists of placing the variable resistance across the coil and capacitor, rather than in series with them. The result of adding the resistor is a reduction of the output impedance, and an increase in selectivity. The circuit behaves oppositely to that of figure 17, however; as the resistance is lowered the selectivity becomes greater. Still another variation of figure 17 is to use the tuning capacitor across the output coil to vary the output impedance. As the output circuit is detuned from resonance, its impedance is lowered, and the selectivity increases. Sometimes a set of fixed capacitors and a multipoint switch are used to give step-by-step variation of the output circuit tuning, and thus of the crystal filter selectivity.

Rejection Notch As previously discussed, a filter crystal has both a resonant (series resonant) and an anti-resonant (parallel resonant) frequency, the impedance of the crystal being quite low at the former frequency, and quite high at the latter frequency. The anti-resonant frequency is just slightly higher than the resonant frequency, the difference depending upon the effective shunt capacitance of the filter crystal and holder. As adjustment of the phasing capacitor controls the effective shunt capacitance of the crystal, it is possible to vary the anti-resonant frequency of the crystal slightly without unbalancing the circuit sufficiently to let undesired signals "leak through" the shunt capacitance in appreciable amplitude. At the exact anti-resonant frequency of the crystal the attenuation is exceedingly high, because of the high impedance of the crystal at this frequency. This is called the "rejection notch," and can be utilized virtually to eliminate the heterodyne image or "repeat tuning" of c-w signals. The beat frequency oscillator can be so adjusted and the phasing capacitor so adjusted that the desired beat note is of such a pitch that the image (the same audio note on the other side of zero beat) falls in the rejection notch and is inaudible. The receiver then is said to be adjusted for "single-signal" operation.

The rejection notch sometimes can be employed to reduce interference from an undesired *phone* signal which is very close in frequency to a desired phone signal. The filter is adjusted to "broad" so as to permit telephony reception, and the receiver tuned so that the carrier frequency of the undesired signal falls in the rejection notch. The modulation sidebands of the undesired signal still will come through, but the carrier heterodyne will be effectively eliminated and interference greatly reduced.

Crystal Filter Considerations A crystal filter, especially when adjusted for "single signal" reception, greatly reduces interference and background noise, the latter feature permitting signals to be copied that would ordinarily be too weak to be heard above the background hiss. However, when the filter is adjusted for maximum selectivity, the pass band is so narrow that the received signal must have a high order of stability in order to stay within the pass band. Likewise, the local oscillator in the receiver must be highly stable, or constant retuning will be required. Another effect that will be noticed with the filter adjusted to "sharp" is a tendency for code characters to produce a ringing sound, and have a hangover or "tails." This effect limits the code speed that can be copied satisfactorily when the filter is adjusted for extreme selectivity.

Beat-Frequency Oscillators The beat-frequency oscillator, usually called the *b.f.o.*, is a necessary adjunct for reception of c-w telegraph signals on superheterodynes which have no other provision for obtaining modulation of an incoming c-w telegraphy signal. The oscillator is coupled into or just ahead of second detector circuit and supplies a signal of nearly the same frequency as that of the desired signal from the i-f amplifier. If the i-f amplifier is tuned to 455 kc., for example, the b.f.o. is tuned to approximately 454 or 456 kc. to produce an audible (1000 cycle) beat note in the output of the second detector of the receiver. The carrier signal itself is, of course, inaudible. The b.f.o. is not used for voice reception, except as an aid in searching for weak stations.

The b-f-o input to the second detector need only be sufficient to give a good beat note on an average signal. Too much coupling into the second detector will give an excessively high hiss level, masking weak signals by the high noise background.

Figure 18 shows a method of manually adjusting the b-f-o output to correspond with the strength of received signals. This type of variable b-f-o output control is a useful adjunct to any superheterodyne, since it allows sufficient b-f-o output to be obtained to "beat" with strong signals or to allow single-sideband reception and at the same time permits the b-f-o output, and consequently the hiss, to be reduced when attempting to receive weak signals. The circuit shown is somewhat better than those in which one of the electrode voltages on the b-f-o tube is changed, as the latter circuits usually change the frequency of the b.f.o. at the same time they change the strength, making it necessary to reset the trimmer each time the output is adjusted.

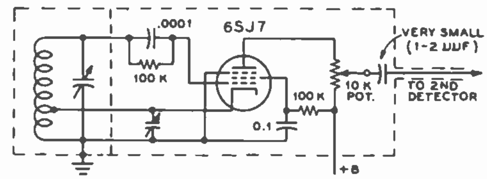


Figure 18.
VARIABLE-OUTPUT B-F-O CIRCUIT.
A beat-frequency oscillator whose output is controllable is of considerable assistance in copying c-w signals over a wide range of levels, and such a control is almost a necessity for satisfactory copying of single-sideband radiophone signals.

The b.f.o. usually is provided with a small trimmer which is adjustable from the front panel to permit adjustment over a range of 5 or 10 kc. For single-signal reception the b.f.o. always is adjusted to the high-frequency side, in order to permit placing the heterodyne image in the rejection notch.

In order to reduce the b-f-o signal output voltage to a reasonable level which will prevent blocking the second detector, the signal voltage is delivered through a low-capacitance (high-reactance) capacitor having a value of 1 to 2 $\mu\mu$fd.

Care must be taken with the b.f.o. to prevent harmonics of the oscillator from being picked up at multiples of the b-f-o frequency. The complete b.f.o. together with the coupling circuits to the second detector, should be thoroughly shielded to prevent pickup of the harmonics by the input end of the receiver.

If b-f-o harmonics still have a tendency to give trouble after complete shielding and isolation of the b-f-o circuit has been accomplished,

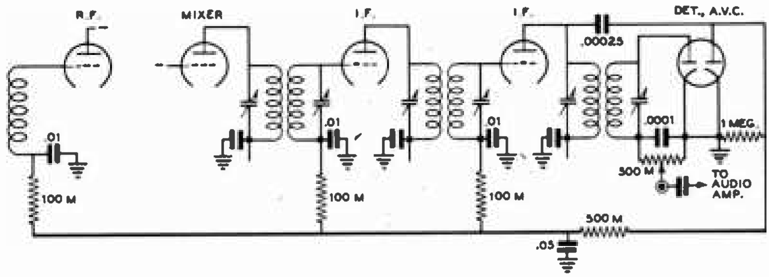


Figure 19.
TYPICAL A-V-C CIRCUIT USING A DOUBLE DIODE.
Any of the small dual-diode tubes may be used in this circuit. Or, if desired, a duo-diode-triode may be used, with the triode acting as the first audio stage. The left-hand diode serves as the detector, while the right-hand side acts as the a-v-c rectifier. The use of separate diodes for detector and a-v-c reduces distortion when receiving an AM signal with a high modulation percentage.

the passage of these harmonics from the b-f-o circuit to the rest of the receiver can be stopped through the use of a low-pass filter in the lead between the output of the b-f-o circuit and the point on the receiver where the b-f-o signal is to be injected. The design of such filters is discussed in Chapter 3.

6-7 Detector, Audio, and Control Circuits

Detectors Second detectors for use in superheterodynes are usually of the diode, plate, or infinite-impedance types. Occasionally, grid-leak detectors are used in receivers using one i-f stage or none at all, in which case the second detector usually is made regenerative.

Diodes are the most popular second detectors because they allow a simple method of obtaining automatic volume control to be used. Diodes load the tuned circuit to which they are connected, however, and thus reduce the selectivity slightly. Special i-f transformers are used for the purpose of providing a low-impedance input circuit to the diode detector.

Automatic Volume Control The elements of an automatic volume control (a.v.c.) system are shown in figure 19. A dual-diode tube is used as a combination diode detector and a-v-c rectifier. The left-hand diode operates as a simple rectifier in the manner described earlier in this chapter. Audio voltage, superimposed on a d-c voltage, appears across the 500,000-ohm potentiometer (the volume control) and the .0001-μfd. capacitor, and is passed on to the audio amplifier. The right-hand diode receives signal voltage directly from the primary of the last i-f amplifier, and acts as the a-v-c rectifier. The pulsating d-c voltage across the 1-megohm a.v.c.-diode load resistor is filtered by a 500,000-ohm resistor and a .05-μfd. capacitor, and applied as bias to the grids of the r-f and i-f amplifier tubes; an increase or decrease in signal strength will cause a corresponding increase or decrease in a-v-c bias voltage, and thus the gain of the receiver is automatically adjusted to compensate for changes in signal strength.

By disassociating the a.v.c. and detecting functions through using separate diodes, as shown, most of the ill effects of a-c shunt loading on the detector diode are avoided. This type of loading causes serious distortion, and the additional components required to eliminate it are well worth their cost. Even with the circuit shown, a-c loading can occur unless a very high (5 megohms, or more) value of grid resistor is used in the following audio amplifier stage.

A.V.C. in B-F-O-Equipped Receivers In receivers having a beat-frequency oscillator for the reception of radiotelegraph signals, the use of a.v.c. can result in a great loss in sensitivity when the b.f.o. is switched on. This is because the beat oscillator output acts exactly like a strong received signal, and causes the a-v-c circuit to put high bias on the r-f and i-f stages, thus greatly reducing the receiver's sensitivity. Due to the above effect, it is necessary to provide a method of making the a-v-c circuit inoperative when the b.f.o. is being used. The simplest method of eliminating the a-v-c action is to short the a-v-c line to ground when the b.f.o. is turned on. A two-circuit switch may be used for the dual purpose of turning on the beat oscillator and shorting out the a.v.c. if desired.

Signal Strength Indicators Visual means for determining whether or not the receiver is properly tuned, as well as an indication of the relative signal strength, are both provided by means of *tuning indicators* (S meters) of the meter or vacuum-tube type. A d-c milliammeter can be connected in the plate supply circuit of one or more r-f or i-f amplifiers, as shown in figure 20A, so that the change in plate current, due to the action of the a-v-c voltage, will be indicated on the instrument. The d-c instrument M should have a full-scale reading approximately equal to the total plate current taken by the stage or stages whose plate current passes through the instrument. The value of this current can be estimated by assuming a plate current on each stage (with no signal input to the receiver) of about 6 ma. However, it will be found to be more satisfactory to measure the actual plate current on the stages with a milliammeter of perhaps 0-100 ma. full scale before purchasing an instrument for use as an S meter. The 50-ohm potentiometer shown in the drawing is used to adjust the meter reading to full scale with no signal input to the receiver.

When an ordinary meter is used in the plate circuit of a stage, for the purpose of indicating signal strength, the meter reads backwards with respect to strength. This is because increased a-v-c bias on stronger signals causes lower plate current through the meter. For this reason, special meters which indicate zero at the right-hand end of the scale are often used for signal strength indicators in commercial receivers using this type of circuit. Alternatively, the meter may be mounted upside down, so that the needle moves toward the right with increased signal strength.

The circuit of figure 20B can frequently be used to advantage in a receiver where the cathode of one of the r-f or i-f amplifier stages runs directly to ground through the cathode

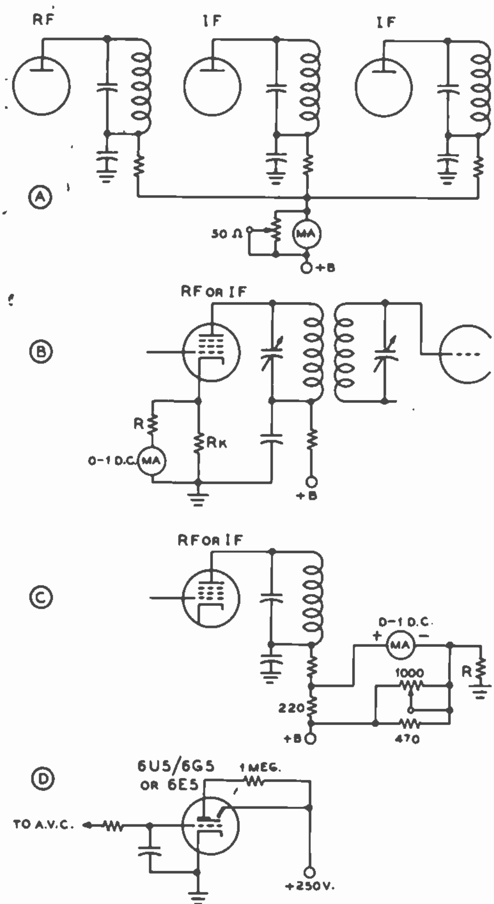


Figure 20.
SIGNAL-STRENGTH-METER CIRCUITS.
Shown above are four circuits for obtaining a signal-strength reading which is a function of incoming carrier amplitude. The circuits are discussed in the accompanying text.

bias resistor instead of running through a cathode-voltage gain control. In this case a 0-1 d-c milliammeter in conjunction with a resistor from 1000 to 3000 ohms can be used as shown as a signal-strength meter. With this circuit the meter will read backwards with increasing signal strength as in the circuit previously discussed.

Figure 20C is the circuit of a forward-reading S meter as is often used in communications receivers. The instrument is used in an unbalanced bridge circuit with the d-c plate resistance of one i-f tube as one leg of the bridge and with resistors for the other three legs. The value of the resistor R must be determined by trial and error and will be somewhere in the vicinity of 50,000 ohms. Sometimes the screen circuits of the r-f and i-f stages are taken from this point along with the screen-circuit voltage divider.

Electron-ray tubes (sometimes called "magic eyes") can also be used as indicators of relative signal strength in a circuit similar to that shown in figure 20D. A 6U5/6G5 tube should be used where the a-v-c voltage will be from 5 to 20 volts and a type 6E5 tube should be used when the a-v-c voltage will run from 2 to 8 volts.

Audio Amplifiers Audio amplifiers are employed in nearly all radio receivers. The audio amplifier stage or stages are usually of the Class A type, although Class AB push-pull stages are used in some receivers. The operation of both of these types of amplifiers was described in detail in Chapter 5. The purpose of the audio amplifier is to bring the relatively weak signal from the detector up to a strength sufficient to operate a pair of headphones or a loud speaker. Either triodes, pentodes, or beam tetrodes may be used, the pentodes and beam tetrodes usually giving greater output. In some receivers, particularly those employing grid leak detection, it is possible to operate the headphones directly from the detector, without audio amplification. In such receivers, a single audio stage with a beam tetrode or pentode tube is ordinarily used to drive the loud speaker.

Most communications receivers, either home-constructed or factory-made, have a single-ended beam tetrode (such as a 6L6 or 6V6) or pentode (6F6 or 6K6-GT) in the audio output stage feeding the loudspeaker. If precautions are not taken such a stage will actually bring about a decrease in the effective signal-to-noise ratio of the receiver due to the rising high-frequency characteristic of such a stage when feeding a loud-speaker. One way of improving this condition is to place a mica or paper capacitor of approximately 0.003 μfd. capacitance across the primary of the output transformer. The use of a capacitor in this manner tends to make the load impedance seen by the plate of the output tube more constant over the audio-frequency range. The speaker and transformer will tend to present a rising impedance to the tube as the frequency increases, and the parallel capacitor will tend to make the total impedance more constant since it will tend to present a decreasing impedance with increasing audio frequency.

A still better way of improving the frequency characteristic of the output stage, and at the same time reducing the harmonic distortion, is to use shunt feedback from the plate of the output tube to the plate of a tube such as a 6SJ7 acting as an audio amplifier stage ahead of the output stage. This circuit is illustrated in figure 27, Chapter 5.

6-8 Noise Suppression

The problem of noise suppression confronts the listener who is located in places where interference from power lines, electrical appliances, and automobile ignition systems is troublesome. This noise is often of such intensity as to swamp out signals from desired stations.

There are two principal methods for reducing this noise:

(1) A-c line filters at the source of interference, if the noise is created by an electrical appliance.

(2) Noise-limiting circuits for the reduction, in the receiver itself, of interference of the type caused by automobile ignition systems.

Power Line Filters Many household appliances, such as electric mixers, heating pads, vacuum sweepers, refrigerators, oil burners, sewing machines, doorbells, etc., create an interference of an intermittent nature. The insertion of a line filter near the source of interference often will effect a complete cure. Filters for small appliances can consist of a 0.1-μfd. capacitor connected across the 110-volt a-c line. Two capacitors in series across the line, with the midpoint connected to ground, can be used in conjunction with ultra-violet ray machines, refrigerators, oil burner furnaces, and other more stubborn offenders. In severe cases of interference, additional filters in the form of heavy-duty r-f choke coils must be connected in series with the 110-volt a-c line on both sides of the line right at the interfering appliance.

Peak Noise Limiters Numerous noise-limiting circuits which are beneficial in overcoming key clicks, automobile ignition interference, and similar noise im-

C_1—0.1-μfd. paper
C_2—50-$\mu\mu$fd. mica
C_3—100-$\mu\mu$fd. mica
C_4, C_5—0.01-μfd. paper
R_1, R_2—1 megohm, ½ watt
R_3, R_4—220,000 ohms, ½ watt
R_5, R_6—1 megohm, ½ watt
R_7—2-megohm potentiometer

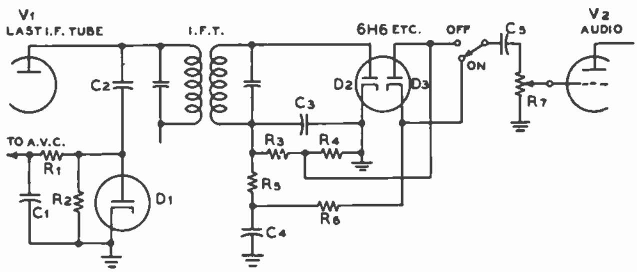


Figure 21.
NOISE LIMITER CIRCUIT, WITH ASSOCIATED A-V-C.
This limiter is of the series type, and is self-adjusting to carrier strength for phone reception. For proper operation several volts should be developed across the secondary of the last i-f transformer (IFT) under carrier conditions.

pulses have become popular. They operate on the principle that each individual noise pulse is of very short duration, yet of very high amplitude. The popping or clicking type of noise from electrical ignition systems may produce a signal having a peak value ten to twenty times as great as the incoming radio signal, but an average power much less than the signal.

As the duration of this type of noise peak is short, the receiver can be made inoperative during the noise pulse without the human ear detecting the total loss of signal. Some noise limiters actually *punch a hole* in the signal, while others merely *limit* the maximum peak signal which reaches the headphones or loudspeaker.

The noise peak is of such short duration that it would not be objectionable except for the fact that it produces an over-loading effect on the receiver, which increases its time constant. A sharp voltage peak will give a kick to the diaphragm of the headphones or speaker, and the momentum or inertia keeps the diaphragm in motion until the dampening of the diaphragm stops it. This movement produces a popping sound which may completely obliterate the desired signal. If the noise pulse can be limited to a peak amplitude equal to that of the desired signal, the resulting interference is practically negligible for moderately low repetition rates, such as ignition noise.

Virtually all of the practicable peak limiters for radio-telephony employ one or two diodes either as shunt or series limiters in the a-f system. When a noise pulse exceeds a certain predetermined threshold value, the limiter diode acts either as a short or open circuit, depending upon whether it is used in a shunt or series circuit. The threshold is made to occur at a level high enough that it will not clip modulation peaks enough to impair voice intelligibility, but low enough to limit the noise peaks effectively.

Because the action of the peak limiter is needed most on very weak signals, and these usually are not strong enough to produce proper a-v-c action, a threshold setting that is correct for a strong phone signal is not correct for optimum limiting on very weak signals. For this reason the threshold control often is tied in with the a-v-c system so as to make the optimum threshold adjustment automatic instead of manual.

Suppression of impulse noise by means of an audio peak limiter is best accomplished at the very front end of the audio system, and for this reason the function of superheterodyne second detector and limiter often are combined in a composite circuit.

The amount of limiting that can be obtained is a function of the audio distortion that can be tolerated. Because excessive distortion will reduce the intelligibility as much as will background noise, the degree of limiting for which the circuit is designed has to be a compromise.

Peak noise limiters working at the second detector are much more effective when the i-f bandwidth of the receiver is broad, because a sharp i-f amplifier will lengthen the pulses by the time they reach the second detector, making the limiter less effective. V-h-f superheterodynes have an i-f bandwidth considerably wider than the minimum necessary for voice sidebands (to take care of drift and instability). Therefore, they are capable of better peak noise suppression than a standard communications receiver having an i-f bandwidth of perhaps 8 kc. Likewise, when a crystal filter is used on the "sharp" position an a-f peak limiter is of little benefit.

Practical Peak Noise Limiter Circuits Noise limiters range all the way from an audio stage running at very low screen or plate voltage, to elaborate affairs employing 5 or more tubes. Rather than attempt to show the numerous types, many of which are quite complex considering the results obtained, only two very similar types will be described. Either is just about as effective as the most elaborate limiter that can be constructed, yet requires the addition of but a single diode and a few resistors and capacitors over what would be employed in a good superheterodyne without a limiter. Both circuits, with but minor modifications in resistance and capacitance values, are incorporated in one form or another in different types of factory-built communications receivers.

Referring to figure 21, the first circuit shows a conventional superheterodyne second detector, a.v.c., and first audio stage with the addition of one tube element, D_3, which may be either a separate diode or part of a twin-diode as illustrated. Diode D_3 acts as a series gate, allowing audio to get to the grid of the a-f tube only so long as the diode is conducting. The diode is biased by a d-c voltage obtained in the same manner as a-v-c control voltage, the bias being such that pulses of short duration no longer conduct when the pulse voltage exceeds the carrier by approximately 60 per cent. This also clips voice modulation peaks, but not enough to impair intelligibility.

It is apparent that the series diode clips only *positive* modulation peaks, by limiting upward modulation to about 60 per cent. Negative or downward peaks are limited automatically to 100 per cent in the detector, because obviously the rectified voltage out of the diode detector cannot be less than zero. Limiting the downward peaks to 60 per cent or so instead of 100 per cent would result in but little improvement in noise reduction, and the results do not justify the additional components required.

It is important that the exact resistance values shown be used, for best results, and that 10 per cent tolerance resistors be used for R_3 and R_4. Also, the rectified carrier voltage developed across C_1 should be at least 5 volts for good limiting.

The limiter will work well on c-w telegraphy if the amplitude of beat frequency oscillator injection is not too high. Variable injection is to be preferred, adjustable from the front panel. If this feature is not provided, the b-f-o injection should be reduced to the lowest value that will give a satisfactory beat. When this is done, effective limiting and a good beat can be obtained by proper adjustment of the r-f and a-f gain controls. It is assumed, of course, that the a.v.c. is cut out of the circuit for c-w telegraphy reception.

Alternative Limiter Circuit The circuit of figure 22 is more effective than that shown in figure 21 under certain conditions and requires the addition of only one more resistor and one more capacitor than the other circuit. Also, this circuit involves a smaller loss in output level than the circuit of figure 21. This circuit can be used with equal effectiveness with a combined diode-triode or diode-pentode tube (6R7, 6SR7, 6Q7, 6SQ7 or similar diode-triodes, or 6B8, 6SF7, or similar diode-pentodes) as diode detector and first audio stage. However, a separate diode must be used for the noise limiter, D_2. This diode may be one-half of a 6H6, 6AL5, 7A6, etc., or it may be a triode connected 6J5, 6C4 or similar type.

Note that the return for the volume control must be made to the cathode of the detector diode (and not to ground) when a dual tube is used as combined second-detector first-audio. This means that in the circuit shown in figure 22 a connection will exist across the points where the "X" is shown on the diagram since a common cathode lead is brought out of the tube for D_1 and V_1. If desired, of course, a single dual diode may be used for D_1 and D_2 in this circuit as well as in the circuit of figure 21. Switching the limiter in and out with the switch S brings about no change in volume.

In any diode limiter circuit such as the ones shown in these two figures it is important that the mid-point of the heater potential for the noise-limiter diode be as close to ground potential as possible. This means that the center-tap of the heater supply for the tubes

This circuit is of the self-adjusting type and gives less distortion for a given degree of modulation than the more common limiter circuits.
R_1, R_2—470K, ½ watt
R_3—100K, ½ watt
R_4, R_5—1 megohm, ½ watt
R_6—2-megohm potentiometer
C_1—0.00025 mica (approx.)
C_2—0.01-μfd. paper
C_3—0.01-μfd. paper
C_4—0.01-μfd. paper
D_1, D_2—6H6, 6AL5, 7A6, or diode sections of a 6S8-GT

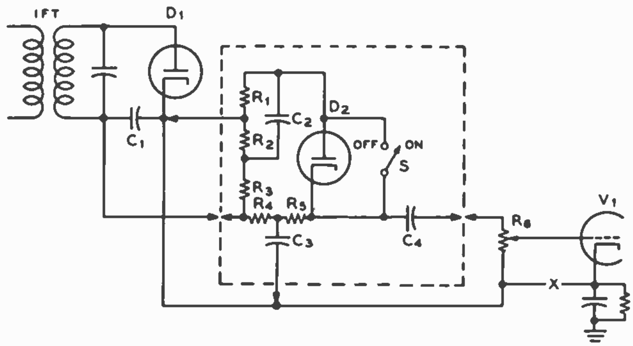


Figure 22.
ALTERNATIVE NOISE LIMITER CIRCUIT.

should be grounded wherever possible rather than grounding one side of the heater supply as is often done. Difficulty with hum pickup in the limiter circuit *may* be encountered when one side of the heater is grounded due to the high values of resistance necessary in the limiter circuit.

The circuit of figure 22 has been used with excellent success in several home-constructed receivers, and in the BC-312/BC-342 and BC-348 series of surplus communications receivers. It is also used in certain manufactured receivers.

Incidentally, an excellent check on the operation of the noise limiter in any communications receiver can be obtained by listening to the Loran signals in the 160-meter band. With the limiter out a sharp rasping buzz will be obtained when one of these stations is tuned in. With the noise limiter switched into the circuit the buzz should be greatly reduced and a low-pitched hum should be heard.

6-9 Special Considerations in U-H-F Receiver Design

Transmission Line Circuits At increasingly higher frequencies, it becomes progressively more difficult to obtain a satisfactory amount of selectivity and impedance from an ordinary coil and capacitor used as a resonant circuit. On the other hand, quarter wavelength sections of parallel conductors or concentric transmission line are not only better but also become of practical dimensions.

Tuning Short Lines Tubes and tuning capacitors connected to the open end of a transmission line provide a capacitance that makes the resonant length less than a quarter wave-length. The amount of shortening for a specified capacitive reactance is determined by the surge impedance of the line section. It is given by the equation for resonance:

$$\frac{1}{2\pi fC} = Z_o \tan l$$

in which $\pi = 3.1416$, f is the frequency, C the capacitance, Z_o the surge impedance of the line, and tan l is the tangent of the electrical length in degrees.

The capacitive reactance of the capacitance across the end is $1/(2\pi f C)$ ohms. For resonance, this must equal the surge impedance of the line times the tangent of its electrical length (in degrees, where 90° equals a quarter wave). It will be seen that twice the capacitance will resonate a line if its surge impedance is halved; also that a given capacitance has twice the loading effect when the frequency is doubled.

Coupling Into Lines It is possible to couple into a parallel-rod line by tapping directly on one or both rods, preferably through blocking capacitors if any d.c. is present. More commonly, however, a "hairpin" is inductively coupled at the shorting bar end, either to the bar or to the two rods, or both. This normally will result in a balanced load. Should a loop unbalanced to ground be coupled in, any resulting unbalance

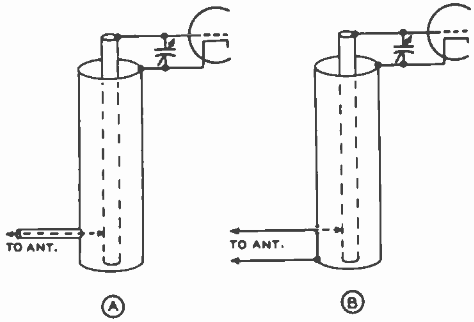


Figure 23.
COUPLING AN ANTENNA TO A COAXIAL RESONANT CIRCUIT.
(A) shows the recommended method for coupling a coaxial line to a coaxial resonant circuit. (B) shows an alternative method for use with an open-wire type of antenna feed line.

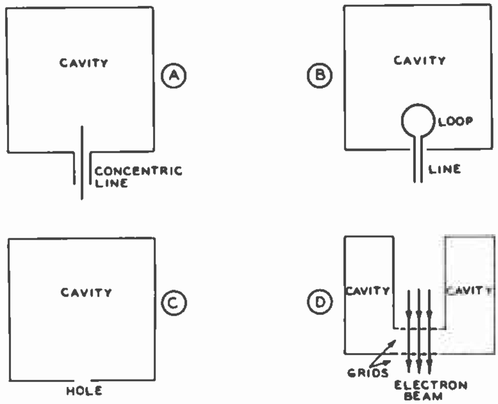


Figure 24.
METHODS OF EXCITING A RESONANT CAVITY.

reflected into the rods can be reduced with a simple Faraday screen, made of a few parallel wires placed between the hairpin loop and the rods. These should be soldered at only one end and grounded.

An unbalanced tap on a coaxial resonant circuit can be made directly on the inner conductor at the point where it is properly matched. For low impedances, such as a concentric line feeder, a small one-half turn loop can be inserted through a hole in the outer conductor of the coaxial circuit, being in effect a half of the hairpin type recommended for coupling balanced feeders to coaxial resonant lines. The size of the loop and closeness to the inner conductor determines the impedance matching and loading. Such loops coupled in near the shorting disc do not alter the tuning appreciably, if not overcoupled.

Resonant Cavities A *cavity* is a closed resonant chamber made of metal. It is known also as a rhumbatron. The cavity, having both inductance and capacitance, supersedes coil-capacitor and capacitance-loaded transmission-line tuned circuits at extremely high frequencies where conventional L and C components, of even the most refined design, prove impractical because of the tiny electrical and physical dimensions they must have. Microwave cavities have high Q factors and are superior to conventional tuned circuits. They may be employed in the manner of an absorption wavemeter or as the tuned circuit in other r-f test instruments and in microwave transmitters and receivers.

Resonant cavities usually are closed on all sides and all of their walls are made of electrical conductor. However, in some forms, small openings are present for the purpose of excitation. Cavities have been produced in several shapes including the plain sphere, dimpled sphere, sphere with reentrant cones of various sorts, cylinder, prism (including cube), ellipsoid, ellipsoid-hyperboloid, doughnut-shape, and various reentrant types. In appearance, they resemble in their simpler forms metal boxes or cans.

The cavity actually is a linear circuit, but one which is superior to a conventional coaxial resonator in the s-h-f range. The cavity resonates in much the same manner as does a barrel or a closed room with reflecting walls.

Because electromagnetic energy, and the associated electrostatic energy, oscillates to and fro inside them in one mode or another, resonant cavities resemble wave guides. The mode of operation in a cavity is affected by the manner in which micro-wave energy is injected. A cavity will resonate to a large number of frequencies, each being associated with a particular mode or standing-wave pattern. The lowest mode (lowest frequency of operation) of a cavity resonator normally is the one used.

The resonant frequency of a cavity may be varied, if desired, by means of movable plungers or plugs, as shown in figure 25A, or a

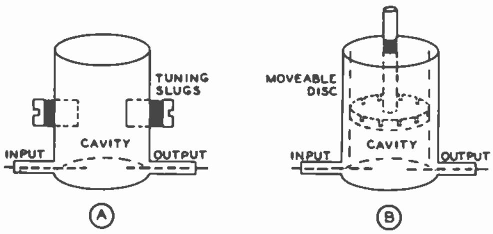


Figure 25.
TUNING METHODS FOR CYLINDRICAL RESONANT CAVITIES.

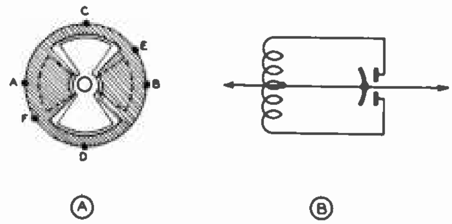


Figure 26.
THE BUTTERFLY RESONANT CIRCUIT.
Shown at (A) is the physical appearance of the butterfly circuit as used in the v-h-f and lower u-h-f range. (B) shows an electrical representation of the circuit.

movable metal disc (See figure 25B). A cavity that is too small for a given wavelength will not oscillate.

The resonant frequencies of simple spherical, cylindrical, and cubical cavities may be calculated simply for one particular mode. Wavelength and cavity dimensions (in centimeters) are related by the following simple resonance formulae:

For Cylinder $\lambda_r = 2.6 \times$ radius
" Cube $\lambda_r = 2.83 \times$ half of 1 side
" Sphere $\lambda_r = 2.28 \times$ radius

Butterfly Circuit Unlike the cavity resonator, which in its conventional form is a device which can tune over a relatively narrow band, the butterfly circuit is a tunable resonator which permits coverage of a fairly wide u-h-f band. The butterfly circuit is very similar to a conventional coil-variable capacitor combination, except that both inductance and capacitance are provided by what appears to be a variable capacitor alone. The Q of this device is somewhat less than that of a concentric-line tuned circuit but is entirely adequate for numerous applications.

Figure 26A shows construction of a single butterfly section. The butterfly-shaped rotor, from which the device derives its name, turns in relation to the unconventional stator. The two groups of stator "fins" or sectors are in effect joined together by a semi-circular metal band, integral with the sectors, which provides the circuit inductance. When the rotor is set to fill the loop opening (the position in which it is shown in figure 26A), the circuit inductance and capacitance are reduced to minimum. When the rotor occupies the position indicated by the dotted lines, the inductance and capacitance are at maximum. The tuning range of practical butterfly circuits is in the ratio of 1.5:1 to 3.5:1.

Direct circuit connections may be made to points A and B. If balanced operation is desired, either point C or D will provide the "center-tap" (electrical mid-point). Coupling may be effected by means of a small single-turn loop placed near point E or F. The butterfly thus permits continuous variation of both capacitance and inductance, as indicated by the equivalent circuit in figure 26B, while at the same time eliminating all pigtails and wiping contacts.

Several butterfly sections may be stacked in parallel in the same way that variable capacitors are built up. In stacking these sections, the effect of adding inductances in parallel is to lower the total circuit inductance, while the addition of stators and rotors raises the total capacitance, as well as the ratio of maximum to minimum capacitance.

Butterfly circuits have been applied specifically to oscillators for transmitters, superheterodyne receivers, and heterodyne frequency meters in the 100-1000-Mc. frequency range.

Receiver Circuits The types of resonant circuits described in the previous paragraphs have largely replaced conventional coil-capacitor circuits in the range above 100 Mc. Tuned short lines and butterfly circuits are used in the range from about 100 Mc. to perhaps 3500 Mc., and above about 3500 Mc. resonant cavities are used almost exclusively. The resonant cavity is also quite generally employed in the 2000-Mc. to 3500-Mc. range.

In a properly designed receiver, thermal agitation in the first tuned circuit is amplified by subsequent tubes and predominates in the output. For good signal-to-set-noise ratio, therefore, one must strive for a high-gain low-noise r-f stage. Hiss can be held down by giving careful attention to this point. A mixer has about 0.3 of the gain of an r-f tube of the same type; so it is advisable to precede a mixer by an efficient r-f stage. It is also of some value to have good r-f selectivity before the first detector in order to reduce noises produced by beating noise at one frequency against noise at another, to produce noise at the intermediate frequency in a superheterodyne.

The frequency limit of a tube is reached when the shortest possible external connections are used as the tuned circuit, except for abnormal types of oscillation. Wires or sizeable components are often best considered as sections of transmission lines rather than as simple resistances, capacitances, or inductances.

So long as small triodes and pentodes will operate normally, they are generally preferred as v-h-f tubes over other receiving methods that have been devised. However, the input capacitance, input conductance, and transit time of these tubes limit the upper frequency at which they may be operated. The input resistance, which drops to a low value at very short wave-lengths, limits the stage gain and broadens the tuning.

V-H-F Tubes The first tube in a v-h-f receiver is most important in raising the signal above the noise generated in successive stages, for which reason small v-h-f types are definitely preferred.

Tubes employing the conventional grid-controlled and diode rectifier principles have been modernized, through various expedients, for operation at frequencies as high, in some new types, as 4000 Mc. Beyond that frequency, electron transit time becomes the limiting factor and new principles must be enlisted. In general, the improvements embodied in existing tubes have consisted of (1) reducing electrode spacing to cut down electron transit time, (2) reducing electrode areas to decrease interelectrode capacitances, and (3) shortening of electrode leads either by mounting the electrode assembly close to the tube base or by bringing the leads out directly through the glass envelope at nearby points. Through reduction of lead inductance and interelectrode capacitances, input and output resonant frequencies due to tube construction have been increased substantially.

Tubes embracing one or more of the features just outlined include the later loctal types, high-frequency acorns, button-base types, and the lighthouse types. Type 6J4 button-base triode will reach 500 Mc. Type 6F4 acorn triode is recommended for use up to 1200 Mc. Type 1A3 button-base diode has a resonant frequency of 1000 Mc., while type 9005 acorn diode resonates at 1500 Mc. Lighthouse type 2C40 can be used at frequencies up to 3500 Mc. as an oscillator.

Crystal Rectifiers More than two decades have passed since the crystal (mineral) rectifier enjoyed widespread use in radio receivers. Low-priced tubes completely supplanted the fragile and relatively insensitive crystal detector, although it did continue for a few years as a simple meter rectifier in absorption wavemeters after its demise as a receiver component.

Today, the crystal detector is of new importance in microwave communication. It is being employed as a detector and as a mixer in receivers and test instruments used at extremely high radio frequencies. At some of the frequencies employed in microwave operations, the crystal rectifier is the only satisfactory detector or mixer. The chief advantages of the crystal rectifier are very low capacitance, relative freedom from transit-time difficulties, and its two-terminal nature. No batteries or a-c power supply are required for its operation.

The crystal detector consists essentially of a small piece of silicon or germanium mounted in a base of low-melting-point alloy and contacted by means of a thin, springy feeler wire known as the *cat whisker*. This arrangement is shown in figure 27A.

The complex physics of crystal rectification is beyond the scope of this discussion. It is sufficient to state that current flows from several hundred to several thousand times more readily in one direction through the contact of cat whisker and crystal than in the opposite direction. Consequently, an alternating current (including one of microwave frequency) will be rectified by the crystal detector. The load, through which the rectified currents flow, may be connected in series or shunt with the crystal, although the former connection is most generally employed. Certain spots on the crystal surface afford more efficient rectification than others and these accordingly are searched for with the cat whisker.

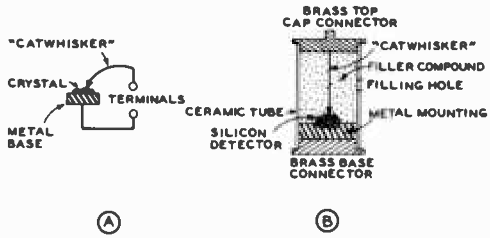


Figure 27.
THE MICROWAVE CRYSTAL DIODE.
(A) shows the essentials of a crystal diode, while (B) is a sketch to illustrate the construction of the common microwave crystal diode.

If the cat whisker is by some means permanently secured in contact with a very sensitive spot, a *fixed crystal detector* is obtained which requires no further adjustment. The basic arrangement of a modern fixed crystal detector developed during World War II for microwave work, particularly radar, is shown in figure 27B. Once the cat whisker of this unit is set at the factory to the most sensitive spot on the surface of the silicon crystal and its pressure is adjusted, a filler compound is injected through the filling hole to hold the cat whisker permanently in position.

6-10 Receiver Adjustment

A simple regenerative receiver requires little adjustment other than those necessary to insure correct tuning and smooth regeneration over some desired range. Receivers of the tuned radio-frequency type and superheterodynes require precise alignment to obtain the highest possible degree of selectivity and sensitivity.

Good results can be obtained from a receiver only when it is properly aligned and adjusted. The most practical technique for making these adjustments is given below.

Instruments A very small number of instruments will suffice to check and align a communications receiver, the most important of these testing units being a modulated oscillator and a d-c and a-c voltmeter. The meters are essential in checking the voltage applied at *each* circuit point from the power supply. If the a-c voltmeter is of the oxide-rectifier type, it can be used, in addition, as an output meter when connected across the receiver output when tuning to a modulated signal. If the signal is a steady tone, such as from a test oscillator, the output meter will indicate the value of the detected signal. In this manner, alignment results may be visually noted on the meter.

T-R-F Receiver Alignment Alignment procedure in a multistage t-r-f receiver is exactly the same as aligning a single stage. If the detector is regenerative, each preceding stage is successively aligned while keeping the detector circuit tuned to the test signal, the latter being a station signal or one locally generated by a test oscillator loosely coupled to the antenna lead. During these adjustments, the r-f amplifier gain control is adjusted for maximum sensitivity, assuming that the r-f amplifier is stable and does not oscillate. Often a sensitive receiver can be roughly aligned by tuning for maximum noise pickup.

Superheterodyne Alignment Aligning a superhet is a detailed task requiring a great amount of care and patience. It should never be undertaken without a thorough understanding of the involved job to be done and then only when there is abundant time to devote to the operation. There are no short cuts; every circuit must be adjusted individually and accurately if the receiver is to give peak performance. The precision of each adjustment is dependent upon the accuracy with which the preceding one was made.

Superhet alignment requires (1) a good signal generator (modulated oscillator) covering the radio and intermediate frequencies and equipped with an attenuator; (2) the necessary socket wrenches, screwdrivers, or "neutralizing tools" to adjust the various i-f and r-f trimmer capacitors; and (3) some convenient type of tuning indicator, such as a copper-oxide or electronic voltmeter.

Throughout the alignment process, unless specifically stated otherwise, the r-f gain control must be set for maximum output, the beat oscillator switched off, and the a.v.c. turned off or shorted out. When the signal output of the receiver is excessive, either the attenuator or the a-f gain control may be turned down, but never the r-f gain control.

I-F Alignment After the receiver has been given a rigid electrical and mechanical inspection, and any faults which may have been found in wiring or the selection and assembly of parts corrected, the i-f amplifier may be aligned as the first step in the checking operations.

With the signal generator set to give a modulated signal on the frequency at which the i-f amplifier is to operate, clip the "hot" output lead from the generator to the last i-f stage through a small fixed capacitor to the control grid. Adjust both trimmer capacitors in the last i-f transformer (the one between the last i-f amplifier and the second detector) to resonance as indicated by maximum deflection of the output meter.

Each i-f stage is adjusted in the same manner, moving the hot lead, stage by stage, back toward the front end of the receiver and backing off the attenuator as the signal strength increases in each new position. The last adjustment will be made to the first i-f transformer, with the hot signal generator lead connected to the control grid of the mixer. Occasionally it is necessary to disconnect the mixer grid lead from the coil, grounding it through a 1,000- or 5,000-ohm resistor, and coupling the signal generator through a small capacitor to the grid.

When the last i-f adjustment has been completed, it is good practice to go back through the i-f channel, re-peaking all of the transformers. It is imperative that this recheck be made in sets which do not include a crystal filter, and where the simple alignment of the i-f amplifier to the generator is final.

I-F with Crystal Filter There are several ways of aligning an i-f channel which contains a crystal-filter circuit. However, the following method is one which has been found to give satisfactory results in every case: An unmodulated signal generator capable of tuning to the frequency of the filter crystal in the receiver is coupled to the grid of the stage which precedes the crystal filter in the receiver. Then, with the crystal filter switched in, the signal generator is tuned *slowly* to find the frequency where the crystal peaks. The receiver "S" meter may be used as the indicator, and the sound heard from the loudspeaker will be of assistance in finding the point. When the frequency at which the crystal peaks has been found, all the i-f transformers in the receiver should be touched up to peak at that frequency.

B-F-O Adjustment Adjusting the beat oscillator on a receiver that has no front panel adjustment is relatively simple. It is only necessary to tune the receiver to resonance with any signal, as indicated by the tuning indicator, and then turn on the b.f.o. and set its trimmer (or trimmers) to produce the desired beat note. Setting the beat oscillator in this way will result in the beat note being stronger on one "side" of the signal than on the other, which is what is desired for c-w reception. The b.f.o. should *not* be set to "zero beat" when the receiver is tuned to resonance with the signal, as this will cause an equally strong beat to be obtained on both sides of resonance.

Front-End Alignment Alignment of the front end of a home-constructed receiver is a relatively simple process, consisting of first getting the oscillator to cover the desired frequency range and then of peaking the various r-f circuits for maximum gain. However, if the frequency range covered by the receiver is very wide a fair amount of cut and try will be required to obtain satisfactory tracking between the r-f circuits and the oscillator. Manufactured communications receivers should always be tuned in accordance with the instructions given in the maintenance manual for the receiver.

CHAPTER SEVEN

Generation of Radio-Frequency Energy

A radio communication or broadcast transmitter consists of a source of radio frequency power, or *carrier;* a system for *modulating* the carrier whereby voice or telegraph keying or other modulation is superimposed upon it; and an antenna system, including feed line, for *radiating* the intelligence-carrying radio frequency power. The power supply employed to convert primary power to the various voltages required by the r-f and modulator portions of the transmitter may also be considered part of the transmitter. Power supplies are treated separately in Chapter Twenty-five.

Voice modulation usually is accomplished by varying either the amplitude or the frequency of the radio frequency carrier in accordance with the components of intelligence to be transmitted. The process of amplitude modulation is covered in detail in Chapter Eight and frequency modulation is covered in Chapter Nine.

Radiotelegraph modulation (keying) normally is accomplished either by interrupting, shifting the frequency of, or superimposing an audio tone on the radio-frequency carrier in accordance with the dots and dashes to be transmitted.

The complexity of the radio-frequency generating portion of the transmitter is dependent upon the power, order of stability, and frequency desired. An oscillator feeding an antenna directly is the simplest form of radio-frequency generator. A modern high-frequency transmitter, on the other hand, is a very complex generator. Such an equipment usually comprises a very stable crystal-controlled or self-controlled oscillator to stabilize the output frequency, a series of frequency multipliers, one or more amplifier stages to increase the power up to the level which is desired for feeding the antenna system, and a filter system for keeping the harmonic energy generated in the transmitter from being fed to the antenna system.

7-1 Self-Controlled Oscillators

In Chapter Five, it was explained that the amplifying properties of a tube having three or more elements give it the ability to generate an alternating current of a frequency determined by the components associated with it. A vacuum tube operated in such a circuit is called an oscillator, and its function is essentially to convert direct current into radio-frequency alternating current of a predetermined frequency.

Oscillators for controlling the frequency of conventional radio transmitters can be divided into two general classes: self-controlled and crystal-controlled.

There are a great many types of self-controlled oscillators, each of which is best suited to a particular application. They can further be subdivided into the classifications of: negative-grid oscillators, electron-orbit oscillators, negative-resistance oscillators, velocity modulation oscillators, and magnetron oscillators.

Negative-Grid Oscillators A negative-grid oscillator is essentially a vacuum-tube amplifier with a sufficient portion of the output energy coupled back into the input circuit to sustain oscillation. The control grid is biased negatively with respect to the cathode. Common types of negative-grid oscillators are diagrammed in figure 1.

The Hartley Illustrated in figure 1(A) is the oscillator circuit which finds the most general application at the present time; this circuit is commonly called the Hartley. The operation of this oscillator will be described as an index to the operation of all negative-grid oscillators; the only real difference between the various circuits is the manner in which energy for excitation is coupled from the plate to the grid circuit.

When plate voltage is applied to the Hartley oscillator shown at (A), the sudden flow of plate current accompanying the application of plate voltage will cause an electro-magnetic field to be set up in the vicinity of the coil. The building-up of this field will cause a potential drop to appear from turn-to-turn along the coil. Due to the inductive coupling between the portion of the coil in which the plate current is flowing and the grid portion, a potential will be induced in the grid portion.

Since the cathode tap is between the grid and plate ends of the coil, the induced grid voltage acts in such manner as to increase further the plate current to the tube. This action will continue for a short period of time determined by the inductance and capacitance of the tuned circuit, until the "flywheel" effect of the tuned circuit causes this action to come to a maximum and then to reverse itself. The plate current then decreases, the magnetic field around the coil also decreasing, until a minimum is reached, when the action starts again in the original direction and at a greater amplitude than before. The amplitude of these oscillations, the frequency of which is determined by the coil capacitor circuit, will increase in a very short period of time to a limit determined by the plate voltage of the oscillator tube.

The Colpitts Figure 1 (B) shows a version of the Colpitts oscillator. It can be seen that this is essentially the same circuit as the Hartley except that the ratio of a pair of capacitances in series determines the effective cathode tap, instead of actually using a tap on the tank coil. Also, the net capacitance of these two capacitors comprises the tank capacitance of the tuned circuit. This oscillator circuit is somewhat less susceptible to parasitic (spurious) oscillations than the Hartley.

For best operation of the Hartley and Colpitts oscillators, the voltage from grid to cathode, determined by the tap on the coil or the setting of the two capacitors, normally should be from 1/3 to 1/5 that appearing between plate and cathode.

The T.P.T.G. The tuned-plate tuned-grid oscillator illustrated at (C) has a tank circuit in both the plate and grid circuits. The feedback of energy from the plate to the grid circuits is accomplished by the plate-to-grid inter-electrode capacitance within the tube. The necessary phase reversal in feedback voltage is provided by tuning the grid tank capacitor to the low side of the desired frequency and the plate capacitor to the high side. A broadly resonant coil may be substituted for the grid tank to form the *T.N.T.* oscillator shown at (D).

Electron-Coupled Oscillators In any of the oscillator circuits just described it is possible to take energy from the oscillator circuit by coupling an external load to the tank circuit. Since the tank circuit determines the frequency of oscillation of the tube, any variations in the conditions of the external circuit will be coupled back into the frequency determining portion of the oscillator. These variations will result in frequency instability.

The frequency determining portion of an oscillator may be coupled to the load circuit only by an electron stream, as illustrated in (E) and (F) of figure 1. When it is considered that the screen of the tube acts as the plate to the oscillator circuit, the plate merely acting as a coupler to the load, then the similarity between the cathode-grid-screen circuit of these oscillators and the cathode-grid-plate circuits of the corresponding prototype can be seen.

The electron-coupled oscillator has good

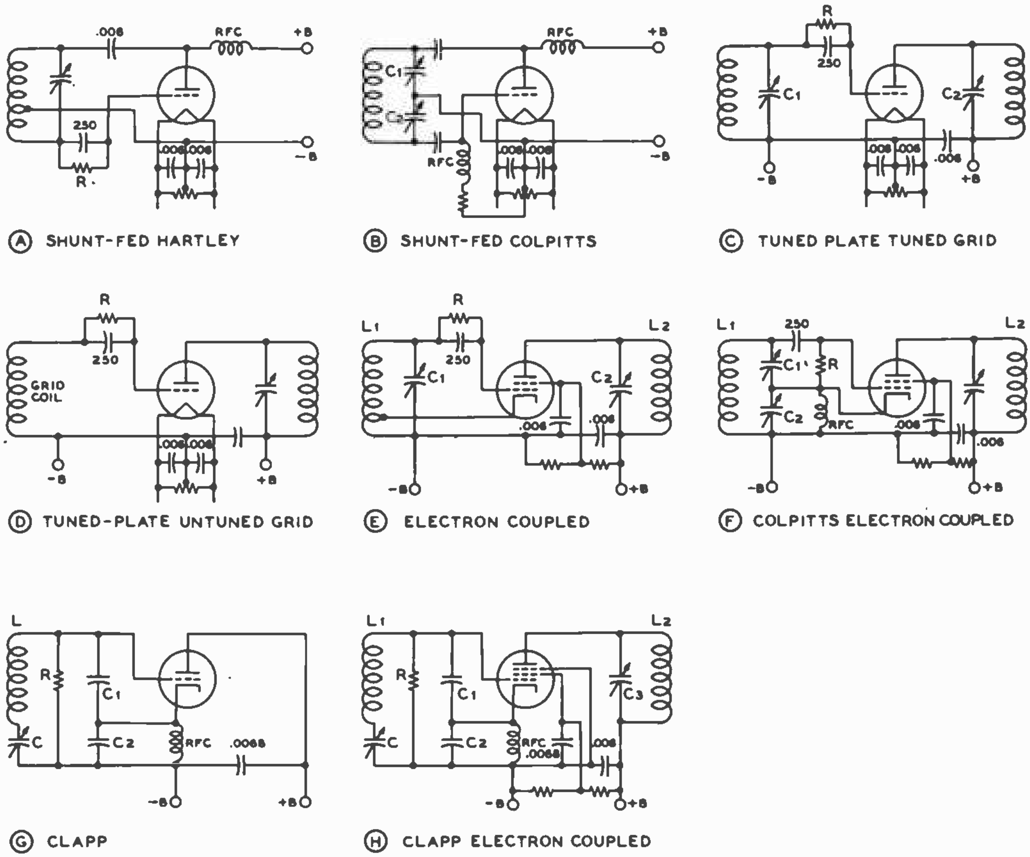


Figure 1.

COMMON TYPES OF SELF-EXCITED OSCILLATORS.

Fixed capacitor values are typical, but will vary somewhat with the application. In the Clapp oscillator circuits of (G) and (H), capacitors C_1 and C_2 should have a reactance of 50 to 100 ohms at the operating frequency of the oscillator. Tuning of these two oscillators is accomplished by capacitor C. In the circuits of (E), (F), and (H), tuning of the tank circuit in the plate of the oscillator tube will have relatively small effect on the frequency of oscillation. The plate tank circuit also may, if desired, be tuned to a harmonic of the oscillation frequency, or a broadly resonant circuit may be used in this circuit position.

stability with respect to load and voltage variation. Load variations have a relatively small effect on the frequency, since the only coupling between the oscillating circuit and the load is through the electron stream flowing through the other elements to the plate. The plate is electrostatically shielded from the oscillating portion by the bypassed screen.

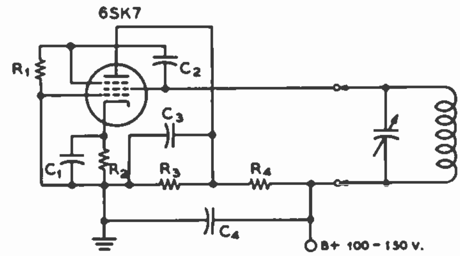
The stability of the e.c.o. with respect to variations in supply voltages is explained as follows: The frequency will shift in one direction with an increase in screen voltage, while an increase in plate voltage will cause it to shift in the other direction. By a proper proportioning of the resistors that comprise the voltage divider supplying screen voltage, it is possible to make the frequency of the oscillator substantially independent of supply voltage variations.

The Clapp Oscillator A relatively new type of oscillator circuit which is capable of giving excellent frequency stability is illustrated in figure 1G. Comparison between the more standard circuits of figure 1A through 1F and the Clapp oscillator circuits of figures 1G and 1H will immediately show one marked difference: the tuned circuit which controls the operating frequency in the Clapp oscillator is *series* resonant, while in all the more standard oscillator circuits the frequency controlling circuit is parallel resonant. Also, the capacitors C_1 and C_2 are relatively large in terms of the usual values for a Colpitts oscillator. In fact, the value of capacitors C_1 and C_2 will be in the vicinity of 0.001 μfd. to 0.0025 μfd. for an oscillator which is to be operated in the 1.8-Mc. band.

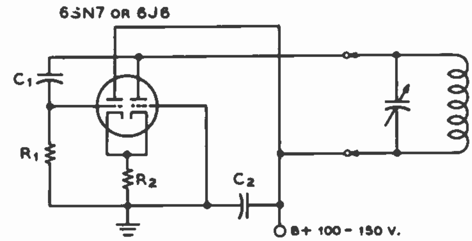
The Clapp oscillator operates in the following manner: at the resonant frequency of the oscillator tuned circuit (L, C) the impedance of this circuit is at minimum (since it operates in series resonance) and maximum current flows through it. Note however, that C_1 and C_2 also are included within the current path for the series resonant circuit, so that at the frequency of resonance an appreciable voltage drop appears across these capacitors. The voltage drop appearing across C_1 is applied to the grid of the oscillator tube as excitation, while the amplified output of the oscillator tube appears across C_2 as the driving power to keep the circuit in oscillation.

Capacitors C_1 and C_2 should be made as large in value as possible, while still permitting the circuit to oscillate over the full tuning range of C. The larger these capacitors are made, the smaller will be the coupling between the oscillating circuit and the tube, and consequently the better will be oscillator stability with respect to tube variations. High G_m tubes such as the 6AC7, 6AG7, and 6CB6 will permit the use of larger values of capacitance at C_1 and C_2 than will more conventional tubes such as the 6SJ7, 6V6, and such types. In general it may be said that the reactance of capacitors C_1 and C_2 should be on the order of 40 to 120 ohms at the operating frequency of the oscillator—with the lower values of reactance going with high-G_m tubes and the higher values being necessary to permit oscillation with tubes having G_m in the range of 2000 micromhos such as the 6SJ7.

It will be found that the Clapp oscillator will have a tendency to vary in power output over the frequency range of tuning capacitor C. The output will be greatest where C is at its largest setting, and will tend to fall off with C at minimum capacitance. In fact, if capacitors C_1 and C_2 have too large a value the circuit will stop oscillation near the ***minimum*** capacitance setting of C. Hence it will be necessary to use a slightly *smaller* value of capacitance at C_1 and C_2 (to provide an increase in the capacitive reactance at this point), or else the frequency range of the oscillator must be restricted by paralleling a fixed capacitor across C so that its effective capacitance at minimum setting will be increased to a value which will sustain oscillation.



(A) TRANSITRON OSCILLATOR



(B) CATHODE COUPLED OSCILLATOR

Figure 2.

TWO-TERMINAL OSCILLATOR CIRCUITS.

Both circuits may be used for an audio oscillator or for frequencies into the v-h-f range simply by placing a tank circuit tuned to the proper frequency where indicated on the drawing. Recommended values for the components are given below for both oscillators.

TRANSITRON OSCILLATOR

C_1—0.01-μfd. mica for r.f. 10-μfd. elect. for a.f.
C_2—0.00005-μfd. mica for r.f. 0.1-μfd. paper for a.f.
C_3—0.003-μfd. mica for r.f. 0.5-μfd. paper for a.f.
C_4—0.01-μfd. mica for r.f. 8-μfd. elect. for a.f.
R_1—220K ½-watt carbon
R_2—1800 ohms ½-watt carbon
R_3—22K 2-watt carbon
R_4—22K 2-watt carbon

CATHODE-COUPLED OSCILLATOR

C_1—0.00005-μfd. mica for r.f. 0.1-μfd. paper for audio
C_2—0.003-μfd. mica for r.f. 8-μfd. elect. for audio
R_1—47K ½-watt carbon
R_2—1K 1-watt carbon

In the triode Clapp oscillator, such as shown at figure 1G, output voltage for excitation of an amplifier, doubler, or isolation stage normally is taken from the cathode of the oscillator tube by capacitive coupling to the grid of the next tube. However, where greater isolation of succeeding stages from the oscillating circuit is desired, the electron-coupled Clapp oscillator diagrammed in figure 1H may be used. Output then may be taken from the plate circuit of the tube by capacitive coupling with either a tuned circuit, as shown, or with an r-f choke or a broadly resonant circuit in the plate return. Alternatively, energy may be coupled from the output circuit L_2-C_3 by link coupling. The considerations with regard to C_1, C_2, and the grid tuned circuit are the same as for the triode oscillator arrangement of figure 1G.

Negative Resistance Oscillators Negative-resistance oscillators often are used when unusually high frequency stability is desired, as in a frequency meter. The *dynatron* of a few years ago and the newer *transitron* are examples of oscillator circuits which make use of the negative resistance characteristic between different elements in some multi-grid tubes.

In the dynatron, the negative resistance is a consequence of secondary emission of electrons from the plate of a tetrode tube. By a proper proportioning of the electrode voltage, an increase in screen voltage will cause a decrease in screen current, since the increased screen voltage will cause the screen to attract a larger number of the secondary electrons emitted by the plate. Since the net screen current flowing from the screen supply will be decreased by an increase in screen voltage, it is said that the screen circuit presents a negative resistance.

If any type of tuned circuit, or even a resistance-capacitance circuit, is connected in series with the screen, the arrangement will oscillate —provided, of course, that the external circuit impedance is greater than the negative resistance. A negative resistance effect similar to the dynatron is obtained in the *transitron* circuit, which uses a pentode with the suppressor coupled to the screen. The negative resistance in this case is obtained from a combination of secondary emission and inter-electrode coupling, and is considerably more stable than that obtained from uncontrolled secondary emission alone in the dynatron. A representative transitron oscillator circuit is shown in figure 2.

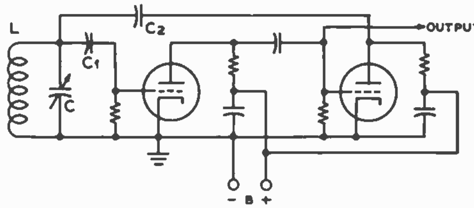


Figure 3.

THE FRANKLIN OSCILLATOR CIRCUIT.

A separate phase inverter tube is used in this oscillator to feed a portion of the output back to the input in the proper phase to sustain oscillation. The values of C_1 and C_2 should be as small as will permit oscillations to be sustained over the desired frequency range.

The chief distinction between a conventional "negative grid" oscillator and a "negative resistance" oscillator is that in the former the tank circuit must act as a phase inverter in order to permit the amplification of the tube to act as a negative resistance, while in the latter the tube acts as its own phase inverter. Thus a negative resistance oscillator requires only an untapped coil and a single capacitor as the frequency determining tank circuit, and is classed as a "two terminal" oscillator. In fact, the time constant of an R/C circuit may be used as the frequency determining element and such an oscillator is rather widely used as a tunable audio frequency oscillator.

The Franklin Oscillator The Franklin oscillator makes use of two cascaded tubes to obtain the negative-resistance effect (figure 3). The tubes may be either a pair of triodes, tetrodes, or pentodes, a dual triode, or a combination of a triode and a multi-grid tube. The chief advantage of this oscillator circuit is that the frequency determining tank only has two terminals, and one side of the circuit is grounded.

The second tube acts as a phase inverter to give an effect similar to that obtained with the dynatron or transitron, except that the effective transconductance is much higher. If the tuned circuit is omitted or is replaced by a resistor, the circuit becomes a relaxation oscillator or a multivibrator.

Oscillator Stability The Clapp oscillator has proved to be inherently the most stable of all the oscillator circuits

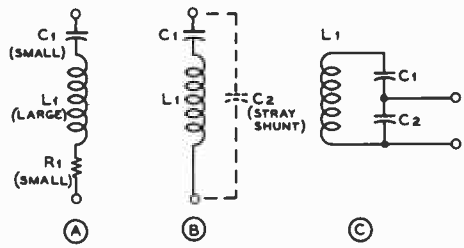


Figure 4.

EQUIVALENT ELECTRICAL CIRCUIT OF QUARTZ PLATE IN A HOLDER.

At (A) is shown the equivalent series-resonant circuit of the crystal itself, at (B) is shown how the shunt capacitance of the holder electrodes and associated wiring affects the circuit to the combination circuit of (C) which exhibits both series resonance and parallel resonance (anti-resonance), the separation in frequency between the two modes being very small and determined by the ratio of C_1 to C_2.

discussed above, since minimum coupling between the oscillator tube and its associated tuned circuit is possible. However, this inherently good stability is with respect to tube variations; instability of the tuned circuit with respect to vibration or temperature will of course have as much effect on the frequency of oscillation as with any other type of oscillator circuit. Solid mechanical construction of the components of the oscillating circuit, along with a small negative-coefficient compensating capacitor included as an element of the tuned circuit, usually will afford an adequate degree of oscillator stability.

V.F.O. Transmitter Controls When used to control the frequency of a transmitter in which there are stringent limitations on frequency tolerance, several precautions are taken to ensure that a variable frequency oscillator will stay on frequency. The oscillator is fed from a voltage regulated power supply, uses a well designed and temperature compensated tank circuit, is of rugged mechanical construction to avoid the effects of shock and vibration, is protected against excessive changes in ambient room temperature, and is isolated from feedback or stray coupling from other portions of the transmitter by shielding, filtering of voltage supply leads, and incorporation of one or more "buffer" amplifier stages. In a high power transmitter a small amount of stray coupling from the final amplifier to the oscillator can produce appreciable degradation of the oscillator stability if both are on the same frequency. Therefore, the oscillator usually is operated on a subharmonic of the transmitter output frequency, with one or more frequency multipliers between the oscillator and final amplifier.

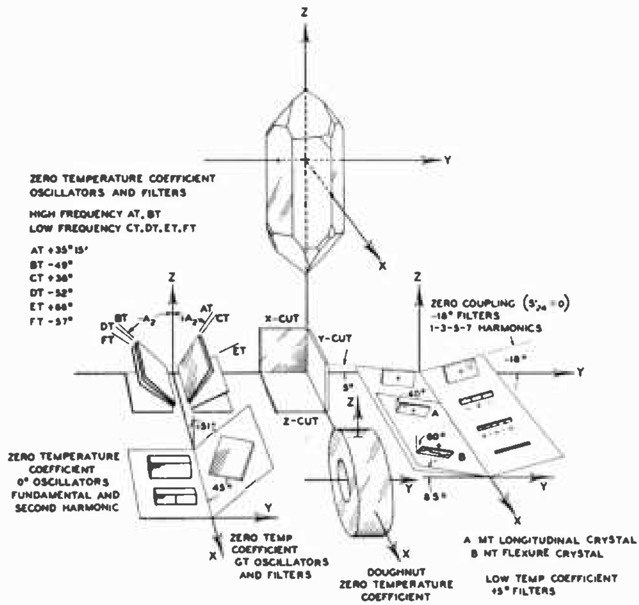
7-2 Quartz Crystal Oscillators

Quartz is a naturally occurring crystal having a structure such that when plates are cut in certain definite relationships to the crystallographic axes, these plates will show the *piezo-electric* effect—the plates will be deformed in the influence of an electric field, and, conversely, when such a plate is compressed or deformed in any way a potential difference will appear upon its opposite sides.

The crystal has mechanical resonance, and will vibrate at a very high frequency because of its stiffness, the natural period of vibration depending upon the dimensions, the method of electrical excitation, and crystallographic orientation. Because of the piezo-electric properties, it is possible to cut a quartz plate which, when provided with suitable electrodes, will have the characteristics of a series resonant circuit with a very high L/C ratio and very high Q. The Q is several times as high as can be obtained with an inductor-capacitor combination in conventional physical sizes. The equivalent electrical circuit is shown in figure 4A, the resistance component simply being an acknowledgment of the fact that the Q, while high, does not have an infinite value.

The shunt capacitance of the electrodes and associated wiring (crystal holder and socket, plus circuit wiring) is represented by the dotted portion of figure 4B. In a high frequency crystal this will be considerably greater than the capacitance component of an equivalent series L/C circuit, and unless the shunt capacitance is balanced out in a bridge circuit, the crystal will exhibit both resonant (series resonant) and anti-resonant (parallel resonant) frequencies, the latter being slightly higher than the series resonant frequency and approaching it as C_2 is increased.

The series resonance characteristic is employed in crystal filter circuits in receivers, as covered in Chapter 6, and also in certain oscillator circuits wherein the crystal is used



Courtesy Bell System Technical Journal

Figure 5.
ORIENTATION OF THE COMMON CRYSTAL CUTS.

as a selective feedback element in such a manner that the phase of the feedback is correct and the amplitude adequate only at or very close to the series resonant frequency of the crystal.

While quartz, tourmaline, Rochelle salts, ADP, and EDT crystals all exhibit the piezoelectric effect, quartz is the material widely employed for frequency control.

As the cutting and grinding of quartz plates has progressed to a high state of development and these plates may be purchased at prices which discourage the cutting and grinding by simple hand methods for one's own use, the procedure will be only lightly touched upon here.

The crystal "blank" is cut from the raw quartz at a predetermined orientation with respect to the optical and electrical axes, the orientation determining the activity, temperature coefficient, thickness coefficient, and other characteristics. Various orientations or "cuts" having useful characteristics are illustrated in figure 5.

The crystal blank is then rough-ground almost to frequency, the frequency increasing in inverse ratio to the oscillating dimension (usually the thickness). It is then finished to exact frequency either by careful lapping, by etching, or plating. The latter process consists of finishing it to a frequency slightly higher than that desired and then silver plating the electrodes right on the crystal, the frequency decreasing as the deposit of silver is increased. If the crystal is not etched, it must be carefully scrubbed and "baked" several times to stabilize it, or otherwise the frequency and activity of the crystal will change with time. Irradiation by X-rays recently has been used in crystal finishing.

Unplated crystals usually are mounted in "pressure" or "clamped" holders, in which two electrodes are held against the crystal faces under slight pressure. Unplated crystals also are sometimes mounted in an "air gap" holder, in which there is a very small gap between the crystal and one or both electrodes. By making this gap variable, the frequency of the crystal may be altered over narrow limits (about 0.3% for certain types).

The temperature coefficient of frequency for various crystal cuts of the "-T" rotated family is indicated in figure 5. These angles are typical, but crystals of a certain cut will vary slightly. By controlling the orientation and dimensioning, the "turning point" (point of zero temperature coefficient) for a BT cut plate may be made either lower or higher than the 75 degrees shown. Also, by careful control of axes and dimensions, it is possible to get AT cut crystals with a very flat temperature-frequency characteristic.

The first quartz plates used were either Y cut or X cut. The former had a very high temperature coefficient which was discontinuous, causing the frequency to jump at certain critical temperatures. The X cut had a moderately bad coefficient, but it was more continuous, and by keeping the crystal in a temperature controlled oven, a high order of stability could be obtained. However, the X cut crystal was considerably less active than the Y cut, especially in the case of poorly ground plates.

For frequencies between 500 kc. and about 6 Mc., the AT cut crystal now is the most widely used. It is active, can be made free from spurious responses, and has an excellent temperature characteristic. However, above about 6 Mc. it becomes quite thin, and a difficult production job. Between 6 Mc. and about 12 Mc., the BT cut plate is widely used. It also works well between 500 kc. and 6 Mc., but the AT cut is more desirable when a high order of stability is desired and no crystal oven is employed.

For low frequency operation on the order of 100 kc., such as is required in a frequency standard, the GT cut crystal is recommended, though CT and DT cuts also are widely used for applications between 50 and 500 kc. The CT, DT, and GT cut plates are known as ***contour*** cuts, as these plates oscillate along the long dimension of the plate or "bar," and are much smaller physically than would be the case for a regular AT or BT cut crystal for the same frequency.

Crystal Holders Crystals normally are purchased ready mounted. The best type mount is determined by the type crystal and its application, and usually an optimum mounting is furnished with the crystal. However, certain features are desirable in all holders. One of these is exclusion of moisture and prevention of electrode oxidization. The best means of accomplishing this is a metal holder, hermetically sealed, with glass insulation and a metal-to-glass bond. However, such holders are more expensive, and a ceramic or phenolic holder with rubber gasket will serve where requirements are not too exacting.

Temperature Control; Crystal Ovens Where the frequency tolerance requirements are not too stringent and the ambient temperature does not include extremes, an AT-cut plate, or a BT-cut plate with optimum (mean temperature) turning point, will often provide adequate stability without resorting to a temperature controlled oven. However, for broadcast stations and other applications where very close tolerances must be maintained, a thermostatically controlled oven, adjusted for a temperature slightly higher than the highest ambient likely to be encountered, must of necessity be employed.

Harmonic Cut Crystals Just as a vibrating string can be made to vibrate on its harmonics, a quartz crystal will exhibit mechanical resonance (and therefore electrical resonance) at harmonics of its fundamental frequency. When employed in the usual holder, it is possible to excite the crystal only on its odd harmonics (overtones).

By grinding the crystal especially for harmonic operation, it is possible to enhance its operation as a harmonic resonator. BT and AT cut crystals designed for optimum operation on the 3d, 5th and even the 7th harmonic are available. The 5th and 7th harmonic types, especially the latter, require special holder and oscillator circuit precautions for satisfactory operation, but the 3d harmonic type needs little more consideration than a regular fundamental type. A crystal ground for optimum operation on a particular harmonic may or may not be a good oscillator on a different harmonic or on the fundamental. One interesting characteristic of a harmonic cut crystal is that its harmonic frequency is not quite an exact multiple of its fundamental, though the disparity is very small.

The harmonic frequency for which the crystal was designed is the "working frequency." It is not the "fundamental," but the crystal itself actually oscillates on this "working frequency" when it is functioning in the proper manner.

When a harmonic-cut crystal is employed, a selective tuned circuit must be employed somewhere in the oscillator in order to discriminate against the fundamental frequency or undesired harmonics. Otherwise the crystal might not always oscillate on the intended frequency. For this reason the Pierce oscillator, later described in this chapter, is not suitable for use with harmonic-cut crystals, because the only tuned element in this oscillator circuit is the crystal itself.

Crystal Current; Heating and Fracture For a given crystal operating as an anti-resonant tank in a given oscillator at fixed load impedance and plate and screen voltages, the r-f current through the crystal will increase as the shunt capacitance C_2 of figure 4 is increased, because this effectively increases the "step up ratio" of C_1 to C_2. For a given shunt capacitance, C_2, the crystal current for a given crystal is directly proportional to the r-f voltage across C_2. This voltage may be measured by means of a vacuum tube voltmeter having a low input capacitance, and such a measurement is a more pertinent one than a reading of r-f current by means of a thermogalvanometer inserted in series with one of the leads to the crystal holder.

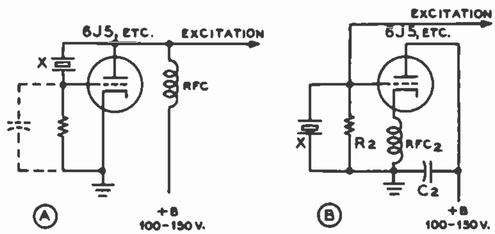
The function of a crystal is to provide accurate frequency control, and unless it is used in such a manner as to take advantage of its inherent high stability, there is no point in using a crystal oscillator. For this reason a crystal oscillator should not be run at high plate input in an attempt to obtain considerable power directly out of the oscillator, as such operation will cause the crystal to heat, with resultant frequency drift and possible fracture.

7-3 Crystal Oscillator Circuits

Considerable confusion exists as to nomenclature of crystal oscillator circuits, due to a tendency to name a circuit after its discoverer. Nearly all the basic crystal oscillator circuits were either first used or else developed independently by G. W. Pierce, but he has not been so credited in all the literature.

Use of the crystal oscillator in master oscillator circuits in radio transmitters dates back to about 1924 when the first application articles appeared.

The Pierce Oscillator The circuit of figure 6A is the simplest crystal oscillator circuit. It is one of those developed by Pierce, and is generally known among amateurs



BASIC "PIERCE" OSCILLATOR HOT-CATHODE "PIERCE" OSCILLATOR

Figure 6.

THE PIERCE CRYSTAL OSCILLATOR CIRCUIT.

Shown at (A) is the basic Pierce crystal oscillator circuit. A capacitance of 10 to 75 $\mu\mu$fd. normally will be required at C_1 for optimum operation. If a plate supply voltage higher than indicated is to be used, RFC$_1$ may be replaced by a 22,000-ohm 2-watt resistor. Shown at (B) is an alternative arrangement with the r-f ground moved to the plate, and with the cathode floating. This alternative circuit has the advantage that the full r-f voltage developed across the crystal may be used as excitation to the next stage, since one side of the crystal is grounded.

as the "Pierce oscillator." The crystal simply replaces the tank circuit in a Colpitts or ultra-audion oscillator. The r-f excitation voltage available to the next stage is low, being somewhat less than that developed across the crystal. Capacitor C_1 will make more of the voltage across the crystal available for excitation, and sometimes will be found necessary to ensure oscillation. Its value is small, usually approximately equal to or slightly greater than the stray capacitance from the plate circuit to ground (including the grid of the stage being driven).

If the r-f choke has adequate inductance, a crystal (even a harmonic cut crystal) will almost invariably oscillate on its fundamental. The Pierce oscillator therefore cannot be used with harmonic cut crystals.

The circuit at (B) is the same as that of (A) except that the plate instead of the cathode is operated at ground r-f potential. All of the r-f voltage developed across the crystal is available for excitation to the next stage, but still is low for reasonable values of crystal current. For best operation a tube with low heater-cathode capacitance is required. Excitation for the next stage may also be taken from the cathode when using this circuit.

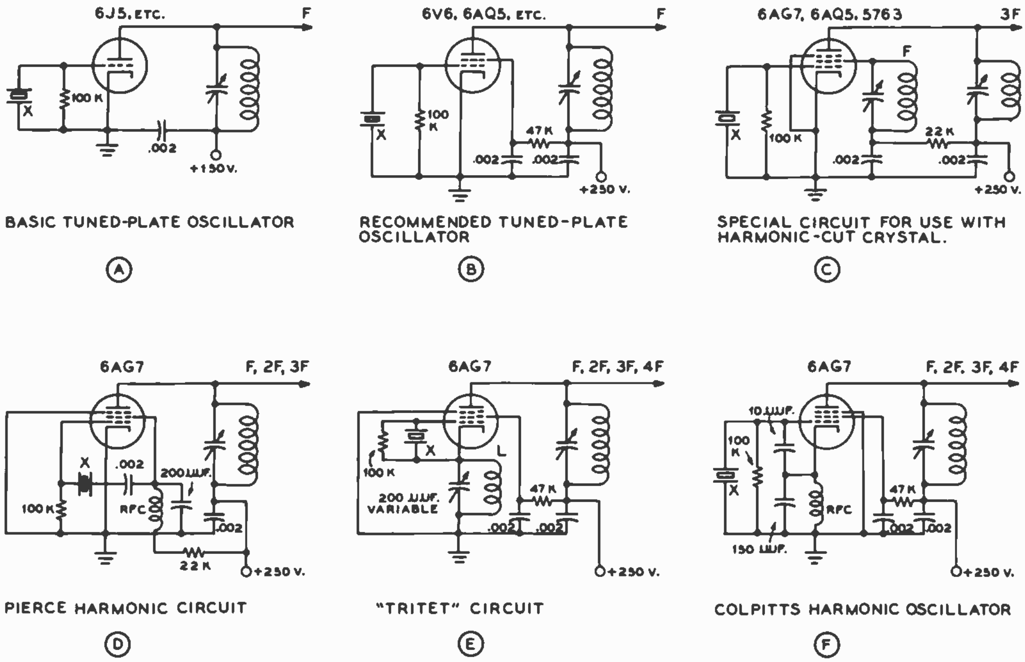


Figure 7.

COMMONLY USED CRYSTAL OSCILLATOR CIRCUITS.

Shown at (A) is the basic tuned-plate crystal oscillator with a triode oscillator tube. The plate tank must be tuned on the low-capacitance side of resonance to sustain oscillation. (B) shows the tuned-plate oscillator as it is normally used, with an a-f power pentode to permit high output with relatively low crystal current.

Schematics (C), (D), (E), and (F) illustrate crystal oscillator circuits which can deliver moderate output energy on harmonics of the oscillating frequency of the crystal. (C) shows a special circuit which will permit use of a harmonic-cut crystal to obtain output energy well into the v-h-f range. (D) is valuable when extremely low crystal current is a requirement, but delivers relatively low output. (E) is commonly used, but is subject to crystal damage if the cathode circuit is mis-tuned. (F) is recommended as the most generally satisfactory from the standpoints of: low crystal current regardless of mis-adjustment, good output on harmonic frequencies, one side of crystal is grounded, will oscillate with crystals from 1.5 to 10 Mc. without adjustment, output tank may be tuned to the crystal frequency for fundamental output without stopping oscillation or changing frequency.

Tuned-Plate Crystal Oscillator The circuit shown in figure 7A is also one used by Pierce, but is more widely referred to as the "Miller" oscillator. To avoid confusion, we shall refer to it as the "tuned-plate" crystal oscillator. It is essentially an Armstrong or "tuned plate-tuned grid" oscillator with the crystal replacing the usual L-C grid tank. The plate tank must be tuned to a frequency slightly higher than the anti-resonant (parallel resonant) frequency of the crystal. Whereas the Pierce circuits of figure 6 will oscillate at (or very close to) the anti-resonant frequency of the crystal, the circuits of figure 7 will oscillate at a frequency a little above the anti-resonant frequency of the crystal.

The diagram shown in figure 7A is the basic circuit. The most popular version of the tuned-plate oscillator employs a pentode or beam tetrode with cathode bias to prevent excessive plate dissipation when the circuit is not oscillating. The cathode resistor is optional. Its omission will reduce both crystal current and oscillator efficiency, resulting in somewhat more output for a given crystal current. The tube usually is an audio or video type beam pentode or tetrode, the plate-grid capacitance of such tubes being sufficient to ensure stable oscillation but not so high as to offer excessive feedback with resulting high crystal current. The 6V6 makes an excellent all-around tube for this type circuit.

Harmonic Crystal Oscillator Circuits The usual type of crystal-controlled h-f transmitter operates, at least part of the time, on a frequency which is an integral multiple of the operating frequency of the controlling crystal. Hence, oscillator circuits which are capable of providing output on the crystal frequency if desired, but which also can deliver output energy on harmonics of the crystal frequency have come into wide use. Four such circuits which have found wide application are illustrated in figures 7C, 7D, 7E, and 7F.

The circuit shown in figure 7C is recommended for use with harmonic-cut crystals when output is desired on a multiple of the oscillating frequency of the crystal. As an example, a 25-Mc. harmonic-cut crystal may be used in this circuit to obtain output on 50 Mc., or a 48-Mc. harmonic-cut crystal may be used to obtain output on the 144-Mc. amateur band. The circuit is not recommended for use with the normal type of fundamental-frequency crystal since more output with fewer variable elements can be obtained with the circuits of 7D and 7F.

The Pierce-harmonic circuit shown in figure 7D is satisfactory for many applications which require very low crystal current, but has the disadvantage that both sides of the crystal are above ground potential. The Tri-tet circuit of figure 7E is widely used and can give excellent output with low crystal current. However, the circuit has the disadvantages of requiring a cathode coil, of requiring careful setting of the variable cathode capacitor to avoid damaging the crystal when changing frequency ranges, and of having both sides of the crystal above ground potential.

The Colpitts harmonic oscillator of figure 7F is recommended as being the most generally satisfactory harmonic crystal oscillator circuit since it has the following advantages: (1) the circuit will oscillate with crystals over a very wide frequency range with no change other than plugging in or switching in the desired crystal; (2) crystal current is extremely low; (3) one side of the crystal is grounded, which facilitates crystal-switching circuits; (4) the circuit will operate straight through without frequency pulling, or it may be operated with output on the second, third, or fourth harmonic of the crystal frequency.

Crystal Oscillator Tuning The tunable circuits of all oscillators illustrated should be tuned for maximum output as indicated by maximum excitation to the following stage, except that the oscillator plate tank of tuned-plate oscillators (figure 7A and figure 7B) should be backed off slightly towards the low capacitance side from maximum output, as the oscillator then is in a more stable condition and sure to "take off" immediately when power is applied. This is especially important when the oscillator is keyed, as for break-in c-w operation.

Crystal Switching It is desirable to keep stray shunt capacitances in the crystal circuit as low as possible, regardless of the oscillator circuit. If a selector switch is used, this means that both switch and crystal sockets must be placed close to the oscillator tube socket. This is especially true of harmonic-cut crystals operating on a comparatively high frequency. In fact, on the highest frequency crystals it is preferable to use a turret arrangement for switching, as the stray capacitances can be kept lower.

Crystal Oscillator Keying When the crystal oscillator is keyed, it is necessary that crystal activity and oscillator-tube transconductance be moderately high, and that oscillator loading and crystal shunt capacitance be low. Below 2500 kc. and above 6 Mc. these considerations become especially important. Keying of the plate voltage (in the negative lead) of a crystal oscillator, with the screen voltage regulated at about 150 volts, has been found to give satisfactory results.

7-4 Radio Frequency Amplifiers

The output of the oscillator stage in a transmitter (whether it be self-controlled or crystal controlled) must be kept down to a fairly low level to maintain stability and to maintain a factor of safety from fracture of the crystal when one is used. The low power output of the oscillator is brought up to the desired power level by means of radio-frequency amplifiers. The two classes of r-f amplifiers that find widest application in radio transmitters are the Class B and Class C types.

Methods for determining the correct operating conditions for various types of radio-frequency amplifiers are discussed in detail, with illustrative examples, in Chapter 5.

The Class B Amplifier Class B amplifiers are used in a radio-telegraph transmitter when maximum power gain and minimum harmonic output is desired in a particular stage. A Class B amplifier operates with cutoff bias and a comparatively small amount of excitation. Power gains of 20 to 200 or so are obtainable in a well-designed Class B amplifier. The plate efficiency of a Class B c-w amplifier will run around 65 per cent.

The Class B Linear Another type of Class B amplifier is the Class B linear stage as employed in radiophone work. This type of amplifier is used to increase the level of a modulated carrier wave, and depends for its operation upon the linear relation between excitation voltage and output voltage. Or, to state the fact in another manner, the power output of a Class B linear stage varies linearly with the square of the excitation voltage.

The Class B linear amplifier is operated with cutoff bias and a small value of excitation, the actual value of exciting power being such that the power output under carrier conditions is one-fourth of the peak power capabilities of the stage. Class B linears are very widely employed in broadcast and commercial installations, but are comparatively uncommon in amateur application, since tubes with high plate dissipation are required for moderate output. The carrier efficiency of such an amplifier will vary from approximately 30 per cent to 35 per cent.

The Class C Amplifier Class C amplifiers are very widely used in all types of transmitters. Good power gain may be obtained (values of gain from 3 to 20 are common) and the plate circuit efficiency may be, under certain conditions, as high as 85 per cent. Class C amplifiers operate with considerably more than cutoff bias and ordinarily with a large amount of excitation as compared to a Class B amplifier. The bias for a normal Class C amplifier is such that plate current on the stage flows for approximately 120° of the 360° excitation cycle. Class C amplifiers are used in transmitters where a fairly large amount of excitation power is available and good plate circuit efficiency is desired.

Plate Modulated Class C The characteristic of a Class C amplifier which makes it linear with respect to changes in plate voltage is that which allows such an amplifier to be *plate modulated* for radiotelephony. Through the use of higher bias than is required for a c-w Class C amplifier and greater excitation, the linearity of such an amplifier may be extended from zero plate voltage to twice the normal value. The output power of a Class C amplifier, adjusted for plate modulation, varies with the square of the plate voltage. This is the same condition that would take place if a resistor equal to the voltage on the amplifier, divided by its plate current, were substituted for the amplifier. Therefore, the stage presents a resistive load to the modulator.

Grid Modulated Class C If the grid current to a Class C amplifier is reduced to a low value, and the plate loading is increased to the point where the plate dissipation approaches the rated value, such an amplifier may be grid modulated for radiotelephony. If the plate voltage is raised to quite a high value and the stage is adjusted carefully, efficiencies as high as 40 to 43 per cent with good modulation capability and comparatively low distortion may be obtained. Fixed bias is required. This type of operation is termed Class C grid-bias modulation.

Grid Excitation Adequate grid excitation must be available for Class B or Class C service. The excitation for a plate-modulated Class C stage must be sufficient to drive a normal value of d-c grid current with rated bias voltage. The bias voltage preferably should be obtained from a combination of grid leak and fixed C-bias supply.

Cutoff bias can be calculated by dividing the amplification factor of the tube into the d-c plate voltage. This is the value normally used for Class B amplifiers (fixed bias, no grid leak). Class C amplifiers use from 1½ to 5 times this value, depending upon the available grid drive, or excitation, and the desired plate efficiency. Less grid excitation is needed for c-w operation, and the values of fixed bias (if greater than cutoff) may be reduced, or the value of the grid leak resistor can be lowered until normal rated d-c grid current flows.

The values of grid excitation listed for each type of tube may be reduced by as much as 50 per cent if only moderate power output and plate efficiency are desired. When consulting the tube tables, it is well to remember that the power lost in the tuned circuits must be taken into consideration when calculating the available grid drive. At very high frequencies, the r-f circuit losses may even exceed the power required for grid drive.

Readjustments in the tuning of the oscillator, buffer, or doubler circuits, will often result in greater grid drive to the final amplifier.

Link coupling between stages, particularly to the final amplifier grid circuit, normally will provide more grid drive than can be obtained from other coupling systems. The number of turns in the coupling link, and the location of the turns on the coil, can be varied with respect to the tuned circuits to obtain the greatest grid drive for allowable values of buffer or doubler plate current. Slight readjustments sometimes can be made after plate voltage has been applied to the driver tube.

Excessive grid current damages tubes by overheating the grid structure; beyond a certain point of grid drive, no increase in power output can be obtained for a given plate voltage.

7-5 Neutralization of R. F. Amplifiers

The plate-to-grid feedback capacitance of triodes makes it necessary that they be neutralized for operation as r-f amplifiers at frequencies above about 500 kc. Those screen-grid tubes, pentodes, and beam tetrodes which have a plate-to-grid capacitance of 0.1 $\mu\mu$fd. or less may ordinarily be operated as an amplifier without neutralization in a well-designed amplifier up to 30 Mc.

Neutralizing Circuits The object of neutralization is to cancel or "neutralize" the capacitive feedback of energy from plate to grid. There are two general methods by which this energy feedback may be eliminated: the first, and the most common method, is through the use of a capacitance bridge, and the second method is through the use of a parallel reactance of equal and opposite polarity to the grid-to-plate capacitance, to nullify the effect of this capacitance.

With the increasing importance of minimizing the production of harmonic energy by

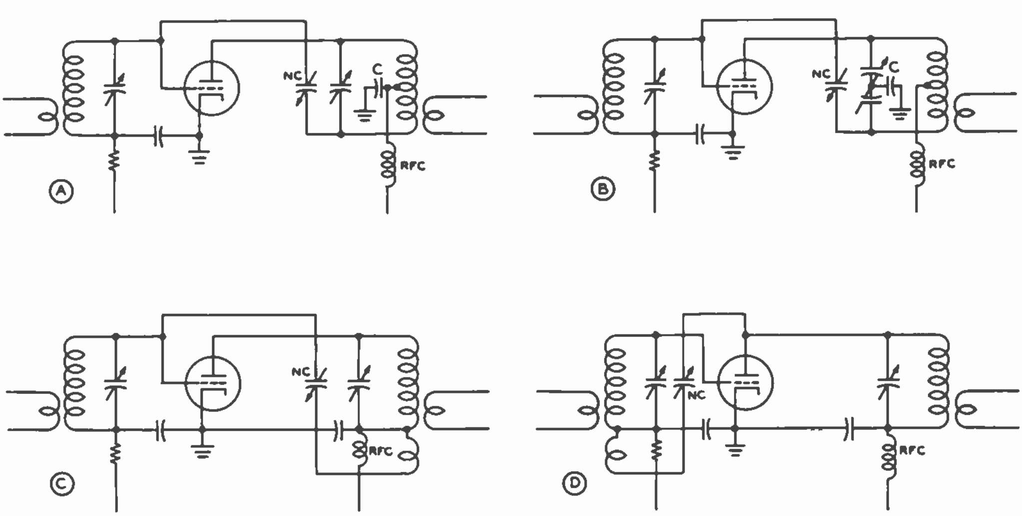


Figure 8.
COMMON NEUTRALIZING CIRCUITS FOR SINGLE-ENDED AMPLIFIERS.

Class C amplifiers, only those plate tank circuits and neutralizing circuits which include a sizeable capacitance directly from plate to ground should be used. This limitation means that the "tapped-coil" method of neutralization, with a single-ended tuning capacitor, should not be used where the radiation of harmonic energy would be undesirable. However, the circuits of figure 8B through 8D are satisfactory from a harmonic radiation standpoint.

Split-Stator Plate Neutralization Figure 8B shows the neutralization circuit which is most widely used in single-ended r-f stages. The use of a split-stator plate capacitor makes the electrical balance of the circuit substantially independent of the mutual coupling within the coil and also makes the balance independent of the place where the coil is tapped. With conventional tubes this circuit will allow one neutralization adjustment to be made on, say, 28 Mc., and this adjustment usually will hold sufficiently close for operation on all lower frequency bands.

Hazeltine Neutralization An alternative system of neutralization, wherein the neutralizating circuit is inductively coupled to one of the tank coils, is shown in figures 8C and 8D. Figure 8C shows the plate-neutralized Hazeltine circuit, while 8D shows the grid-neutralized arrangement. In either case, it will be noticed that there is no tank current flowing through the neutralizing coil L.

In this circuit arrangement, the size of the neutralizing capacitor NC is determined by the coefficient of coupling between the tank coil and L, and upon their relative inductances. It is possible, by proper proportioning of the neutralizing coil L on each band, to make one setting of NC correct for all bands.

Push-Pull Neutralization Two tubes of the same type can be connected for *push-pull* operation so as to obtain twice as much output as that of a single tube. A push-pull amplifier, such as that shown in figure 9, also has an advantage in that the circuit can more easily be balanced than a single-tube r-f amplifier. The various inter-electrode capacitances and the neutralizing capacitors are connected in such a manner that the reactances on one side of the tuned circuits are exactly equal to those on the opposite side. For this reason, push-pull r-f amplifiers can be more easily neutralized in very-high-frequency transmitters; also, they usually remain in perfect neutralization when tuning the amplifier to different bands.

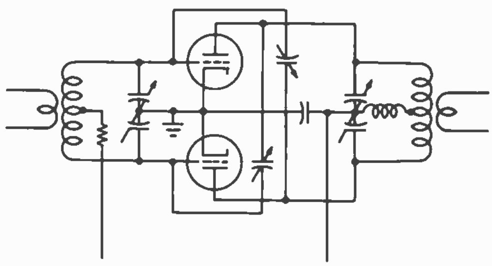


Figure 9.
STANDARD CROSS-NEUTRALIZED PUSH-PULL TRIODE AMPLIFIER.

The circuit shown in figure 9 is perhaps the most commonly used arrangement for a push-pull r-f amplifier stage. The rotor of the grid capacitor is grounded, and the rotor of the plate tank capacitor is by-passed to ground.

Shunt or Coil Neutralization The feedback of energy from grid to plate in an unneutralized r-f amplifier is a result of the grid-to-plate capacitance of the amplifier tube. A neutralization circuit is merely an electrical arrangement for nullifying the effect of this capacitance. All the previous neutralization circuits have made use of a bridge circuit for balancing out the grid-to-plate energy feedback by feeding back an equal amount of energy of opposite phase.

Another method of eliminating the feedback effect of this capacitance, and hence of neutralizing the amplifier stage, is shown in figure 10. The grid-to-plate capacitance in the triode amplifier tube acts as a capacitive reactance, coupling energy back from the plate to the grid circuit. If we parallel this capacitance with an inductance having the same value of reactance of opposite sign the reactance of one will cancel the reactance of the other and we will have a high-impedance tuned circuit from grid to plate.

This neutralization circuit can be used on ultra-high frequencies where other neutralization circuits are unsatisfactory. This is true because the lead length in the neutralization circuit is practically negligible. The circuit can also be used with push-pull r-f amplifiers. In this case, each tube will have its own neutralizing inductor connected from grid to plate.

The big advantage of this arrangement is that it allows the use of single-ended tank circuits with a single-ended amplifier.

The chief disadvantage of the shunt neutralized arrangement is that the stage must be re-neutralized each time the stage is retuned to a new frequency sufficiently removed that the grid and plate tank circuits must be retuned to resonance. However, by the use of plug-in coils it is possible to change to a different band of operation by changing the neutralizing coil at the same time that the grid and plate coils are changed.

The 0.0001-µfd. capacitor in series with the neutralizing coil is merely a blocking capacitor to isolate the plate voltage from the grid circuit. The coil L will have to have a very large number of turns for the band of operation in order to be resonant with the comparatively small grid-to-plate capacitance. But since, in all ordinary cases with tubes operating on frequencies for which they were designed, the L/C ratio of the tuned circuit will be very high, the coil can use comparatively small wire, although it must be wound on air or very low-loss dielectric, and must be insulated for the sum of the plate r-f voltage and the grid r-f voltage.

7-6 Neutralizing Procedure

An r-f amplifier is neutralized to prevent self-oscillation or regeneration. A neon bulb, a flashlight lamp and loop of wire, or an r-f galvanometer can be used as a *null indicator* for neutralizing low-power stages. *The plate voltage lead is disconnected from the r-f amplifier stage while it is being neutralized.* Normal grid drive then is applied to the r-f stage, the neutralizing indicator is coupled to the plate coil, and the plate tuning capacitor is tuned to resonance. The neutralizing capacitor (or capacitors) then can be adjusted until *minimum* r.f. is indicated for resonant settings of both grid and plate tuning capacitors. Both neutralizing capacitors are adjusted simultaneously and to approximately the same value of capacitance when a physically symmetrical push-pull stage is being neutralized.

A final check for neutralization should be made with a d-c milliammeter connected in

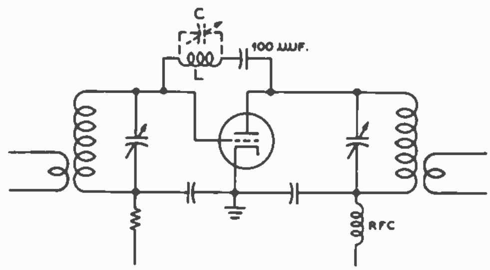


Figure 10.
COIL NEUTRALIZED AMPLIFIER.

This neutralization circuit is very effective with triode tubes on any frequency, but is particularly effective in the v-h-f range. The coil L is adjusted so that it resonates at the operating frequency with the grid-to-plate capacitance of the tube. Capacitor C may be a very small unit of the low-capacitance neutralizing type and is used to trim the circuit to resonance at the operating frequency. If some means of varying the inductance of the coil a small amount is available, the trimmer capacitor is not needed.

the grid leak or grid-bias circuit. There will be no movement of the meter reading as the plate circuit is tuned through resonance (without plate voltage being applied) when the stage is completely neutralized. The milliammeter check is more accurate than any other means for indicating complete neutralization and it also is suitable for neutralizing the stages of a high-power transmitter.

Plate voltage should be *completely* removed by actually opening the d-c plate circuit. (Turning off the *filaments* of the rectifier tubes will do the trick.) If there is a d-c return through the plate supply, a small amount of plate current will flow when grid excitation is applied, even though no primary a-c voltage is being fed to the plate transformer.

A further check on the neutralization of any r-f amplifier can be made by noting whether maximum grid current on the stage comes at the same point of tuning on the plate tuning capacitor as minimum plate current. This check is made with plate voltage on the amplifier and with normal antenna coupling. As the *plate* tuning capacitor is detuned *slightly* from resonance on either side the grid current on the stage should decrease the same amount and without any sudden jumps on either side of resonance. This will be found to be a very precise indication of accurate neutralization in either a triode or beam-tetrode r-f amplifier

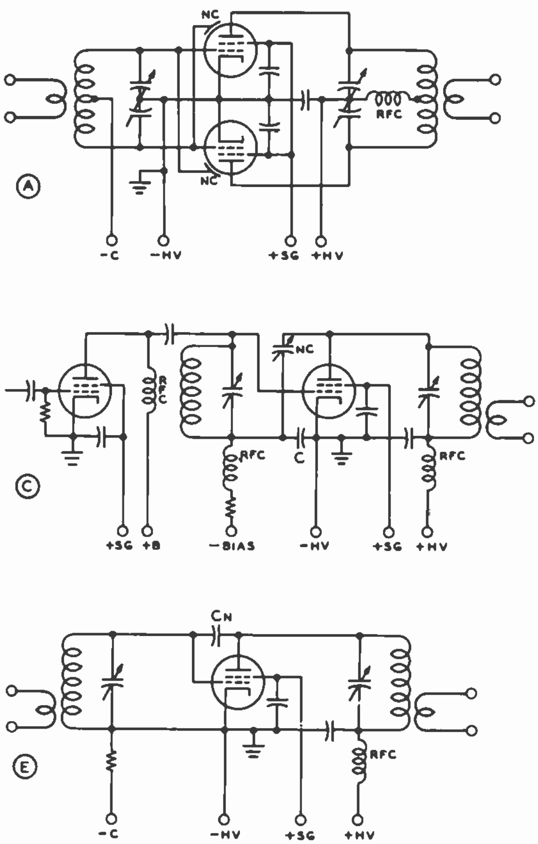


Figure 11.

NEUTRALIZING CIRCUITS FOR BEAM TETRODES.

A conventional cross neutralized circuit for use with push-pull beam tetrodes is shown at (A). The neutralizing capacitors (NC) usually consist of small plates or rods mounted alongside the plate elements of the tubes. (B) and (C) show "grid neutralized" circuits for use with a single-ended tetrode stage having either link coupling or capacitive coupling into the grid tank. (D) shows a method of tuning the screen-lead inductance to accomplish neutralization in a single-frequency v-h-f tetrode amplifier, while (E) shows a method of neutralization by increasing the grid-to-plate capacitance on a tetrode when the operating frequency is higher than that frequency where the tetrode is "self-neutralized" as a result of series resonance in the screen lead. Methods (D) and (E) normally are not practicable at frequencies below about 50 Mc. with the usual types of beam tetrode tubes.

stage, so long as the stage is feeding a load which presents a resistive impedance at the operating frequency.

Push-pull circuits usually can be more completely neutralized than single-ended circuits at very high frequencies. In the intermediate range of from 3 to 15 Mc., single-ended circuits will give satisfactory results.

Neutralization of Screen-Grid R-F Amplifiers Radio-frequency amplifiers using screen-grid tubes can normally be operated without any additional provision for neutralization at frequencies up to about 15 Mc., provided adequate shielding has been provided between the input and output circuits. Special v-h-f screen-grid and beam tetrode tubes such as the 2E26, 807W, and 5516 in the low-power category and HK-257B, 4E27/8001, 4-125A, and 4-250A in the medium-power category can frequently be operated at frequencies as high as 100 Mc. without any additional provision for neutralization. Tubes such as the 807, 2E22, HY-69, and 813 can normally be operated with good circuit design at frequencies up to 30 Mc. without any additional provision for neutralization. The 815 has been found to require neutralization in many cases at frequencies above 30 Mc., although the 829B will normally operate quite stably at 148 Mc. without neutralization.

At frequencies above those listed in the previous paragraph for each tube type, some additional provision for neutralization will quite frequently be required. Also, it has been found by experience that a single-ended beam-tetrode r-f amplifier will often operate stably at a frequency somewhat higher than a push-pull r-f amplifier using the same tube type.

In most push-pull tetrode amplifiers the simplest method of accomplishing neutralization is to use the cross-neutralized capacitance bridge arrangement as normally employed with triode tubes. The neutralizing capacitances, however, must be very much smaller than used with triode tubes, values of the order of 0.2 $\mu\mu$fd. normally being required with beam tetrode tubes. This order of capacitance is far less than can be obtained with a conventional neutralizing capacitor at minimum setting, so the neutralizing arrangement is most commonly made especially for the case at hand. Most common procedure is to bring a conductor (connected to the opposite grid) in the vicinity of the plate itself or of the plate tuning capacitor of one of the tubes. Either one or two such "capacitors" may be used, two being normally used on a higher frequency amplifier in order to maintain balance within the stage.

Neutralizing Single-Ended Tetrode Stages A single-ended tetrode r-f amplifier stage may be neutralized in the same manner as illustrated for a push-pull stage in figure 11, provided a split-stator tank capacitor is in use in the plate circuit. However, in the majority of single-ended tetrode r-f amplifier stages a single-section capacitor is used in the plate tank. Hence, other neutralization procedures must be employed when neutralization is found necessary.

The circuit shown in figure 11B is not a true neutralizing circuit, in that the plate-to-grid capacitance is not balanced out. However, the circuit can afford the equivalent effect by isolating the high resonant impedance of the grid tank circuit from the energy fed back from plate to grid. When NC and C are adjusted to bear the following ratio to the grid-to-plate capacitance and the total capacitance from grid-to-ground in the output tube:

$$\frac{NC}{C} = \frac{C_{gp}}{C_{gk}}$$

both ends of the grid tank circuit will be at the same voltage with respect to ground as a result of r-f energy fed back to the grid circuit. This means that the impedance from grid to ground will be effectively equal to the reactance of the grid-to-cathode capacitance in parallel with the stray grid-to-ground capacitance, since the high resonant impedance of the tuned circuit in the grid has been effectively isolated from the feedback path. It is important to note that the effective grid-to-ground capacitance of the tube being neutralized includes the rated grid-to-cathode or input capacitance of the tube, the capacitance of the socket, wiring capacitances and other strays, but it does *not* include the capacitances associated with the grid tuning capacitor. Also, if the tube is being excited by capacitive coupling from a preceding stage (as in figure 11C), the effective grid-to-ground capacitance includes the output capacitance of the preceding stage and its associated socket and wiring capacitances.

Cancellation of Screen-Lead Inductance The provisions discussed in the previous paragraphs are for neutralization of the small, though still important at the higher frequencies, grid-to-plate capacitance of beam-tetrode tubes. However, in the vicinity of the upper-frequency limit of each tube type the inductance of the screen lead of the tube becomes of considerable importance. With a tube operating at a frequency where the inductance of the screen lead is appreciable, the screen will allow a considerable amount of energy leak-through from plate to grid even though the socket terminal on the tube is carefully by-passed to ground. This condition takes place because, even though the socket pin is by-passed, the reactance of the screen lead will allow a moderate amount of r-f potential to appear on the screen itself inside the electrode assembly in the tube. This effect has been reduced to a very low amount in such tubes as the Hytron 5516, the 829B, and the Eimac 4X150A and 4X500A but it is still quite appreciable in most beam-tetrode tubes.

The effect of screen-lead inductance on the stability of a stage can be eliminated at any particular frequency by one of two methods

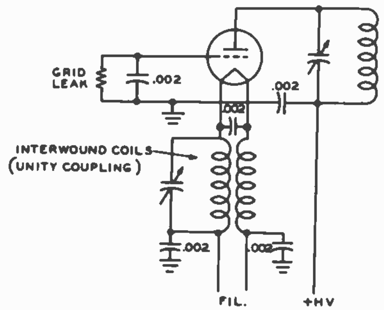


Figure 12.
GROUNDED-GRID AMPLIFIER.
This type of triode amplifier requires no neutralization, but can be used only with tubes having a relatively low plate-to-cathode capacitance. An alternative arrangement is illustrated in figure 21C, and operating conditions are discussed in Chapter Five.

These methods are: (1) Tuning out the screen-lead inductance by series resonating the screen lead inductance with a capacitor to ground. This method is illustrated in figure 11D and is commonly employed in commercially-built equipment for operation on a narrow frequency band in the range above about 75 Mc. The other method (2) is illustrated in figure 11E and consists in feeding back additional energy from plate to grid by means of a small capacitor connected between these two elements. Note that this capacitor is connected in such a manner as to *increase* the effective grid-to-plate capacitance of the tube. This method has been found to be effective with 807 tubes in the range above 50 Mc. and with tubes such as the 4-125A and 4-250A in the vicinity of their upper frequency limits.

Note that both these methods of stabilizing a beam-tetrode v-h-f amplifier stage by cancellation of screen-lead inductance are suitable only for operation over a relatively narrow band of frequencies in the v-h-f range. At lower frequencies both these expedients for reducing the effects of screen-lead inductance will tend to increase the tendency toward oscillation of the amplifier stage.

Neutralizing Problems When a stage cannot be completely neutralized, the difficulty usually can be traced to one or more of the following causes: (1) Filament leads not by-passed to the common ground point of that particular stage. (2) Ground lead from the rotor connection of the split-stator tuning capacitor to filament open or too long. (3) Neutralizing capacitors in a field of excessive r.f. from one of the tuning coils. (4) Electromagnetic coupling between grid and plate coils, or between plate and preceding buffer or oscillator circuits. (5) Insufficient shielding or spacing between stages, or between grid and plate circuits in compact transmitters. (6) Shielding placed too close to plate circuit coils, causing induced currents in the shields. (7) Parasitic oscillations when plate voltage is applied. The cure for the latter is mainly a matter of cut and try—rearrange the parts, change the length of grid or plate or neutralizing leads, insert a parasitic choke in the grid lead or leads, or eliminate the grid r-f chokes which may be the cause of a low-frequency parasitic (in conjunction with plate r-f chokes). See *Parasitic Oscillation in R-F Amplifiers* in Section 7-11 of this chapter.

7-7 Grounded Grid Amplifiers

Certain triodes have a grid configuration and lead arrangement which results in very low plate to filament capacitance when the control grid is grounded, the grid acting as an effective shield much in the manner of the screen in a screen-grid tube.

By connecting such a triode in the circuit of figure 12, taking the usual precautions against stray capacitive and inductive coupling between input and output leads and components, a stable power amplifier is realized which requires no neutralization.

A detailed discussion of the operation of grounded-grid r-f power amplifiers, along with the method of design and design data on a typical stage, has been given in Chapter 5, Section 5-14.

At ultra-high frequencies, where it is difficult to obtain satisfactory neutralization with conventional triode circuits (particularly when a wide band of frequencies is to be covered), the grounded-grid arrangement is about the only practicable means of employing a triode amplifier.

Because of the large amount of degeneration inherent in the circuit, considerably more excitation is required than if the same tube

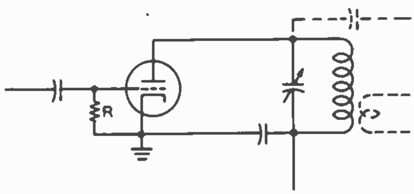


Figure 13.
CONVENTIONAL TRIODE FREQUENCY MULTIPLIER.
Small triodes such as the 6C4 operate satisfactorily as frequency multipliers, and can deliver output well into the v-h-f range. Resistor R normally will have a value in the vicinity of 100,000 ohms.

were employed in a conventional grounded-cathode circuit. The additional power required to drive a triode in a grounded-grid amplifier is not lost, however, as it shows up in the output circuit and adds to the power delivered to the load. But nevertheless it means that a larger driver stage is required for an amplifier of given output, because a moderate amount of power is delivered to the amplifier load by the driver stage of a grounded-grid amplifier.

7-8 Frequency Multipliers

Quartz crystals and variable-frequency oscillators are not ordinarily used for direct control of the output of high-frequency transmitters. *Frequency multipliers* are usually employed to multiply the frequency to the desired value. These multipliers operate on exact multiples of the excitation frequency; a 3.6-Mc. crystal oscillator can be made to control the output of a transmitter on 7.2 or 14.4 Mc., or on 28.8 Mc., by means of one or more frequency multipliers. When used at twice frequency, they are often termed *frequency doublers*. A simple doubler circuit is shown in figure 13. It consists of a vacuum tube with its plate circuit tuned to *twice* the frequency of the grid driving circuit. This doubler can be excited from a crystal oscillator or another multiplier or amplifier stage.

Doubling is best accomplished by operating the tube with high grid bias. The grid circuit is driven approximately to the normal value of d-c grid current through the r-f choke and grid-leak resistor, shown in figure 13. The resistance value generally is from two to five times as high as that used with the same tube for straight amplification. Consequently, the grid bias is several times as high for the same value of grid current.

Neutralization is seldom necessary in a doubler circuit, since the plate is tuned to twice the frequency of the grid circuit. The impedance of the grid driving circuit is very low at the doubling frequency, and thus there is little tendency for self-excited oscillation.

Frequency doublers require bias of several times cutoff; high-μ tubes therefore are desirable for this type of service. Tubes which have amplification factors from 20 to 200 are suitable for doubler circuits. Tetrodes and pentodes make excellent doublers. Low-μ triodes, having amplification constants of from 3 to 10, are not applicable for doubler service. In extreme cases the grid voltage must be as high as the plate voltage for efficient doubling action.

Angle of Flow in Frequency Multipliers The angle of plate current flow in a frequency multiplier is a very important factor in determining the efficiency. As the angle of flow is decreased for a given value of grid current, the efficiency increases. To reduce the angle of flow, higher grid bias is required so that the grid excitation voltage will exceed the cutoff value for a shorter portion of the exciting-voltage cycle. For a high order of efficiency, frequency doublers should have an angle of flow of 90 degrees or less, triplers 60 degrees or less, and quadruplers 45 degrees or less. Under these conditions the efficiency will be on the same order as the reciprocal of the harmonic on which the stage operates. In other words the efficiency of a doubler will be approximately ½ or 50 per cent, the efficiency of a tripler will be approximately ⅓ or 33 per cent and that of a quadrupler will be about 25 per cent. With good stage design the efficiency can be somewhat greater than these values, but as the angle of flow is made greater than these limiting values, the efficiency falls off rapidly. The reason is apparent from a study of figure 15.

The pulses ABC, EFG, JKL illustrate 180-degree excitation pulses under Class B operation, the solid straight line indicating cutoff bias. If the bias is increased by N times, to the value indicated by the dotted straight line, and the excitation increased until the peak r-f voltage with respect to ground is the same as before, then the excitation frequency can be cut in half and the effective excitation pulses

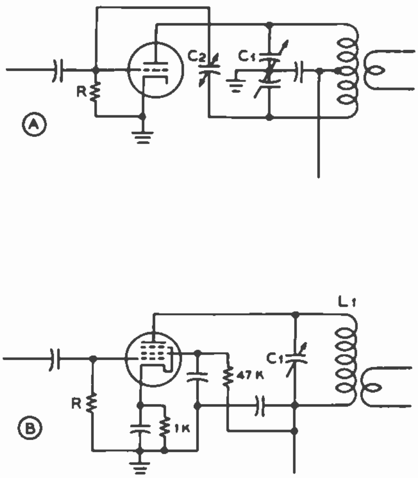


Figure 14.

FREQUENCY MULTIPLIER CIRCUITS.

The output of a triode v-h-f frequency multiplier often may be increased by neutralization of the grid-to-plate capacitance as shown at (A) above. Such a stage also may be operated as a straight amplifier when the occasion demands. A pentode frequency multiplier is shown at (B). Conventional power tetrodes operate satisfactorily as multipliers so long as the output frequency is below about 100 Mc. Above this frequency special v-h-f tetrodes must be used to obtain satisfactory output.

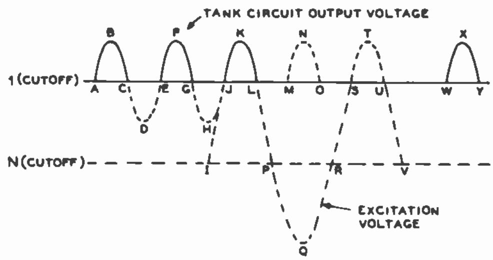


Figure 15.

ILLUSTRATING THE ACTION OF A FREQUENCY DOUBLER.

will have almost the same shape as before. The only difference is that every other pulse is missing; MNO simply shows where the missing pulse would go. However, if the Q of the plate tank circuit is high, it will have sufficient "flywheel" effect to carry over through the missing pulse, and the only effect will be that the plate input and r-f output at optimum loading drop to approximately half. As the input frequency is half the output frequency, an efficient frequency doubler is the result.

By the same token, a tripler or quadrupler can be analyzed, the tripler skipping two excitation pulses and the quadrupler three. In each case the excitation pulse ideally should be short enough that it does not exceed 180 degrees at the output frequency; otherwise the excitation actually is *bucking* the output over a portion of the cycle.

In actual practice, it is found uneconomical to provide sufficient excitation to run a tripler or quadrupler in this fashion. Usually the excitation pulses will be at least 90 degrees at the exciting frequency, with correspondingly low efficiency, but it is more practicable to accept the low efficiency and build up the output in succeeding amplifier stages. The efficiency can become quite low before the power gain becomes less than unity.

Push-Push Multipliers Two tubes can be connected in parallel to give twice the output of a single-tube doubler. If the grids are driven *out* of phase instead of *in* phase, the tubes then no longer work simultaneously, but rather one at a time. The effect is to fill in the missing pulses (figure 15). Not only is the output doubled, but several advantages accrue which cannot be obtained by straight parallel operation.

Chief among these is the effective neutralization of the fundamental and all *odd* harmonics, an advantage when spurious emissions must be minimized. Another advantage is that when the available excitation is low and excitation pulses exceed 90 degrees, the output and efficiency will be greater than for the same tubes connected in parallel.

The same arrangement may be used as a quadrupler, with considerably better efficiency than for straight parallel operation, because seldom is it practicable to supply sufficient excitation to permit 45 degree excitation pulses. As pointed out above, the push-push arrangement exhibits better efficiency than a single ended multiplier when excitation is inadequate for ideal multiplier operation.

A typical push-push doubler is illustrated in figure 16. When high transconductance tubes are employed, it is necessary to employ a split-stator grid tank capacitor to prevent self oscillation; with well screened tetrodes or pentodes having medium values of transconduct-

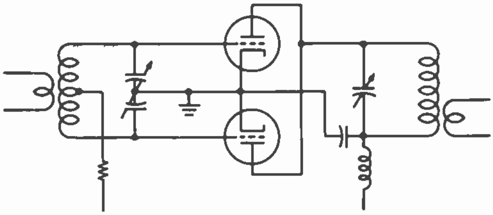


Figure 16.
PUSH-PUSH FREQUENCY DOUBLER.
The output of a doubler stage may be materially increased through the use of a push-push circuit such as illustrated above.

ance, a split-coil arrangement with a single-section capacitor may be employed (the center tap of the grid coil being by-passed to ground).

Push-Pull Frequency Triplers It is frequently desirable in the case of u-h-f and v-h-f transmitters that frequency multiplication stages be balanced with respect to ground. Further it is just as easy in most cases to multiply the crystal or v-f-o frequency by powers of three rather than multiplying by powers of two as is frequently done on lower frequency transmitters. Hence the use of push-pull triplers has become quite prevalent in both commercial and amateur v-h-f and u-h-f transmitter designs. Such stages are balanced with respect to ground and appear in construction and on paper essentially the same as a push-pull r-f amplifier stage with the exception that the output tank circuit is tuned to three times the frequency of the grid tank circuit. A circuit for a push-pull tripler stage is shown in figure 17.

A push-pull tripler stage has the further advantage in amateur work that it can also be used as a conventional push-pull r-f amplifier merely by changing the grid and plate coils so that they tune to the same frequency. This is of some advantage in the case of operating in the 50-Mc. band with 50-Mc. excitation, and then changing the plate coil to tune to 144 Mc. for operation of the stage as a tripler from excitation on 48 Mc. This circuit arrangement is excellent for operation with push-pull beam tetrodes such as the 815 and 829B, although a pair of tubes such as the 2E25, 2E26, or 5516 could just as well be used if proper attention were given to the matter of screen-lead inductance.

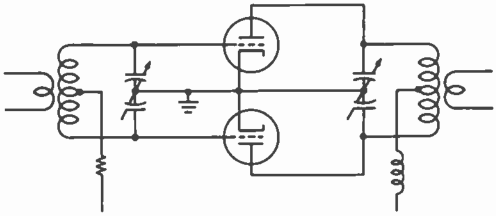


Figure 17.
PUSH-PULL FREQUENCY TRIPLER.
The push-pull tripler is advantageous in the v-h-f range since circuit balance is maintained both in the input and output circuits. If the circuit is neutralized it may be used either as a straight amplifier or as a tripler. Either triodes or tetrodes may be used; dual-unit tetrodes such as the 815, 832A, and 829B are particularly effective in the v-h-f range.

7-9 Tank Circuit Capacitances

It is necessary that the proper value of Q be used in the plate tank circuit of any r-f amplifier. A brief discussion of this matter has been given in Section 5-12 of Chapter 5. However, the following section has been devoted to a more thorough treatment of the subject, and charts and curves are given to assist the reader in the determination of the proper L/C ratio to be used in a radio-frequency amplifier stage.

A Class C amplifier draws plate current in the form of very distorted pulses of short duration. Such an amplifier is always operated into a tuned inductance-capacitance or tank circuit which tends to smooth out these pulses, by its storage or "tank" action, into a sine wave of radio-frequency output. Any waveform distortion of the carrier frequency results in harmonic interference in higher-frequency channels.

A Class A r-f amplifier would produce a sine wave of radio-frequency output if its exciting waveform were also a sine wave. However, a Class A amplifier stage converts its d-c input to r-f output by acting as a variable resistance, and therefore heats considerably. A Class C amplifier when driven hard with short pulses at the peak of the exciting waveform acts more as an electronic switch, and therefore can convert its d-c input to r-f output with relatively good efficiency. Values of plate circuit efficiency from 65 to 85 per cent are common in Class C amplifiers operating under optimum conditions of excitation, grid bias, and loading.

Tank Circuit Q As stated before, the tank circuit of a Class C amplifier receives energy in the form of short pulses of plate current which flow in the amplifier tube. But the tank circuit must be able to store enough energy so that it can deliver a current essentially sine wave in form to the load. The ability of a tank to store energy in this manner may be designated as the effective Q of the tank circuit. The effective circuit Q may be stated in any of several ways, but essentially the Q of a tank circuit is the *ratio of the energy stored to 2π times the energy lost per cycle*. Further, the *energy lost per cycle* must, by definition, be equal to *the energy delivered* to the tank circuit by the Class C amplifier tube or tubes.

The Q of a tank circuit at resonance is equal to its parallel resonant impedance (the resonant impedance is resistive at resonance) divided by the reactance of either the capacitor or the inductor which go to make up the tank. The inductive reactance is equal to the capacitive reactance, by definition, at resonance. Hence we may state:

$$Q = \frac{R_L}{X_C} = \frac{R_L}{X_L}$$

where R_L is the resonant impedance of the tank and X_C is the reactance of the tank capacitor and X_L is the reactance of the tank coil. This value of resonant impedance, R_L, is the load which is presented to the Class C amplifier tube in a single-ended circuit such as shown in figure 18.

The value of load impedance, R_L, which the Class C amplifier tube sees may be obtained, looking in the other direction from the tank coil, from a knowledge of the operating conditions on the Class C tube. This load impedance may be obtained from the following expression, which is true in the general case of any Class C amplifier:

$$R_L = \frac{E_{pm}^2}{2\ N_p\ I_b\ E_{bb}}$$

where the values in the equation have the characteristics listed at the beginning of Chapter 5.

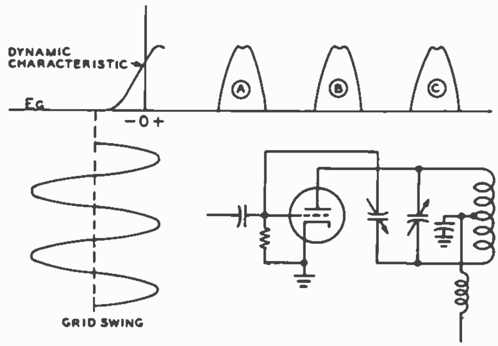


Figure 18.

CLASS C AMPLIFIER OPERATION.

Plate current pulses are shown at (A), (B), and (C). The dip in the top of the plate current waveform will occur when the excitation voltage is such that the minimum plate voltage dips below the maximum grid voltage. A detailed discussion of the operation of Class C amplifiers is given in Chapter Five.

The expression above is all very fine, except that the peak value of the fundamental component of plate voltage swing, E_{pm}, is not ordinarily known unless a high-voltage peak a-c voltmeter is available for checking. Also, the decimal value of plate circuit efficiency is not ordinarily known with any degree of accuracy. However, in a *normally operated* Class C amplifier the plate voltage swing will be approximately equal to 0.85 to 0.9 times the d-c plate voltage on the stage, and the plate circuit efficiency will be from 70 to 80 per cent (N_p of 0.7 to 0.8), the higher values of efficiency normally being associated with the higher values of plate voltage swing. With these two assumptions as to the normal Class C amplifier, the expression for the plate load impedance can be greatly simplified to the following approximate but useful expression:

$$R_L \simeq \frac{R_{d.c.}}{2}$$

which means simply that the resistance presented by the tank circuit to the Class C tube is approximately equal to one-half the d-c load resistance which the Class C stage presents to the power supply (and also to the modulator in case high-level modulation of the stage is to be used).

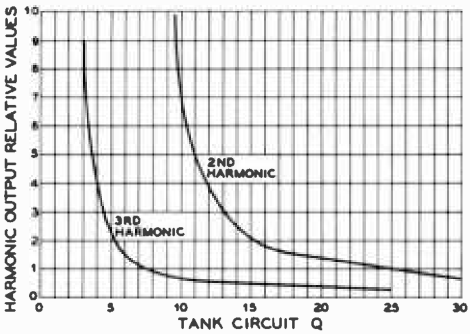


Figure 19.
RELATIVE HARMONIC OUTPUT PLOTTED AGAINST TANK CIRCUIT Q.

Combining the above simplified expression for the r-f impedance presented by the tank to the tube, with the expression for tank Q given in a previous paragraph we have the following expression which relates the reactance of the tank capacitor or coil to the d-c input to the Class C stage:

$$X_C = X_L \simeq \frac{R_{d.c.}}{2Q}$$

The above expression is the basis of the usual charts giving tank capacitance for the various bands in terms of the d-c plate voltage and current to the Class C stage, including the chart of figure 20.

Harmonic Radiation vs. Q The problem of harmonic radiation from transmitters has long been present, but it has become critical only relatively recently along with the extensive occupation of the v-h-f range. Television signals are particularly susceptible to interference from other signals falling within the pass band of the receiver, so that the TVI problem has received the major emphasis of all the services in the v-h-f range which are susceptible to interference from harmonics of signals in the h-f or lower v-h-f range.

Inspection of figure 19 will show quickly that the tank circuit of a Class C amplifier should have an operating Q of 15 or greater to afford satisfactory rejection of second harmonic energy. The curve begins to straighten out above a Q of about 15, so that a considerable increase in Q must be made before an appreciable reduction in second-harmonic energy is obtained. Above a circuit Q of about 10 any increase will not afford appreciable reduction in the third-harmonic energy, so that

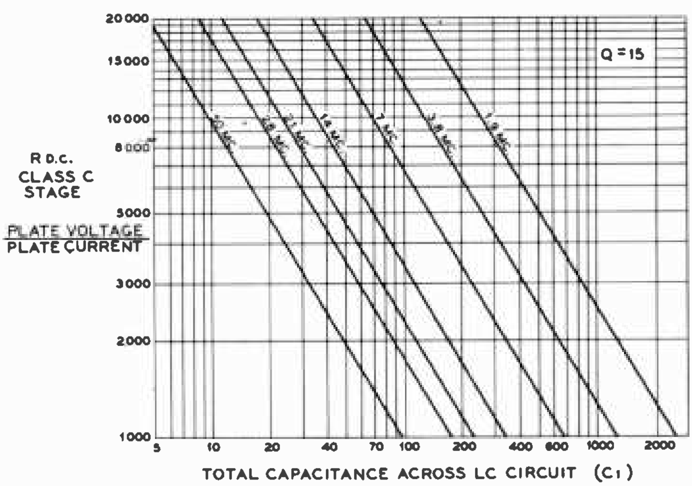


Figure 20.
CORRECT VALUES OF TANK CIRCUIT CAPACITANCE FOR AN OPERATING Q OF APPROXIMATELY 15.
Capacitance values in terms of this chart for various tank circuit arrangements are given in figure 21.

additional harmonic filtering circuits external to the amplifier proper must be called into play.

Effect of Loading on Q The Q of a circuit depends upon the resistance in series with the capacitance and inductance. This series resistance is very low for a low-loss coil not loaded by an antenna circuit. The value of Q may be from 100 to 600 under these conditions. Coupling an antenna circuit has the effect of increasing the series resistance, though in this case the power is consumed as useful radiation by the antenna. Mathematically, the antenna increases the value of R in the expression $Q = \omega L/R$ where L is the coil inductance in microhenrys and ω is the term $2\pi f$, f being in megacycles.

The coupling from the final tank circuit to the antenna or antenna transmission line can be varied to obtain values of Q from perhaps 3 at maximum coupling to a value of Q equal to the unloaded Q of the circuit at zero antenna coupling. This value of unloaded Q can be as high as 500 or 600, as mentioned in the preceding paragraph. However, the value of $Q = 15$ will not be obtained at values of normal d-c plate current in the Class C amplifier stage unless the C-to-L ratio in the tank circuit is correct for that frequency of operation.

The capacitance values in figure 20 are for the total capacitance across the tank coil in a circuit such as shown in figure 21A. This value includes the tube interelectrode capacitance (output, or plate-to-ground), coil distributed capacitance, wiring capacitances, and the value of any low-inductance plate-to-ground by-pass capacitor as used for reducing harmonic generation, in addition to the actual capacitance of the plate tuning capacitor. Total circuit stray capacitance may vary from perhaps 5$\mu\mu$fd. for a v-h-f stage to 30 $\mu\mu$fd. for a medium power h-f stage. When a split-stator tuning capacitor is being used, as in figures 21D and 21E, the total circuit capacitance may be cut to one-fourth, which means that each section of the split-stator capacitor should have one-half the capacitance specified for a single-ended stage.

The tank circuit operates in the same manner whether the tube feeding it is a pentode, beam tetrode, neutralized triode, grounded-grid triode, whether it is single ended or push-pull, or whether it is shunt fed or series fed. The

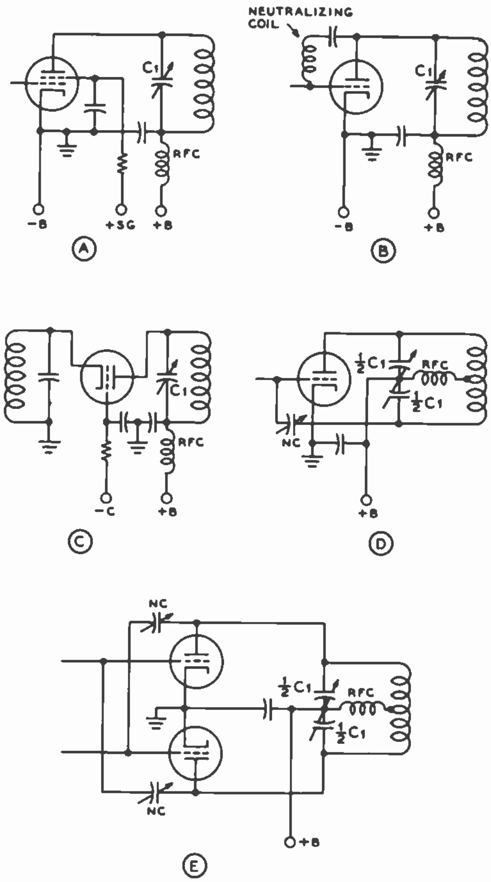


Figure 21.
PLATE-TANK CIRCUIT ARRANGEMENTS.

Shown above in the case of each of the tank circuit types is the recommended tank circuit capacitance at the operating frequency in terms of the values given in figure 20 for an operating Q of 15. (A) is a conventional tetrode amplifier, (B) is a coil-neutralized triode amplifier, (C) is a grounded-grid triode amplifier, (D) is a single-ended neutralized triode stage with a split-stator plate tank, and (E) represents a push-pull amplifier which may be either of the triode or tetrode type.

important thing in establishing the operating Q of the tank circuit is the ratio of the loaded resonant impedance across its terminals to the reactance of the L and the C which make up the tank.

Due to the unknowns involved in determining circuit stray capacitances it is sometimes more convenient to determine the value of L required for the proper circuit Q (by the

TABLE I

USUAL BREAKDOWN RATINGS OF COMMON PLATE SPACINGS

Air-gap in inches	Peak voltage breakdown
.030	1,000
.050	2,000
.070	3,000
.100	4,000
.125	4,500
.150	5,200
.170	6,000
.200	7,500
.250	9,000
.350	11,000
.500	15,000
.700	20,000

method discussed earlier in this Section) and then to vary the tuned circuit capacitance until resonance is reached. This method is most frequently used in obtaining proper circuit Q in commercial transmitters.

The values of R_p for using the chart of figure 20 are easily calculated by dividing the d-c plate supply voltage by the total d-c plate current (expressed in amperes). Correct values of total tuning capacitance are shown in the chart for the different amateur bands. The shunt stray capacitance can be estimated closely enough for all practical purposes. The coil inductance should then be chosen which will produce resonance at the desired frequency with the total calculated tuning capacitance.

Tuning Capacitor Air Gap To determine the required tuning capacitor air gap for a particular amplifier circuit it is first necessary to estimate the peak r-f voltage which will appear between the plates of the tuning capacitor. Then, using the chart of Table I, it is possible to estimate the plate spacing which will be required.

The instantaneous r-f voltage in the plate circuit of a Class C amplifier tube varies from nearly zero to nearly twice the d-c plate voltage. If the d-c voltage is being 100 per cent modulated by an audio voltage, the r-f peaks will reach nearly four times the d-c voltage.

These rules apply to a loaded amplifier or buffer stage. If either is operated without an r-f load, the peak voltages will be greater and can *exceed* the d-c plate supply voltage. For this reason no amplifier should be operated without load when anywhere near normal d-c plate voltage is applied.

The fixed capacitors, usually of the mica type, shown in figure 21, must be rated to withstand the d-c plate voltage plus any audio voltage. This capacitor should be rated at a d-c working voltage of at least *twice the d-c plate supply in a plate modulated amplifier,* and at least *equal to the d-c supply* in any other type of r-f amplifier.

TABLE II

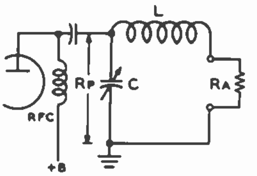
Recommended air gap (approx. 100% factor of safety) for the circuits of figure 21.

D.C. PLATE VOLTAGE	C.W.	PLATE MOD.
400	.030	.050
600	.050	.070
750	.050	.084
1000	.070	.100
1250	.070	.144
1500	.078	.200
2000	.100	.250
2500	.175	.375
3000	.200	.500
3500	.250	.600

7-10 L and Pi Matching Networks

The L and pi networks often can be put to advantageous use in accomplishing an impedance match between two differing impedances. Common applications are the matching between a transmission line and an antenna, or between the plate circuit of a single-ended amplifier stage and an antenna transmission line. Such networks may be used to accomplish a match between the plate tank circuit of an amplifier and a transmission line, or they may be used to match directly from the plate circuit of an amplifier to the line without the requirement for a tank circuit — provided the network is designed in such a manner that it has sufficient operating Q for accomplishing harmonic attenuation.

The L Matching Network The L network is of limited utility in impedance matching since its ratio of impedance transformation is fixed at a value equal to (Q^2+1). The operating Q may be relatively low — perhaps from 3 to 6 — in a matching network between the plate *tank circuit* of an amplifier and a transmission line; hence impedance transformation ratios of 10 to 1 and even lower may be attained. But when the network also acts as the plate tank circuit of the amplifier stage, as in figure 22, the operating Q should be at least 12 and preferably 15. An operating Q of 15 represents an impedance transformation of 225; this value nor-



$R_P = R_A(Q^2+1)$ (EXACT)

$R_P = Q^2 R_A$ (APPROX.)

$Q = \frac{X_L}{R_A} = \frac{X_C}{R_A} = \frac{R_P}{X_C} = \frac{R_P}{X_L}$

$X_L = X_C$

FOR OPERATING CIRCUIT Q OF 15:

$R_P =$ APPROX. $\frac{\text{PLATE VOLTAGE}}{\text{PLATE CURRENT}}$

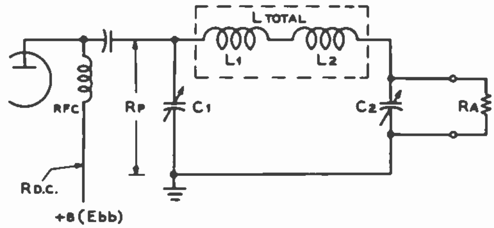
$R_P = 225\, R_A$

$X_C = \frac{R_P}{15}$

$X_L = \frac{R_P}{15}$

Figure 22.
THE L NETWORK IMPEDANCE TRANSFORMER.

The L network is useful with a moderate operating Q for high values of impedance transformation, and it may be used for applications other than in the plate circuit of a tube with relatively low values of operating Q for moderate impedance transformations. Exact and approximate design equations are given.



$R_{D.C.} = \frac{E_{bb}}{I_b}$

$R_P \approx \frac{R_{D.C.}}{2}$

$X_{C_1} = \frac{R_P}{Q}$

$X_{L_1} = \frac{R_P}{Q}$

$X_{C_2} = -R_A \sqrt{\frac{R_P}{R_A(Q^2+1)-R_P}}$

$X_{L_2} = -\frac{R_A^2\, X_{C_2}}{R_A^2+X_{C_2}^2}$

$X_{L\,TOT.} = X_{L_1} + X_{L_2}$

Figure 23.
THE PI NETWORK.

The pi network is valuable for use as an impedance transformer over a wide ratio of transformation values. The operating Q should be at least 12 and preferably 15 to 20 when the circuit is to be used in the plate circuit of a Class C amplifier. Design equations are given above. The inductor L_{tot} represents a single inductance, usually variable, with a value equal to the sum of L_1 and L_2.

mally will be too high even for transforming from the 2000 to 10,000 ohm plate impedance of a Class C amplifier stage down to a 50-ohm transmission line.

However, the L network is interesting since it forms the basis of design for the pi network. Inspection of figure 22 will show that the L network in reality may be considered as a parallel-resonant tank circuit in which R_A represents the coupled-in load resistance; only in this case the load resistance is directly coupled into the tank circuit rather than being inductively coupled as in the conventional arrangement where the load circuit is coupled to the tank circuit by means of a link. When R_A is shorted, L and C comprise a conventional parallel-resonant tank circuit, since for proper operation L and C must be resonant in order for the network to present a resistive load to the Class C amplifier.

The Pi Network The pi impedance matching network, illustrated in figure 23, is much more general in its application than the L network since it offers greater harmonic attenuation, and since it can be used to match a relatively wide range of impedances while still maintaining any desired operating Q. The values of C_1 and L_1 in the pi network of figure 23 can be thought of as having the same values as the L network in figure 22 for the same operating Q, but what is more important from the comparison standpoint these values will be the *same as in a conventional tank circuit* as in figure 21A. So the value of the capacitance may be determined by calculation, with the operating Q and the load impedance which should be reflected to the plate of the Class C amplifier as the two knowns — or the actual values of the capacitance may be obtained for an operating Q of 15 by reference to figure 20.

The inductive arm in the pi network can be thought of as consisting of two inductances in series, as illustrated in figure 23. The first portion of this inductance, L_1, is that value of inductance which would resonate with C_1 at the operating frequency — the same as in a conventional tank circuit. However, the actual value of inductance in this arm of the pi network, L_{tot}, will be greater than L_1 for normal values of impedance transformation. For high transformation ratios L_{tot} will be only slightly greater than L_1; for a transformation ratio of 1.0, L_{tot} will be twice as great as L_1. The amount of inductance which must be added to L_1 to restore resonance and maintain circuit Q is obtained through use of the expression for X_{L2} in figure 23.

The peak voltage rating of the main tuning capacitor C_1 should be the normal value for a

Class C amplifier operating at the plate voltage to be employed. The inductor L_{tot} may be a plug-in coil which is changed for each band of operation, or some sort of variable inductor may be used. A continuously variable slider-type of variable inductor, such as used in certain items of surplus military equipment, may be used to good advantage if available, or a tapped inductor such as used in the ART-13 may be employed. However, to maintain good circuit Q on the higher frequencies when a variable or tapped coil is used on the lower frequencies, the tapped or variable coil should be removed from the circuit and replaced by a smaller coil which has been especially designed for the higher frequency ranges.

The peak voltage rating of the output or loading capacitor, C_2, is determined by the power level and the impedance to be fed. If a 50-ohm coaxial line is to be fed from the pi network, receiving-type capacitors will be satisfactory even up to the power level of a plate-modulated kilowatt amplifier. In any event, the peak voltage which will be impressed across the output capacitor is expressed by: $E_{pk}^2 = 2\ R_a\ W_o$, where E_{pk} is the peak voltage across the capacitor, R_a is the value of resistive load which the network is feeding, and W_o is the maximum value of the average power output of the stage. The harmonic attenuation of the pi network is quite good, although an external low-pass filter will be required to obtain harmonic attenuation value upward of 100 db such as normally required. The attenuation to second harmonic energy will be approximately 40 db for an operating Q of 15 for the pi network; the value increases to about 45 db for a 1:1 transformation and falls to about 38 db for an impedance step-down of 80:1, assuming that the operating Q is maintained at 15.

7-11 Parasitic Oscillation in R-F Amplifiers

Parasitics (as distinguished from *self-oscillation* on the normal tuned frequency of the amplifier) are undesirable oscillations either of very high or very low frequencies which may occur in radio-frequency amplifiers.

They may cause spurious signals (which are often rough in tone) other than normal harmonics, hash on each side of a modulated carrier, key clicks, voltage breakdown or flash-over, instability or inefficiency, and shortened life or failure of the tubes. They may be damped and stop by themselves after keying or modulation peaks, or they may be undamped and build up during ordinary unmodulated transmission, continuing if the excitation is removed. They may result from series or parallel resonant circuits of all types. Due to neutralizing lead length and the nature of most parasitic circuits, the amplifier usually is not neutralized for the parasitic frequency.

Sometimes the fact that the plate supply is keyed will obscure parasitic oscillations in a final amplifier stage that might be very severe if the plate voltage were left on and the excitation were keyed.

In some cases, an all-wave receiver will prove helpful in locating v-h-f spurious oscillations, but it may be necessary to check from several hundred megacycles downward in frequency to the operating range. A normal harmonic is weaker than the fundamental but of good tone; a strong harmonic or a rough note at any frequency generally indicates a parasitic.

Low-Frequency Parasitics One type of unwanted oscillation often occurs in shunt-fed circuits in which the grid and plate chokes resonate, coupled through the tube's inter-electrode capacitance. It also can happen with series feed. This oscillation is generally at a much lower frequency than the operating frequency and will cause additional carriers to appear, spaced from perhaps twenty to a few hundred kilocycles on either side of the main wave. One cure is to change the type of feed in either the grid or plate circuit or to eliminate one choke. Another is to use much less inductance in the grid choke than in the plate choke, or to replace the grid choke by a wire-wound resistor if the grid is series fed. In a Class C stage with grid-leak bias, no r-f choke is required if the bias is series fed.

Low-frequency parasitics often take place in push-pull circuits. In such cases the tubes are effectively in parallel for the parasitic and hence the neutralization is not effective. The grids or plates can be connected together without affecting the undesired oscillation; this is a simple test for this type of parasitic.

Parallel Tubes A very high frequency inter-tube oscillation often occurs when tubes are operated in parallel. Non-inductive damping resistors or parasitic suppressors in the grid circuit and very short interconnecting leads between the tube elements, plus the usual measures for attacking parasitics in a single-tube stage, will prove helpful.

Tapped Inductances When capacitive coupling is used between stages, particularly when one of the stages is tapped down from the end of the coil, additional parasitic circuits are formed because of the multiple resonant effects of the complex interstage circuit. Inductive or link coupling permits making adjustments to interstage coupling without forming these undesired circuits. A capacitor tapped across only part of an inductance, for bandspread tuning or capacitive loading, also can give rise to parasitics.

Multi-Element Tubes The high transconductance and power sensitivity of multi-element tubes, particularly beam tetrodes, tends to aggravate any tendency toward parasitic oscillations. It is particularly important that the by-pass circuit from the additional elements in these tubes to the filament must be short and effective, particularly at the higher frequencies, to prevent undesired internal coupling. At very high frequencies, a certain critical value of screen bypass capacitor may improve the internal shielding without causing a new parasitic oscillation. The capacitance should be such as to series resonate the screen lead inductance at the operating frequency of the amplifier.

Crystal Stages Crystal oscillators are seldom suspected of parasitic oscillation troubles, but are often guilty. The same remedial measures as recommended for amplifiers should be employed.

Parasitic Suppressors The most common type of parasitic is of the v-h-f type, which fortunately can usually be dampened by inserting a parasitic suppressor of appropriate characteristics in series with the grid or the plate, or both elements, of the tube which is oscillating.

Chasing Parasitics The preceding paragraphs give a short introduction to the subject of parasitic oscillations. However, in order to insure that an amplifier is operating in a perfectly stable manner without any tendency toward parasitics, and also to eliminate any parasitics should they occur, it is wise to follow an orderly and set procedure. Such a procedure is discussed in detail in Chapter Eleven, *Transmitter Adjustment and Loading*.

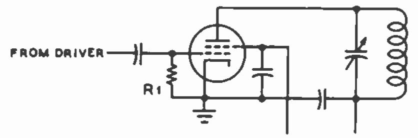


Figure 24.
GRID-LEAK BIAS.
The grid leak on an amplifier or multiplier stage may also be used as the shunt feed impedance to the grid of the tube when a high value of grid leak (greater than perhaps 20,000 ohms) is used. When a lower value of grid leak is to be employed, an r-f choke should be used between the grid of the tube and the grid leak to reduce r-f losses in the grid leak resistance.

7-12 Grid Bias

Radio-frequency amplifiers require some form of *grid bias* for proper operation. Practically all r-f amplifiers operate in such a manner that plate current flows in the form of short pulses which have a duration of only a fraction of an r-f cycle. To accomplish this with a sinusoidal excitation voltage, the operating grid bias must be at least sufficient to cut off the plate current. In very high efficiency Class C amplifiers the operating bias may be many times the cutoff value. Cutoff bias, it will be recalled, is that value of grid voltage which will reduce the plate current to zero at the plate voltage employed. The method for calculating it has been indicated previously. This theoretical value of cutoff will not reduce the plate current completely to zero, due to the variable-μ tendency or "knee" which is characteristic of all tubes as the cutoff point is approached.

Class C Bias Amplitude modulated Class C amplifiers should be operated with the grid bias adjusted to a value greater than twice cutoff at the operating plate voltage. This procedure will insure that the tube is operating at a bias greater than cutoff when the plate voltage is doubled on positive modulation peaks. C-w telegraph and FM transmitters can be operated with bias as low as

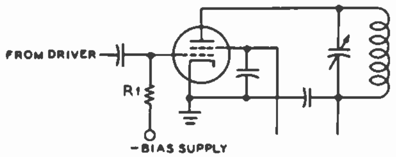


Figure 25.
COMBINATION GRID-LEAK AND FIXED BIAS.

Grid-leak bias often is used in conjunction with a fixed minimum value of power supply bias. This arrangement permits the operating bias to be established by the excitation energy, but in the absence of excitation the electrode currents to the tube will be held to safe values by the fixed-minimum power supply bias. If a relatively low value of grid leak is to be used, an r-f choke should be connected between the grid of the tube and the grid leak as discussed in figure 24.

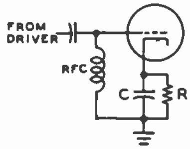


Figure 26.
R-F STAGE WITH CATHODE BIAS.

Cathode bias sometimes is advantageous for use in an r-f stage that operates with a relatively small amount of r-f excitation.

cutoff, if only limited excitation is available and moderate plate efficiency is satisfactory. In a c-w transmitter, the bias supply or resistor should be adjusted to the point which will allow normal grid current to flow for the particular amount of grid driving r-f power available. This form of adjustment will allow more output from the under-excited r-f amplifier than when higher bias is used with correspondingly lower values of grid current. In any event, the operating bias should be set at as low a value as will give satisfactory operation, since harmonic generation in a stage increases rapidly as the bias is increased.

Grid-Leak Bias A resistor can be connected in the grid circuit of a Class C amplifier to provide grid-leak bias. This resistor, R_1 in figure 24, is part of the d-c path in the grid circuit.

The r-f excitation applied to the grid circuit of the tube causes a pulsating direct current to flow through the bias supply lead, due to the rectifying action of the grid, and any current flowing through R_1 produces a voltage drop across that resistor. The grid of the tube is positive for a short duration of each r-f cycle, and draws electrons from the filament or cathode of the tube during that time. These electrons complete the circuit through the d-c *grid return*. The voltage drop across the resistance in the grid return provides a *negative bias* for the grid.

Grid-leak bias automatically adjusts itself over fairly wide variations of r-f excitation. The value of grid-leak resistance should be such that normal values of grid current will flow at the maximum available amount of r-f excitation. Grid-leak bias cannot be used for grid-modulated or linear amplifiers in which the average d-c grid current is constantly varying with modulation.

Safety Bias Grid-leak bias alone provides no protection against excessive plate current in case of failure of the source of r-f grid excitation. A *C-battery* or *C-bias* supply can be connected in series with the grid leak, as shown in figure 25. This fixed "protective" bias will protect the tube in the event of failure of grid excitation. "Zero-bias" tubes do not require this bias source in addition to the grid leak, since their plate current will drop to a safe value when the excitation is removed.

Cathode Bias A resistor can be connected in series with the cathode or center-tapped filament lead of an amplifier to secure *automatic bias*. The plate current flows through this resistor, then back to the cathode or filament, and the voltage drop across the resistor can be applied to the grid circuit by connecting the grid bias lead to the grounded or power supply end of the resistor R, as shown in figure 26.

The grounded (B-minus) end of the cathode resistor is negative relative to the filament by an amount equal to the voltage drop across the resistor. The value of resistance must be so chosen that the sum of the desired grid and plate current flowing through the resistor will bias the tube for proper operation.

This type of bias is used more extensively in audio-frequency than in radio-frequency amplifiers. The voltage drop across the resistor must be subtracted from the total plate supply

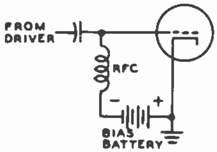


Figure 27.
R-F STAGE WITH BATTERY BIAS.
Battery bias is seldom used, due to deterioration of the cells by the reverse grid current. However, it may be used in certain special applications, or the fixed bias voltage may be supplied by a bias power supply.

voltage when calculating the power input to the amplifier, and this loss of plate voltage in an r-f amplifier may be excessive. A Class A audio amplifier is biased only to approximately one-half cutoff, whereas an r-f amplifier may be biased to twice cutoff, or more, and thus the plate supply voltage loss may be a large percentage of the total available voltage when using low or medium μ tubes.

Oftentimes just enough cathode bias is employed in an r-f amplifier to act as safety bias to protect the tubes in case of excitation failure, with the rest of the bias coming from a grid leak.

Separate Bias Supply An external supply often is used for grid bias, as shown in figure 27. Battery bias gives very good voltage regulation and is satisfactory for grid-modulated or linear amplifiers, which operate at low grid current. In the case of Class C amplifiers which operate with high grid current, battery bias is not satisfactory. This direct current has a charging effect on the dry batteries; after a few months of service the cells will become unstable, bloated, and noisy.

A separate a-c operated power supply is commonly used for grid bias. The bleeder resistance across the output of the filter can be made sufficiently low in value that the grid current of the amplifier will not appreciably change the amount of negative grid-bias voltage. Alternately, a voltage regulated grid-bias supply as described in Chapter Twenty-five can be used. This type of bias supply is used in Class B audio and Class B r-f linear amplifier service where the voltage regulation in the C-bias supply is important. For a Class C amplifier, regulation is not so important, and an economical design of components in the power supply, therefore, can be utilized. In this case, the bias voltage must be adjusted with normal grid current flowing, as the grid current will raise the bias considerably when it is flowing through the bias-supply bleeder resistance.

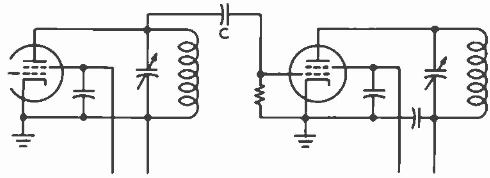


Figure 28.
CAPACITIVE INTERSTAGE COUPLING.

7-13 Interstage Coupling

Energy is usually coupled from one circuit of a transmitter into another either by *capacitive coupling, inductive coupling,* or *link coupling.* The latter is a special form of inductive coupling. The choice of a coupling method depends upon the purpose for which it is to be used.

Capacitive Coupling Capacitive coupling between an amplifier or doubler circuit and a preceding driver stage is shown in figure 28. The coupling capacitor, C, isolates the d-c plate supply from the next grid and provides a low impedance path for the r-f energy between the tube being driven and the driver tube. This method of coupling is simple and economical for low power amplifier or exciter stages, but has certain disadvantages, particularly for high frequency stages. The grid leads in an amplifier should be as short as possible, but this is difficult to attain in the physical arrangement of a high power amplifier with respect to a capacitively-coupled driver stage.

Disadvantages of Capacitive Coupling One significant disadvantage of capacitive coupling is the difficulty of adjusting the load on the driver stage. Impedance adjustment can be accomplished by tapping the coupling lead a part of the way down on the plate coil of the tuned stage of the driver circuit; but often when this is done

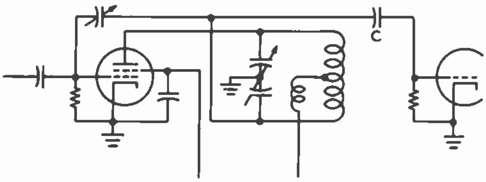


Figure 29.
BALANCED CAPACITIVE COUPLING.
Balanced capacitive coupling sometimes is useful when it is desirable to use a relatively large inductance in the interstage tank circuit, or where the exciting stage is neutralized as shown above.

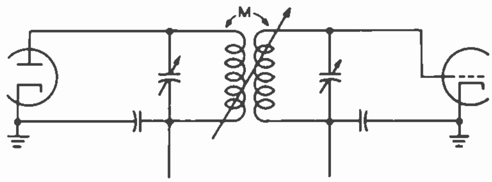


Figure 30.
INDUCTIVE INTERSTAGE COUPLING.

a parasitic oscillation will take place in the stage being driven.

One main disadvantage of capacitive coupling lies in the fact that the grid-to-filament capacitance of the driven tube is placed directly across the driver tuned circuit. This condition sometimes makes the r-f amplifier difficult to neutralize, and the increased minimum circuit capacitance makes it difficult to use a reasonable size coil in the v-h-f range. Difficulties from this source can be partially eliminated by using a center-tapped or split-stator tank circuit in the plate of the driver stage, and coupling capacitively to the opposite end from the plate. This method places the plate-to-filament capacitance of the driver across one-half of the tank and the grid-to-filament capacitance of the following stage across the other half. This type of coupling is shown in figure 29.

Capacitive coupling can be used to advantage in reducing the total number of tuned circuits in a transmitter so as to conserve space and cost. It also can be used to advantage between stages for driving beam tetrode or pentode amplifier or doubler stages.

Inductive Coupling *Inductive coupling* (figure 30) results when two coils are electromagnetically coupled to one another. The degree of coupling is controlled by varying the mutual inductance of the two coils, which is accomplished by changing the spacing or the relationship between the axes of the coils.

Inductive coupling is used extensively for coupling r-f amplifiers in radio receivers. However, the mechanical problems involved in adjusting the degree of coupling limit the usefulness of direct inductive coupling in transmitters. Either the primary or the secondary or both coils may be tuned.

Unity Coupling If the grid tuning capacitor of figure 30 is removed and the coupling increased to the maximum practicable value by interwinding the turns of the two coils, the circuit insofar as r.f. is concerned acts like that of figure 28, in which one tank serves both as plate tank for the driver and grid tank for the driven stage. The interwound grid winding serves simply to isolate the d-c plate voltage of the driver from the grid of the driven stage, and to provide a return for d-c grid current. This type of coupling, illustrated in figure 31, is commonly known as "unity coupling."

Because of the high mutual inductance, both primary and secondary are resonated by the one tuning capacitor.

Link Coupling A special form of inductive coupling which is widely employed in radio transmitter circuits is known as *link coupling*. A low impedance r-f transmission line couples the two tuned circuits together. Each end of the line is terminated in

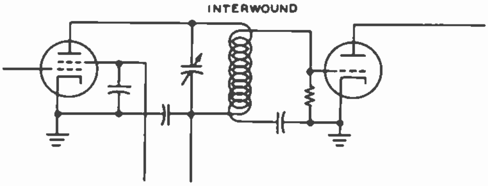


Figure 31.
"UNITY" INDUCTIVE COUPLING.
Due to the high value of coupling between the two coils, one tuning capacitor tunes both circuits. This arrangement often is useful in coupling from a single-ended to a push-pull stage.

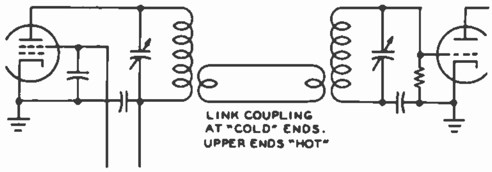


Figure 32.
INTERSTAGE COUPLING BY MEANS OF A "LINK".
Link interstage coupling is very commonly used since the two stages may be separated by a considerable distance, since the amount of coupling between the two stages may be easily varied, and since the capacitances of the two stages may be isolated to permit use of larger inductances in the v-h-f range.

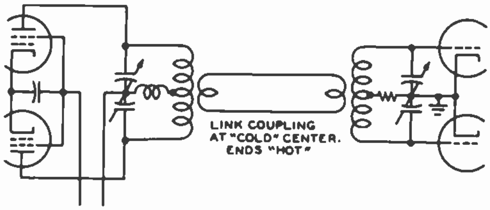


Figure 33.
PUSH-PULL LINK COUPLING.

one or more turns of wire, or *loops,* wound around the coils which are being coupled together. These loops should be coupled to each tuned circuit at the point of zero r-f potential, or *nodal point.* A ground connection to one side of the link usually is used to reduce harmonic coupling, or where capacitive coupling between two circuits must be minimized. Coaxial line is commonly used to transfer energy between the two coupling loops, although Twin-Lead may be used where harmonic attenuation is not so important.

Typical link coupled circuits are shown in figures 32 and 33. Some of the advantages of link coupling are the following:

(1) It eliminates coupling taps on tuned circuits.

(2) It permits the use of series power supply connections in both tuned grid and tuned plate circuits, and thereby eliminates the need of shunt-feed r-f chokes.

(3) It allows considerable separation between transmitter stages without appreciable r-f losses or stray chassis currents.

(4) It reduces capacitive coupling and thereby makes neutralization more easily attainable in r-f amplifiers.

(5) It provides semi-automatic impedance matching between plate and grid tuned circuits, with the result that greater grid drive can be obtained in comparison to capacitive coupling.

(6) It effectively reduces the coupling of harmonic energy.

The link-coupling line and loops can be made of no. 18 push-back wire for coupling between low-power stages. For coupling between higher powered stages the 150-ohm Twin-Lead transmission line is quite effective and has very low loss. Coaxial transmission is most satisfactory between high powered amplifier stages, and should always be used where harmonic attenuation is important.

7-14 Radio-Frequency Chokes

Radio-frequency chokes are connected in circuits for the purpose of stopping the passage of r-f energy while still permitting a direct current or audio-frequency current to pass. They consist of inductances wound with a large number of turns, either in the form of a solenoid, a series of solenoids, a single universal pie winding, or a series of pie windings. These inductors are designed to have as much inductance and as little distributed or shunt capacitance as possible. The unavoidable small amount of distributed capacitance resonates the inductance, and this frequency normally should be much lower than the frequency at which the transmitter or receiver circuit is operating. R-f chokes for operation on several bands must be designed carefully so that the impedance of the choke will be extremely high (several hundred thousand ohms) in each of the bands.

The direct current which flows through the r-f choke largely determines the size of wire to be used in the winding. The inductance of r-f chokes for the v-h-f range is much less than for chokes designed for broadcast and ordinary short-wave operation. A very high inductance r-f choke has more distributed capacitance than a smaller one, with the result

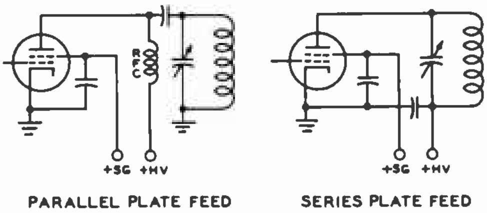


Figure 34.
ILLUSTRATING PARALLEL AND SERIES PLATE FEED.

Parallel plate feed is desirable from a safety standpoint since the tank circuit is at ground potential with respect to d.c. However, a high-impedance r-f choke is required, and the r-f choke must be able to withstand the peak r-f voltage output of the tube. Series plate feed eliminates the requirement for a high-performance r-f choke, but requires the use of a relatively large value of by-pass capacitance at the bottom end of the tank circuit, as contrasted to the moderate value of coupling capacitance which may be used at the top of the tank circuit for parallel plate feed.

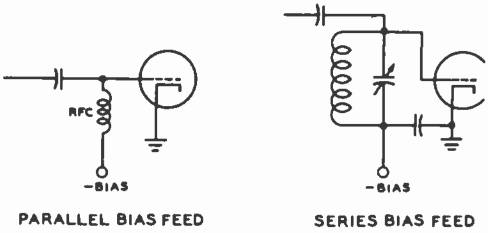


Figure 35.
ILLUSTRATING SERIES AND PARALLEL BIAS FEED.

that it will actually offer *less* impedance at very high frequencies.

Another consideration, just as important as the amount of d.c. the winding will carry, is the r-f voltage which may be placed across the choke without its breaking down. This is a function of insulation, turn spacing, frequency, number and spacing of pies and other factors.

Some chokes which are designed to have a high impedance over a very wide range of frequency are, in effect, really two chokes: a u-h-f choke in series with a high-frequency choke. A choke of this type is polarized; that is, it is important that the correct end of the combination choke be connected to the "hot" side of the circuit.

Shunt and Series Feed Direct-current grid and plate connections are made either by *series* or *parallel feed* systems. Simplified forms of each are shown in figures 34 and 35.

Series feed can be defined as that in which the d-c connection is made to the grid or plate circuits at a point of very low r-f potential. Shunt feed always is made to a point of high r-f voltage and always requires a high impedance r-f choke or a relatively high resistance to prevent waste of r-f power.

7-15 Parallel and Push-Pull Tube Circuits

The comparative r-f power output from parallel or push-pull operated amplifiers is the same if proper impedance matching is accomplished, if sufficient grid excitation is available in both cases, and if the frequency of measurement is considerably lower than the frequency limit of the tubes.

Parallel Operation Operating tubes in parallel has some advantages in transmitters designed for operation below 10 Mc., particularly when tetrode or pentode tubes are to be used. Only one neutralizing capacitor is required for parallel operation of triode tubes, as against two for push-pull. Above about 10 Mc., depending upon the tube type, parallel tube operation is ordinarily not recommended with triode tubes. However, parallel operation of grounded-grid stages and stages using low-C beam tetrodes often will give excellent results well into the v-h-f range.

Push-Pull Operation The push-pull connection provides a well-balanced circuit insofar as miscellaneous capacitances are concerned; in addition, the circuit can be neutralized more completely, especially in high-frequency amplifiers. The L/C ratio in a push-pull amplifier can be made higher than in a plate-neutralized parallel-tube operated amplifier. Push-pull amplifiers, when perfectly balanced, have less second-harmonic output than parallel or single-tube amplifiers, but in practice undesired capacitive coupling and circuit unbalance more or less offset the theoretical harmonic-reducing advantages of push-pull r-f circuits.

CHAPTER EIGHT

Amplitude Modulation

If the output of a c-w transmitter is by some means varied in amplitude at an audio frequency rate instead of interrupted in accordance with code characters, a tone will be heard on a receiver tuned to the signal. If the audio signal consists of a band of audio frequencies comprising voice or music intelligence, then the voice or music which is superimposed on the radio frequency carrier will be heard on the receiver.

When voice, music, video, or other intelligence is superimposed on a radio frequency carrier by means of a corresponding variation in the *amplitude* of the radio frequency output of a transmitter, *amplitude modulation* is the result. Telegraph keying of a c-w transmitter is the simplest form of amplitude modulation, while video modulation in a television transmitter represents a highly complex form. Systems for modulating the amplitude of a carrier envelope in accordance with voice, music, or similar types of complicated audio waveforms are many and varied, and will be discussed later on in this chapter.

Sidebands Modulation is essentially a form of *mixing* or *combining* already covered in a previous chapter. To transmit voice at radio frequencies by means of amplitude modulation, the voice frequencies are mixed with a radio frequency carrier so that the voice frequencies are converted to radio frequency *sidebands*. Though it may be difficult to visualize, *the amplitude of the radio frequency carrier does not vary during conventional amplitude modulation.*

Even though the amplitude of radio frequency voltage representing the composite signal (resultant of the carrier and sidebands, called the "envelope") will vary from zero to twice the unmodulated signal value during full modulation, the amplitude of the *carrier* component does not vary. Also, so long as the amplitude of the modulating voltage does not vary, the amplitude of the sidebands will remain constant. For this to be apparent, however, it is necessary to measure the amplitude of each component with a highly selective filter. Otherwise, the measured power or voltage will be a *resultant* of two or more of the components, and the amplitude of the resultant will vary at the modulation rate.

If a carrier frequency of 5000 kc. is modulated by a pure tone of 1000 cycles, or 1 kc., two sidebands are formed: one at 5001 kc. (the sum frequency) and one at 4999 kc. (the difference frequency). The frequency of each sideband is independent of the amplitude of the modulating tone, or *modulation percentage;* the frequency of each sideband is determined only by the frequency of the modulating tone. This assumes, of course, that the transmitter is not modulated in excess of its linear capability.

When the modulating signal consists of multiple frequencies, as is the case with voice or music modulation, two sidebands will be formed by each modulating frequency (one on each side of the carrier), and the radiated signal will consist of a *band* of frequencies. The *band width*, or *channel* taken up in the frequency spectrum by a conventional double-sideband amplitude-modulated signal, is equal to twice the highest modulating frequency. For example, if the highest modulating frequency is 5000 cycles, then the signal (assuming modulation of complex and varying waveform) will occupy a band extending from 5000 cycles below the carrier to 5000 cycles above the carrier.

Frequencies up to at least 2500 cycles, and preferably 3500 cycles, are necessary for good speech intelligibility. If a filter is incorporated in the audio system to cut out all frequencies above approximately 3000 cycles, the band width of a radio-telephone signal can be limited to 6 kc. without a significant loss in intelligibility. However, if harmonic distortion is introduced subsequent to the filter, as would happen in the case of an overloaded modulator or overmodulation of the carrier, new frequencies will be generated and the signal will occupy a band wider than 6 kc.

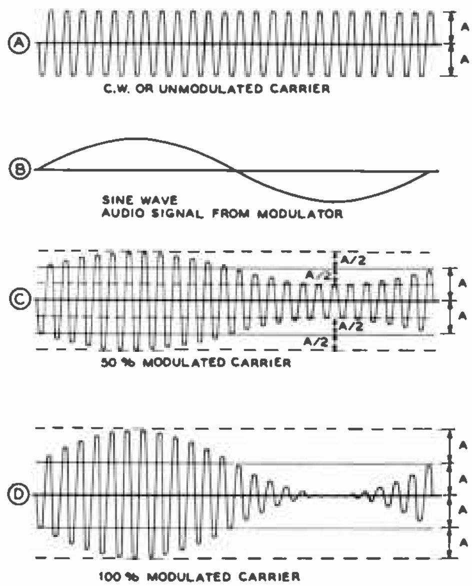


Figure 1.
AMPLITUDE MODULATED CARRIER WAVE.

Top drawing (A) represents an unmodulated carrier wave; (B) shows the audio output of the modulator. Drawing (C) shows the audio signal impressed on the carrier wave to the extent of 50 per cent modulation; (D) shows the carrier with 100 per cent amplitude modulation.

8-1 Mechanics of Modulation

A c-w or unmodulated r-f carrier wave is represented in figure 1A. An audio frequency sine wave is represented by the curve of figure 1B. When the two are combined or "mixed," the carrier is said to be amplitude modulated, and a resultant similar to 1C or 1D is obtained. It should be noted that under modulation, each half cycle of r-f voltage differs slightly from the preceding one and the following one; therefore at no time during modulation is the r-f waveform a pure sine wave. This is simply another way of saying that during modulation, the transmitted r-f energy no longer is confined to a single radio frequency.

It will be noted that the *average* amplitude of the peak r-f voltage, or modulation envelope, is the same with or without modulation. This simply means that the modulation is symmetrical (assuming a symmetrical modulating wave) and that for distortionless modulation the upward modulation is limited to a value of twice the unmodulated carrier wave amplitude because the amplitude cannot go below zero on downward portions of the modulation cycle. Figure 1D illustrates the maximum obtainable distortionless modulation with a sine modulating wave, the r-f voltage at the peak of the r-f cycle varying from zero to twice the unmodulated value, and the r-f power varying from zero to four times the unmodulated value (because the power varies as the square of the voltage).

While the average r-f *voltage* of the modulated wave over a modulation cycle is the same as for the unmodulated carrier, the average *power* increases with modulation. If the radio frequency power is integrated over the audio cycle, it will be found that with 100 per cent sine wave modulation the average r-f power has increased 50 per cent. This additional power is represented by the sidebands, because as previously mentioned, the carrier power does not vary under modulation. Thus, when a 100-watt carrier is modulated 100 per cent by a sine wave, the total r-f power is 150 watts; 100 watts in the carrier and 25 watts in each of the two sidebands.

Modulation Percentage So long as the *relative proportion* of the various sidebands making up voice modulation is maintained, the signal may be received and detected without distortion. However, the higher the average amplitude of the sidebands, the greater the audio signal produced at the receiver. For this reason it is desirable to increase the *modulation percentage,* or degree of modulation, to the point where maximum peaks just hit 100 per cent. If the modulation percentage is increased so that the peaks exceed this value, distortion is introduced, and if carried very far, bad interference to signals on nearby channels will result.

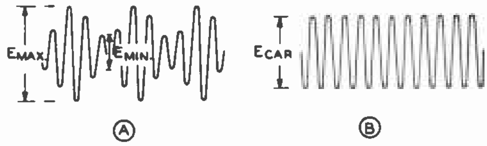


Figure 2.
GRAPHICAL DETERMINATION OF MODULATION PERCENTAGE.
The procedure for determining modulation percentage from the peak voltage points indicated is discussed in the text.

Modulation Measurement The amount by which a carrier is being modulated may be expressed either as a modulation factor, varying from zero to 1.0 at maximum modulation, or as a percentage. The percentage of modulation is equal to 100 times the modulation factor. Figure 2A shows a carrier wave modulated by a sine-wave audio tone. A picture such as this might be seen on the screen of a cathode-ray oscilloscope with sawtooth sweep on the horizontal plates and the modulated carrier impressed on the vertical plates. The same carrier without modulation would appear on the oscilloscope screen as figure 2B.

The percentage of modulation of the positive peaks and the percentage of modulation of the negative peaks can be determined separately from two oscilloscope pictures such as shown.

The modulation factor of the positive peaks may be determined by the formula:

$$M = \frac{E_{max} - E_{car}}{E_{car}}$$

The factor for negative peaks may be determined from this formula:

$$M = \frac{E_{car} - E_{min}}{E_{car}}$$

In the two above formulas E_{max} is the maximum carrier amplitude with modulation and E_{min} is the minimum amplitude; E_{car} is the steady-state amplitude of the carrier without modulation. Since the deflection of the spot on a cathode-ray tube is linear with respect to voltage, the relative voltages of these various amplitudes may be determined by measuring the deflections, as viewed on the screen, with a rule calibrated in inches or centimeters. The percentage of modulation of the carrier may be had by multiplying the modulation factor thus obtained by 100. The above procedure assumes that there is no *carrier shift,* or change in average amplitude, with modulation.

If the modulating voltage is symmetrical, such as a sine wave, and modulation is accomplished without the introduction of distortion, then the percentage modulation will be the same for both negative and positive peaks. However, the distribution and phase relationships of harmonics in voice and music waveforms are such that the percentage modulation of the negative modulation peaks may exceed the percentage modulation of the positive peaks, and vice versa. The percentage modulation when referred to without regard to polarity is an indication of the average of the negative and positive peaks.

Modulation Capability The modulation capability of a transmitter is the maximum percentage to which that transmitter may be modulated before spurious sidebands are generated in the output or before the distortion of the modulating waveform becomes objectionable. The highest modulation capability which *any* transmitter may have on the *negative* peaks is 100 per cent. The maximum permissible modulation of many transmitters is less than 100 per cent, especially on positive peaks. The modulation capability of a transmitter may be limited by tubes with insufficient filament emission, by insufficient excitation or grid bias to a plate-modulated stage, too light loading of any type of amplifier carrying modulated r.f., insufficient power output capability in the modulator, or too much excitation to a grid-modulated stage or a Class B linear amplifier. In any case, the FCC regulations specify that no transmitter be modulated in excess of its modulation capability. Hence, it is desirable to make the modulation capability of a transmitter as near as possible to 100 per cent so that the carrier power may be used most effectively.

Speech Waveform Dissymmetry The manner in which the human voice is produced by the vocal cords gives rise to a certain dissymmetry in the waveform of voice sounds when they are picked up by a good-quality microphone. This is especially pronounced in the male voice, and more so on certain voiced sounds than on others. The result of this dissymmetry in the waveform is that the voltage peaks on one side of the average value of the wave will be considerably greater, often two or three times as great, as the voltage excursions on the other side of the zero axis. The *average* value of voltage on both sides of the wave is, of course, the same.

As a result of this dissymmetry in the male voice waveform, there is an optimum polarity of the modulating voltage that must be observed if maximum sideband energy is to be obtained without negative peak clipping and generation of "splatter" on adjacent channels.

A double-pole double-throw "phase reversing" switch in the input or output leads of any transformer in the speech amplifier system will permit poling the extended peaks in the direction of maximum modulation capability. The optimum polarity may be determined easily by listening on a selective receiver tuned to a frequency 30 to 50 kc. removed from the desired signal and adjusting the phase reversing switch to the position which gives the least "splatter" when the transmitter is modulated rather heavily. If desired, the switch then may be replaced with permanent wiring, so long as the microphone and speech system are not to be changed.

A more conclusive illustration of the lopsidedness of a speech waveform may be obtained by observing the modulated waveform of a radiotelephone transmitter on an oscilloscope. A portion of the carrier energy of the transmitter should be coupled by means of a link directly to the vertical plates of the 'scope, and the horizontal sweep should be a sawtooth or similar wave occurring at a rate of approximately 30 to 70 sweeps per second.

With the speech signal from the speech amplifier connected to the transmitter in one polarity it will be noticed that negative-peak clipping—as indicated by bright "spots" in the center of the 'scope pattern whenever the carrier amplitude goes to zero—will occur at a considerably lower level of average modulation than with the speech signal being fed to the transmitter in the other polarity. When the input signal to the transmitter is polarized in such a manner that the "fingers" of the speech wave extend in the direction of positive modulation these fingers usually will be clipped in the plate circuit of the modulator at an acceptable peak modulation level.

The use of the proper polarity of the incoming speech wave in modulating a transmitter can afford an increase of approximately two to one in the amount of speech audio power which may be placed upon the carrier of an amplitude-modulated transmitter for the same amount of sideband splatter. More effective methods for increasing the amount of audio power on the carrier of an AM phone transmitter are discussed in Section 8-5 at the end of this chapter.

Single-Sideband Transmission Because the same intelligibility is contained in each of the sidebands associated with a modulated carrier, it is not necessary to transmit sidebands on both sides of the carrier. Also, because the carrier is simply a single radio frequency wave of unvarying amplitude, it is not necessary to transmit the carrier if some means is provided for inserting a locally generated carrier at the receiver.

When the carrier is suppressed but both upper and lower sidebands are transmitted, it is necessary to insert a locally generated carrier at the receiver of *exactly* the same frequency and phase as the carrier which was suppressed. For this reason, suppressed-carrier double-sideband systems have little practical application.

When the carrier is suppressed and only the upper *or* the lower sideband is transmitted, a highly intelligible signal may be obtained at the receiver even though the locally generated carrier differs a few cycles from the frequency of the carrier which was suppressed at the transmitter. A communications system utilizing but one group of sidebands with carrier suppressed is known as a "single sideband" system. Such systems are widely used for commercial point to point work, and are being used to an increasing extent in amateur communication. The two chief advantages of the system are: (1) an effective power gain of about 9 db results from putting all the radiated power in intelligence carrying sideband frequencies instead of mostly into radiated carrier, and (2) elimination of the selective fading and distortion that normally occurs in a conventional double-sideband system when the carrier fades and the sidebands do not, or the sidebands fade differently.

SYSTEMS OF AMPLITUDE MODULATION

There are many different systems and methods for amplitude modulating a carrier, but most may be grouped under three general classifications: (1) *variable efficiency* systems in which the average input to the stage remains constant with and without modulation and the variations in the efficiency of the stage in accordance with the modulating signal accomplish the modulation; (2) *constant efficiency* systems in which the input to the stage is varied by an external source of modulating energy to accomplish the modulation; and (3) so-called *high-efficiency* systems in which circuit complexity is increased to obtain high plate circuit efficiency in the modulated stage without the requirement of an external high-level modulator. The various systems under each classification have individual characteristics which make certain ones best suited to particular applications.

8-2 Variable Efficiency Modulation

Since the *average* input remains constant in a stage employing variable efficiency modulation, and since the average power output of the stage increases with modulation, the additional average power output from the stage *with* modulation must come from the plate dissipation of the tubes in the stage. Hence, for the best relation between tube cost and power output the tubes employed should have as high a plate dissipation rating per dollar as possible.

The plate efficiency in such an amplifier is doubled when going from the unmodulated condition to the peak of the modulation cycle. Hence, the unmodulated efficiency of such an amplifier must always be less than 45 per cent, since the maximum peak efficiency obtainable in a conventional amplifier is in the vicinity of 90 per cent. Since the peak efficiency in certain types of amplifiers will be as low as 60 per cent, the unmodulated efficiency in such amplifiers will be in the vicinity of 30 per cent.

Assuming a typical amplifier having a peak efficiency of 70 per cent, the following figures give an idea of the operation of an idealized efficiency-modulated stage adjusted for 100 per cent sine-wave modulation. It should be kept in mind that the plate voltage is constant at all times, even over the audio cycles.

Plate input without modulation	100 watts
Output without modulation	35 watts
Efficiency without modulation	35%
Input on 100% positive modulation peak (plate current doubles)	200 watts
Efficiency on 100% positive peak	70%
Output on 100% positive modulation peak	140 watts
Input on 100% negative peak	0 watts
Efficiency on 100% negative peak	0%
Output on 100% negative peak	0 watts
Average input with 100% modulation	100 watts
Average output with 100% modulation (35 watts carrier plus 17.5 watts sideband)	52.5 watts
Average efficiency with 100% modulation	52.5%

Systems of Efficiency Modulation There are many systems of efficiency modulation, but they *all* have the general limitation discussed in the previous paragraph—so long as the carrier amplitude is to remain constant with and without modulation; the efficiency at carrier level must be not greater than one-half the peak modulation efficiency if the stage is to be capable of 100 per cent modulation.

The classic example of efficiency modulation is the Class B linear r-f amplifier, to be discussed below. The other three common forms of efficiency modulation are control-grid modulation, screen-grid modulation, and suppressor-grid modulation. In each case, including that of the Class B linear amplifier, note that the modulation, or the modulated signal, is impressed on a control electrode of the stage.

The Class B Linear Amplifier This is the simplest practicable type of amplifier for an amplitude-modulated wave or a single-sideband signal. The system possesses the disadvantage that excitation, grid bias, and loading must be carefully controlled to preserve the linearity of the stage. Also, the grid circuit of the tube, in the usual application where grid current is drawn on peaks, presents a widely varying value of load impedance to the source of excitation. Hence it is necessary to include some sort of "swamping" resistor to reduce the effect of grid-impedance variations with modulation. If such a swamping resistance across the grid tank is not included, or is too high in value, the positive modulation peaks of the incoming modulated signal will tend to be flattened with resultant distortion of the wave being amplified.

The Class B linear amplifier has long been used in broadcast transmitters, but recently has received much more general usage in the h-f range for two significant reasons: (a) the Class B linear is an excellent way of increasing the power output of a single-sideband transmitter, since the plate efficiency with full signal will be in the vicinity of 70 per cent, while with no modulation the input to the stage drops to a relatively low value; and (b) the Class B linear amplifier operates with relatively low harmonic output since the grid bias on the stage normally is slightly less than the value which will cut off plate current to the stage in the absence of excitation.

Since a Class B linear amplifier is biased to "extended" cutoff with no excitation (the grid bias at extended cutoff will be approximately equal to the plate voltage divided by the amplification factor for a triode, and will be approximately equal to the screen voltage divided by the grid-screen mu factor for a tetrode or pentode) the plate current will flow essentially in 180-degree pulses. Due to the relatively large operating angle of plate current flow the theoretical peak plate efficiency is limited to 78.5 per cent, with 65 to 70 per cent representing a range of efficiency normally attainable, and the harmonic output will be low.

The carrier power output from a Class B linear amplifier of a normal 100 per cent modulated AM signal will be about one-half the rated plate dissipation of the stage, with optimum operating conditions. The peak output from a Class B linear, which represents the maximum-signal output as a single-sideband amplifier, or peak output with a 100 per cent AM signal, will be about twice the plate dissipation of the tubes in the stage. Thus the carrier-level input power to a Class B linear should be about 1.5 times the rated plate dissipation of the stage.

The schematic circuit of a Class B linear amplifier is the same as a conventional single-ended or push-pull stage, whether triodes or beam tetrodes are used. However, a swamping resistor, as mentioned before, must be placed across the grid tank of the stage if the operating conditions of the tube are such that appreciable grid current will be drawn on modulation peaks. Also, a *fixed* source of grid bias must be provided for the stage. A regulated grid-bias power supply is the usual source of negative bias voltage.

Adjustment of a Class B Linear Amplifier With grid bias adjusted to the correct value, and with provision for varying the excitation voltage to the stage and the loading of the plate circuit, a fully modulated signal is applied to the grid circuit of the stage. Then with an oscilloscope coupled to the output of the stage, excitation and loading are varied until the stage is drawing the normal plate input and the output waveshape is a good replica of the input signal. The adjustment procedure normally will require a succession of approximations, until the optimum set of adjustments is attained. Then the modulation being applied to the input signal should be removed to check the linearity. With modulation removed, in the case of a 100 per cent AM signal, the input to the stage should remain constant, and the peak output of the r-f envelope should fall to half the value obtained on positive modulation peaks.

Class C Grid Modulation One widely used system of efficiency modulation for communications work is Class C control-grid bias modulation. The distortion is slightly higher than for a properly operated Class B linear amplifier, but the efficiency is also higher, and the distortion can be kept within tolerable limits for communications work.

Class C grid modulation requires high plate voltage on the modulated stage, if maximum output is desired. The plate voltage is normally run about 50 per cent higher than for maximum output with plate modulation.

The driving power required for operation of a grid-modulated amplifier under these conditions is somewhat more than is required for operation at lower bias and plate voltage, but the increased power output obtainable overbalances the additional excitation requirement. Actually, almost half as much excitation is required as would be needed if the same stage were to be operated as a Class C plate-modulated amplifier. The resistor R across the grid tank of the stage serves as "swamping" to stabilize the r-f driving voltage. At least 50 per cent of the output of the driving stage should be dissipated in this swamping resistor under carrier conditions.

A comparatively small amount of audio power will be required to modulate the amplifier stage 100 per cent. An audio amplifier having 20 watts output will be sufficient to modulate an amplifier with one kilowatt input. Proportionately smaller amounts of audio will be required for lower powered stages. However, the audio amplifier that is being used as the grid modulator should, in any case, either employ low plate resistance tubes such as 2A3's, employ degenerative feedback from the output stage to one of the preceding stages of the speech amplifier, or be resistance loaded with a resistor across the secondary of the modulation transformer. This provision of low driving impedance in the grid modulator is to insure good regulation in the audio driver for the grid modulated stage. Good regulation of both the audio and the r-f drivers of a grid-modulated stage is quite important if distortion-free modulation approaching 100 per cent is desired, because the grid impedance of the modulated stage varies widely over the audio cycle.

Two circuits for obtaining grid-bias modulation are shown in figure 3. Figure 3A illustrates the conventional method utilizing a regulated grid-bias power supply and a separate audio amplifier as the grid-bias modulator. This circuit is satisfactory and gives excellent results. The circuit of figure 3B is somewhat simpler than that illustrated in (A) since the separate modulator stage has been dispensed with and the function combined with that of bias regulation in the single 6B4G tube shown. The regulator-modulator tube operates as a cathode-follower. The average d-c voltage on the control grid is controlled by the 70,000-ohm wire-wound potentiometer and hence this potentiometer adjusts the average grid bias on the modulated stage. However, a-c signal voltage is also impressed on the control-grid of the tube and since the cathode follows this a-c wave the incoming speech wave is superimposed on the average grid bias, thus effecting grid-bias modulation of the r-f amplifier stage. An audio voltage swing is required on the grid of the 6B4G of approximately the same peak value as will be required as bias-voltage swing on the grid-bias modulated stage. This voltage swing will normally be in the region from 50 to 200 peak volts. Up to about 100 volts peak swing can be obtained from a 6SJ7 tube as a conventional speech amplifier stage. The higher voltages may be obtained from a tube such as a 6J5 through an audio transformer of 2:1 or 2½:1 ratio.

With the normal amount of comparatively tight antenna coupling to the modulated stage, a non-modulated carrier efficiency of 40 per cent can be obtained with substantially distortion-free modulation up to practically 100 per cent. If the antenna coupling is decreased slightly from the condition just described, and the excitation is increased to the point where the amplifier draws the same input, carrier efficiency of 50 per cent is obtainable with tolerable distortion at 90 per cent modulation.

Tuning the Grid-Bias Modulated Stage It will be noticed, by reference to figures 3A and 3B, that a special type of bias supply for the grid-modulated stage has been incorporated as a part of the schematic of the stage in each case. This was done purposely to make it clear that a

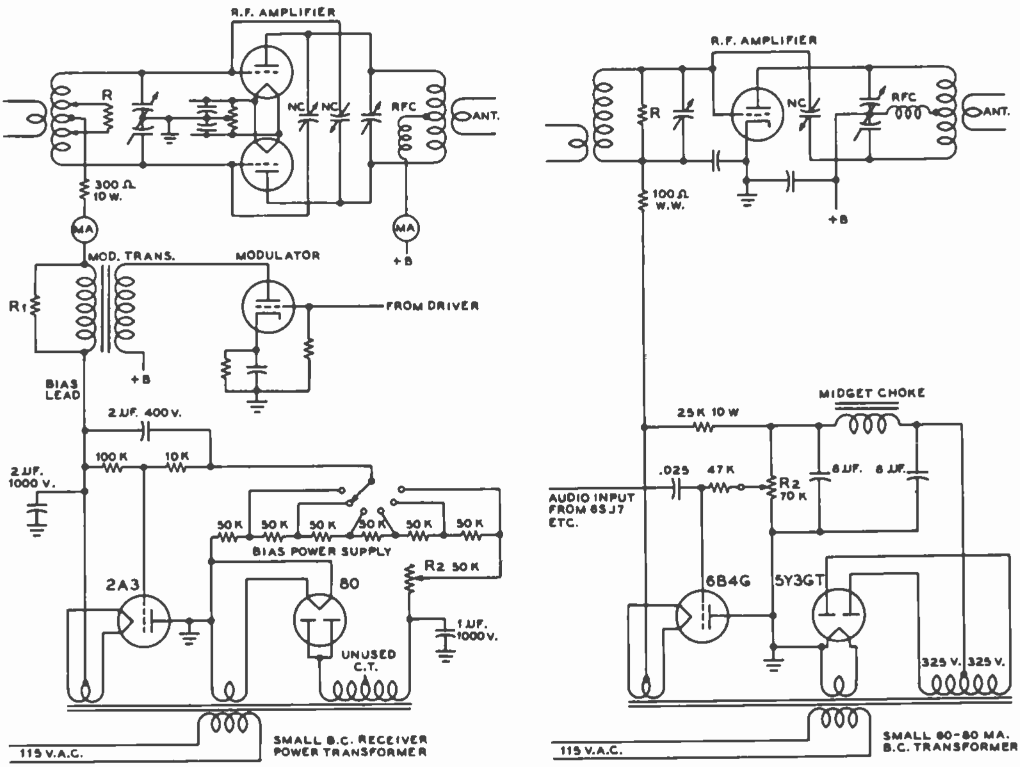


Figure 3.

CIRCUITS FOR GRID-BIAS MODULATION.

The circuits of (A) and (B) show two practical circuits for obtaining grid-bias modulation of a Class C amplifier. Both arrangements will give satisfactory results but the new circuit arrangement shown at (B) is recommended since fewer components are required, the audio power amplifier stage is eliminated, and no modulation transformer is required. The grid of the 6B4-G modulator-bias regulator tube may be excited from a low level speech stage such as a 6J5 or a 6SJ7 since this tube requires no grid current for full modulation.

special type of high-voltage bias supply is required for best operation of such an amplifier. The supply of figure 3A has very good regulation up to about 75 ma. of grid current (the maximum capability of a single 2A3) and the voltage may be varied from nearly zero to about 700 volts; also, this particular supply may be constructed quite inexpensively.

The most satisfactory procedure for tuning a stage for grid-bias modulation of the Class C type is as follows. The amplifier should first be neutralized, and any possible tendency toward parasitics under any condition of operation should be eliminated. Then the antenna should be coupled to the plate circuit, the grid bias should be run up to the maximum available value, and the plate voltage and excitation should be applied. The grid bias voltage should then be reduced until the amplifier draws the approximate amount of plate current it is desired to run, and modulation corresponding to about 80 per cent is then applied. If the plate current kicks up when modulation is applied, the grid bias should be reduced; if the plate meter kicks down, increase the grid bias.

When the amount of bias voltage has been found (by adjusting the fine control, R_2, on the bias supply) where the plate meter remains constant with modulation, it is more than probable that the stage will be drawing either too much or too little input. The an tenna coupling should then be either increased or decreased (depending on whether the input was too little or too much, respectively) until the input is more nearly the correct value. The bias should then be readjusted until the plate meter remains constant with modulation as before. By slight jockeying back and forth of antenna coupling and grid bias, a point can be reached where the tubes are running at rated plate dissipation, and where the plate milliammeter on the modulated stage remains substantially constant with modulation.

The linearity of the stage should then be checked by any of the conventional methods; the trapezoidal pattern method employing a cathode-ray oscilloscope is probably the most satisfactory. The check with the trapezoidal pattern will allow the determination of the proper amount of gain to employ on the speech amplifier. Incidentally, too much audio power on the grid of the modulated stage should not be used in the tuning-up process, as the plate meter will kick erratically and it will be impossible to make a satisfactory adjustment.

Screen-Grid Modulation Amplitude modulation may be accomplished by varying the screen-grid voltage in a Class C amplifier which employs a pentode, beam tetrode, or other type of screen-grid tube. The modulation obtained in this way is not especially linear, but screen-grid modulation does offer other advantages and the linearity is quite adequate for communications work.

There are two significant and worthwhile advantages of screen-grid modulation for communications work: (1) The excitation requirements for an amplifier which is to be modulated in the screen are not at all critical, and good regulation of the excitation voltage is not required. The normal rated grid-circuit operating conditions specified for Class C c-w operation are quite adequate for screen-grid modulation. (2) The audio modulating power requirements for screen-grid modulation are relatively low.

A screen-grid modulated r-f amplifier operates as an efficiency-modulated amplifier, the same as does a Class B linear amplifier and a grid-modulated stage. Hence, *plate circuit* loading is relatively critical as in any efficiency-modulated stage, and must be adjusted to the correct value if normal power output with full modulation capability is to be obtained. As in the case of any efficiency-modulated stage, the operating efficiency at the peak of the modulation cycle will be between 70 and 80 per cent, with efficiency at the carrier level (if the stage is operating in the normal manner with full carrier) about half the peak-modulation value.

There are two main disadvantages of screen-grid modulation, and several factors which must be considered if satisfactory operation of the screen-grid modulated stage is to be obtained. The disadvantages are: (1) As mentioned before, the linearity of modulation with respect to screen-grid voltage of such a stage is satisfactory only for communications work, unless carrier-rectified degenerative feedback is employed around the modulated stage to straighten the linearity of modulation. (2) The impedance of the screen grid to the modulating signal is non-linear. This means that the modulating signal must be obtained from a source of quite low impedance if audio distortion of the signal appearing at the screen grid is to be avoided.

Screen-Grid Impedance Instead of being linear with respect to modulating voltage, as is the plate circuit of a plate-modulated Class C amplifier, the screen grid presents approximately a square-law impedance to the modulating signal over the region of signal excursion where the screen is positive with respect to ground. This non-linearity may be explained in the following manner: At the carrier level of a conventional screen-modulated stage the plate-voltage swing of the modulated tube is one-half the voltage swing at peak-modulation level. This condition must exist in any type of conventional efficiency-modulated stage if 100 per cent positive modulation is to be attainable. Since the plate-voltage swing is at half amplitude, and since the screen voltage is at half its full-modulation value, the screen current is relatively low. But at the positive modulation peak the screen voltage is approximately doubled, and the plate-voltage swing also is at twice the carrier amplitude. Due to the increase in plate-voltage swing with increasing screen voltage, the screen current increases more than linearly with increasing screen voltage.

In a test made on an amplifier with an 813 tube, the screen current at carrier level was about 6 ma. with screen potential of 190 volts; but under conditions which represented a

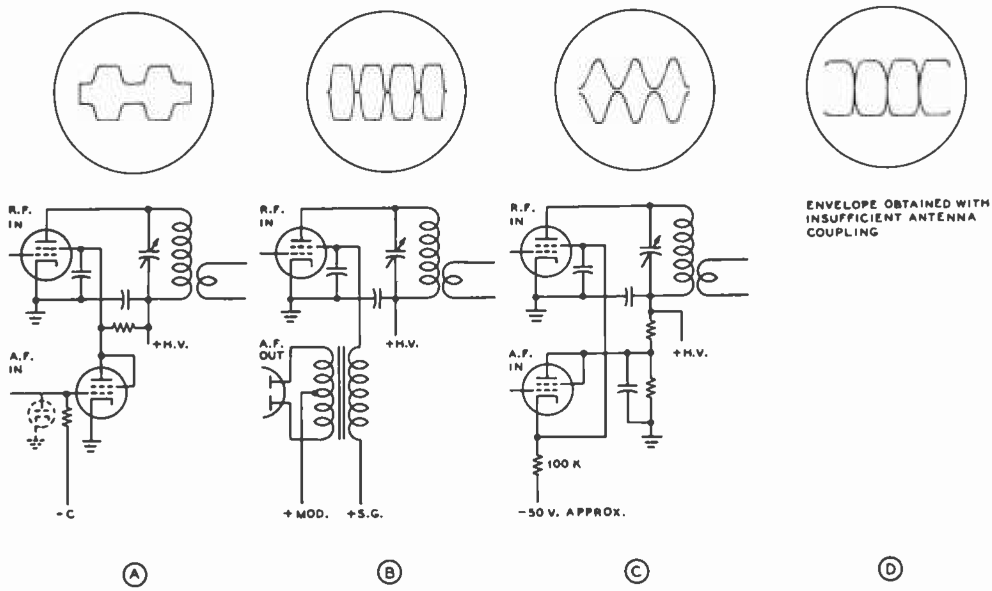


Figure 4.

SCREEN-MODULATION CIRCUITS.

Three common screen modulation circuits are illustrated above. All three circuits are capable of giving intelligible voice modulation although the waveform distortion in the circuits of (A) and (B) is likely to be rather severe. The arrangement at (A) is often called "clamp tube" screen modulation; by returning the grid leak on the clamp tube to ground the circuit will give controlled-carrier screen modulation. This circuit has the advantage that it is simple and is well suited to use in mobile transmitters. (B) is an arrangement using a transformer coupled modulator, and offers no particular advantages. The arrangement at (C) is capable of giving good modulation linearity due to the low impedance of the cathode-follower modulator. However, due to the relatively low heater-cathode ratings on tubes suited for use as the modulator, a separate heater supply for the modulator tube normally is required. This limitation makes application of the circuit to the mobile transmitter a special problem, since an isolated heater supply normally is not available. Shown at (D) as an assistance in the tuning of a screen-modulated transmitter (or any efficiency-modulated transmitter for that matter) is the type of modulation envelope which results when loading to the modulated stage is insufficient.

positive modulation peak the screen current measured 25 ma. at a potential of 400 volts. Thus instead of screen current doubling with twice screen voltage as would be the case if the screen presented a resistive impedance, the screen current became about four times as great with twice the screen voltage.

Another factor which must be considered in the design of a screen-modulated stage, if full modulation is to be obtained, is that the power output of a screen-grid stage with zero screen voltage is still relatively large. Hence, if anything approaching full modulation on negative peaks is to be obtained, the screen potential must be made negative with respect to ground on negative modulation peaks. In the usual types of beam tetrode tubes the screen potential must be 20 to 50 volts negative with respect to ground before cut-off of output is obtained. This condition further complicates the problem of obtaining good linearity in the audio modulating voltage for the screen-modulated stage, since the screen voltage must be driven negatively with respect to ground over a portion of the cycle. Hence the screen draws *no* current over a portion of the modulating cycle, and over the major portion of the cycle when the screen does draw current, it presents approximately a square-law impedance.

Circuits for Screen-Grid Modulation Laboratory analysis of a large number of circuits for accomplishing screen modulation has led to the conclusion that the audio modulating voltage *must* be obtained from a low-impedance source if low-distortion modulation is to be obtained. Figure 4 shows a group of sketches of the modulation envelope obtained with various types of modulators and also with insufficient antenna coupling. The result of this laboratory work led to the conclusion that the cathode-follower modulator of the basic circuit shown in figure 5 *is* capable of giving good-quality screen-grid modulation, and in addition the circuit provides convenient adjustments for the carrier level and the output level on *negative* modulation peaks. This latter control, P_2 in figure 5, allows the amplifier to be adjusted in such a manner that negative-peak clipping cannot take place, yet the negative modulation peaks may be adjusted to a level just above that at which sideband splatter will occur.

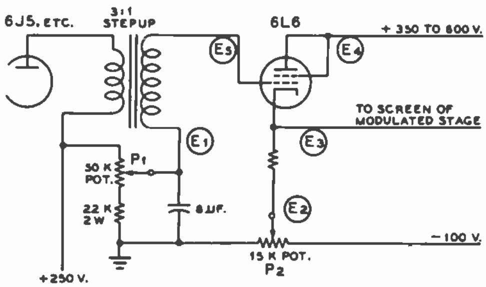


Figure 5.
CATHODE-FOLLOWER SCREEN-MODULATION CIRCUIT.
A detailed discussion of this circuit, which also is represented in figure 4C, is given in the accompanying text.

The Cathode-Follower Modulator The cathode follower is ideally suited for use as the modulator for a screen-grid stage since it acts as a relatively low-impedance source of modulating voltage for the screen-grid circuit. In addition the cathode-follower modulator allows the supply voltage both for the modulator and for the screen grid of the modulated tube to be obtained from the high-voltage supply for the plate of the screen-grid tube or beam tetrode. In the usual case the plate supply for the cathode follower, and hence for the screen grid of the modulated tube, may be taken from the bleeder on the high-voltage power supply. A tap on the bleeder may be used, or two resistors may be connected in series to make up the bleeder, with appropriate values such that the voltage applied to the plate of the cathode follower is appropriate for the tube to be modulated. It is important that a by-pass capacitor be used from the plate of the cathode-follower modulator to ground.

The voltage applied to the plate of the cathode follower should be about 100 volts greater than the rated screen voltage for the tetrode tube as a c-w Class C amplifier. Hence the cathode-follower plate voltage should be about 350 volts for an 815, 2E26, or 829B, about 400 volts for an 807 or 4-125A, about 500 volts for an 813, and about 600 volts for a 4-250A or a 4E27. Then potentiometer P_1 in figure 5 should be adjusted until the carrier-level screen voltage on the modulated stage is about one-half the rated screen voltage specified for the tube as a Class C c-w amplifier. The current taken by the screen of the modulated tube under carrier conditions will be about one-fourth the normal screen current for c-w operation.

Incidentally, the only current taken by the cathode follower itself will be that which will flow through the 100,000-ohm resistor between the cathode of the 6L6 modulator and the negative supply. Hence, the current taken from the bleeder on the high-voltage supply will be the carrier-level screen current of the tube being modulated (which current passes of course through the cathode follower) plus that current which will pass through the 100,000-ohm resistor.

The loading to the modulated stage should be adjusted until the input to the tube is about 50 per cent greater than the rated plate dissipation of the tube or tubes in the stage. If the carrier-level screen voltage value is correct for linear modulation of the stage, the loading will have to be somewhat *greater* than that amount of loading which gives maximum output from the stage. The stage may then be modulated by applying an audio signal to the grid of the cathode-follower modulator, while observing the modulated envelope on an oscilloscope.

If good output is being obtained, and the modulation envelope appears as shown in figure 4C, all is well, except that P_2 in figure 5 should be adjusted until negative modulation peaks, even with excessive modulating signal, do not cause carrier cutoff with its attendant sideband splatter. If the envelope appears as at figure 4D, antenna coupling should be increased while the carrier level is backed down by potentiometer P_1 in figure 5 until a set of adjustments is obtained which will give a satisfactory modulation envelope as shown in figure 4C.

Changing Bands After a satisfactory set of adjustments has been obtained, it is *not* difficult to readjust the amplifier for operation on different bands. Potentiometers P_1 (carrier level), and P_2 (negative peak level) may be left fixed after a satisfactory adjustment, with the aid of the scope, has once been found. Then when changing bands it is only necessary to adjust excitation until the correct value of grid current is obtained, and then to adjust antenna coupling until correct plate current is obtained. Note that the correct plate current for an efficiency-modulated amplifier is only slightly less than the out-of-resonance plate current of the stage. Hence carrier-level screen voltage must be low so that the out-of-resonance plate current will not be too high, and relatively heavy antenna coupling must be used so that the operating plate current will be near the out-of-resonance value, and so that the operating input will be slightly greater than 1.5 times the rated plate dissipation of the tube or tubes in the stage. Since the carrier efficiency of the stage will be only 35 to 40 per cent, the tubes will be operating with plate dissipation of approximately the rated value without modulation.

Speech Clipping in the Modulated Stage The maximum r-f output of an efficiency-modulated stage is limited by the maximum possible plate voltage swing on positive modulation peaks. In the modulation circuit of figure 5 the *minimum* output is limited by the minimum voltage which the screen will reach on a negative modulation peak, as set by potentiometer P_2. Hence the screen-grid-modulated stage, when using the modulator of figure 5 acts effectively as a speech clipper, provided the modulating signal amplitude is not too much more than that value which will accomplish full modulation. With correct adjustment of the operating conditions of the stage it can be made to clip positive and negative modulation peaks symmetrically. However, the inherent peak clipping ability of the stage should *not* be relied upon as a means of obtaining a large amount of speech compression, since excessive audio distortion and excessive screen current on the modulated stage will result.

Characteristics of a Typical Screen-Modulated Stage An important characteristic of the screen-modulated stage, when using the cathode-follower modulator, is that excessive plate voltage on the modulated stage is not required. In fact, full output usually may be obtained with the larger tubes at an operating plate voltage from one-half to two-thirds the maximum rated plate voltage for c-w operation. This desirable condition is the natural result of using a low-impedance source of modulating signal for the stage.

As an example of a typical screen-modulated stage, full output of 75 watts of carrier may be obtained from an 813 tube operating with a plate potential of only 1250 volts. No increase in output from the 813 may be obtained by increasing the plate voltage, since the tube may be operated with full rated plate dissipation of 125 watts, with normal plate efficiency for a screen-modulated stage, 37.5 per cent, at the 1250-volt potential.

The operating conditions of the screen-modulated 813 stage (which is illustrated in Chapter Twenty-Two) are as follows:

Plate voltage—1250 volts
Plate current—160 ma.
Plate input—200 watts
Grid current—11 ma.
Grid bias—110 volts
Carrier screen voltage—190 volts
Carrier screen current—6 ma.
Power output—approx. 75 watts

With full 100 per cent modulation the plate current decreases about 2 ma. and the screen current increases about 1 ma.; hence plate, screen, and grid current remain essentially constant with modulation. Referring to figure 5, which was the circuit used as modulator for the 813, (E_1) measured plus 155 volts, (E_2) measured −50 volts, (E_3) measured plus 190 volts, (E_4) measured plus 500 volts, and the r.m.s. swing at (E_5) for full modulation measured 210 volts, which reprsents a peak swing of about 296 volts. Due to the high positive voltage, and the large audio swing, on the cathode of the 6L6 (triode connected) modulator tube, it is important that the heater of this tube be fed from a separate filament transformer or filament winding. Note also that the operating plate-to-cathode voltage on the 6L6 modulator tube does not exceed the 360-volt rating of the tube, since the operating potential of the cathode is considerably above ground potential.

Controlled-Carrier Screen Modulation Controlled-carrier modulation, in which only sufficient carrier is transmitted to sustain the modulating signal at that moment of transmission, attained a fair degree of popularity among amateur operators during the middle 1930's. However, it never attained commercial acceptance (single-sideband transmission, having all the advantages of controlled carrier and many more, being used instead) and had substantially dropped from the amateur picture by 1940. Recently, however, controlled carrier has attained some degree of use, primarily because it offers the advantage of reduced average power drain for mobile operation.

The advantages of controlled-carrier modulation are: (1) Relatively large power output may be obtained from an efficiency modulated stage (such as a screen-modulated amplifier) since the average plate efficiency for such a stage when carrying only a modulated signal is about 55 per cent, and since the stage is operating with full input only on modulation peaks. And (2), the average power input to a controlled-carrier screen-modulated stage is much less for a given peak power output than the more usual arrangement of a Class C amplifier modulated by a Class B or Class AB_2 modulator. Both these advantages are significant for mobile work, where average power drain is of the greatest importance.

The controlled-carrier modulation system has serious disadvantages, however, all of which were discussed at length more than a decade ago. The first disadvantage is that a controlled-carrier signal is difficult to tune in the presence of interference, simply because there is no full-strength carrier to tune in on the receiver. It is necessary to tune until the voice is of greatest strength; a procedure which has proved to be quite difficult in the case of fading signals in the presence of interference. Second, severe distortion of the modulating signal will be obtained when receiving with the usual communications receiver with the a.v.c active. This distortion takes place because the a-v-c action attempts to follow the rapid excursions in carrier amplitude, but these syllabic variations normally take place at a rate greater than can be followed by the time constant of the a-v-c filter. Hence distortion and cross modulation takes place as a result of the lag between the a-v-c diode bias voltage and the instantaneous carrier amplitude. The effect can be alleviated by operating the receiver with the a.v.c. inoperative.

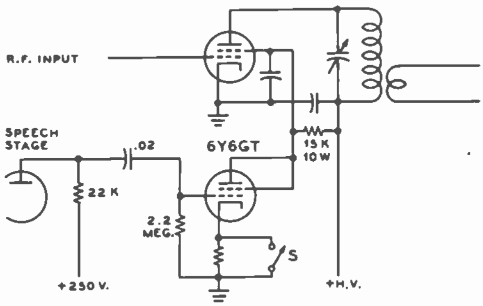


Figure 6.
CLAMP-TUBE SCREEN-MODULATION CIRCUIT.

A third disadvantage in controlled-carrier operation, which is of particular significance in mobile work, lies in the fact that the controlled-carrier stage is relatively critical as to antenna loading. This condition makes the antenna coupling adjustment of a controlled-carrier mobile transmitter more difficult than with the conventional mobile transmitter.

Circuits for Controlled-Carrier Screen Modulation Nevertheless, the advantages of controlled-carrier operation outweigh the disadvantages for certain applications. Shown in figure 6 and figure 7 are two circuits for obtaining controlled-carrier screen modulation. Figure 6 shows a circuit for obtaining "clamp-tube controlled-carrier screen modulation," with a 6Y6-GT as the clamp tube. With this arrangement the screen voltage is held down to a value in the vicinity of 50 volts in the absence of signal. Then, when signal is applied the grid of the 6Y6-GT rectifies the positive peaks of the incoming audio signal with the resultant production of a *negative* d-c bias on the grid of the 6Y6-GT. This bias, which varies in proportion to the average modulation level, permits the screen voltage on the modulated stage to rise to a value which will accommodate the modulating signal being applied.

The time constant of carrier variation is determined by the grid coupling capacitor and the grid leak on the 6Y6-GT. An additional diode may be connected between the grid of the 6Y6-GT and ground, but is not necessary since the grid of the 6Y6-GT acts satisfactorily as the audio-signal rectifier. The switch S is opened to provide cathode bias on the 6Y6-GT so that the screen voltage on the modulated stage will rise to a value which will allow tuning of the stage at the carrier level of output.

The circuit of figure 6 operates satisfactorily for some types of communication work, such as mobile operation, but the distortion is quite high (as shown in figure 4A) and negative modulation peaks do not go below the no-modulation output level of the modulated stage.

An alternative circuit which can give much better linearity, and consequently low distortion, is illustrated in figure 7. The circuit is substantially the same as that of figure 5, except that the grid of the cathode follower is returned to the negative supply instead of to a positive potential, and a diode is included to control the average bias on the cathode follower in accordance with the incoming audio signal.

Note that in each of the controlled-carrier modulation systems the diode rectifier acts as a sort of "d.c. restorer" as used in TV practice. In the circuit of figure 6 the diode (which in this case may be the grid of the tube) acts to keep the grid from going positive with respect to the cathode. In the circuit of figure 7 the diode serves to stop the grid from going negatively with respect to the negative return voltage of the cathode follower.

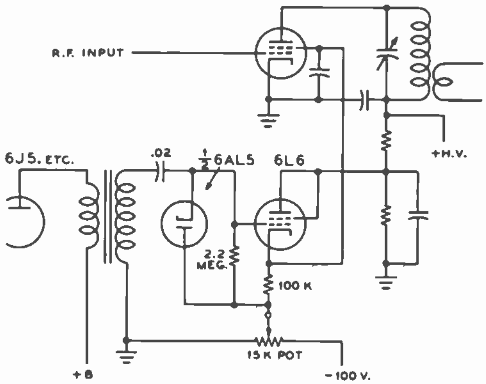


Figure 7.
CONTROLLED-CARRIER SCREEN-MODULATION CIRCUIT.

Suppressor-Grid Modulation Still another form of efficiency modulation may be obtained by applying the audio modulating signal to the suppressor grid of a pentode Class C r-f amplifier. Basically, suppressor-grid modulation operates in the same general manner as other forms of efficiency modulation; carrier plate circuit efficiency is about 35 per cent, and antenna coupling must be rather tight. However, suppressor-grid modulation has one sizeable disadvantage, in addition to the fact that pentode tubes are not nearly so widely used as beam tetrodes which of course do not have the suppressor element. This disadvantage is that the screen-grid current to a suppressor-grid modulated amplifier is rather high. The high screen current is a natural consequence of the rather high negative bias on the suppressor grid, which reduces the plate-voltage swing and plate current with a resulting increase in the screen current.

In tuning a suppressor-grid modulated amplifier, the grid bias, grid current, screen voltage, and plate voltage are about the same as for Class C c-w operation of the stage. But the suppressor grid is biased negatively to a value which reduces the plate-circuit efficiency to about one-half the maximum obtainable from the particular amplifier, with antenna coupling adjusted until the plate input is about 1.5 times the rated plate dissipation of the stage. It is important that the input to the screen grid be measured to

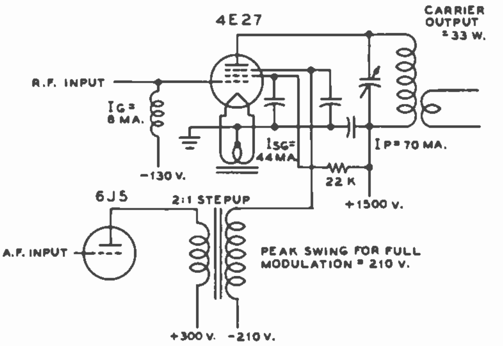


Figure 8.

AMPLIFIER WITH SUPPRESSOR-GRID MODULATION.

Recommended operating conditions for linear suppressor-grid modulation of a 4E27/257B/8001 stage are given on the drawing.

make sure that the rated screen dissipation of the tube is not being exceeded. Then the audio signal is applied to the suppressor grid. In the normal application the audio voltage swing on the suppressor will be somewhat greater than the negative bias on the element. Hence suppressor-grid current will flow on modulation peaks, so that the source of audio signal voltage must have good regulation. Tubes suitable for suppressor-grid modulation are: 2E22, HK-57, HK-257B, 4E27/8001, 5-125A, 804 and 803. A typical suppressor-grid modulated amplifier is illustrated in figure 8.

8-3 Input Modulation Systems

Constant efficiency variable-input modulation systems operate by virtue of the addition of external power to the modulated stage to effect the modulation. There are two general classifications that come under this heading; those systems in which the additional power is supplied as audio frequency energy from a modulator, usually called plate modulation systems, and those systems in which the additional power to effect modulation is supplied as direct current from the plate supply.

Under the former classification comes Heising modulation (probably the oldest type of modulation to be applied to a continuous carrier), Class B plate modulation, and series modulation. These types of plate modulation are by far the easiest to get into operation, and they give a very good ratio of power input to the modulated stage to power output; 65 to 80 per cent efficiency is the general rule. It is for these two important reasons that these modulation systems, particularly Class B plate modulation, are at present the most popular for communications work.

Modulation systems coming under the second classification are of comparatively recent development but have been widely applied to broadcast work. There are quite a few systems in this class. Two of the more widely used are the Doherty linear amplifier, and the Terman-Woodyard high-efficiency grid-modulated amplifier. Both systems operate by virtue of a carrier amplifier and a peak amplifier connected together by electrical quarter-wave lines. They will be described later in this section.

Plate Modulation Plate modulation is the application of the audio power to the *plate circuit* of an r-f amplifier. The r-f amplifier must be operated Class C for this type of modulation in order to obtain a radio-frequency output which changes in exact accordance with the variation in plate voltage. *The r-f amplifier is 100 per cent modulated when the peak a-c voltage from the modulator is equal to the d-c voltage applied to the r-f tube.* The positive peaks of audio voltage increase the instantaneous plate voltage on the r-f tube to *twice* the d-c value, and the negative peaks reduce the voltage to zero.

The instantaneous plate *current* to the r-f stage also varies in accordance with the modulating voltage. The peak alternating current in the output of a modulator must be equal to the d-c plate current of the Class C r-f stage at the point of 100 per cent modulation. This combination of change in audio voltage and current can be most easily referred to in terms of *audio power in watts.*

In a sinusoidally modulated wave, the antenna current increases approximately 22 per cent for 100 per cent modulation with a pure tone input; an r-f meter in the antenna circuit indicates this increase in antenna current. The *average power* of the r-f wave increases 50 per cent for 100 per cent modulation, the efficiency remaining constant.

This indicates that in a plate-modulated radiotelephone transmitter, the audio-frequency channel must supply this additional 50 per cent increase in average power for sine-wave modulation. If the power input to the modulated stage is 100 watts, for example, the *average power* will increase to 150 watts at 100 per cent modulation, and this additional 50 watts of power must be supplied by the *modulator* when plate modulation is used. The actual antenna power is a constant percentage of the total value of input power.

One of the advantages of plate (or power) modulation is the ease with which proper adjustments can be made in the transmitter. Also, there is less plate loss in the r-f amplifier for a given value of carrier power than with other forms of modulation, because the plate efficiency is higher.

By properly matching the plate impedance of the r-f tube to the output of the modulator, the ratio of voltage and current swing to d-c voltage and current is automatically obtained. The modulator should have a peak voltage output equal to the average d-c plate voltage on the modulated stage. The modulator should also have a *peak power* output equal to the d-c plate input power to the modulated stage. The *average* power output of the modulator will depend upon the type of waveform. If the amplifier is being Heising modulated by a Class A stage, the modulator must have an average power output capability of one-half the input to the Class C stage. If the modulator is a Class B audio amplifier, the average power required of it may vary from one-quarter to more than one-half the Class C input depending upon the waveform. However, the *peak* power output of any modulator must be equal to the Class C input to be modulated. This subject is completely covered in the section *Speech Waveforms*.

Heising Modulation Heising modulation is the oldest system of plate modulation, and usually consists of a Class A audio amplifier coupled to the r-f amplifier by means of a modulation choke coil, as shown in figure 9.

The d-c plate voltage and plate current in the r-f amplifier must be adjusted to a value which will cause the plate impedance to match the output of the modulator, since the modulation choke gives a 1-to-1 coupling ratio. A series resistor, by-passed for audio frequencies by means of a capacitor, must be connected in series with the plate of the r-f amplifier to obtain modulation up to 100 per cent. The peak output voltage of a Class A amplifier does not reach a value equal to the d-c voltage applied to the amplifier and, consequently, the d-c plate voltage impressed across the r-f tube must be reduced to a value equal to the maximum available a-c peak voltage if 100% modulation is to be obtained.

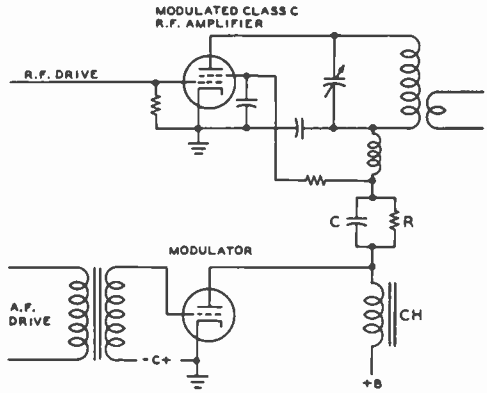


Figure 9.
HEISING PLATE MODULATION.
This type of modulation was the first form of plate modulation. It is sometimes known as "constant current" modulation. Because of the effective 1:1 ratio of the coupling choke, it is impossible to obtain 100 per cent modulation unless the plate voltage to the modulated stage is dropped slightly by resistor R. The capacitor C merely bypasses the audio around R, so that the full a-f output voltage of the modulator is impressed on the Class C stage.

A higher degree of distortion can be tolerated in low-power emergency phone transmitters which use a pentode modulator tube, and the series resistor and by-pass capacitor are usually omitted in such transmitters.

Class B Plate Modulation High-level Class B plate modulation is the least expensive method of plate modulation. Figure 10 shows a conventional Class B plate-modulated Class C amplifier.

The statement that the modulator output power must be one-half the Class C input for 100 per cent modulation is correct only

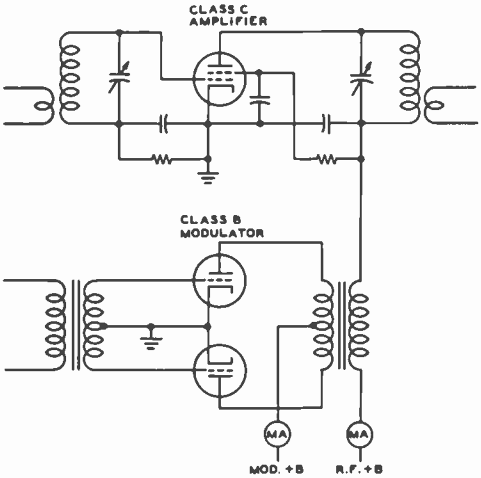


Figure 10.
CLASS B PLATE MODULATION.
This type of modulation is the most flexible in that the loading adjustment can be made in a short period of time and without elaborate test equipment after a change in operating frequency of the Class C amplifier has been made.

if the waveform of the modulating power is a *sine wave.* Where the modulator waveform is unclipped speech, the average modulator power for 100 per cent modulation is considerably less than one-half the Class C input.

Power Relations in Speech Waveforms It has been determined experimentally that the ratio of peak to average power in a speech waveform is approximately 4 to 1 as contrasted to a ratio of 2 to 1 in a sine wave. This is due to the high harmonic content of such a waveform, and to the fact that this high harmonic content manifests itself by making the wave unsymmetrical and causing sharp peaks or "fingers" of high energy content to appear. Thus for unclipped speech, the *average* modulator plate current, plate dissipation, and power output are approximately one-half the sine wave values for a given *peak* output power.

Both peak power and average power are necessarily associated with waveform. *Peak* power is just what the name implies: the power at the peak of a wave. Peak power, although of the utmost importance in modulation, is of no great significance in a-c power work, except insofar as the *average* power may be determined from the peak value of a known wave form.

There is no time element implied in the definition of peak power; peak power may be instantaneous—and for this reason average power, which is definitely associated with time, is the important factor in plate dissipation. It is possible that the peak power of a given wave form be several times the average value; for a sine wave, the peak power is twice the average value, and for unclipped speech the peak power is approximately four times the *average* value. For 100 per cent modulation, the *peak* (instantaneous) audio power must equal the Class C input, although the average power for this value of peak varies widely depending upon the modulator wave form, being greater than 50 per cent for speech that has been clipped and filtered, 50 per cent for a sine wave, and about 25 per cent for typical unclipped speech tones.

Class B Modulators A detailed discussion of the operating conditions for Class B a-f modulators has been given in Section 5-8 of Chapter Five. In addition, Table III in Chapter Five lists recommended operating conditions for a large number of tubes which are commonly used in Class B modulator stages. Data is also given in Section 5-8 for the calculation of operating conditions for tubes as Class B modulators when it is desired to operate a pair of tubes under conditions different from those normally specified.

Modulation Transformer Calculations The modulation transformer is a device for matching the load impedance of the Class C amplifier to the recommended load impedance of the Class B modulator tubes. Modulation transformers intended for communications work are usually designed to carry the Class C plate current through their secondary windings, as shown in figure 10. The manufacturer's ratings should be consulted to insure that the d-c plate current being pulled through the secondary winding does not exceed the maximum rating.

A detailed discussion of the method of making modulation transformer calculations has been given in Chapter Five Section 5-8. However, to emphasize the method of making the calculation, an additional example will be given.

Suppose we take the case of a Class C amplifier operating at a plate voltage of 2000 with 225 ma. of plate current. This amplifier would present a load resistance of 2000 divided by 0.225 amperes or 8888 ohms. The plate power input would be 2000 times 0.225 or 450 watts. By reference to Table III in Chapter Five we see that a pair of 811 tubes operating at 1500 plate volts will deliver 225 watts of audio output. The plate-to-plate load resistance for these tubes under the specified operating conditions is 18,000 ohms. Hence our problem is to match the Class C amplifier load resistance of 8888 ohms to the 18,000-ohm load resistance required by the modulator tubes.

A 200-to-300 watt modulation transformer will be required for the job. If the taps on the transformer are given in terms of impedances it will only be necessary to connect the secondary for 8888 ohms (or a value approximately equal to this such as 9000 ohms) and the primary for 18,000 ohms. If it is necessary to determine the proper turns ratio required of the transformer it can be determined in the following manner. The square root of the impedance ratio is equal to the turns ratio, hence:

$$\sqrt{\frac{8888}{18000}} = \sqrt{0.494} = 0.703$$

The transformer must have a turns ratio of approximately 1-to-0.7 step down, total primary to total secondary. The greater number of turns always goes with the higher impedance, and vice versa.

Plate-and-Screen Modulation When *only* the plate of a screen-grid tube is modulated, it is impossible to obtain high-percentage linear modulation under ordinary conditions. The plate current of such a stage is not linear with plate voltage. However, if the screen is modulated simultaneously with the plate, the instantaneous screen voltage drops in proportion to the drop in the plate voltage, and linear modulation can then be obtained. Four satisfactory circuits for accomplishing combined plate and screen modulation are shown in figure 12.

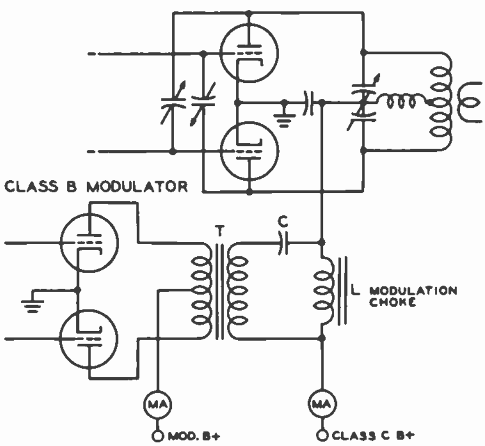


Figure 11.
SHUNT-FED CLASS B PLATE MODULATION.

The arrangement shown above feeds the plate current to the Class C modulated stage through a modulation choke as contrasted to running this current through the secondary of the modulation transformer as shown in figure 10. When an adequate-size choke is used for L and a capacitor of moderate size for C, this arrangement will give improved low-frequency response over the circuit of figure 10. It is for this reason that this circuit is commonly used for broadcast stations. The choke L should have an inductance high enough so that its inductive reactance will be at least equal to the Class C amplifier load impedance at the lowest frequency to be modulated. The capacitor C should have a capacitive reactance much lower than the Class C load impedance at the lowest frequency to be transmitted. The arrangement shown will give improved phase-shift characteristics for clipped speech waves over the simpler system shown in figure 10.

The screen r-f by-pass capacitor, C_2, should not have a greater value than 0.005 μfd., preferably not larger than 0.001 μfd. It should be large enough to bypass effectively all r-f voltage without short-circuiting high-frequency audio voltages. The plate by-pass capacitor can be of any value from 0.002 μfd. to 0.005 μfd. The screen-dropping resistor, R_1, should reduce the applied high voltage to the value specified for operating the particular tube in the circuit. Capacitor C_1, is seldom required, yet some tubes may require this capacitor in order to keep C_2 from attenuating the high audio frequencies. Different values between .002 and .0002 μfd. should be tried for best results.

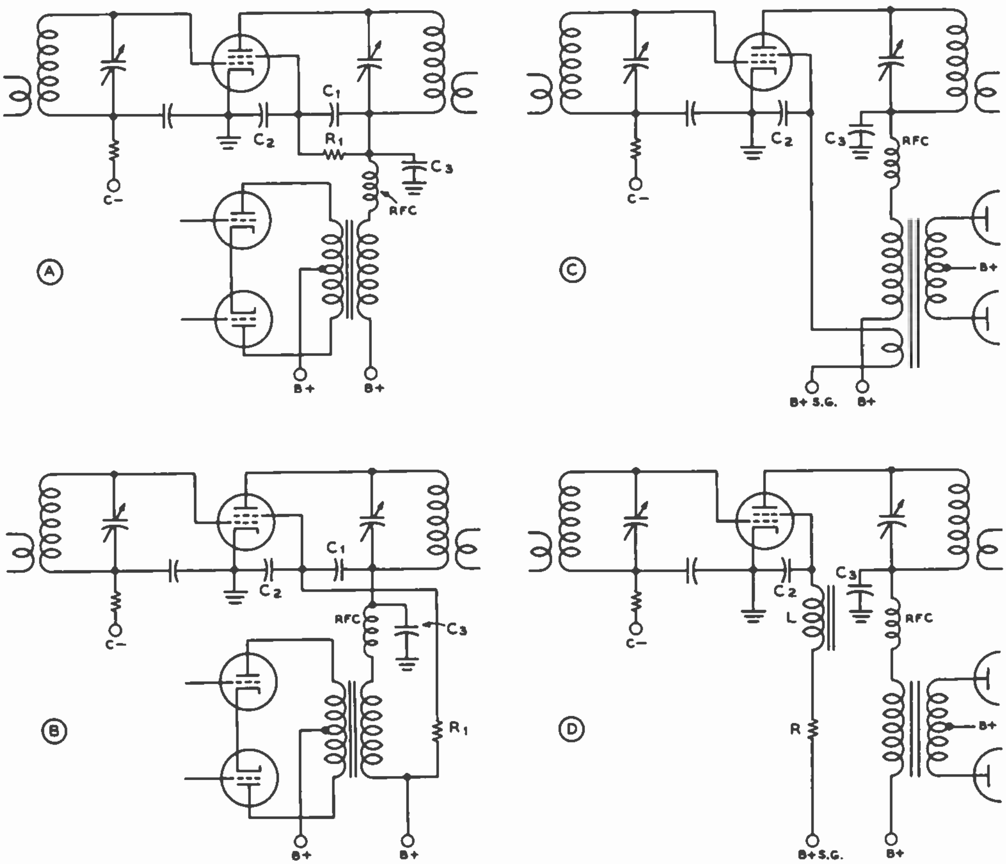


Figure 12.

PLATE MODULATION OF A BEAM TETRODE OR SCREEN-GRID TUBE.

These alternative arrangements for plate modulation of tetrodes or pentodes are discussed in detail in the text. The arrangements shown at (B) or (D) are recommended for most applications.

Figure 12C shows another method which uses a third winding on the modulation transformer, through which the screen-grid is connected to a low-voltage power supply. The ratio of turns between the two output windings depends upon the type of screen-grid tube which is being modulated. Normally it will be such that the screen voltage is being modulated 60 per cent when the plate voltage is receiving 100 per cent modulation.

If the screen voltage is derived from a dropping resistor (*not* a divider) that is by-passed for r.f but not a.f., it is possible to secure quite good modulation by applying modulation only to the plate. Under these conditions, the screen tends to modulate itself, the screen voltage varying over the audio cycle as a result of the screen impedance increasing with plate voltage, and decreasing with a decrease in plate voltage. This circuit arrangement is illustrated in figure 12B.

A similar application of this principle is shown in figure 12D. In this case the screen voltage is fed directly from a low-voltage supply of the proper potential through a choke L. A conventional filter choke having an inductance from 10 to 20 henries will be satisfactory for L.

To afford protection of the tube when plate voltage is not applied but screen voltage is supplied from the exciter power supply, when using the arrangement of figure 12D, a resistor of 3000 to 10,000 ohms can be connected in series with the choke L. In this case the screen supply voltage should be at least 1½ times as much as is required for actual screen voltage, and the value of resistor is chosen such that with normal screen current the drop through the resistor and choke will be such that normal screen voltage will be applied to the tube. When the plate voltage is removed the screen current will increase greatly and the drop through resistor R will increase to such a value that the screen voltage will be lowered to the point where the screen dissipation on the tube will not be exceeded. However, the supply voltage and value of resistor R must be chosen carefully so that the maximum rated screen dissipation cannot be exceeded. The maximum possible screen dissipation using this arrangement is equal to: $W = E^2/4R$ where E is the screen supply voltage and R is the combined resistance of the resistor R in figure 12D and the d-c resistance of the choke L. It is wise, when using this arrangement to check, using the above formula, to see that the value of W obtained is less than the maximum rated screen dissipation of the tube or tubes used in the modulated stage. This same system can of course also be used in figuring the screen supply circuit of a pentode or tetrode amplifier stage where modulation is not to be applied.

The modulation transformer for plate-and-screen-modulation, when utilizing a dropping resistor as shown in figure 12A, is similar to the type of transformer used for any plate-modulated phone. The combined screen and plate current is divided into the plate voltage in order to obtain the Class C amplifier load impedance. The peak audio power required to obtain 100 per cent modulation is equal to the d-c power input to the screen, screen resistor, and plate of the modulated r-f stage.

8-4 Cathode Modulation

Cathode modulation offers a workable compromise between the good plate efficiency but expensive modulator of high-level plate modulation, and the poor plate efficiency but inexpensive modulator of grid modulation. Cathode modulation consists essentially of an admixture of the two.

The efficiency of the average well-designed plate-modulated transmitter is in the vicinity of 75 to 80 per cent, with a compromise perhaps at 77.5 per cent. On the other hand, the efficiency of a good grid-modulated transmitter may run from 28 to maybe 40 per cent, with the average falling at about 34 per cent. Now since cathode modulation consists of simultaneous grid and plate modulation, in phase with each other, we can theoretically obtain any efficiency from about 34 to 77.5 per cent from our cathode-modulated stage, depending upon the relative percentages of grid and plate modulation.

Since the system is a compromise between the two fundamental modulation arrangements, a value of efficiency approximately half way between the two would seem to be the best compromise. Experience has proved this to be the case. A compromise efficiency of about 56.5 per cent, roughly half way between the two limits, has proved to be optimum. Calculation has shown that this value of efficiency can be obtained from a cathode-modulated amplifier when the audio-frequency modulating power is approximately 20 per cent of the d-c input to the cathode-modulated stage.

Cathode-Modulation Operating Curves Figure 13 shows a set of operating curves for cathode-modulated r-f amplifier stages. The chart is a plot of the percentage of plate modulation (m) against plate circuit efficiency, audio power required, plate input wattage in per cent of the plate-modulated Class C rating, and output power in percentage of the Class C phone output rating. These last two curves are not of as great importance in designing new transmitters as are the curves showing the relationship between per cent plate modulation and plate circuit efficiency.

Optimum Operating Conditions As was mentioned before, the optimum operating condition for a normal cathode-modulated amplifier is that at which the audio power output of the cathode modulator is about 20 per cent of the d-c input to the modulated stage. Under these conditions the plate efficiency will be in

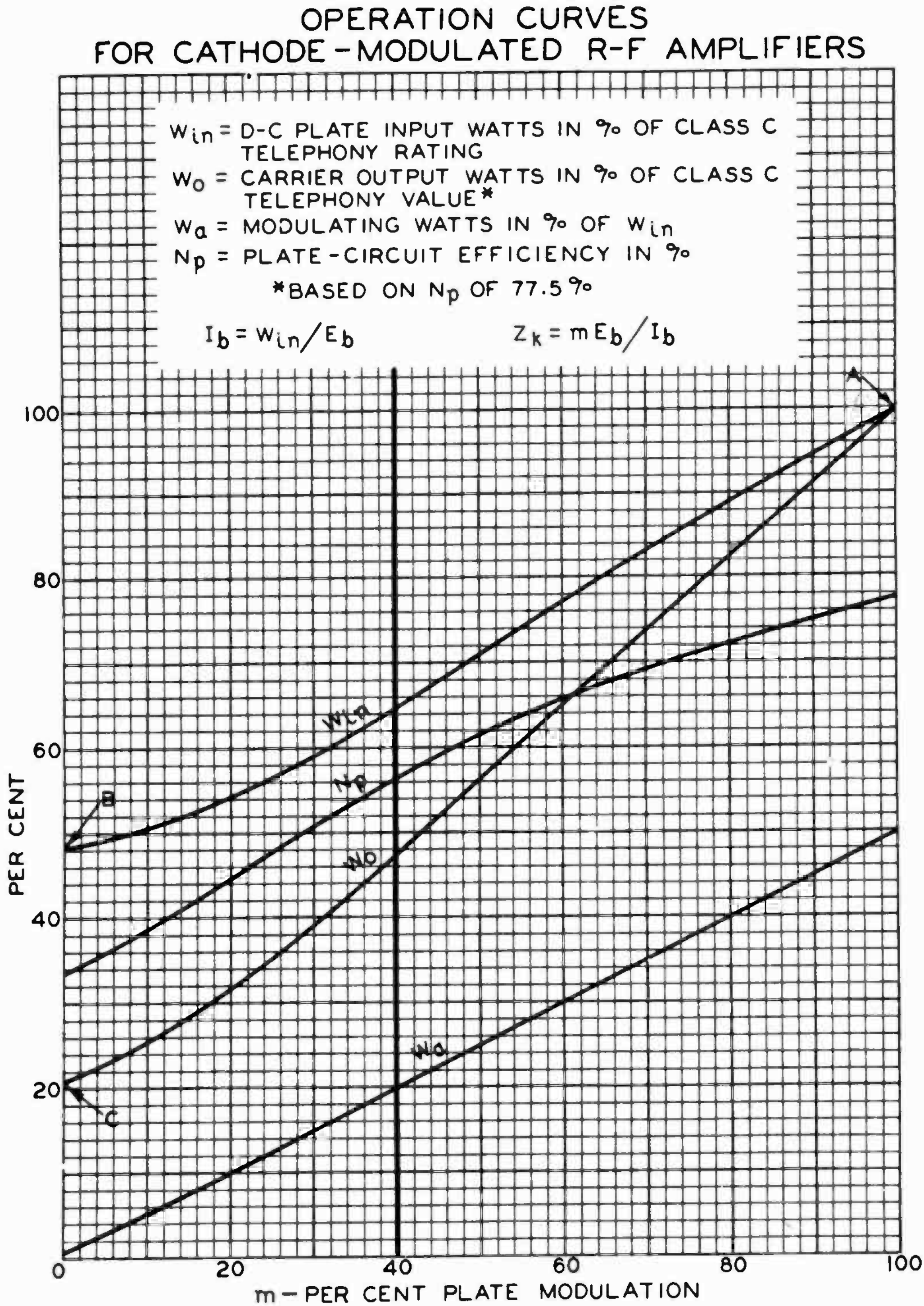

OPERATION CURVES
FOR CATHODE-MODULATED R-F AMPLIFIERS
Win = D-C PLATE INPUT WATTS IN % OF CLASS C TELEPHONY RATING
Wo = CARRIER OUTPUT WATTS IN % OF CLASS C TELEPHONY VALUE*
Wa = MODULATING WATTS IN % OF Win
Np = PLATE-CIRCUIT EFFICIENCY IN %
*BASED ON Np OF 77.5 %
Ib = Win/Eb
Zk = mEb/Ib
Win
Np
Wo
Wa
B
C
PER CENT
100
80
60
40
20
0
20
40
60
80
100
m—PER CENT PLATE MODULATION

FIGURE 13.

the vicinity of 56.5 per cent (between 54 and 58 per cent in a practical transmitter). The limiting factor in an efficiency-modulated amplifier of this type is, to a large extent, plate dissipation. If, under the conditions given above, the plate dissipation of the tube under carrier conditions is less than the rated value, the plate input can be increased until rated plate dissipation is reached. The plate dissipation for any condition of operation can easily be determined by reference to figure 13 and a little calculation. Determine the input, and from the efficiency value given, figure the power output from the stage. Subtract this from the plate input, and the result is the amount that the tube will be required to dissipate.

Cathode Impedance The impedance of the cathode circuit of an amplifier which is being cathode modulated is an important consideration in the selection of the transformer which is to be used to couple the modulator. The cathode impedance of an amplifier is equal to the peak *modulating* voltage divided by the peak a-f component of the plate current of the stage. The peak *modulating* voltage is equal to the plate voltage times *m* (the per cent plate modulation).

$$\text{Hence: } Z_k = m \frac{E_p}{I_p}$$

Or, simply, the cathode impedance is equal to the per cent plate modulation (expressed as a decimal; e.g., 0.4 for 40%) times the plate voltage, divided by the plate current.

Cathode Modulator A typical cathode-modulated r-f amplifier is shown in figure 14. The modulator which is used to feed the audio into the cathode circuit of the modulated stage should preferably have a power output of 20 per cent of the d-c input to the stage, for 40 per cent plate modulation. Although this is the recommended percentage of plate modulation, satisfactory operation may be had with other percentage values than this provided the proper operating values are taken from figure 13. The modulator tubes may be operated Class A, Class AB, or Class B, but it is recommended that some form of degenerative

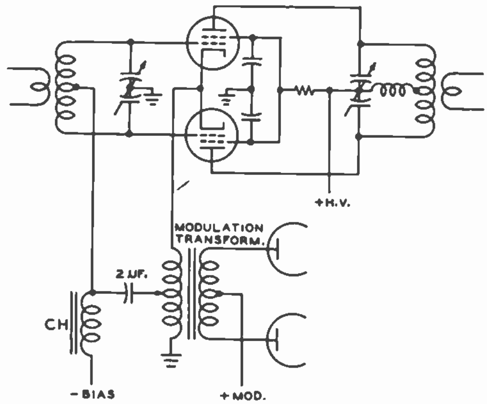


Figure 14.
CATHODE-MODULATION CIRCUIT.
The modulation transformer in series with the cathode return of the modulated stage must match the cathode impedance of this stage. The choke in series with the grid return of the stage should have from 15 to 40 henrys inductance and should be capable of carrying the full grid current of the stage. The grid tap on the modulation transformer is varied, after the stage has been placed into operation, to give the best modulation pattern.

feedback be employed around the modulator tubes when they are to be operated in any manner other than Class A. This is particularly true of beam tetrodes when used as modulators; if some form of feedback is not used around them the harmonic distortion can easily be serious enough to be objectionable, since the cathode modulated stage does not present a strictly linear impedance.

The transformer which couples the modulator to the cathode circuit of the modulated amplifier should match the cathode impedance, as calculated by the formula above, and in addition should have a number of taps so that the proper amount of audio voltage will be impressed upon the grid of the stage. In most cases one of the conventional multi-match output transformers will be satisfactory for the job, the cathode lead and the ground terminal of the stage being connected to the proper taps to give the desired value of impedance. The stage is then coupled to a cathode-ray oscilloscope so that the modulated waveform is shown on the screen. As the stage is being modulated, the grid is tapped varying amounts up and down on the modulation transformer until the best waveform is obtained on the screen of the oscilloscope. The more closely the grid is tapped to the cathode, the less will be the amount of audio voltage upon the grid. On the other hand, if the grid return is grounded, the full cathode swing will be placed upon the grid. It will be found that low-μ tubes will require a larger percentage of the total cathode swing upon them than will tubes with a higher μ factor. Hence, high-μ tubes will be tapped closer to cathode; low-μ tubes will be tapped more closely to ground.

Excitation The r-f driver for a cathode-modulated stage should have about the same power output capabilities as would be required to drive a c-w amplifier to the same input as it is desired to drive the cathode-modulated stage. However, some form of excitation control should be available since the amount of excitation power has a direct bearing on the linearity of a cathode-modulated amplifier stage. If link coupling is used between the driver and the modulated stage, variation in the amount of link coupling will afford ample excitation variation. If much less than 40% plate modulation is employed, the stage begins to resemble a grid-bias modulated stage, and the necessity for good r-f regulation will apply.

Biasing Systems Any of the conventional biasing arrangements which are suitable for use on a Class C amplifier are also suitable for use with a cathode-modulated stage. Battery bias, grid leak bias, and power supply bias all are usable in their conventional fashion; cathode bias may be used if the bias resistor is by-passed with a high capacitance electrolytic capacitor. In any case the bias voltage should be variable or adjustable so that the optimum value for distortionless modulation can be found. If grid-leak or cathode bias is used, the value of the grid leak or cathode resistor should be adjustable. Grid-leak bias is not recommended if the per cent plate modulation is less than 30%, as the stage then is essentially a grid-modulated amplifier, requiring a well-regulated bias source.

Screen-Plate Modulation Cathode modulation has not proved too satisfactory for use with beam tetrode tubes. This is a result of the small excitation and grid swing requirements for such tubes, plus the fact that some means for holding the screen potential at cathode potential as far as audio is concerned is usually necessary. However, an equivalent method, which has been called "screen-plate modulation" may be used for those applications where a compromise between high-level plate modulation and low-level efficiency modulation is desired.

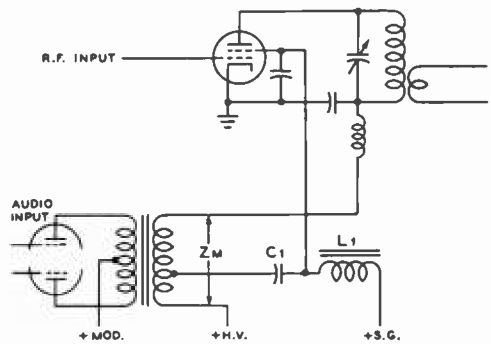


Figure 15.
SCREEN-PLATE MODULATION CIRCUIT.
The operating conditions for this type of modulation are quite similar to those of cathode modulation.

The circuit for screen-plate modulation is illustrated in figure 15. Inspection of the circuit will show that the modulation system is an admixture of screen modulation and plate modulation, the same as cathode modulation is a combination of grid modulation and plate modulation. Hence the curves of figure 13 for cathode modulation apply also to the determination of the operating conditions for screen-plate modulation. Also, the secondary impedance of the modulation transformer, Z_m is determined by using the same formula as is employed for determining the cathode impedance when using cathode modulation: $Z_m = m\, E_p / I_p$.

Example of Screen-Plate Modulation As an example of the application of screen-plate modulation, let us take the case of an 813 stage to be operated at a plate potential of 1500 volts. A Class AB_2 modulator with an 815, with rated output of 54 watts is available. The 54-watt output of the 815 amplifier will modulate an input of 270 watts to the 813, with 40 per cent plate modulation as determined from the chart of figure 13. With 40 per cent plate modulation the plate-

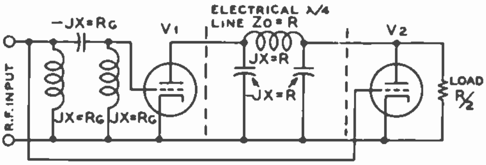


Figure 16.
DIAGRAMMATIC REPRESENTATION OF THE DOHERTY LINEAR.

circuit efficiency of the 813 stage will be about 56 per cent, so that the power output will be about 150 watts, with the remaining 120 watts of input to the stage appearing as plate dissipation on the 813.

An input of 270 watts at 1500 volts will be obtained with a plate current of 180 ma. to the 813. Under these conditions the secondary impedance of the modulation transformer should be: Z_m = 0.4 1500/0.180, or 3330 ohms. Thus the modulator should be adjusted so that it will deliver 54 watts into a 3330-ohm load. The d-c screen voltage on the 813 should be adjusted to a value about 2/3 the rated value for operation of the tube as a Class C amplifier, and the tap on the modulation transformer should be adjusted until full amplitude linear modulation of the carrier is obtained, as viewed on an oscilloscope. Some adjustment of the d-c screen voltage, the a-c screen audio signal, and the plate loading of the amplifier should be made until an optimum adjustment of operating conditions is obtained. Grid circuit operating conditions for the stage will be approximately the same as for Class C plate modulation, and are relatively non-critical for satisfactory operation of the screen-plate modulated stage.

The Doherty and the Termon-Woodyard Modulated Amplifiers These two amplifiers will be described together since they operate upon very similar principles. Figure 16 shows a greatly simplified schematic diagram of the operation of both types. Both systems operate by virtue of a carrier tube (V_1 in both figures 16 and 17) which supplies the unmodulated carrier, and whose output is reduced to supply negative peaks, and a peak tube (V_2) whose function is to supply approximately half the positive peak of the modulation cycle and whose additional function is to lower the load impedance on the carrier tube so that it will be able to supply the other half of the positive peak of the modulation cycle.

The peak tube is enabled to increase the output of the carrier tube by virtue of an impedance inverting line between the plate circuits of the two tubes. This line is designed to have a characteristic impedance of one-half the value of load into which the carrier tube operates under the carrier conditions. Then a load of one-half the characteristic impedance of the quarter-wave line is coupled into the output. By experience with quarter-wave lines in antenna-matching circuits we know that such a line will vary the impedance at one end of the line in such a manner that the geometric mean between the two terminal impedances will be equal to the characteristic impedance of the line. Thus, if we have a value of load of *one-half* the characteristic impedance of the line at one end, the other end of the line will present a value of *twice* the characteristic impedance of the line to the carrier tube V_1.

This is the situation that exists under the carrier conditions when the peak tube merely floats across the load end of the line and contributes no power. Then as a positive peak of modulation comes along, the peak tube starts to contribute power to the load until at the peak of the modulation cycle it is contributing enough power so that the impedance at the load end of the line is equal to R, instead of the R/2 that is presented under the carrier conditions. This is true because at a positive modulation peak (since it is delivering full power) the peak tube subtracts a negative resistance of R/2 from the load end of the line.

Now, since under the peak condition of modulation the load end of the line is terminated in R ohms instead of R/2, the impedance at the *carrier-tube* will be *reduced* from 2R ohms to R ohms. This again is due to the impedance inverting action of the line. Since the load resistance on the carrier tube has been reduced to half the carrier value, its output at the peak of the modulation cycle will be doubled. Thus we have the necessary condition for a 100 per cent positive modulation peak; the amplifier will deliver four times as much power as it does under the carrier conditions.

On negative modulation peaks the peak tube does not contribute; the output of the carrier tube is reduced until on a 100 per cent negative peak its output is zero.

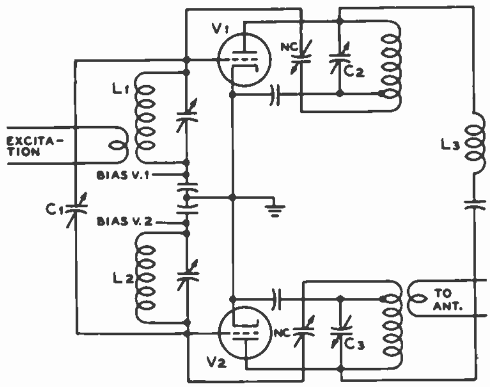


Figure 17.
SIMPLIFIED SCHEMATIC OF A "HIGH EFFICIENCY" AMPLIFIER.
The basic system, comprising a "carrier" tube and a "peak" tube interconnected by lumped-constant quarter-wave lines, is the same for either grid-bias modulation or for use as a linear amplifier of a modulated wave.

The Electrical Quarter-Wave Line While an electrical quarter-wave line (consisting of a pi network with the inductance and capacitance lags having a reactance equal to the characteristic impedance of the line) does have the desired impedance-inverting effect, it also has the undesirable effect of introducing a 90° phase shift across such a line. If the shunt elements are capacitances, the phase shift across the line leads by 90°; if they are inductances, the phase shift lags by 90°. Since there is an undesirable phase shift of 90° between the plate circuits of the carrier and peak tubes, an equal and opposite phase shift must be introduced in the exciting voltage to the grid circuits of the two tubes so that the resultant output in the plate circuit will be in phase. This additional phase shift has been indicated in figure 16 and a method of obtaining it has been shown in figure 17.

Comparison Between Linear and Grid Modulator The difference between the Doherty linear amplifier and the Terman-Woodyard grid-modulated amplifier is the same as the difference between any linear and grid-modulated stages. Modulated r.f. is applied to the grid circuit of the Doherty linear amplifier with the carrier tube biased to cutoff and the peak tube biased to the point where it draws substantially zero plate current at the carrier condition.

In the Terman-Woodyard grid-modulated amplifier the carrier tube runs Class C with comparatively high bias and high plate efficiency, while the peak tube again is biased so that it draws almost no plate current. Unmodulated r.f. is applied to the grid circuits of the two tubes and the modulating voltage is inserted in series with the fixed bias voltages. From one-half to two-thirds as much *audio* voltage is required at the grid of the peak tube as is required at the grid of the carrier tube.

Operating Efficiencies The resting carrier efficiency of the grid-modulated amplifier may run as high as is obtainable in any Class C stage, 80 per cent or better. The resting carrier efficiency of the linear will be about as good as is obtainable in any Class B amplifier, 60 to 70 per cent. The overall efficiency of the bias-modulated amplifier at 100 per cent modulation will run about 75 per cent; of the linear, about 60 per cent.

In figure 17 the plate tank circuits are detuned enough to give an effect equivalent to the shunt elements of the quarter-wave "line" of figure 16. At resonance, the coils L_1 and L_2 in the grid circuits of the two tubes have each an inductive reactance equal to the capacitive reactance of the capacitor C_1. Thus we have the effect of a pi network consisting of shunt inductances and series capacitance. In the plate circuit we want a phase shift of the same magnitude but in the opposite direction; so our series element is the inductance L_3 whose reactance is equal to the characteristic impedance desired of the network. Then the plate tank capacitors of the two tubes C_2 and C_3 are increased an amount past resonance, so that they have a capacitive reactance equal to the inductive reactance of the coil L_3. It is quite important that there be no coupling between the inductors.

Although both these types of amplifiers are highly efficient and require no high-level audio equipment, they are difficult to adjust — particularly so on the higher frequencies — and it would be an extremely difficult problem to design a multiband transmitter employing the circuit. However, the grid-bias modulation system has advantages for the high-power transmitter which will be operated on a single frequency band.

Other High-Efficiency Modulation Systems Many other high-efficiency modulation systems have been described since about 1936. The majority of these, however, have received little application either by commercial interests or by amateurs. In most cases the circuits are difficult to adjust, or they have other undesirable features which make their use impracticable alongside the more conventional modulation systems. Nearly all these circuits have been published in the *I.R.E. Proceedings* and the interested reader can refer to them in back copies of that journal.

8-5 Speech Clipping

Speech waveforms are characterized by frequently recurring high-intensity peaks of very short duration. These peaks will cause overmodulation if the "average" level of modulation on loud syllables exceeds approximately 30 per cent. Careful checking into the nature of speech sounds has revealed that these high-intensity peaks are due primarily to the vowel sounds. Further research has revealed that the vowel sounds add little to intelligibility, the major contribution to intelligibility coming from the consonant sounds such as *v, b, k, s, t,* and *l.* Measurements have shown that the power contained in these consonant sounds may be down 30 db or more from the energy in the vowel sounds in the same speech passage. Obviously, then, if we can increase the relative energy content of the consonant sounds with respect to the vowel sounds it will be possible to understand a signal modulated with such a waveform in the presence of a much higher level of background noise and interference. Experiment has shown that it is possible to accomplish this desirable result simply by cutting off or *clipping* the high-intensity peaks and thus building up in a relative manner the effective level of the weaker sounds.

Such clipping theoretically can be accomplished simply by increasing the gain of the speech amplifier until the average level of modulation on loud syllables approaches 90 per cent. This is equivalent to increasing the speech power of the consonant sounds by

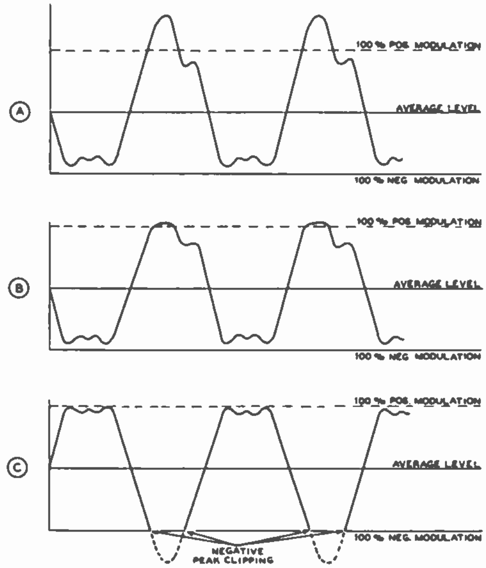


Figure 18.

SPEECH-WAVEFORM AMPLITUDE MODULATION.

Showing the effect of using the proper polarity of a speech wave for modulating a transmitter. (A) shows the effect of proper speech polarity on a transmitter having an upward modulation capability of greater than 100 per cent. (B) shows the effect of using proper speech polarity on a transmitter having an upward modulation capability of only 100 per cent. Both these conditions will give a clean signal without objectionable splatter. (C) shows the effect of the use of improper speech polarity. This condition will cause serious splatter due to negative-peak clipping in the modulated-amplifier stage.

about 10 times or, conversely, we can say that 10 db of clipping has been applied to the voice wave. However, the clipping when accomplished in this manner *will produce higher order sidebands known as "splatter,"* and the transmitted signal would occupy a relatively tremendous slice of spectrum. So another method of accomplishing the desirable effects of clipping must be employed.

A considerable reduction in the amount of splatter caused by a moderate increase in the gain of the speech amplifier can be obtained by poling the signal from the speech amplifier to the transmitter such that the high-intensity peaks occur on *upward* or positive modulation. Overloading on positive modulation peaks produces far less splatter than the negative-peak clipping which occurs with overloading on

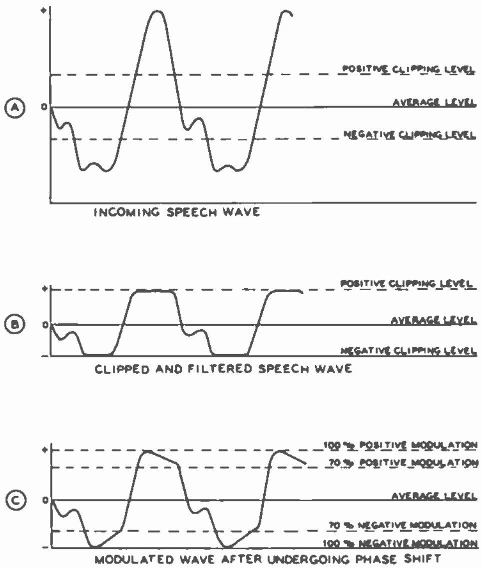


Figure 19.

ACTION OF A CLIPPER-FILTER ON A SPEECH WAVE.

The drawing (A) shows the incoming speech wave before it reaches the clipper stage. (B) shows the output of the clipper-filter, illustrating the manner in which the peaks are clipped and then the sharp edges of the clipped wave removed by the filter. (C) shows the effect of phase shift in the stages following the clipper-filter. (C) also shows the manner in which the transmitter may be adjusted for 100 per cent modulation on the "canted" peaks of the wave, the sloping top of the wave reaching about 70 per cent modulation.

the negative peaks of modulation. This aspect of the problem has been discussed in more detail in the section on *Speech Waveform Dissymmetry* earlier in this chapter. The effect of feeding the proper speech polarity from the speech amplifier to the modulator is shown in figure 18.

A much more desirable and effective method of obtaining speech clipping is actually to employ a clipper circuit in the earlier stages of the speech amplifier, and then to filter out the objectionable distortion components by means of a sharp low-pass filter having a cut-off frequency of approximately 3000 cycles. Tests on *clipper-filter* speech systems have shown that 6 db of clipping on voice is just noticeable, 12 db of clipping is quite acceptable, and values of clipping from 20 to 25 db are tolerable under such conditions that a high degree of clipping is necessary to get through heavy QRM or QRN. A signal with 12 db of clipping doesn't sound quite "natural" but it is not unpleasant to listen to and is much more readable than an unclipped signal in the presence of strong interference.

The use of a clipper-filter in the speech amplifier, to be completely effective, requires that phase shift between the clipper-filter stage and the final modulated amplifier be kept to a minimum. However, if there is phase shift after the clipper-filter the system does not completely break down. The presence of phase shift merely requires that the audio gain following the clipper-filter be reduced to the point where the "cant" applied to the clipped speech waves still cannot cause overmodulation. This effect is illustrated in figure 19.

The "cant" appearing on the tops of the square waves leaving the clipper-filter centers about the clipping level. Hence, as the frequency being passed through the system is lowered, the amount by which the peak of the "canted" wave exceeds the clipping level is increased.

In a normal transmitter having a moderate amount of phase shift the cant applied to the tops of the waves will cause overmodulation on frequencies below those for which the gain following the clipper-filter has been adjusted unless remedial steps have been taken. The following steps are advised:

(1) Introduce bass *suppression* into the speech amplifier *ahead* of the clipper-filter.

(2) *Improve* the low-frequency response characteristic insofar as it is possible in the stages *following* the clipper-filter. Feeding the plate current to the final amplifier through a choke rather than through the secondary of the modulation transformer will help materially.

Even with the normal amount of improvement which can be attained through the steps mentioned above there will still be an amount of wave cant which must be compensated in some manner. This compensation can be done in either of two ways. The first and simplest way is as follows:

(1) Adjust the speech gain *ahead* of the clipper-filter until with normal talking into the microphone the distortion being introduced by the clipper-filter circuit is

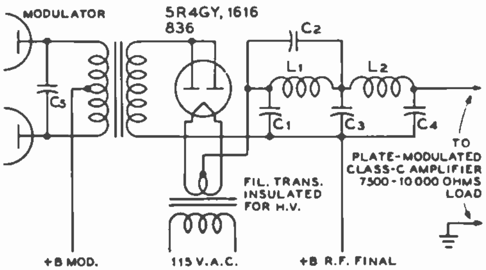


Figure 20.
HIGH-LEVEL SPLATTER SUPPRESSOR.
This circuit is effective in reducing splatter caused by negative-peak clipping in the modulated amplifier stage. The use of a two-section filter as shown is recommended, although either a single m-derived or a constant-k section may be used for greater economy. Suitable chokes, along with recommended capacitor values, are available from several manufacturers.

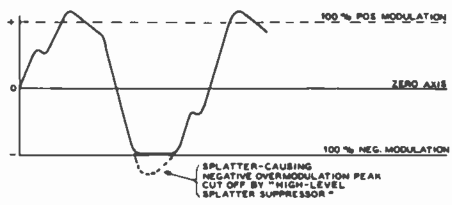


Figure 21.
ACTION OF HIGH-LEVEL SPLATTER SUPPRESSOR.
A high-level splatter suppressor may be used in a transmitter without a clipper-filter to reduce negative-peak clipping, or such a unit may be used following a clipper-filter to allow a higher average modulation level by eliminating the negative-peak clipping which the wave-cant caused by phase shift might produce.

quite apparent but not objectionable. This amount of distortion will be apparent to the normal listener when 10 to 15 db of clipping is taking place.

(2) Tune a selective communications receiver about 15 kc. to one side or the other of the frequency being transmitted. Use a short antenna or no antenna at all on the receiver so that the transmitter is not blocking the receiver.

(3) Again with normal talking into the microphone adjust the gain *following* the clipper-filter to the point where sideband splatter is being heard, and then slightly back off the gain after the clipper-filter until the splatter disappears.

If the phase shift in the transmitter or modulator is not excessive the adjustment procedure given above will allow a clean signal to be radiated regardless of any reasonable voice level being fed into the microphone.

If a cathode-ray oscilloscope is available the modulated envelope of the transmitter should be checked with 30 to 70 cycle saw-tooth waves on the horizontal axis. If the upper half of the envelope appears in general the same as the drawing of figure 19C, all is well and phase-shift is not excessive. However, if much more slope appears on the tops of the waves than is illustrated in this figure, it will be well to apply the second step in compensation in order to insure that sideband splatter cannot take place and to afford a still higher average percentage of modulation. This second step consists of the addition of a high-level "splatter suppressor" such as is illustrated in figure 20.

The use of a high-level splatter suppressor after a clipper-filter system will afford the result shown in figure 21 since such a device will not permit the negative-peak clipping which the wave cant caused by audio-system phase shift can produce. The high-level splatter suppressor operates by virtue of the fact that it will not permit the plate voltage on the modulated amplifier to go completely to zero regardless of the incoming signal amplitude. Hence negative-peak clipping with its attendant splatter cannot take place. Such a device can, of course, also be used in a transmitter which does not incorporate a clipper-filter system. However, the full increase in average modulation level without serious distortion, afforded by the clipper-filter system, will not be obtained.

A word of caution should be noted at this time in the case of tetrode final modulated amplifier stages which afford screen voltage modulation by virtue of a tap or a separate winding on the modulation transformer such as is shown in figure 12C of this chapter. If such a system of modulation is in use, the high-level splatter suppressor shown in figure 20 will not operate satisfactorily since negative-peak clipping in the stage can take place when the *screen* voltage goes too low. There are several remedies which can be employed:

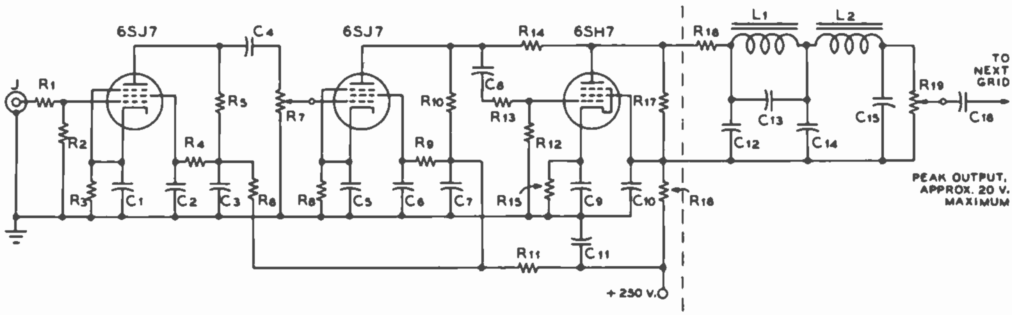


Figure 22.

CLIPPER-FILTER CIRCUIT USING AMPLIFIER-CLIPPER STAGE.

C_1—25-μfd. 25-volt elect.
C_2—0.5-μfd. 400-volt paper
C_3—1.0-μfd. 400-volt paper
C_4—0.003-μfd. mica
C_5—25-μfd. 25-volt elect.
C_6—0.5-μfd. 400-volt paper
C_7—8-μfd. 450-volt elect.
C_8—0.003-μfd. mica
C_9—25-μfd. 25-volt elect.
C_{10}—8-μfd. 450-volt elect.
C_{11}—0.003-μfd. mica by-pass
C_{12}—200-μμfd. mica
C_{13}—175-μμfd. mica
C_{14}—500-μμfd. mica
C_{15}—330-μμfd. mica
C_{16}—0.1-μfd. 400 volt paper
R_1—47K ½ watt
R_2—1 meg. ½ watt
R_3—1800-ohms ½ watt
R_4—2.2 meg. ½ watt
R_5—470K ½ watt
R_6—47K 1 watt
R_7—1-megohm potentiometer
R_8—1000 ohms ½ watt
R_9—1 meg. ½ watt
R_{10}—220 K ½ watt
R_{11}—22K 2 watt
R_{12}, R_{13}, R_{14}—1 meg. ½ watt
R_{15}—470 ohms 1 watt
R_{16}—22K 2 watts
R_{17}—15K 2 watts
R_{18}—100K ½ watt
R_{19}—100,000-ohm pot.
L_1, L_2—Stancor C-1080 chokes
J—Microphone jack

(1) Introduce an additional high-level splatter suppressor of the type shown in figure 20 in the screen feed circuit of the tube—on the modulated-voltage side of the screen winding of the modulation transformer.

(2) Use a different screen-voltage modulation circuit. The circuits shown in figures 12A, 12B, and 12D will not give this difficulty.

(3) Do *not* use a "high-level splatter suppressor" but reduce the gain following the clipper-filter system until a pattern such as shown in figure 19C is obtained, checking for splatter at the same time on a communications receiver by the method given in three steps in a preceding paragraph.

Clipper Circuits There are three satisfactory methods whereby clipping may be obtained in the low-level stages of the speech amplifier. These methods involve the use of a series-clipper diode system, a shunt-clipper diode, or an amplifier-clipper. In order for a clipper system to introduce the least amount of distortion into the wave being passed, it should be quite linear up to the point where clipping takes place. The amplifier-clipper system with degenerative feedback from the plate of the clipper back to the preceding stage has proved to be the most linear and distortion-free of the various methods used. Next in desirability from the standpoint of effectiveness and simplicity is the shunt-clipper diode system. The series-clipper system is the most complicated and least stable of the three systems but has the advantage that sharp clipping is obtained.

Figure 22 shows a front end for a speech amplifier utilizing an amplifier-clipper and figure 23 shows a speech amplifier front end using a shunt-clipper arrangement. In both cases a filter system has been shown following the clipper stage. Recommended component values have also been given in both circuits.

Filter Circuits for Clippers Recommended filters have been shown in the circuits of both figure 22 and figure 23.

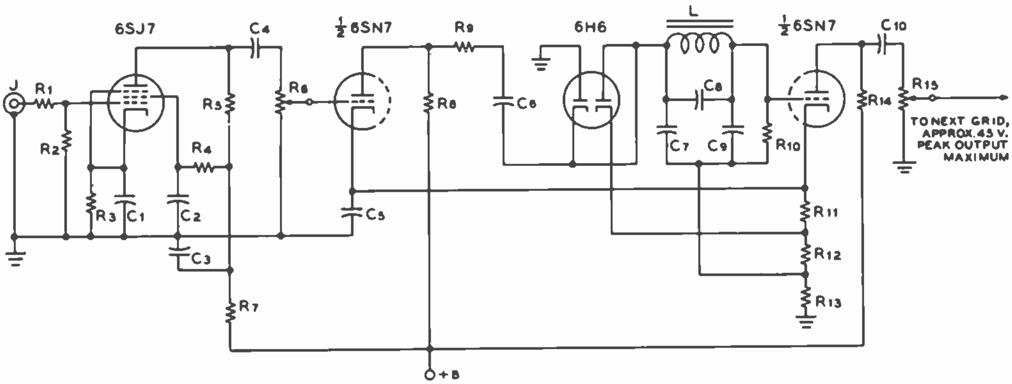


Figure 23.

CLIPPER-FILTER CIRCUIT USING DOUBLE DIODE.

C_1—25-μfd. 25-volt elect.
C_2—0.5-μfd. 400-volt paper
C_3—8-μfd. 450-volt elect.
C_4—0.003-μfd. mica
C_5—25-μfd. 25-volt elect.
C_6—0.01-μfd. 400-volt paper
C_7—200-μμfd. mica
C_8—175-μμfd. mica
C_9—200-μμfd. mica
C_{10}—0.1-μfd. 400-volt paper
R_1—47K ½ watt
R_2—1 meg. ½ watt
R_3—1800 ohms ½ watt
R_4—2.2 meg. ½ watt
R_5—470K ½ watt
R_6—500,000-ohm pot.
R_7—47K ½ watt
R_8—100K 1 watt
R_9—100K ½ watt
R_{10}—100K ½ watt
R_{11}—330 ohms ½ watt
R_{12}—620 ohms ½ watt
R_{13}—620 ohms ½ watt
R_{14}—47K 1 watt
R_{15}—500,000-ohm pot.
L—Stancor C-1080 choke (approx. 4 hy. at no d.c.)
J—Microphone jack

A two-section filter has been used with the circuit of figure 22 and a single-section filter is shown with the circuit of figure 23. The filters for the two circuits may be interchanged since both are designed for a characteristic impedance of 100,000 ohms and a cutoff frequency of 3000 cycles. Should it be desired to employ a filter different from the ones diagrammed, a low-pass filter of any desired characteristic may be designed with the aid of the filter design chart in Chapter Three.

Inspection of the characteristic filter curves shown in figure 20 in Chapter Three will show the attenuation/frequency characteristic of the *m*-derived (m equal 0.6) and constant-*k* types of filters. The *m*-derived filter gives a much more rapid attenuation up to a certain frequency past the cutoff point than does the constant-*k* type of filter. However, after this point of maximum attenuation has been passed in the case of the *m*-derived filter the attenuation begins to decrease. But in the case of the constant-*k* type of filter the attenuation decreases indefinitely. Therefore a combination of an *m*-derived filter section followed by a constant-*k* section will give a more rapid overall rate of attenuation than two sections of either type of filter. This is the type of filter that has been employed in the circuit of figure 22.

High-Level Filters Even though we may have cut off all frequencies above 3000 or 3500 cycles through the use of a filter system such as is shown in the circuit of figure 22, higher frequencies may again be introduced into the modulated wave by distortion in stages following the speech amplifier. Harmonics of the incoming audio frequencies may be generated in the driver stage for the modulator; they may be generated in the plate circuit of the modulator; or they may be generated by non-linearity in the modulated amplifier itself.

Regardless of the point in the system following the speech amplifier where the high audio frequencies may be generated, these frequencies can still cause a broad signal to be transmitted even though all frequencies above 3000 or 3500 cycles have been cut off in the speech amplifier. The effects of distortion in the audio system *following* the speech ampli-

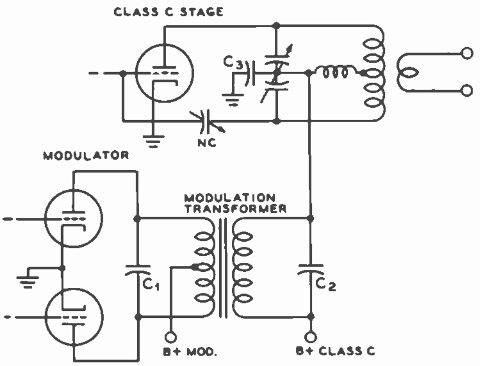


Figure 24.

"BUILDING-OUT" THE MODULATION TRANSFORMER.

This expedient utilizes the leakage reactance of the modulation transformer in conjunction with the capacitors shown to make up a single-section low-pass filter. In order to determine exact values for C_1 and C_2 plus C_3, it is necessary to use a measurement setup such as is shown in figure 25. However, experiment has shown in the case of a number of commercially available modulation transformers that a value for C_1 of 0.002-μfd. and C_2 plus C_3 of 0.004 μfd. will give satisfactory results.

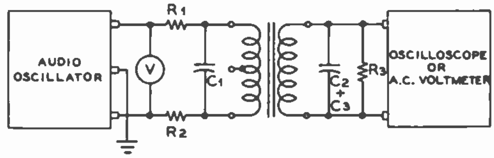


Figure 25.

TEST SETUP FOR BUILDING-OUT MODULATION TRANSFORMER.

Through the use of a test setup such as is shown and the method described in the text it is possible to determine the correct values for a specified filter characteristic in the built-out modulation transformer.

fier can be eliminated quite effectively through the use of a *post-modulator* filter. Such a filter must be used between the modulator plate circuit and the r-f amplifier which is being modulated.

This filter may take three general forms in a normal case of a Class C amplifier plate modulated by a Class B modulator. The best method is to use a "high-level splatter suppressor" of the type shown in figure 20 in which a filter network follows the rectifier tube. The next best arrangement is to use a high-level filter of the type shown without the negative-peak rectifier tube. All the constants for a Class C amplifier load of 7500 to 10,000 ohms can be the same as for the filter shown in figure 20. The third method, which will give excellent results in some cases and poor results in others, dependent upon the characteristics of the modulation transformer, is to "build out" the modulation transformer into a filter section. This is accomplished as shown in figure 24 by placing mica capacitors of the correct value across the primary and secondary of the modulation transformer. The proper values for the capacitors C_1 and C_2 must, in the ideal case, be determined by trial and error. Experiment with a number of modulators has shown, however, that if a 0.002 μfd. capacitor is used for C_1, and if the sum of C_2 and C_3 is made 0.004 μfd. (0.002 μfd. for C_2 and 0.002 for C_3) the ideal condition of cutoff above 3000 cycles will be approached in most cases with the "multiple-match" type of modulation transformer.

If it is desired to determine the optimum values of the capacitors across the transformer this can be determined in several ways, all of which require the use of a calibrated audio oscillator. One way is diagrammed in figure 25. The series resistors R_1 and R_2 should each be equal to ½ the value of the recommended plate-to-plate load resistance for the Class B modulator tubes. Resistor R_3 should be equal to the value of load resistance which the Class C modulated stage will present to the modulator. The meter V can be any type of a-c voltmeter. The indicating instrument on the secondary of the transformer can be either a cathode-ray oscilloscope or a high-impedance a-c voltmeter of the vacuum-tube or rectifier type.

With a set-up as shown in figure 25 a plot of output voltage against frequency is made, at all times keeping the voltage across V constant, using various values of capacitance for C_1 and C_2 plus C_3. When the proper values of capacitance have been determined which give substantially constant output up to about 3000 or 3500 cycles and decreasing output at all frequencies above, high-voltage mica capacitors can be substituted if receiving types were used in the tests and the transformer connected to the modulator and Class C amplifier.

With the transformer reconnected in the transmitter a check of the modulated-wave output of the transmitter should be made using an audio oscillator as signal generator and an oscilloscope coupled to the transmitter output. With an input signal amplitude fed to the speech amplifier of such amplitude that limiting does not take place, a substantially clean sine wave should be obtained on the carrier of the transmitter at all input frequencies up to the cutoff frequency of the filter system in the speech amplifier and of the filter which includes the modulation transformer. Above these cutoff frequencies very little modulation of the carrier wave should be obtained. To obtain a check on the effectiveness of the "built out" modulation transformer, the capacitors across the primary and secondary should be removed for the test. In most cases a marked deterioration in the waveform output of the modulator will be noticed with frequencies in the voice range from 500 to 1500 cycles being fed into the speech amplifier.

A filter system similar to that shown in figure 20 may be used between the modulator and the modulated circuit in a grid-modulated or screen-modulated transmitter. Lower-voltage capacitors and low-current chokes may of course be employed.

Modulated Amplifier Distortion The systems described in the preceding paragraphs will have no effect in reducing a broad signal caused by non-linearity in the modulated amplifier. Even though the modulating waveform impressed upon the modulated stage may be distortion free, if the modulated amplifier is non-linear distortion will be generated in the amplifier. The only way in which this type of distortion may be corrected is by making the modulated amplifier more linear. Degenerative feedback which includes the modulated amplifier in the loop will help in this regard.

Plenty of grid excitation and high grid bias will go a long way toward making a plate-modulated Class C amplifier linear, although such operating conditions will make more difficult the problem of TVI reduction. If this still does not give adequate linearity, the preceding buffer stage may be modulated 50 per cent or so at the same time and in the same phase as the final amplifier. The use of a grid leak to obtain the majority of the bias for a Class C stage will improve its linearity.

The linearity of a grid-bias modulated r-f amplifier can be improved, after proper adjustments of excitation, grid bias, and antenna coupling have been made by modulating the stage which excites the grid-modulated amplifier. The preceding driver stage may be grid-bias modulated or it may be plate modulated. Modulation of the driver stage should be in the same phase as that of the final modulated amplifier.

FM and Single-Sideband Transmission

Exciter systems for FM and single-sideband transmission are basically similar in that modification of the signal in accordance with the intelligence to be transmitted is normally accomplished at a relatively low level. Then the intelligence-bearing signal is amplified to the desired power level for ultimate transmission. True, amplifiers for the two types of signals are basically different; linear amplifiers of the Class A or Class B type being used for ssb signals, while Class C or non-linear Class B amplifiers may be used for FM amplification. But the principle of low-level generation and subsequent amplification is standard for both types of transmission.

Frequency Modulation

The use of frequency modulation and the allied system of phase modulation has become of increasing importance in recent years. For amateur communication frequency and phase modulation offer important advantages in the reduction of broadcast and TV interference and in the elimination of the costly high-level modulation equipment most commonly employed with amplitude modulation. For broadcast work FM offers an improvement in signal-to-noise ratio for the high field intensities available in the local-coverage area of FM and TV broadcast stations.

In this chapter various points of difference between FM and amplitude modulation transmission and reception will be discussed and the advantages of FM for certain types of communication pointed out. Since the distinguishing features of the two types of transmission lie entirely in the modulating circuits at the transmitter and in the detector and limiter circuits in the receiver, these parts of the communication system will receive the major portion of attention.

Modulation As described in Chapter 8, *modulation* is the process of altering a radio wave in accordance with the intelligence to be transmitted. The nature of the intelligence is of little importance as far as the process of modulation is concerned; it is the *method* by which this intelligence is made to give a distinguishing characteristic to the radio wave which will enable the receiver to convert it back into intelligence that determines the type of modulation being used.

Figure 1 is a drawing of an r-f carrier amplitude modulated by a sine-wave audio voltage. After modulation the resultant modulated r-f wave is seen still to vary about the zero axis at a constant rate, but the strength of the individual r-f cycles is proportional to the amplitude of the modulation voltage.

In figure 2, the carrier of figure 1 is shown frequency modulated by the same modulating voltage. Here it may be seen that modulation voltage of one polarity causes the carrier frequency to decrease, as shown by the fact that

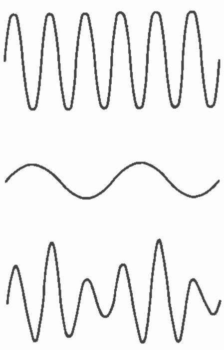


FIGURE 1

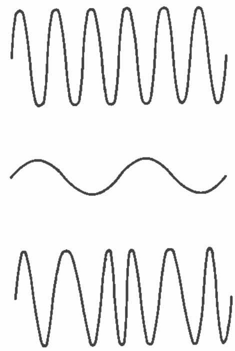


FIGURE 2

AM AND FM WAVES.

Figure 1 shows a sketch of the scope pattern of an amplitude modulated wave at the bottom. The center sketch shows the modulating wave and the upper sketch shows the carrier wave.

Figure 2 shows at the bottom a sketch of a frequency modulated wave. In this case the center sketch also shows the modulating wave and the upper sketch shows the carrier wave. Note that the carrier wave and the modulating wave are the same in either case, but that the waveform of the modulated wave is quite different in the two cases.

the individual r-f cycles of the carrier are spaced farther apart. A modulating voltage of the opposite polarity causes the frequency to increase, and this is shown by the r-f cycles being squeezed together to allow more of them to be completed in a given time interval.

Figures 1 and 2 reveal two very important characteristics about amplitude- and frequency-modulated waves. First, it is seen that while the amplitude (power) of the signal is varied in AM transmission, no such variation takes place in FM. In many cases this advantage of FM is probably of equal or greater importance than the widely publicized noise reduction capabilities of the system. When 100 per cent amplitude modulation is obtained, the average power output of the transmitter must be increased by 50 per cent. This additional output must be supplied either by the modulator itself, in the high-level system, or by operating one or more of the transmitter stages at such a low output level that they are capable of producing the additional output without distortion, in the low-level system. On the other hand, a frequency-modulated transmitter requires an insignificant amount of power from the modulator and needs no provision for in-

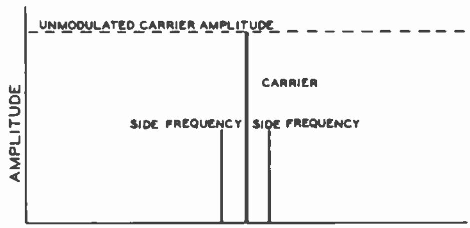


Figure 3.

AM SIDE FREQUENCIES.

For each AM modulating frequency, a pair of side frequencies is produced. The side frequencies are spaced away from the carrier by an amount equal to the modulation frequency, and their amplitude is directly proportional to the amplitude of the modulation. The amplitude of the carrier does not change under modulation.

creased power output on modulation peaks. All of the stages between the oscillator and the antenna may be operated as high-efficiency Class B or Class C amplifiers or frequency multipliers.

Carrier-Wave Distortion The second characteristic of FM and AM waves revealed by figures 1 and 2 is that both types of modulation result in distortion of the r-f carrier. That is, after modulation, the r-f cycles are no longer sine waves, as they would be if no frequencies other than the fundamental carrier frequency were present. It may be shown in the amplitude modulation case illustrated, that there are only two additional frequencies present, and these are the familiar "side frequencies," one located on each side of the carrier, and each spaced from the carrier by a frequency interval equal to the modulation frequency. In regard to frequency and amplitude, the situation is as shown in figure 3. The strength of the carrier itself does not vary during modulation, but the strength of the side frequencies depends upon the percentage of modulation. At 100 per cent modulation the power in the side frequencies is equal to half that of the carrier.

Under frequency modulation, the carrier wave again becomes distorted, as shown in figure 2. But, in this case, many more than two additional frequencies are formed. The first two of these frequencies are spaced from the carrier by the modulation frequency, and the additional side frequencies are located out

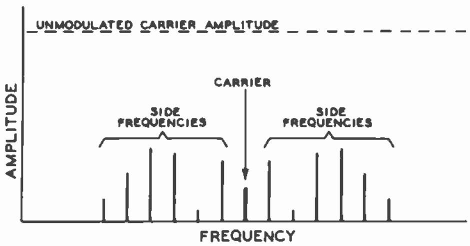


Figure 4.
FM SIDE FREQUENCIES.

With FM each modulation frequency component causes a large number of side frequencies to be produced. The side frequencies are separated from each other and the carrier by an amount equal to the modulation frequency, but their amplitude varies greatly as the amount of modulation is changed. The carrier strength also varies greatly with frequency modulation. The side frequencies shown represent a case where the deviation each side of the "carrier" frequency is equal to five times the modulating frequency. Other amounts of deviation with the same modulation frequency would cause the relative strengths of the various sidebands to change widely.

on each side of the carrier and are also spaced from each other by an amount equal to the modulation frequency. Theoretically, there are an infinite number of side frequencies formed, but, fortunately, the strength of those beyond the frequency "swing" of the transmitter under modulation is relatively low.

One set of side frequencies that might be formed by frequency modulation is shown in figure 4. Unlike amplitude modulation, the strength of the component at the carrier frequency varies widely in FM and it may even disappear entirely under certain conditions. The variation of strength of the carrier component is useful in measuring the amount of frequency modulation, and will be discussed in detail later in this chapter.

One of the great advantages of FM over AM is the reduction in noise at the receiver which the system allows. If the receiver is made responsive only to changes in frequency, a considerable increase in signal-to-noise ratio is made possible through the use of FM, when the signal is of greater strength than the noise. The noise reducing capabilities of FM arise from the inability of noise to cause appreciable frequency modulation of the noise-plus-signal voltage which is applied to the detector in the receiver.

FM Terms Unlike amplitude modulation, the term "percentage modulation" means little in FM practice, unless the receiver characteristics are specified. There are, however, three terms, *deviation, modulation index,* and *deviation ratio,* which convey considerable information concerning the character of the FM wave.

Deviation is the amount of frequency shift each side of the unmodulated or "resting" carrier frequency which occurs when the transmitter is modulated. Deviation is ordinarily measured in kilocycles, and in a properly operating FM transmitter it will be directly proportional to the amplitude of the modulating signal. When a symmetrical modulating signal is applied to the transmitter, equal deviation each side of the resting frequency is obtained during each cycle of the modulating signal, and the total frequency range covered by the FM transmitter is sometimes known as the "swing." If, for instance, a transmitter operating on 1000 kc. has its frequency shifted from 1000 kc. to 1010 kc., back to 1000 kc., then to 990 kc., and again back to 1000 kc. during one cycle of the modulating wave, the *deviation* would be 10 kc. and the *swing* 20 kc.

The *modulation index* of an FM signal is the ratio of the deviation to the audio modulating frequency, when both are expressed in the same units. Thus, in the example above, if the signal is varied from 1000 kc. to 1010 kc. to 990 kc, and back to 1000 kc. at a rate (frequency) of 2000 times a second, the modulation index would be 5, since the deviation (10 kc.) is 5 times the modulating frequency (2000 cycles, or 2 kc.).

The relative strengths of the FM carrier and the various side frequencies depend directly upon the modulation index, these relative strengths varying widely as the modulation index is varied. In the preceding example, for instance, side frequencies occur on the high side of 1000 kc. at 1002, 1004, 1006, 1008, 1010, 1012, etc., and on the low frequency side at 998, 996, 994, 992, 990, 988, etc. In proportion to the unmodulated carrier strength (100 per cent), these side frequencies have the following strengths, as indicated by a modulation index of 5: 1002 and 998—33 per cent, 1004 and 996—5 per cent, 1006 and 994—36 per cent, 1008 and 992—39 per cent, 1010 and 990—26 per cent, 1012 and 988—13 per cent. The carrier strength (1000 kc.) will be 18 per cent of its unmodulated value. Changing the amplitude of the modulating signal will change the deviation, and thus the modulation index will be changed, with the result that the side frequencies, while still located in the same places, will have different strength values from those given above.

The *deviation ratio* is similar to the modulation index in that it involves the ratio between a modulating frequency and deviation. In this case, however, the deviation in question is the peak frequency shift obtained under full modulation, and the audio frequency to be considered is the maximum audio frequency to be transmitted. When the maximum audio frequency to be transmitted is 5000 cycles, for example, a deviation ratio of 3 would call for a peak deviation of 3 x 5000, or 15 kc. at full modulation. The noise-suppression capabilities of FM are directly related to the deviation ratio. As the deviation ratio is increased, the noise suppression becomes better *if* the signal is somewhat stronger than the noise. Where the noise approaches the signal in strength, however, low deviation ratios allow communication to be maintained in many cases where high-deviation-ratio FM and conventional AM are incapable of giving service. This assumes that a narrow-band FM receiver is in use. For each value of r-f signal-to-noise ratio at the receiver, there is a maximum deviation ratio which may be used, beyond which the output audio signal-to-noise ratio decreases. Up to this critical deviation ratio, however, the noise suppression becomes progressively better as the deviation ratio is increased.

For high-fidelity FM broadcasting purposes, a deviation ratio of 5 is ordinarily used, the maximum audio frequency being 15,000 cycles, and the peak deviation at full modulation being 75 kc. Since a swing of 150 kc. is covered by the transmitter, it is obvious that wideband FM transmission must necessarily be confined to the v-h-f range or higher, where room for the signals is available.

In the case of television sound, the deviation ratio is 1.67; the maximum modulation frequency is 15,000 cycles, and the transmitter deviation for full modulation is 25 kc. The sound carrier frequency in a standard TV signal is located exactly 4.5 Mc. higher than the picture carrier frequency. In the *intercarrier* TV sound system, which recently has become quite widely used, this constant difference between the picture carrier and the sound carrier is employed within the receiver to obtain an FM sub-carrier at 4.5 Mc. This 4.5 Mc. sub-carrier then is demodulated by the FM detector to obtain the sound signal which accompanies the picture.

Narrow-Band FM Transmission Narrow-band FM transmission has become standardized for use by the mobile services such as police, fire, and taxicab communication, and also on the basis of a temporary authorization for amateur work in portions of each of the amateur radiotelephone bands. A maximum deviation of 15 kc. has been standardized for the mobile and commercial communication services, while a maximum deviation of 3 kc. is authorized for amateur NBFM communication.

Bandwidth Required by FM As the above discussion has indicated, many side frequencies are set up when a radio-frequency carrier is frequency modulated; theoretically, in fact, an infinite number of side frequencies is formed. Fortunately, however, the amplitudes of those side frequencies falling outside the frequency range over which the transmitter is "swung" are so small that most of them may be ignored. In FM transmission, when a complex modulating wave (speech or music) is used, still additional side frequencies resulting from a beating together of the various frequency components in the modulating wave are formed. This is a situation that does not occur in amplitude modulation and it might be thought that the large number of side frequencies thus formed might make the frequency spectrum produced by an FM transmitter prohibitively wide. Analysis shows, however, that the additional side frequencies are of very small amplitude, and, instead of increasing the bandwidth, modulation by a complex wave actually reduces the effective bandwidth of the FM wave. This is especially true when speech modulation is used, since most of the power in voiced sounds is concentrated at low frequencies in the vicinity of 400 cycles.

The bandwidth required in an FM receiver is a function of a number of factors, both theoretical and practical. Basically, the bandwidth required is a function of the deviation ratio and the maximum frequency of modulation, although the practical consideration of drift and ease of receiver tuning also must be considered. Shown in figure 5 are the frequency spectra (carrier and sideband frequencies) associated with the standard FM broadcast signal, the TV sound signal, and an amateur-band narrow-band FM signal with full modulation using the highest permissible modulating frequency in each case. It will be seen that for low deviation ratios the receiver bandwidth should be at least four times the maximum frequency deviation, but for a deviation ratio of 5 the receiver bandwidth need be only about 2.5 times the maximum frequency deviation.

9-1 Direct FM Circuits

Frequency modulation may be obtained either by the direct method, in which the frequency of an oscillator is changed directly by the modulating signal, or by the indirect method which makes use of phase modulation. Phase-modulation circuits will be discussed in section 9-2.

A successful frequency modulated transmitter must meet two requirements: (1) The frequency deviation must be symmetrical about a fixed frequency, for symmetrical modulation voltage. (2) The deviation must be directly proportional to the amplitude of the modulation, and independent of the modulation frequency. There are several methods of direct frequency modulation which will fulfill these requirements. Some of these methods will be described in the following paragraphs.

Reactance-Tube Modulators One of the most practical ways of obtaining direct frequency modulation is through the use of a *reactance-tube modulator*. In this arrangement the modulator plate-cathode circuit is connected across the oscillator tank circuit, and made to appear as either a capacitive or inductive reactance by exciting the modulator grid with a voltage which either leads or lags the oscillator tank voltage by 90 degrees. The leading or lagging grid voltage causes a corresponding leading or lagging plate current, and the plate-cathode circuit appears as a capacitive or inductive reactance across the oscillator tank circuit. When the transconductance of the modulator tube is varied, by varying one of the element voltages, the magnitude of the reactance across the oscillator tank is varied. By applying audio modulating voltage to one of the elements, the transconductance, and hence the frequency, may be varied at an audio rate. When properly designed and operated, the reactance-tube modulator gives linear frequency modulation, and is capable of producing large amounts of deviation.

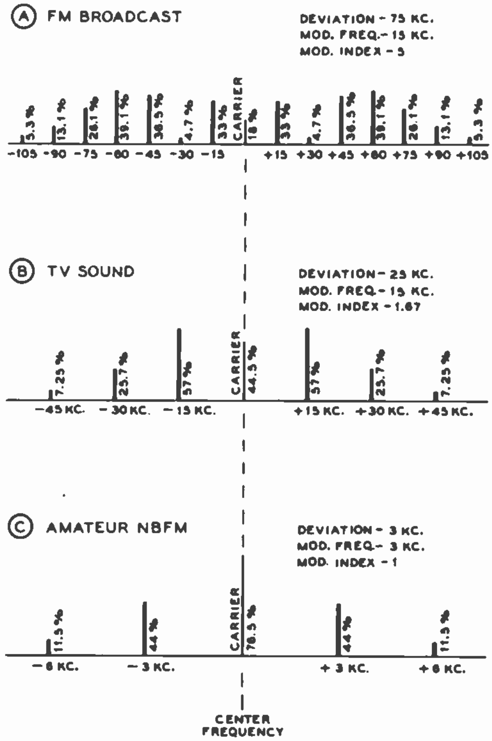


Figure 5.
EFFECT OF FM MODULATION INDEX.
Showing the side-frequency amplitude and distribution for the three most common modulation indices used in FM work. The maximum modulating frequency and maximum deviation are shown in each case.

There are numerous possible configurations of the reactance-tube modulator circuit. The difference in the various arrangements lies principally in the type of phase-shifting circuit used to give a grid voltage which is in phase quadrature with the r-f voltage at the modulator plate.

Figure 6 is a diagram of one of the most popular forms of reactance-tube modulators. The modulator tube, which is usually a pentode

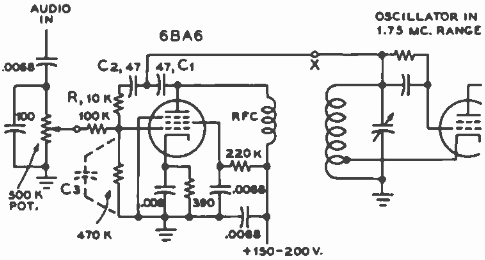


Figure 6.
REACTANCE-TUBE MODULATOR.
This circuit is convenient for direct frequency modulation of an oscillator in the 1.75-Mc. range. Capacitor C_3 may be only the input capacitance of the tube, or a small trimmer capacitor may be included to permit a variation in the sensitivity of the reactance tube.

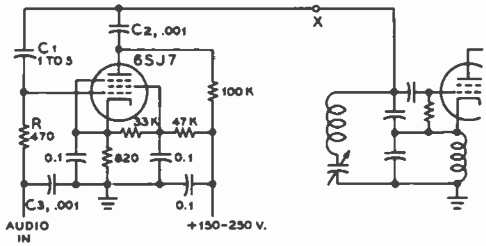


Figure 7.
ALTERNATIVE REACTANCE-TUBE MODULATOR.
This circuit often is preferable for use in the lower frequency range, although it may be used at 1.75 Mc. and above if desired. In the schematic above the reactance tube is shown connected across the voltage-divider capacitors of a Clapp oscillator, although the modulator circuit may be used with any common type of oscillator.

such as a 6BA6, 6AU6, or 6SJ7, has its plate coupled through a blocking capacitor, C_1, to the "hot" side of the oscillator grid circuit. Another blocking capacitor, C_2, feeds r.f. to the phase shifting network R-C_3 in the modulator grid circuit. If the resistance of R is made large in comparison with the reactance of C_3 at the oscillator frequency, the current through the R-C_3 combination will be nearly in phase with the voltage across the tank circuit, and the voltage across C_3 will lag the oscillator tank voltage by almost 90 degrees. The result of the 90-degree lagging voltage on the modulator grid is that its plate current lags the tank voltage by 90 degrees, and the reactance tube appears as an inductance in shunt with the oscillator inductance, thus raising the oscillator frequency.

The phase-shifting capacitor C_3 can consist of the input capacitance of the modulator tube and stray capacitance between grid and ground. However, better control of the operating conditions of the modulator may be had through the use of a variable capacitor as C_3. Resistance R will usually have a value of between 4700 and 100,000 ohms. Either resistance or transformer coupling may be used to feed audio voltage to the modulator grid. When a resistance coupling is used, it is necessary to shield the grid circuit adequately, since the high impedance grid circuit is prone to pick up stray r-f and low frequency a-c voltage, and cause undesired frequency modulation.

An alternative reactance modulator circuit is shown in figure 7. The operating conditions are generally the same, except that the r-f excitation voltage to the grid of the reactance tube is obtained effectively through reversing the R and C_3 of figure 6. In this circuit a small capacitance is used to couple r.f. into the grid of the reactance tube, with a relatively small value of resistance from grid to ground. This circuit has the advantage that the grid of the tube is at a relatively low impedance with respect to r.f. However, the circuit normally is not suitable for operation above a few megacycles due to the shunting capacitance within the tube from grid to ground.

Either of the reactance-tube circuits may be used with any of the common types of oscillators. The reactance modulator of figure 6 is shown connected to the high-impedance point of a conventional hot-cathode Hartley oscillator, while that of figure 7 is shown connected across the low-impedance capacitors of a series-tuned Clapp oscillator.

There are several possible variations of the basic reactance-tube modulator circuits shown in figures 6 and 7. The audio input may be applied to the suppressor grid, rather than the control grid, if desired. Another modification is to apply the audio to a grid other than the control grid in a mixer or pentagrid converter tube which is used as the modulator. Generally, it will be found that the transconductance variation per volt of control-element voltage variation will be greatest when the control (audio) voltage is applied

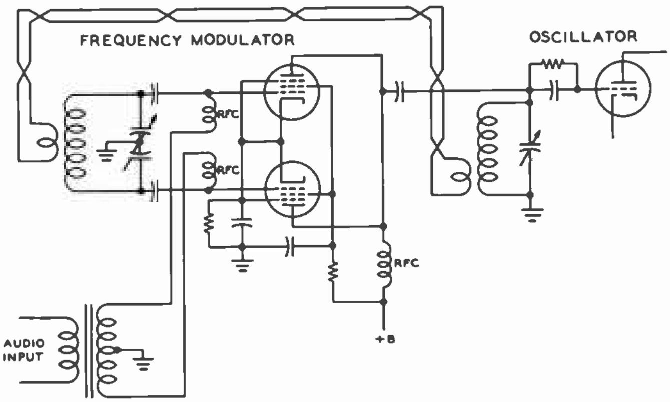


Figure 8.
BALANCED REACTANCE-TUBE MODULATOR.
Frequency shift due to voltage changes on the modulator may be greatly reduced by the use of this circuit. Changes in element voltages cause equal and opposite changes in reactance in the two modulators, thus minimizing the frequency shift. The reactance-tubes' grids receive excitation from a balanced tuned circuit so that one tube receives voltage lagging the oscillator tank voltage by 90°, while the other tube is excited with a voltage that leads the tank voltage by 90°.

to the control grid. In cases where it is desirable to separate completely the audio and r-f circuits, however, applying audio voltage to one of the other elements will often be found advantageous despite the somewhat lower sensitivity.

Adjusting the Phase Shift One of the simplest methods of adjusting the phase shift to the correct amount is to place a pair of earphones in series with the oscillator cathode-to-ground circuit and adjust the phase-shift network until minimum sound is heard in the 'phones when frequency modulation is taking place. If an electron-coupled or Hartley oscillator is used, this method requires that the cathode circuit of the oscillator be inductively or capacitively coupled to the grid circuit, rather than tapped on the grid coil. The 'phones should be adequately by-passed for r.f. of course.

Stabilization Due to the presence of the reactance-tube frequency modulator, the stabilization of an FM oscillator in regard to voltage changes is considerably more involved than in the case of a simple self-controlled oscillator for transmitter frequency control. If desired, the oscillator itself may be made perfectly stable under voltage changes, but the presence of the frequency modulator destroys the beneficial effect of any such stabilization. It thus becomes desirable to apply the stabilizing arrangement to the modulator as well as the oscillator. If the oscillator itself is stable under voltage changes, it is only necessary to apply voltage-frequency compensation to the modulator.

A circuit in which automatic stabilization of the effects of voltage variations on the modulator is obtained, is shown in figure 8. In this circuit, the reactance-tube grids are connected in push-pull across the phase-shifting tank circuit, while the plates are connected in parallel and tied to the oscillator tank in the usual manner. Any variation in the plate-supply voltage to the reactance tubes causes equal and opposite effects in their reactance, and there is no net reactance variation.

Another method of oscillator stabilization makes use of a discriminator circuit. This arrangement stabilizes the frequency against changes arising from any cause (except the desired modulation) by comparing the oscillator frequency with a crystal controlled standard and applying the proper compensating voltages. A block diagram of this method is shown in figure 9. Output from one of the stages of the transmitter is mixed with the output of a crystal oscillator to give an "intermediate frequency" output which is applied to a conventional discriminator. The discriminator, which will be more completely described later in this chapter, is a circuit arrangement to produce an output voltage which depends on the frequency of the r.f. applied to it.

The d-c voltage produced by the discriminator is applied to a reactance tube tied across the oscillator tank circuit. As the average or "center" frequency varies one way or the other from the correct value, a positive or negative voltage appears across the discriminator load resistors. When this voltage is placed on the control element of the reactance tube, it attempts to restore the center (mid-modulation or unmodulated) radio frequency to a value which gives zero voltage output from the discriminator. The oscillator can never be fully restored to its correct frequency, however, since the discriminator output voltage would then be zero, and no frequency correction would be taking place. The frequency is actually shifted back to a value somewhere between what it should be and what it would have been without stabilization. The reactance tube which takes care of the frequency correction may also be used as the modulator, and the frequency stabilizing voltage may be applied in series with the audio voltage or, alternatively, it may be applied to another of the tube elements.

The audio output of the discriminator must be removed by a simple R-C filter so that the compensating voltage is direct current without superimposed audio. The audio output of the discriminator may be used for monitoring purposes, if desired. Obviously the stability of the complete arrangement is dependent upon the stability of the discriminator components under temperature and humidity changes, and upon the stability of the crystal oscillator. Ordinarily the stability of the crystal oscillator will be sufficiently great that the discriminator will be the limiting factor in the amount of stabilization obtainable, making it necessary to use discriminator components (especially the tuned input transformer) of good quality.

The frequency of the crystal used in the stabilizing circuit will depend upon the frequency at which the discriminator operates, and the frequency of the stage in the transmitter from which the stabilizer signal is taken. If a b.c. replacement-type discriminator transformer designed for a frequency in the 400-500 kc. range is used, the r-f input for the stabilizer may be obtained from the transmitter oscillator stage, or if more sensitivity is desired, from the plate circuit of the frequency multiplier following the oscillator. The crystal oscillator must operate on a frequency such that its fundamental, or one of its harmonics, falls at a frequency which differs from that of the transmitter stage applied to the stabilizer by an amount equal to the discriminator frequency. If the required crystal frequency falls higher than is easily obtainable with a crystal, it may be necessary to use a frequency multiplier after the crystal stage.

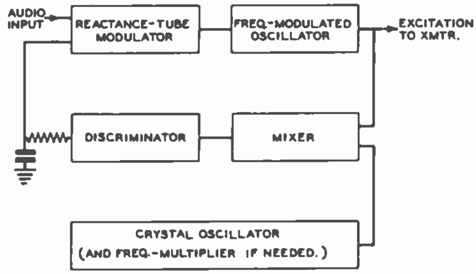


Figure 9.
STABILIZATION SYSTEM.
A frequency-modulated oscillator may be stabilized against undesired frequency shift by comparing the transmitter frequency with that of a crystal oscillator. The difference between the two frequencies is applied to a discriminator circuit, and any change from a predetermined difference will cause the discriminator to restore the transmitter to its correct frequency. An R-C filter is used to remove the audio modulation from the discriminator output.

Linearity Test It is almost a necessity to run a static test on the reactance-tube frequency modulator to determine its linearity and effectiveness, since small changes in the values of components, and in stray capacitances will almost certainly alter the modulator characteristics. A frequency-versus-control-voltage curve should be plotted to ascertain that equal increments in control voltage, both in a positive and a negative direction, cause equal changes in frequency. If the curve shows that the modulator has an appreciable amount of non-linearity, changes in bias, electrode voltages, r-f excitation, and resistance values may be made to obtain a straight-line characteristic.

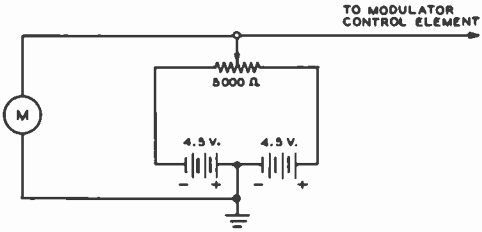


Figure 10.
REACTANCE-TUBE LINEARITY CHECKER.

Figure 10 shows a method of connecting two 4½-volt C batteries and a potentiometer to plot the characteristic of the modulator. It will be necessary to use a zero-center voltmeter to measure the grid voltage, or else reverse the voltmeter leads when changing from positive to negative grid voltage. When a straight-line characteristic for the modulator is obtained by the static test method, the capacitances of the various by-pass capacitors in the circuit must be kept small to retain this characteristic when an audio voltage is used to vary the frequency in place of the d-c voltage with which the characteristic was plotted.

9-2 Phase Modulation

By means of phase modulation (PM) it is possible to dispense with self-controlled oscillators and to obtain directly crystal-controlled FM. In the final analysis, PM is simply frequency modulation in which the deviation is directly proportional to the modulation frequency. If an audio modulating signal of 1000 cycles causes a deviation of ½ kc., for example, a 2000-cycle modulating signal of the same amplitude will give a deviation of 1 kc., and so on. To produce an FM signal, it is necessary to make the deviation independent of the modulation frequency, and proportional only to the amplitude of the modulating signal. With PM this is done by including a frequency correcting network in the audio system of the transmitter. The audio correction network must have an attenuation that varies directly with frequency, and this requirement is easily met by a very simple resistance-capacity network.

The only disadvantage of PM, as compared to direct FM such as is obtained through the use of a reactance-tube modulator, is the fact that very little frequency deviation is produced directly by the phase modulator. The deviation produced by a phase modulator is independent of the actual carrier frequency on which the modulator operates, but is dependent only upon the phase deviation which is being produced and upon the modulation frequency. Expressed as an equation:

$$F_d = M_p \text{ modulating frequency}$$

Where F_d is the frequency deviation one way from the mean value of the carrier, and M_p is the phase deviation accompanying modulation expressed in radians (a radian is approximately 57.3°). Thus, to take an example, if the phase deviation is ½ radian and the modulating frequency is 1000 cycles, the frequency deviation applied to the carrier being passed through the phase modulator will be 500 cycles.

It is easy to see that an enormous amount of multiplication of the carrier frequency is required in order to obtain from a phase modulator the frequency deviation of 75 kc. required for commercial FM broadcasting. However, for amateur and commercial narrow-band FM work (NBFM) only a quite reasonable number of multiplier stages are required to obtain a deviation ratio of approximately one. Actually, phase modulation of approximately one-half radian on the output of a crystal oscillator in the 80-meter band will give adequate deviation for 29-Mc. NBFM radiotelephony. For example; if the crystal frequency is 3700 kc., the deviation in phase produced is ½ radian, and the modulating frequency is 500 cycles, the deviation in the 80-meter band will be 250 cycles. But when the crystal frequency is multiplied on up to 29,600 kc. the frequency deviation will also be multiplied by 8 so that the resulting deviation on the 10-meter band will be 2 kc. either side of the carrier for a total swing in carrier frequency of 4 kc. This amount of deviation is quite adequate for NBFM work.

Odd-harmonic distortion is produced when FM is obtained by the phase-modulation method, and the amount of this distortion that can be tolerated is the limiting factor in determining the amount of PM that can be used. Since the aforementioned frequency-correcting network causes the lowest modulating frequency to have the greatest amplitude, maximum phase modulation takes place at the lowest modulating frequency, and the amount of dis-

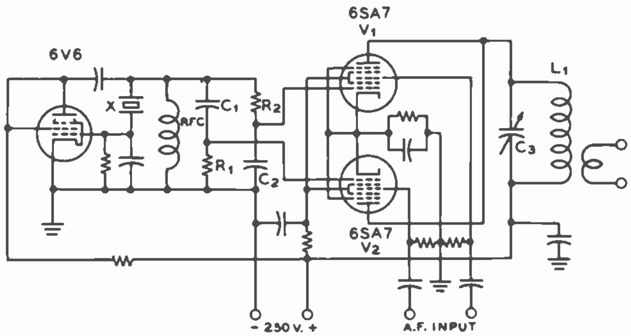


Figure 11.
SIMPLE PHASE-MODULATION CIRCUIT.
The operation of this phase-modulation circuit for obtaining FM is described in detail in the text. R_1C_1 and R_2C_2 comprise the phase-splitting network for the two 6SA7 phase-modulator tubes. The tank circuit L_1C_3 is tuned to the frequency of operation of the crystal.

tortion that can be tolerated at this frequency determines the maximum deviation that can be obtained by the PM method. For high-fidelity broadcasting, the deviation produced by PM is limited to an amount equal to about one-third of the lowest modulating frequency. But for NBFM work the deviation may be as high as 0.6 of the modulating frequency before distortion becomes objectionable on voice modulation. In other terms this means that phase deviations as high as 0.6 radian may be used for amateur and commercial NBFM transmission.

Phase-Modulation Circuits A large number of circuits for producing phase modulation have been developed and introduced. Most of the circuits use conventional tubes such as shown in figures 11, 12, and 13, but some make use of special tubes which have been designed expressly for accomplishing phase modulation of a carrier wave. One stable and relatively simple PM circuit is shown in figure 11.

A 6V6 tube, triode connected, is used as a Pierce crystal oscillator in the circuit of figure 11 to feed the two phase-splitting networks R_1C_1 and R_2C_2. R_1C_1 effectively advances the phase 45° while R_2C_2 effectively retards the phase 45° from the phase of the voltage generated by the crystal oscillator. Hence, the grids of the two 6SA7 phase-modulator tubes are fed with voltages 90° out of phase. The plates of the two 6SA7 tubes are connected in parallel and thence to the tank circuit L_1C_1. This tank circuit is tuned to the frequency of the crystal oscillator. Then, as we apply push-pull audio voltage to the signal grids of the two 6SA7 tubes the G_m of one tube is effectively decreased while the G_m of the other is increased on one half of the input audio cycle, and the reverse effect takes place on the other half of the audio cycle. When the G_m of V_1 is increased, the G_m of V_2 is decreased and the phase of the voltage across the output tank circuit tends to take on the phase of the voltage generated by V_1. The converse takes place on the other half of the audio cycle.

Figure 12 is a block diagram of a phase modulator which is capable of producing about twice the phase deviation of the circuit of figure 11 for the same amount of non-linearity or distortion. The circuit of figure 11 is cap-

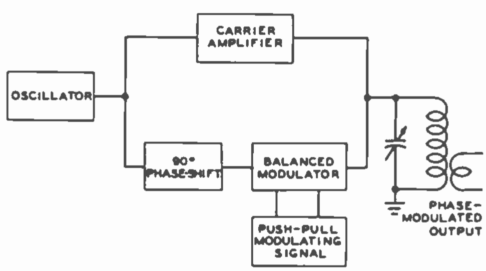


Figure 12.
BALANCED-MODULATOR PHASE-MODULATION CIRCUIT.
This type of circuit, although more complex than the usual phase-modulation circuits, is capable of greater phase deviation with the same distortion tolerance.

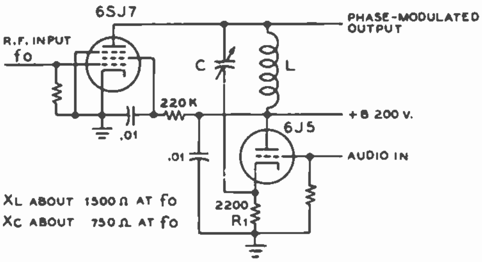


Figure 13.
CATHODE-FOLLOWER PHASE MODULATOR.
The phase modulator illustrated above is quite satisfactory when the stage is to be operated on a single frequency or over a narrow range of frequencies.

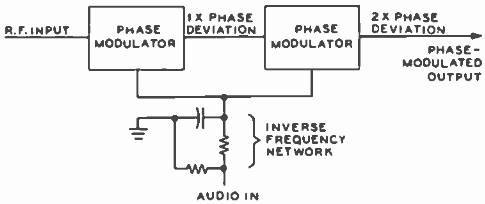


Figure 14.
CASCADE PHASE MODULATOR.
The phase deviation attainable with phase modulation may be increased, without changing the center frequency, through the use of cascaded phase modulator stages. Although two stages are illustrated above, many more stages may be used, as has been done in commercial equipment. The polarity of audio signal fed to the phase modulators must be poled in the correct phase so that the phase modulation introduced by each successive stage will be additive to that produced by the preceding stage.

able of about ½ radian phase deviation with distortion tolerable for communications work, while the circuit of figure 12 can produce about 1 radian deviation. However, the circuit of figure 12 requires one additional tube and several additional components.

Still another PM circuit, which is quite widely used commercially, is shown in figure 13. In this circuit L and C are made resonant at a frequency which is 0.707 times the operating frequency. Hence at the operating frequency the inductive reactance is twice the capacitive reactance. A cathode follower tube acts as a variable resistance in series with the L and C which go to make up the tank circuit. The operating point of the cathode follower should be chosen so that the effective resistance in series with the tank circuit (made up of the resistance of the cathode-follower tube in parallel with the cathode bias resistor of the cathode follower) is equal to the capacitive reactance of the tank capacitor at the operating frequency. The circuit is capable of about plus or minus ½ radian deviation with tolerable distortion, the same as the circuit of figure 11.

Methods for Obtaining Greater Phase Deviation When it is desired to utilize phase modulation for obtaining wide-band FM, some method of obtaining phase multiplication greater than that which would be obtained by simple multiplication from the crystal frequency down to the output frequency must be employed. One method is to cascade a series of phase modulator stages in the manner shown in figure 14. Another method is to use a moderately high-frequency oscillator followed by a small amount of frequency multiplication, and then to "heterodyne back" the signal by means of a heterodyne oscillator and a mixer to another moderately high-frequency, whence it may be multiplied in the usual manner to the output frequency.

An example of this method is the use of a crystal oscillator, followed by the phase modulator, on 1800 kc. The PM output is tripled to 5400 kc., where the deviation is then 3 times what it originally was. Beating the 5400-kc. output with another crystal oscillator on 7350 kc. gives a difference of frequency of 1950 kc., with the deviation still tripled from its original value. By a series of triplers the 1950-kc. signal may be multiplied 27 times to reach a frequency of 52.65 Mc., which is in the 52.5 to 54 Mc. amateur band. The increase in deviation will be equal to the product of the two frequency multiplications (3 x 27) or 81 times.

Measurement of Deviation When a single-frequency modulating voltage is used with an FM transmitter, the relative amplitudes of the various sidebands and the carrier vary widely as the deviation is varied by increasing or decreasing the amount of modulation. Since the relationship between the amplitudes of the various sidebands and carrier to the audio modulating frequency and the deviation is known, a simple method of measuring the deviation of a frequency modulated transmitter is possible. In making the measurement, the result is given in the form of the modulation index for a certain amount of audio input. As previously described, the modulation index is the ratio of the peak frequency deviation to the frequency of the audio modulation.

The measurement is made by applying a sine-wave audio voltage of known frequency to the transmitter, and increasing the modulation until the amplitude of the carrier component of the frequency modulated wave reaches zero. The modulation index for zero carrier may then be determined from the table below. As may be seen from the table, the first point of zero carrier is obtained when the modulation index has a value of 2.405,—in other words, when the deviation is 2.405 times the modulation frequency. For example, if a modulation frequency of 1000 cycles is used, and the modulation is increased until the first carrier null is obtained, the deviation will then be 2.405 times the modulation frequency, or 2.405 kc. If the modulating frequency happened to be 2000 cycles, the deviation at the first null would be 4.810 kc. Other carrier nulls will be obtained when the index is 5.52, 8.654, and at increasing values separated approximately by π. The following is a listing of the modulation index at successive carrier nulls up to the tenth:

Zero carrier point no.	Modulation index
1	2.405
2	5.520
3	8.654
4	11.792
5	14.931
6	18.071
7	21.212
8	24.353
9	27.494
10	30.635

The only equipment required for making the measurements is a calibrated audio oscillator of good wave form, and a communication receiver equipped with a beat oscillator and crystal filter. The receiver should be used with its crystal filter set for minimum bandwidth to exclude sidebands spaced from the carrier by the modulation frequency. The unmodulated carrier is accurately tuned in on the receiver with the beat oscillator operating. Then modulation from the audio oscillator is applied to the transmitter, and the modulation is increased until the first carrier null is obtained. This first carrier null will correspond to a modulation index of 2.405, as previously mentioned. Successive null points will correspond to the indices listed in the table.

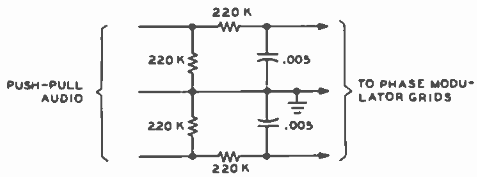


Figure 15.
INVERSE-FREQUENCY CIRCUIT.
Through the use of a circuit such as this between the audio signal source and the grids of the phase modulator, phase modulation will be effectively converted to frequency modulation at all frequencies above about 200 cycles.

A volume indicator in the transmitter audio system may be used to measure the audio level required for different amounts of deviation, and the indicator thus calibrated in terms of frequency deviation. If the measurements are made at the fundamental frequency of the oscillator, it will be necessary to multiply the frequency deviation by the harmonic upon which the transmitter is operating, of course. It will probably be most convenient to make the determination at some frequency intermediate between that of the oscillator and that at which the transmitter is operating, and then to multiply the result by the frequency multiplication between that frequency and the transmitter output frequency.

9-3 Reception of FM Signals

A conventional communications receiver may be used to receive narrow-band FM transmissions, although performance will be much poorer than can be obtained with an NBFM receiver or adapter. However, a receiver specifically designed for FM reception must be used when it is desired to receive high deviation FM such as used by FM broadcast stations, TV sound, and mobile communications FM.

The FM receiver must have, first of all, a bandwidth sufficient to pass the range of fre-

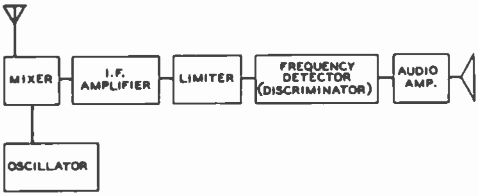


Figure 16.
FM RECEIVER BLOCK DIAGRAM.
Up to the amplitude limiter stage, the FM receiver is similar to an AM receiver, except for a somewhat wider i-f bandwidth. The limiter removes any amplitude modulation, and the frequency detector following the limiter converts frequency variations into amplitude variations.

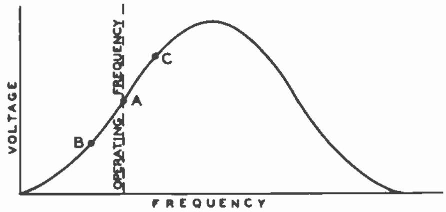


Figure 17.
"OFF TUNE" FREQUENCY DETECTOR.
One side of the response characteristic of a tuned circuit or of an i-f amplifier may be used as shown to convert frequency variations of an incoming signal into amplitude variations.

quencies generated by the FM transmitter. And since the receiver must be a superheterodyne if it is to have good sensitivity at the frequencies to which FM is restricted, i-f bandwidth is an important factor in its design.

The second requirement of the FM receiver is that it incorporate some sort of device for converting frequency changes into amplitude changes, in other words, a detector operating on frequency variations rather than amplitude variations. The third requirement, and one which is necessary if the full noise reducing capabilities of the FM system of transmission are desired, is a limiting device to eliminate amplitude variations before they reach the detector. A block diagram of the essential parts of an FM receiver is shown in figure 16.

The Frequency Detector The simplest device for converting frequency variations to amplitude variations is an "off-tune" resonant circuit, as illustrated in figure 17. With the carrier tuned in at point "A," a certain amount of r-f voltage will be developed across the tuned circuit, and, as the frequency is varied either side of this frequency by the modulation, the r-f voltage will increase and decrease to points "C" and "B" in accordance with the modulation. If the voltage across the tuned circuit is applied to an ordinary detector, the detector output will vary in accordance with the modulation, the amplitude of the variation being proportional to the deviation of the signal, and the rate being equal to the modulation frequency. It is obvious from figure 17 that only a small portion of the resonance curve is usable for linear conversion of frequency variations into amplitude variations, since the linear portion of the curve is rather short. Any frequency variation which exceeds the linear portion will cause distortion of the recovered audio. It is also obvious by inspection of figure 17 that an AM receiver used in this manner is wide open to signals on the peak of the resonance curve and also to signals on the other side of the resonance curve. Further, no noise limiting action is afforded by this type of reception. This system, therefore, is not recommended for FM reception, although widely used by amateurs for occasional NBFM reception.

Travis Discriminator Another form of frequency detector or *discriminator,* is shown in figure 18. In this arrangement two tuned circuits are used, one tuned on each side of the i-f amplifier frequency, and with their resonant frequencies spaced slightly more than the expected transmitter "swing." Their outputs are combined

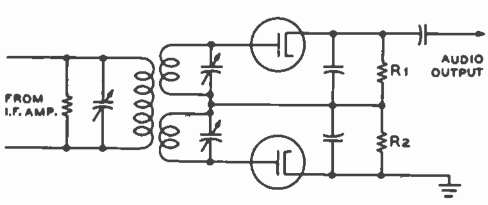
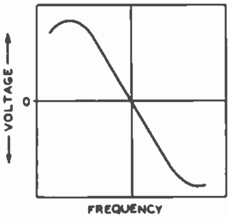


Figure 18.
TRAVIS DISCRIMINATOR.
This type of discriminator makes use of two off-tuned resonant circuits coupled to a single primary winding. The circuit is capable of excellent linearity, but is difficult to align.



At its "center" frequency the discriminator produces zero output voltage. On either side of this frequency it gives a voltage of a polarity and magnitude which depend on the direction and amount of frequency shift.

Figure 19.
DISCRIMINATOR VOLTAGE-FREQUENCY CURVE.

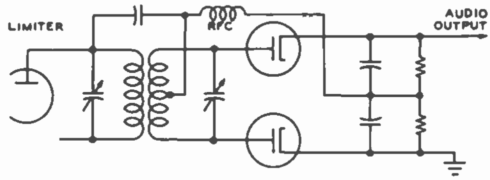


Figure 20.
FOSTER-SEELEY DISCRIMINATOR.
This discriminator is the most widely used circuit since it is capable of excellent linearity and is relatively simple to align when proper test equipment is available.

in a differential rectifier so that the voltage across the series load resistors, R_1 and R_2, is equal to the algebraic sum of the individual output voltages of each rectifier. When a signal at the i-f mid-frequency is received, the voltages across the load resistors are equal and opposite, and the sum voltage is zero. As the r-f signal varies from the mid-frequency, however, these individual voltages become unequal, and a voltage having the polarity of the larger voltage and equal to the difference between the two voltages appears across the series resistors, and is applied to the audio amplifier. The relationship between frequency and discriminator output voltage is shown in figure 19. The separation of the discriminator peaks and the linearity of the output voltage vs. frequency curve depend upon the discriminator frequency, the Q of the tuned circuits, and the value of the diode load resistors. As the intermediate (and discriminator) frequency is increased, the peaks must be separated further to secure good linearity and output. Within limits, as the diode load resistance or the Q is reduced, the linearity improves, and the separation between the peaks must be greater.

Foster-Seeley Discriminator The most widely used form of discriminator is that shown in figure 20. This type of discriminator yields an output-voltage-versus-frequency characteristic similar to that shown in figure 19. Here, again, the output voltage is equal to the algebraic sum of the voltages developed across the load resistors of the two diodes, the resistors being connected in series to ground. However, this *Foster-Seeley* discriminator requires only two tuned circuits instead of the three used in the previous discriminator.

The operation of the circuit results from the phase relationships existing in a transformer having a tuned secondary. In effect, as a close examination of the circuit will reveal, the primary circuit is in series, for r.f., with each half of the secondary to ground. When the received signal is at the resonant frequency of the secondary, the r-f voltage across the secondary is 90 degrees out of phase with that across the primary. Since each diode is connected across one half of the secondary winding and the primary winding in series, the resultant r-f voltages applied to each are equal, and the voltages developed across each diode load resistor are equal and of opposite polarity. Hence, the net voltage between the top of the load resistors and ground is zero. This is shown vectorially in figure 21A where the resultant voltages R and R′ which are applied to the two diodes are shown to be equal when the phase angle between primary and secondary voltages is 90 degrees. If, however, the signal varies from the resonant frequency, the 90-degree phase relationship no longer exists between primary and secondary. The result of this effect is shown in figure 21B, where the secondary r-f voltage is no longer 90 degrees out of phase with respect to the primary voltage. The resultant voltages applied to the two diodes are now no longer equal, and a d-c voltage proportional to the difference between the r-f voltages applied to the two diodes will exist across the series load resistors. As the signal frequency varies back and forth across the resonant frequency of the discriminator, an a-c voltage of the same frequency as the original modulation, and proportional to the deviation, is developed and passed on to the audio amplifier.

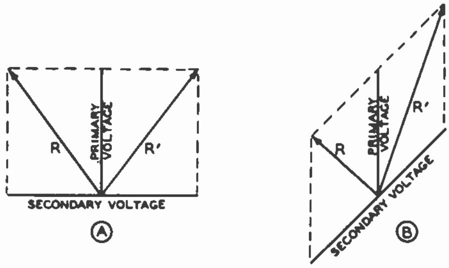


Figure 21.
DISCRIMINATOR VECTOR DIAGRAM.
A signal at the resonant frequency of the secondary will cause the secondary voltage to be 90 degrees out of phase with the primary voltage, as shown at A, and the resultant voltages R and R' are equal. If the signal frequency changes, the phase relationship also changes, and the resultant voltages are no longer equal, as shown at B. A differential rectifier is used to give an output voltage proportional to the difference between R and R'.

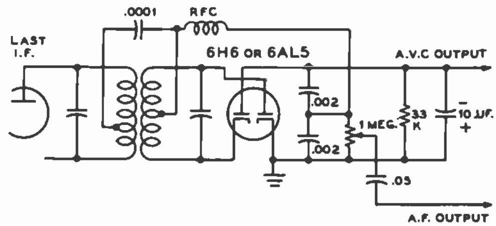


Figure 22.
RATIO DETECTOR CIRCUIT.
The parallel voltage to the diodes in a ratio detector may be obtained from a tap on the primary winding of the transformer or from a third winding. Note that one of the diodes is reversed from the system used with the Foster-Seeley discriminator, and that the output circuit is completely different. The ratio detector does not have to be preceded by a limiter, but is more difficult to align for distortion-free output than the conventional discriminator.

Ratio Detector One of the more recent types of FM detector circuits, called the "ratio detector" is diagrammed in figure 22. The input transformer can be designed so that the "parallel" input voltage to the diodes can be taken from a tap on the primary of the transformer, or this voltage may be obtained from a tertiary winding coupled to the primary. The r-f choke used must have high impedance at the intermediate frequency used in the receiver, although this choke is not needed if the transformer has a tertiary winding.

The circuit of the ratio detector appears very similar to that of the more conventional discriminator arrangement. However, it will be noted that the two diodes in the ratio detector are poled so that their d-c output voltages add, as contrasted to the Foster-Seeley circuit wherein the diodes are poled so that the d-c output voltages buck each other. At the center frequency to which the discriminator transformer is tuned the voltage appearing at the top of the 1-megohm potentiometer will be one-half the d-c voltage appearing at the "a-v-c output" terminal—since the contribution of each diode will be the same. However, as the input frequency varies to one side or the other of the tuned value (while remaining within the pass band of the i-f amplifier feeding the detector) the relative contributions of the two diodes will be different. The voltage appearing at the top of the 1-megohm volume control will increase for frequency deviations in one direction and will decrease for frequency deviations in the other direction from the mean or tuned value of the transformer. The audio output voltage is equal to the ratio of the relative contributions of the two diodes, hence the name "ratio detector."

The ratio detector offers several advantages over the simple discriminator circuit. The circuit does not require the use of a limiter preceding the detector since the circuit is inherently insensitive to amplitude modulation on an incoming signal. This factor alone means that the r-f and i-f gain ahead of the detector can be much less than with the conventional discriminator for the same overall sensitivity. Further, the circuit provides a-v-c voltage for controlling the gain of the preceding r-f and i-f stages. The ratio detector is, however, susceptible to variations in the amplitude of the incoming signal as is any other detector circuit except the discriminator *with* a limiter preceding it, so that a-v-c should be used on the stages preceding the detector.

Limiters The limiter in an FM receiver using a conventional discriminator serves to remove amplitude modulation and pass on to the discriminator a frequency modulated signal of constant amplitude; a typical circuit is shown in figure 23. The limiter tube

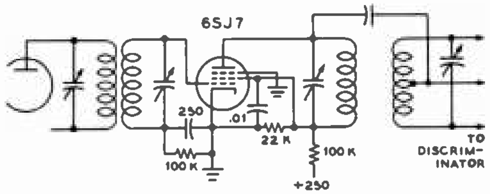


Figure 23.
LIMITER CIRCUIT.
One, or sometimes two, limiter stages normally precede the discriminator so that a constant signal level will be fed to the FM detector. This procedure eliminates amplitude variations in the signal fed to the discriminator, so that it will respond only to frequency changes.

is operated as an i-f stage with very low plate voltage and with grid leak bias, so that it overloads quite easily. Up to a certain point the output of the limiter will increase with an increase in signal. Above this point, however, the limiter becomes overloaded, and further large increases in signal will not give any increase in output. To operate successfully, the limiter must be supplied with a large amount of signal, so that the amplitude of its output will not change for rather wide variations in amplitude of the signal. Noise, which causes little frequency modulation but much amplitude modulation of the received signal, is virtually wiped out in the limiter.

The voltage across the grid resistor varies with the amplitude of the received signal. For this reason, conventional amplitude modulated signals may be received on the FM receiver by connecting the input of the audio amplifier to the top of this resistor, rather than to the discriminator output. When properly filtered by a simple R-C circuit, the voltage across the grid resistor may also be used as a-v-c voltage for the receiver. When the limiter is operating properly, a.v.c. is neither necessary nor desirable, however, for FM reception alone.

Receiver Design Considerations One of the most important factors in the design of an FM receiver is the frequency swing which it is intended to handle. It will be apparent from figure 19 that if the straight portion of the discriminator circuit covers a wider range of frequencies than those generated by the transmitter, the audio output will be reduced from the maximum value of which the receiver is capable.

In this respect, the term "modulation percentage" is more applicable to the FM receiver than it is to the transmitter, since the "modulation capability" of the communication system is limited by the receiver bandwidth and the discriminator characteristic; full utilization of the linear portion of the characteristic amounts, in effect, to "100 per cent" modulation. This means that some sort of standard must be agreed upon, for any particular type of communication, to make it unnecessary to vary the transmitter swing to accommodate different receivers.

Two considerations influence the receiver bandwidth necessary for any particular type of communication. These are the maximum audio frequency which the system will handle, and the deviation ratio which will be employed. For voice communication, the maximum audio frequency is more or less fixed at 3000 to 4000 cycles. In the matter of deviation ratio, however, the amount of noise suppression which the FM system will provide is influenced by the ratio chosen, since the improvement in signal-to-noise ratio which the FM system shows over amplitude modulation is equivalent to a constant *multiplied by the deviation ratio*. This assumes that the signal is somewhat stronger than the noise at the receiver, however, as the advantages of wide-band FM in regard to noise suppression disappear when the signal-to-noise ratio approaches unity.

On the other hand, a low deviation ratio is more satisfactory for strictly communication work, where readability at low signal-to-noise ratios is more important than additional noise suppression when the signal is already appreciably stronger than the noise.

As mentioned previously, broadcast FM practice is to use a deviation ratio of 5. When this ratio is applied to a voice-communication system, the total "swing" becomes 30 to 40 kc. With lower deviation ratios, such as are most frequently used for voice work, the swing becomes proportionately less, until at a deviation ratio of 1 the swing is equal to twice the highest audio frequency. Actually, however, the receiver bandwidth must be slightly greater than the expected transmitter swing, since for distortionless reception the receiver must pass the complete band of energy generated by the transmitter, and this band will always cover a range somewhat wider than the transmitter swing.

Pre-Emphasis and De-Emphasis Standards in FM broadcast and TV sound work call for the pre-emphasis of all audio modulating frequencies above about 2000 cycles, with a rising slope such as would be produced by a 75-microsecond RL network. Thus the FM receiver should include a compensating de-emphasis RC network with a time constant of 75 microseconds so that the overall frequency response from microphone to loudspeaker will approach linearity. The use of pre-emphasis and de-emphasis in this manner results in a considerable improvement in the overall signal-to-noise ratio of an FM system. Appropriate values for the de-emphasis network, for different values of circuit impedance, are given in figure 24.

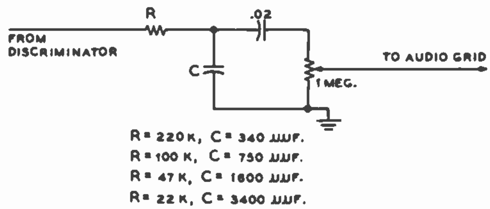


Figure 24.
75-MICROSECOND DE-EMPHASIS CIRCUITS.
The audio signal transmitted by FM and TV stations has received high-frequency pre-emphasis, so that a de-emphasis circuit should be included between the output of the FM detector and the input of the audio system.

SINGLE-SIDEBAND TELEPHONY

With continuously expanding requirements for channels in the high-frequency spectrum, the single-sideband method of telephony has become of increasing importance. The use of single-sideband transmission effectively reduces by a factor of two the channel space required for a radiotelephone signal. In very crowded portions of the spectrum, such as the amateur h-f radiotelephone bands, the use of single-sideband transmission permits two stations to communicate—with less actual interference to other channels—in the amount of channel space occupied by a single AM radiotelephone station.

9-4 Derivation of Single Sideband Signals

The single-sideband method of communication is, essentially, a procedure for obtaining more efficient use of available frequency spectrum and of available transmitter capability. As a starting point for the discussion of single-sideband signals, let us take a conventional AM signal, such as shown in figure 25, as representing the most common and the simplest method for transmitting complex intelligence such as voice or music.

It will be noted in figure 25 that there are three distinct portions to the signal: the carrier, and the upper and the lower sideband group. These three portions always are present in a conventional AM signal. Of all these portions the carrier is the least necessary and the most expensive to transmit. It is an actual fact, and it can be proved mathematically (and physically with a highly selective receiver) that the carrier of an AM signal remains unchanged in amplitude, whether it is being modulated or not. Of course the carrier *appears* to be modulated when we observe the modulated signal on a receiving system or indicator which passes a sufficiently wide band that the carrier and the modulation sidebands are viewed at the same time. This apparent change in the amplitude of the carrier with modulation is simply the result of the sidebands beating with

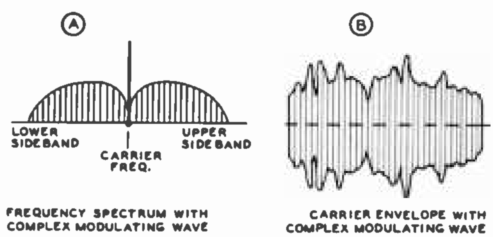


Figure 25.
REPRESENTATION OF A CONVENTIONAL AM SIGNAL.

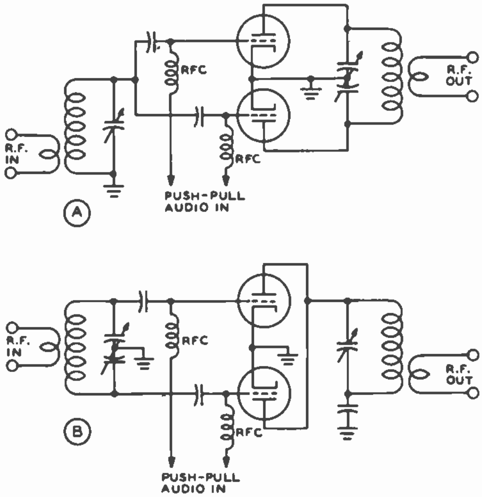


Figure 26.
SHOWING TWO COMMON TYPES OF BALANCED MODULATORS.

Notice that a balanced modulator changes the circuit condition from single ended to push-pull, or vice versa. Choice of circuit depends upon external circuit conditions since both the (A) and (B) arrangements can give satisfactory generation of a double-sideband suppressed-carrier signal.

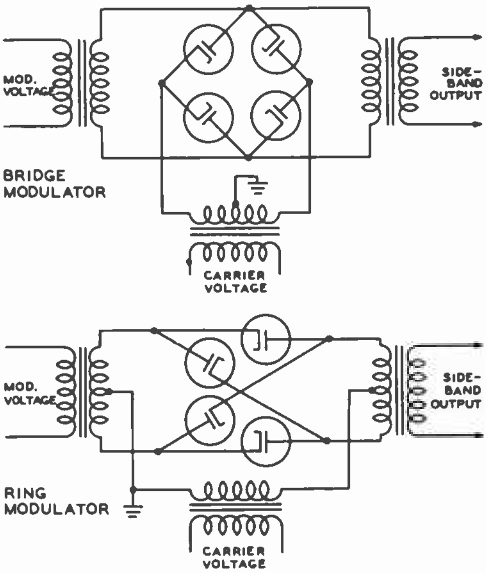


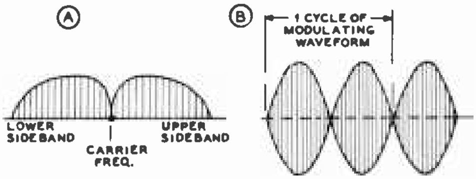
Figure 27.
TWO TYPES OF DIODE BALANCED MODULATOR.

Such balanced modulator circuits are commonly used in carrier telephone work and in single-sideband systems where the carrier frequency and modulating frequency are relatively close together. Vacuum diodes, copper-oxide rectifiers, or crystal diodes may be used in the circuits.

the carrier. However, if we receive the signal on a highly selective receiver (such as a communications receiver equipped with a BC-453 accessory i-f channel), and we modulate the carrier with a sine wave of 3000 to 5000 cycles, we will readily see that the carrier, or either of the sidebands can be tuned in separately; the carrier amplitude, as observed on a signal strength meter, will remain constant, while the amplitude of the sidebands will vary in direct proportion to the modulation percentage.

Elimination of the Carrier From the discussion in the previous paragraphs it is obvious that the carrier is superfluous so far as transmitting intelligence is concerned. It obviously is a *convenience,* however, since it provides a signal at the receiving end for the sidebands to beat with and thus to reproduce the original modulating signal. Elimination of the carrier (or carrier suppression as it is usually called) is quite readily accomplished at a low level in the transmitter through the use of a balanced modulator of one type or another. The balanced modulator simply balances out the carrier, so that in the absence of modulating signal there is no signal leaving the balanced modulator. But as the modulating signal is fed to the balanced modulator it produces both sidebands with the suppressed carrier as the reference frequency, but the carrier is not transmitted. Two types of triode balanced modulators are illustrated in figure 26, and two diode modulators are shown in figure 27.

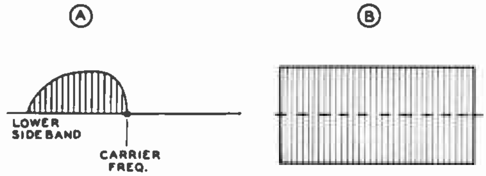
Re-Insertion of the Carrier A representation of a double-sideband suppressed-carrier signal, such as would appear at the output of a balanced modulator, is shown in figure 28. Such signals can be, and have been used for communication on the amateur bands and other frequency ranges. When such a signal is being received on a conventional communications receiver, with



FREQUENCY SPECTRUM WITH COMPLEX MODULATING WAVE

DOUBLE SIDE-BAND OUTPUT FROM BALANCED MODULATOR WITH SINE-WAVE MODULATION

Figure 28.
DOUBLE-SIDEBAND SUPPRESSED-CARRIER SIGNAL.
The envelope shown at B also is obtained on the oscilloscope when two audio frequencies of the same amplitude are fed to the input of a single-sideband transmitter.



FREQUENCY SPECTRUM WITH COMPLEX MODULATING WAVE

SIGNAL ENVELOPE WITH SINGLE-FREQUENCY MODULATING WAVE.

Figure 29.
SINGLE-SIDEBAND SUPPRESSED-CARRIER SIGNAL.
Note that the signal envelope, as viewed on an oscilloscope, appears as an unmodulated carrier with a single-frequency modulating tone; the frequency of the signal will be either the sum or the difference between the suppressed carrier frequency and the modulating frequency. In the absence of modulation the signal output is negligible.

a diode detector for example, the signal occupies the same amount of spectrum as a conventional AM signal, except that the carrier interference is not present — however, the rectified signal will have 100 per cent second harmonic distortion. This is to say that each of the original modulating frequencies will be doubled by the rectifying action of the diode second detector in the communications receiver.

If the carrier is re-inserted at the communications receiver, through use of the beat oscillator on the receiver or some other signal source such as a frequency-meter signal generator, the harmonic distortion will be reduced proportionately until the re-inserted carrier has twice the peak amplitude of the received signal; when the re-inserted carrier is more than twice the amplitude of the double-sideband suppressed-carrier received the distortion will be eliminated and the received signal will be quite normal in all respects. *However,* the re-inserted carrier must have exactly the correct frequency and *phase* to obtain distortion-free demodulation. This is not easily accomplished by any manual means. But if a small amount of carrier is transmitted with the double-sideband signal, say 2 to 5 per cent, the locally generated carrier at the receiver may be *locked* in frequency and phase to the received pilot carrier by electrical means so that normal distortion-free reception is possible. The General Electric YRS-1 adapter is such an automatic-locking device, and circuits for accomplishing automatic locking have been described in the literature.*

Suppression of One Sideband The superfluity of the carrier, except as a convenience, has been discussed. But it is equally true that the transmission of both sidebands also is superfluous since identically the same intelligence is contained in both sidebands. And further, the transmission of both sidebands is an *in*-convenience since the re-inserted carrier must have identically the correct frequency and phase if both sidebands are transmitted; but if only one sideband is transmitted the re-inserted carrier need have a frequency only approximately correct (within a few cycles) for satisfactory demodulation of the single-sideband signal. This means that a single-sideband signal may be received on any good communications receiver whose high-frequency oscillator and beat-frequency oscillator have good stability.

Single-Sideband Power Gain and Signal Effectiveness A sizeable increase in the effectiveness of a transmitter with a specified peak power capability is attainable through the use of single-sideband transmission. Actually, when transmitters are compared on the basis of peak power capability alone, the effective signal improvement at a receiver is about 9 db for

* GE Ham News, Nov.-Dec., 1948; QST, July, 1948, p. 12.

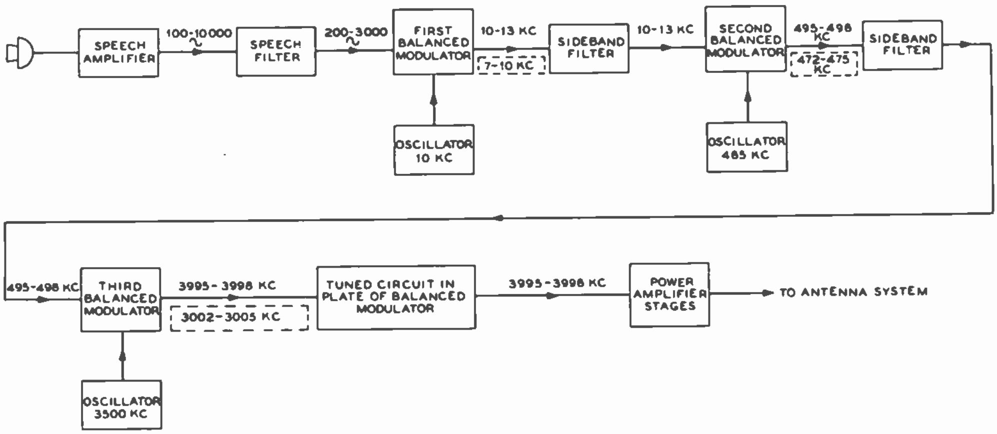


Figure 30.

BLOCK DIAGRAM OF THE "FILTER" METHOD.

The filter method of obtaining a single-sideband signal is quite satisfactory provided an adequate first sideband filter (which in this case passes 10 to 13 kc. and rejects the sideband below 10 kc.) is purchased or constructed.

a single-sideband signal as compared to a conventional AM signal.

A further advantage of single-sideband transmission is that the undesirable effects of selective fading are greatly reduced as compared to AM. Completely satisfactory ssb communication can be had when propagation conditions are such that the intelligibility of conventional AM signals over the same signal path is almost unusable. Also, the average power input to a ssb transmitter is a very small fraction of the power input to a conventional AM transmitter of the same power rating. This is true since a signal is being emitted from the ssb transmitter only during the instants of modulation, and the amount of power being taken from the line, as well as the signal being transmitted, is in direct proportion to the level of signal being transmitted. Further, since no signal is transmitted between speech passages, it is possible to operate duplex transmission on the same frequency (This is true in general only of the filter method, unless extremely good shielding of the local carrier generator of the transmitter is being employed.) with several stations being in contact at the same time.

9-5 Generation of Single-Sideband Signals

In general, there are two methods by which a single-sideband signal may be generated. These systems are: (1) The Filter Method, and (2) The Phasing Method. The systems may be used singly or in combination, and either method, in theory, may be used at the operating frequency of the transmitter or at some other frequency with the signal at the operating frequency being obtained through the use of frequency changers (mixers).

The Filter Method The filter method for obtaining a ssb signal is the classic method which has been in use by the telephone companies for many years both for land-line and radio communications. The mode of operation of the filter method is diagrammed in figure 30, in terms of components and filters which normally would be available to the amateur or experimenter. The output of the speech amplifier passes through a conventional speech filter to limit the frequency range of the speech to about 200 to 3000 cycles. This signal then is fed to a balanced modulator along with a 10,000-cycle first car-

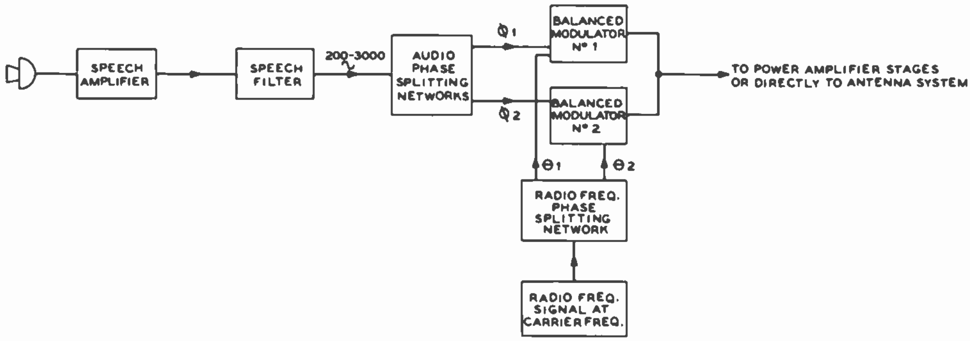


Figure 31.

BLOCK DIAGRAM OF THE "PHASING" METHOD.

The phasing method of obtaining a single-sideband signal is simpler than the filter system in regard to the number of tubes and circuits required. The system is also less expensive in regard to the components required, but is more critical in regard to adjustments for the transmission of a pure single-sideband signal.

rier from a self-excited oscillator. A low-frequency balanced modulator of this type most conveniently may be made up of four diodes of the vacuum or crystal type cross connected in a balanced bridge or ring modulator circuit. Such a modulator passes only the sideband components resulting from the sum and difference between the two signals being fed to the balanced modulator. The audio signal and the 10-kc. carrier signal from the oscillator both cancel out in the balanced modulator so that a band of frequencies between 7 and 10 kc. and another band of frequencies between 10 and 13 kc. appear in the output.

The signals from the first balanced modulator are then fed through the most critical component in the whole system—the first sideband filter. It is the function of this first sideband filter to separate the desired 10 to 13 kc. sideband from the unneeded and undesired 7 to 10 kc. sideband. Hence this filter must have low attenuation in the region between 10 and 13 kc., a very rapid slope in the vicinity of 10 kc., and a very high attenuation to the sideband components falling between 7 and 10 kc. This filter may be purchased (the National Company F-22 filter has excellent characteristics) or it may be constructed and tested at home if the test equipment is available (See "An Inexpensive Sideband Filter," by David O. Mann, *QST*, March 1949, page 21, for a description of an adequate filter which may be constructed from inexpensive and easily available components.).

The balance of the components in the block diagram of the ssb transmitter shown in figure 30 are readily available. Transformers of the type used in the i-f amplifier of a receiver are used in the sideband filter following the second balanced modulator, while the tuned circuit in the plate of the third balanced modulator serves to separate the desired from the undesired sideband.

The Phasing System There are a number of points of view from which the operation of the phasing system of ssb generation may be described. We may state that we generate two double-sideband suppressed carrier signals, each in its own balanced modulator, that both the r-f phase and the audio phase of the two signals differ by 90 degrees, and that the outputs of the two balanced modulators are added with the result that one sideband is increased in amplitude and the other one is cancelled. This, of course, is a true description of the action that takes place. But it is much easier to consider the phasing system as a method simply of adding (or of subtracting) the desired modulation frequency and the nominal carrier frequency. The carrier frequency of course is not transmitted, as is the case with all ssb transmissions, but only the sum or the difference of the modulation band from the nominal carrier is transmitted.

The phasing system has the obvious advantage that all the electrical circuits which

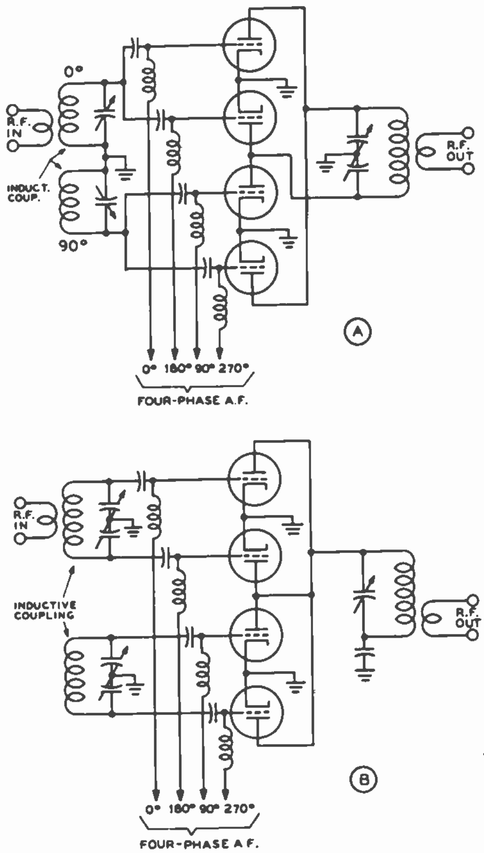


Figure 32.
TWO CIRCUITS FOR SINGLE-SIDEBAND GENERATION BY THE PHASING METHOD.
The circuit at (A) offers the advantages of simplicity in the single-ended input circuits plus a push-pull output circuit. Circuit (B) requires double-ended input circuits but allows all the plates to be connected in parallel for the output circuit.

give rise to the single sideband can operate in a practical transmitter at the nominal output frequency of the transmitter. That is to say that if we desire to produce a single sideband whose nominal carrier frequency is 3.9 Mc., the balanced modulators are fed with a 3.9-Mc. signal and with the audio signal from the phase splitters. It is not necessary to go through several frequency conversions in order to obtain a sideband at the desired output frequency, as is the case with the filter method of sideband generation.

Assuming that we feed a speech signal to the balanced modulators along with the 3900-kc. carrier (3.9 Mc.) we will obtain in the output of the balanced modulators a signal which is either the sum of the carrier signal and the speech band, or the difference between the carrier and the speech band. Thus if our speech signal covers the band from 200 to 3000 cycles, we will obtain in the output a band of frequencies from 3900.2 to 3903 kc. (the sum of the two, or the "upper" sideband), or a band from 3897 to 3899.8 kc. (the difference between the two or the "lower" sideband). A further advantage of the phasing system of sideband generation is the fact that it is a very simple matter to select either the upper sideband or the lower sideband for transmission. A simple double-pole double-throw reversing switch in two of the four audio leads to the balanced modulators is all that is required.

High-Level Phasing Vs. Low-Level Phasing The plate-circuit efficiency of the four tubes usually used to make up the two balanced modulators of the phasing system may run as high as 50 to 70 per cent, depending upon the operating angle of plate current flow. Hence it is practicable to operate the double balanced modulator directly into the antenna system as the output stage of the transmitter.

The alternative arrangement is to generate the ssb signal at a lower level and then to amplify this signal to the level desired by means of class A or class B r-f power amplifiers. If the ssb signal is generated at a level of a few milliwatts it is most common to make the first stage in the amplifier chain a class A amplifier, then to use one or more class B linear amplifiers to bring the output up to the desired level.

Balanced Modulator Circuits Illustrated in figure 26 are the two basic balanced modulator circuits which give good results with a radio frequency carrier and an audio modulating signal. Note that one push-pull and one single ended tank circuit is required, but that the push-pull circuit may be placed either in the plate or the grid circuit. Also, the audio modulating voltage always is fed into the stage in push-pull, and the tubes normally are operated Class A.

When combining two balanced modulators to make up a double balanced modulator as used in the generation of an ssb signal by the phasing system, only one plate circuit is required for the two balanced modulators. However, separate grid circuits are required since the grid circuits of the two balanced modulators operate at an r-f phase difference of 90 degrees. Shown in figure 32 are the two types of double balanced modulator circuits used for generation of an ssb signal. Note that the circuit of figure 32A is derived from the balanced modulator of figure 26A, and similarly figure 32B is derived from figure 26B.

Radio-Frequency Phasing A single-sideband generator of the phasing type requires that the two balanced modulators be fed with r-f signals having a 90-degree phase difference. This r-f phase difference may be obtained through the use of two loosely coupled resonant circuits, such as illustrated in figures 32A and 32B. The r-f signal is coupled directly or inductively to one of the tuned circuits, and the coupling between the two circuits is varied until, at resonance of both circuits, the r-f voltages developed across each circuit have the same amplitude and a 90-degree phase difference

The 90-degree r-f phase difference also may be obtained through the use of a low-Q phase shifting network, such as illustrated in figure 33; or it may be obtained through the use of a lumped-constant quarter-wave line. The low-Q phase-shifting system has proved quite practicable for use in single-sideband systems, particularly on the lower frequencies. In such an arrangement the two resistances R have the same value, usually in the range between 100 and a few thousand ohms. Capacitor C, in shunt with the input capacitances of the tubes and circuit capacitances, has a reactance at the operating frequency equal to the value of the resistor R. Also, inductor L has a *net* inductive reactance equal in value at the operating frequency to resistance R.

The inductance chosen for use at L must take into account the cancelling effect of the input capacitance of the tubes and the circuit capacitance; hence the inductance should be variable and should have a *lower* value of inductance than that value of inductance which would have the same reactance as resistor R.

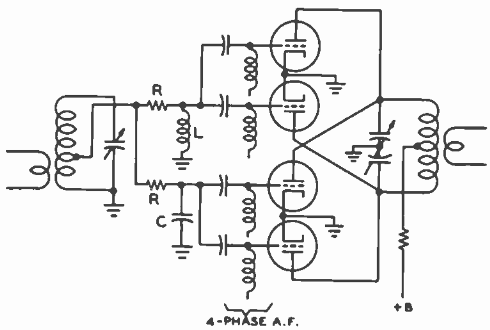


Figure 33.
LOW-Q R-F PHASE-SHIFT NETWORK.
The r-f phase-shift system illustrated above is convenient in a case where it is desired to make small changes in the operating frequency of the system without the necessity of being precise in the adjustment of two coupled circuits as used for r-f phase shift in the circuit of figure 32.

Inductor L may be considered as being made up of two values of inductance in parallel; (a) a value of inductance which will resonate at the operating frequency with the circuit and tube capacitances, and (b) the value of inductance which is equal in reactance to the resistance R. In a network such as shown in figure 33, equal and opposite 45-degree phase shifts are provided by the RL and RC circuits, thus providing a 90-degree phase difference between the excitation voltages applied to the two balanced modulators.

Audio-Frequency Phasing The audio-frequency phase-shifting networks used in generating a single-sideband signal by the phasing method usually are based on those described by Dome in an article in the December, 1946, *Electronics*. A relatively simple network for accomplishing the 90-degree phase shift over the range from 160 to 3500 cycles is illustrated in figure 34. The values of resistance and capacitance must be carefully checked to insure minimum deviation from a 90-degree phase shift over the 200 to 3000 cycle range.

Two alternative types of phase-shifting networks, which make use of electron tubes within the actual network, have been described. The more complex network, which is capable of 90 degrees plus or minus one degree phase shift over the range from 70 to 7000 cycles and uses four dual triodes, has been described

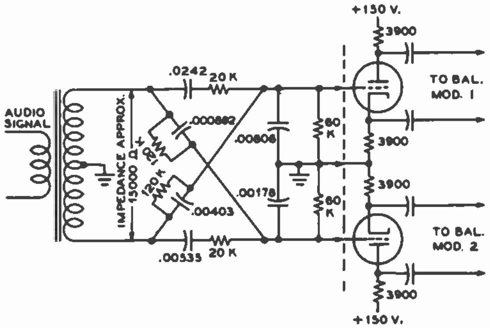


Figure 34.
DOME AUDIO-PHASE-SHIFT NETWORK.
This circuit arrangement is convenient for obtaining the audio phase shift when it is desired to use a minimum of circuit components and tube elements.

in detail in the Nov.-Dec., 1948 issue of *GE Ham News*. The simpler network, which covers the range between 300 and 3000 cycles with the same phase tolerance and uses three dual triodes, was described in the Jan., 1950, issue of *QST*, on page 42.

A convenient low-power ssb exciter unit using a very simple phase-shift network is described in the Nov.-Dec. 1950, *GE Ham News*. Other representative ssb exciter units are described in Chapter Twenty-One of this handbook and in recent issues of the amateur periodicals.

Comparison of Filter and Phasing Methods of SSB Generation Either the filter or the phasing method of single-sideband generation is theoretically capable of a high degree of performance. For operation well within a frequency band the phasing method is less complex and less critical in adjustment. For operation near the edge of a band where *none* of the other sideband beyond a few hundred cycles may be transmitted, the filter method is to be preferred. The filter method has the advantage for this type of application that selective circuits at a relatively low frequency may be employed to reject within the exciter portion any signals of appreciable amplitude which might fall more than a few hundred cycles on the undesired side of the suppressed carrier.

In general it may be said that a *high* degree of unwanted signal rejection may be attained with less expense and circuit complexity with the filter method. But the filter method is less flexible with respect to frequency change and band change. Where a relatively simple ssb transmitter for operation well within a band is desired, the phasing method is to be preferred due to its lesser circuit complexity.

Where the very highest degree of performance in the ssb exciter is desired, the filter method and the phasing method may be combined. Through the use of the phasing method in the first balanced modulator those undesired sideband components lying within 1000 cycles of the carrier may be given a much higher degree of rejection than is attainable with the filter method alone, with any reasonable amount of complexity in the sideband filter. Then the sideband filter may be used in its normal way to attain very high attenuation of all undesired sideband components lying perhaps further than 500 cycles away from the carrier, and to restrict the sideband width on the *desired* side of the carrier to the specified frequency limit.

Requirement for Linearity In any type of single-sideband system there is a requirement for a high degree of linearity in the amplification and frequency conversion stages. This means that the low-level stages should be operated Class A, with the operating bias in the center of the dynamic characteristic. If this is not done, undesired cross-modulation products will be generated, and these undesired signals will be amplified along with the desired-signal output of the exciter unit. High level stages may be operated in the Class B region, but careful attention to establishing operating conditions which give good linearity is necessary.

9-6 Reception of Single-Sideband Signals

Single-sideband signals may be received, after a certain degree of practice in the technique, in a quite adequate and satisfactory manner with a good communications receiver. However, the receiver must have quite good frequency stability both in the high-frequency oscillator and in the beat oscillator. For this reason, receivers which use a crystal-controlled first oscillator are likely to offer a greater degree of satisfaction than the more common type which uses a self-controlled oscillator.

Beat oscillator stability in most receivers is usually quite adequate, but many receivers do not have a sufficient amplitude of beat oscillator injection to allow reception of strong ssb signals without distortion. In such receivers it is necessary either to increase the amount of beat-oscillator injection into the diode detector, or the manual gain control of the receiver must be turned down quite low.

The tuning procedure for ssb signals is as follows: The ssb signals may first be located by tuning over the band with the receiver set for the reception of c.w.; that is, with the manual gain at a moderate level and with the beat oscillator operating. By tuning in this manner ssb signals may be *located* when they are far below the amplitude of conventional AM signals on the frequency band. Then after a signal has been located, the beat oscillator should be turned off and the receiver put on a.v.c. Following this the receiver should be tuned for maximum swing of the S meter with modulation of the ssb signal. It will not be possible to understand the ssb signal at this time, but the receiver may be tuned for maximum deflection. Then the receiver is put back on manual gain control, the beat oscillator is turned on again, the manual gain is turned down until the background noise level is quite low, and the *beat oscillator* control is varied until the signal sounds natural.

The procedure in the preceding paragraph may sound involved, but actually all the steps except the last one can be done in a moment. However, the last step is the one which will require some practice. In the first place, it is not known in advance whether the upper or lower sideband is being transmitted. So it will be best to start tuning the beat oscillator from one side of the pass band of the receiver to the other, rather than starting with the beat oscillator near the center of the pass band as is normal for c-w reception.

With the beat oscillator on the wrong side of the sideband, the speech will sound inverted; that is to say that low-frequency modulation tones will have a high pitch and high-frequency modulation tones will have a low pitch—and the speech will be quite unintelligible. With the beat oscillator on the correct side of the sideband but too far from the correct position, the speech will have some intelligibility but the voice will sound quite high pitched. Then as the correct setting for the beat oscillator is approached the voice will begin to sound natural but will have a background growl on each syllable. At the correct frequency for the beat oscillator the speech will clear completely and the voice will have a clean, crisp quality. It should also be mentioned that there is a narrow region of tuning of the beat oscillator a small distance on the wrong side of the sideband where the voice will sound quite bassy and difficult to understand.

With a little experience it will be possible to identify the sound associated with improper settings of the beat-oscillator control so that corrections in the setting of the control can be made. Note that the main tuning control of the receiver is not changed after the sideband once is tuned into the pass band of the receiver. All the fine tuning should be done with the beat oscillator control. Also, is it very important that the r-f gain control be turned to quite a low level during the tuning process. Then after the signal has been tuned properly the r-f gain may be increased for good signal level, or until the point is reached where best oscillator injection becomes insufficient and the signal begins to distort.

Single-Sideband Receivers and Adapters Greatly simplified tuning, coupled with strong attenuation of undesired signals, can be obtained through the use of a single-sideband receiver or receiver adapter. The exalted carrier principle usually is employed in such receivers, with a phase-sensitive system sometimes included for locking the local oscillator to the frequency of the carrier of the incoming signal. In order for the locking system to operate, some carrier must be transmitted along with the ssb signal. The GE YRS-1 adapter is an example of such a unit, and references for the home construction of such devices have been given on a previous page. Such receivers and adapters include a means for selecting the upper or lower sideband by the simple operation of a switch. For the reception of a single-sideband signal the switch obviously must be placed in the correct position. But for the reception of a conventional AM or phase-modulated signal, either sideband may be selected, allowing the sideband with the least interference to be used.

CHAPTER TEN

Transmitter Design, Keying, and Control

Communications receivers are usually designed as an integral unit, but there is an almost unlimited number of combinations of tubes, exciter circuits, amplifier circuits, power supply arrangements, and control provisions which one may incorporate in a "250-watt" transmitter. For this reason the bulk of the transmitter section of this book has been devoted to units and more or less minor assemblies. However, Chapter Twenty-Four has been devoted to illustrating the manner in which the various units which make up a transmitter may be grouped together into a major assembly — a complete transmitter.

If a tube requires 25 watts r-f driving power for a certain application, it is obvious that it makes little difference just what exciter circuit is used so long as it puts out 25 watts on the desired bands. Because of its characteristics one exciter may be preferred by one amateur, another exciter by another amateur.

It is fortunate that there is this flexibility with regard to transmitter design, because it makes it easy for an amateur to start out with a low-power transmitter and then add to it from time to time. It also permits one a certain degree of "custom tailoring" of his transmitter to suit his particular requirements.

10-1 Exciters and Amplifiers

A 5-watt crystal oscillator may be accurately referred to as a transmitter *when it is used to feed an antenna.* On the other hand a multi-tube r-f unit ending with a 150-watt power amplifier may be properly termed an exciter *when it is used to drive a higher power amplifier.* Thus we see that any r-f unit, even a simple oscillator, may be either an exciter or a transmitter depending upon how it is used.

The requirements for a low-power (15 to 75 watt) transmitter are practically the same as for an exciter of the same output: the over-all efficiency should be good, the unit should cover all the desired bands with a minimum of coil changing and retuning, shielding and provisions for harmonic reduction should be included, and both initial cost and upkeep should be low in proportion to the power output.

For this reason, several low power r-f units and several medium- and high-power amplifiers are described in other chapters, and the reader is permitted to use his own ingenuity in working out the combination which appears to fit his requirements. If one is designing a complete transmitter, to which no additions are to be made, it is probably best to decide first upon the final amplifier and then work backwards from there, the driving requirements of the particular tubes used determining the exciter.

Choosing Tubes Low-power exciters invariably use receiving tubes or "modified" receiving tubes for the sake of economy. Large scale production brings the cost of 6AG7's, 6L6's, etc., down to a price that would be impossible were they designed for and purchased only by amateurs. Some tubes, like the T21 and 807, resemble standard receiving tubes in one or more respects, and while costing more than a standard receiving tube equivalent (6L6G in this case), are still obtainable at a price below that which would be necessary were they not outgrowths of receiving tubes.

The tubes in the high-power amplifier and in the Class B modulator (if used) should be chosen with care. While in general there is little to choose between tubes by reliable manufacturers, some are better adapted than others for certain applications. Also, the more recently released tubes of a particular manufacturer are usually better and less expensive than older tubes of the same general type.

Beam-tetrode tubes such as the 807, 814, 813, 4-65A, 4-125A and 4-250A have much to commend them both for r-f amplifier use and for use as modulators. Tubes of this type require far less excitation both for r-f and audio use than triodes of equivalent plate dissipation and purchase price. Also, since less excitation power is required by a beam-tetrode amplifier stage, the problem of TVI elimination from the exciter unit is much simplified as compared to an exciter design for a high-power triode stage. However, it must always be kept in mind, when planning a transmitter, that much more attention must be given to shielding and elimination of parasitic circuits when using beam tetrodes than when using triodes. This condition is a natural consequence of the much greater power sensitivity of the beam-tetrode type tube.

Tubes for modulator service should have good emission and adequate plate dissipation. Interelectrode capacitances are relatively unimportant. For triode Class B modulator service the usual practice is to use high-μ ("zero bias") tubes so that little or no bias will be required. In the interest of economy it is often wise to work out a tube combination such that the modulator may be operated from the same plate supply as the final amplifier. It is *much* less expensive to increase the power capability of one power supply by 50 to 75 per cent than to build a separate power supply for the Class B modulator. Also, there is nothing whatever objectionable about running both the modulator and modulated amplifier from the same power supply provided that the supply has been designed to have adequate current capability for both stages with reasonably good regulation.

For service as frequency multipliers, beam tetrodes and pentodes will give the best results. These types of tubes will operate at higher plate-circuit efficiency with less excitation and bias than will a triode of equivalent plate dissipation.

For triode Class C or Class B r-f amplifier service, the amplification factor is not too important, though tubes with a medium-high μ (20 to 40) are most popular.

In Class A or Class AB audio/driver service, triodes with low amplification factor are to be preferred, though pentodes or beam tetrodes may be used provided they are included within an inverse-feedback loop. Shunt feedback from the plate of the beam tube to the plate of the preceding audio stage is quite effective and has been shown in several of the audio systems illustrated in this Handbook.

Driving Power It is always advisable to have a slight reserve of driving power in order to be on the safe side. Therefore, the potential output of an exciter on the band upon which its output is least (usually the highest frequency band) should be slightly greater than the excitation requirements of the following stage as determined from the manufacturer's tube data.

It has become more or less standard practice both in amateur and in commercial work to accomplish the frequency control and frequency multiplication functions in a separate or at least self-contained exciter unit with a power output of 10 to 50 watts. The usual exciter unit ends in an 807 or similar tube for the 25 to 50 watt range, or a 2E26 for the 10 to 25 watt range. V-h-f exciters usually end in a 2E26, an 832A, or an 829B tube, depending upon the power requirements.

Plate-modulated Class C amplifiers require the most excitation, the stage normally requiring full grid current and approximately 2½ times cutoff bias. For these reasons, high excitation and large bias, the Class C amplifier which has been adjusted for plate modulation normally will be a prolific generator of harmonic signals. Hence the Class C amplifier which is to be plate modulated will usually require the most stringent shielding and filtering procedures before TVI may be eliminated.

C-w, FM, and Class B linear amplifiers such as used for conventional AM or for single-sideband amplification, may be operated approximately at cutoff bias with a relatively small amount of excitation. Plate circuit efficiency will be reduced something like ten per cent by such operating conditions, but the efficiency of the amplifier as a harmonic generator will be reduced enormously. And since the excitation requirements of the stage will be reduced materially, the TVI-proofing problem with the exciter also will be eased.

Also to be taken into consideration, when tentatively planning a transmitter, are such things as the limiting factor in tube characteristics. For instance, in a grid-modulated or other type of efficiency-modulated transmitter, the output is always limited by the plate dissipation, while for plate-modulated phone work either the plate voltage or plate current rating is exceeded first. Thus we see that for grid modulation, a tube with high plate dissipation is of prime importance, while for plate-modulated operation the matters of filament emission and insulation are of greatest importance.

Care should be taken to make sure that the tubes chosen are capable of efficient and safe operation on the highest frequency used.

10-2 Design Considerations

Transmitter Wiring At the higher frequencies, solid enamelled copper wire is most efficient for r-f leads. Tinned or stranded wire will show greater losses at these frequencies. Tank coil and tank capacitor leads should be of heavier wire than other r-f leads, though there is little point in using wire heavier than is used for the tank coil itself.

All grounds and by-passes in an r-f stage should be made to a common point, and the grounding points for several stages bonded together with heavy wire.

The best type of flexible lead from the envelope of a tube to a terminal is thin copper strip, cut from thin sheet copper. Heavy, rigid leads to these terminals may crack the envelope glass when a tube heats or cools.

Wires carrying only a.f. or d.c. should be chosen with the voltage and current in mind. Some of the low-filament-voltage transmitting tubes draw heavy current, and heavy wire must be used to avoid voltage drop. The voltage is low, and hence not much insulation is required. Filament and heater leads are usually twisted together. An initial check should be made on the filament voltage of all tubes of 25 watts or more plate dissipation rating. This voltage should be measured right at the tube sockets. If it is low, the filament transformer voltage should be raised. If this is impossible, heavier or paralleled wires should be used for filament leads, cutting down their length if possible.

Spark-plug-type high-tension ignition cable makes the best wire for high-voltage leads. This cable will safely withstand 10,000 volts peak. If this cable is used, the high-voltage leads may be cabled with filament and other low-voltage leads. For high-voltage leads in low-power exciters, where the plate voltage is not over 450 volts, ordinary radio hookup wire of good quality will serve the purpose.

No r-f leads should be cabled; in fact it is better to use enamelled or bare copper wire for r-f leads and rely upon spacing for insulation. All r-f joints should be soldered, and the joint should be a good mechanical junction before solder is applied.

Coil Placement For best Q a coil should be in the form of a solenoid with length from one to two times the diameter. For minimum interstage coupling, coils should be made as small physically as is practicable. The coils should then be placed so that adjoining coils are oriented for minimum mutual coupling. To determine if this condition exists, apply the following test: the axis of one of the two coils must lie in the plane formed by the center turn of the other coil. If this condition is not met, there will be appreciable coupling unless the unshielded coils are very small in diameter or are spaced a considerable distance from each other.

Variable Capacitors The question of optimum C/L ratio and capacitor plate spacing is covered in Chapter Seven. For all-band operation of a high-power stage, it is recommended that a capacitor just large enough for 40-meter c-w operation be chosen. (This will have sufficient capacitance for 'phone operation on all higher frequency bands.) Then use fixed padding capacitors for operation on 80 meters. Such padding capacitors are available in air, ceramic, and vacuum types.

Specially designed variable capacitors are recommended for u-h-f work; ordinary capacitors often have "loops" in the metal frame which may resonate near the operating frequency.

Insulation On frequencies above 7 Mc., ceramic, polystyrene, or Mycalex insulation is to be recommended. Cold flow must be considered when using polystyrene (Amphenol 912, etc.). Bakelite has low losses on the lower frequencies but should never be used in the field of high-frequency tank circuits.

Lucite (or Plexiglas), which is available in rods, sheets, or tubing, is excellent for use at all radio frequencies where the r-f voltages are not especially high. It is very easy to work with ordinary tools and is not expensive. The loss factor depends to a considerable extent upon the amount and kind of plasticizer used.

The most important thing to keep in mind regarding insulation is that the best insulation is air. If it is necessary to reinforce air-wound coils to keep turns from vibrating or touching, use strips of Lucite or polystyrene cemented in place with Amphenol 912 coil dope. This will result in lower losses than the commonly used celluloid ribs and Duco cement.

Metering The ideal transmitter would have an individual meter in every circuit requiring measurement. However, for the sake of economy, it is common practice to measure filament and plate voltages by means of a test set or universal meter during the initial tryout of the transmitter, and then to assume that these voltages will be maintained. Further economies can be effected by doubling up on meters when measuring current in various circuits in which the current is variable, and as an index of transmitter tuning.

By a system of plugs and jacks, or a selector switch, one or two milliammeters can be used to make all the measurements necessary to tune up a transmitter properly. However, it often is of considerable advantage to be able to observe the current of several circuits or stages simultaneously. Obviously one would not be justified in buying $100 worth of meters for a transmitter containing other parts totaling $75. On the other hand, the purchase of a filament voltmeter to keep careful tab on the filament voltage of a pair of 250-watt tubes is a good investment.

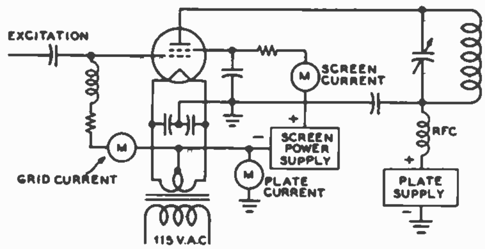


Figure 1.
CATHODE METERING OF PLATE CURRENT.

Through use of the circuit illustrated above it is possible to meter the plate current of an r-f amplifier stage in the cathode (or center tap) return. In order that the cathode milliammeter read only the plate current, the grid-bias supply (or grid leak) and the screen power supply must return to the filament center tap (or cathode) rather than to ground. If the screen voltage is taken from the plate supply through a dropping resistor, screen current plus plate current will be indicated on the cathode milliammeter.

Probably the most popular arrangement calls for meter switching in the low power stages and individual meters in the last stage. Ordinarily, r-f meters are not used except in certain antenna coupling circuits. Where line voltage does not fluctuate appreciably, one can get by very nicely with just d-c milliammeters: plate-current meters in the low power stages, and a grid and a plate meter in the final stage.

D-c meters should be bypassed with small 0.004-μfd. or larger capacitors directly at the meter terminals. The capacitor is placed across the terminals, not from one terminal to ground. Such capacitors are a wise precaution in all cases, because even though meter and meter leads are kept away from r-f components, the meter may be subjected to considerable r-f current because of an r-f choke not doing an adequate job of blocking r.f. from the meter, or as a result of induced currents.

Most meters now come with bakelite cases. If the "zero adjuster" screw is well insulated, such meters can be placed in positive high voltage leads where the voltage does not exceed 1000 volts. When the voltage is higher than 1000 volts, the meter should preferably be placed behind a protective glass or screen. The meter should not be mounted directly on a grounded metal panel when the plate voltage exceeds 2000 volts, as the metal portions of the meter may arc through the bakelite case to the grounded metal panel, particularly when plate modulation is used.

One highly recommended method of arranging meters in a high-powered rack and panel transmitter is to group all meters on a recessed bakelite meter panel with a glass front or 1/8" mesh brass screen cover at the top of the rack, near eye level of the operator and not close to any of the tuning dials. With the bakelite meter panel, there is no danger of meters arcing to ground, and because of the protection afforded by the glass or wire-mesh cover there is little likelihood of an operator accidentally coming in contact with the meters.

An alternative method is to place all meters at low-potential points in the circuits to be measured. This means that plate milliammeters should be placed in the lead between the center tap of the filament transformer and ground. The obstacle to this method is the fact that the milliammeters in the cathode return will read total *cathode* current (sum of plate, screen, grid, and even suppressor current in a pentode) unless special provisions are included in the returns of the screen and grid power supplies so that their return circuits go to the filament center tap of the tube rather than to ground. Figure 1 shows a system using a beam tetrode whereby only the plate current of the tube passes through the cathode milliammeter. But if the grid bias supply or the screen voltage supply returns to *ground* instead of to the filament center tap, the currents to these electrodes also will pass through the cathode milliammeter. Of course, if the currents to these electrodes are metered in addition to the cathode current, they may be subtracted from the total cathode current to obtain the true plate current.

Meter Switching The metering method which uses one meter and a circuit selector switch can be used to advantage where the voltages on the leads which carry the current to be measured are not greater than about 500 volts to ground. Fifty-ohm resistors are inserted in the leads, and because the resistance of the meter is so low compared to the 50-ohm resistors, the meter can be considered as being inserted in series with the circuit when it is tapped across the resistor. Thus, with a double-pole selector switch of the *non-shorting* type having sufficient positions, one can use a single meter to measure the current in several circuits.

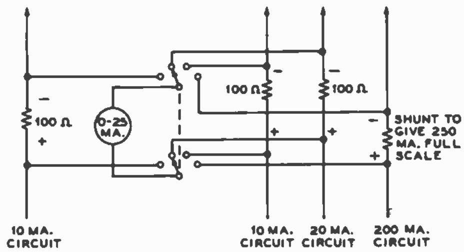


Figure 2.
MULTIPLE METERING CIRCUIT.
Showing the use of a single milliammeter for measuring currents, with different full-scale ranges, in several circuits.

The resistor should be made 25 ohms where the current to be measured runs over 200 ma., and the resistor increased to 200 ohms when the current to be measured is less than 15 ma. It is necessary to minimize the resistance where heavy current is present, in order to avoid excessive voltage drop when the meter is not shunting the resistor. It is necessary to increase the value of resistance when the current is so low that a low-range meter must be used to measure the current. Low range milliammeters begin to show appreciable resistance themselves, and their calibration will be thrown off when shunted by too low a value of resistor.

Meter switching is not practicable in high voltage circuits (over 1200 volts). For measuring plate current in high power stages, the resistor should be placed either in the B minus lead or in the filament return (center tap).

It is possible, by means of systems of shunts, to use a single low-range milliammeter for measuring widely different values of current in different circuits. For instance, a 0-25 d-c milliammeter could be used to measure grid current in several stages. Then the same instrument could be used as a 0-250 ma. full-scale instrument when switched into the cath-

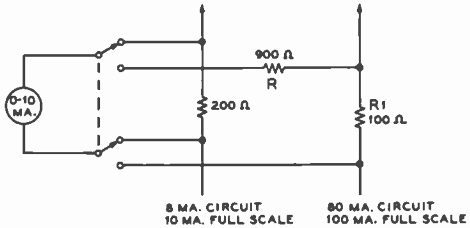


Figure 3.
ALTERNATIVE METERING CIRCUIT.
Showing the use of a high-impedance shunt system for obtaining different full-scale ranges when switching a low-range milliammeter to different circuits.

ode or plate circuit of a stage by the incorporation of a shunt across the switch position of the higher-current stage. This procedure is illustrated in figure 2. The shunt value can be determined by cut and try, using a short length of constantan wire across the switch contacts. It is a great convenience to use a wire material for the shunt which can be soldered directly. Some shunt materials will not solder readily.

An alternative meter switching arrangement, which does not require the use of low-resistance shunts, is illustrated in figure .3. The series resistor R should be much greater in resistance than the d-c resistance of the instrument movement. The current multiplication obtained through the use of this circuit is equal to: $(R/R_1)+1$. The advantage of the circuit of figure 3 is that the resistance of switch contacts and meter leads is not of significance in the re-scaling of the instrument. Also, conventional ½-watt resistors may be used in suitable combination as R and R_1.

10-3 Power Systems

It is probable that the average amateur station that has been in operation for a number of years will have at least two transmitters available for operation on different frequency bands, at least two receivers or one receiver and a converter, at least one item of monitoring or frequency measuring equipment and probably two, a v.f.o., a speech amplifier, a desk light, and a clock. In addition to the above 8 or 10 items, there must be an outlet available for a soldering iron and there should be one or two additional outlets available for plugging in one or two pieces of equipment which are being worked upon.

It thus becomes obvious that 10 or 12 outlets connected to the 115-volt a-c line should be available at the operating desk. It may be practicable to have this number of outlets installed as an outlet strip along the baseboard at the time that a new home is being planned and constructed. Or you might wish to install the outlet strip on the operating desk so as to have the flexibility of moving the operating desk from one position to another. Alternatively, the outlet strip might be wall mounted just below the desk top.

Power Drain Per Outlet When the power drain of all the items of equipment, other than transmitters, used at the operating position is totalled, you probably will find that 350 to 600 watts will be required. Since the usual home outlet is designed to handle only about 600 watts maximum, the transmitter, unless it is of relatively low power, should be powered from another source. This procedure is desirable in any event so that the voltage supplied to the receiver, frequency control, and frequency monitor will be substantially constant with the transmitter on or off the air.

So we come to two general alternative plans with their variations. Plan (A) is the more desirable and also the most expensive since it involves the installation of two separate lines from the meter box to the operating position either when the house is constructed or as an alteration. One line, with its switch, is for the transmitters and the other line and switch is for receivers and auxiliary equipment. Plan (B) is the more practicable for the average amateur, but its use requires that all cords be removed from the outlets whenever the station is not in use in order to comply with the electrical codes.

Figure 4 shows a suggested arrangement for carrying out Plan (A). In most cases an installation such as this will require approval of the plans by the city or county electrical inspector. Then the installation itself will also require inspection after it has been completed. It will be necessary to use approved outlet boxes at the rear of the transmitter where the cable is connected, and also at the operating bench where the other BX cable connects to the outlet strip. Also, the connectors at the

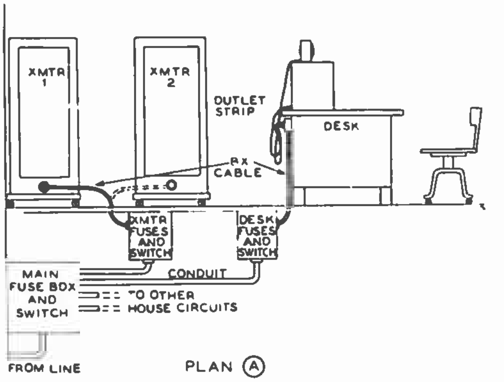


Figure 4.
THE PLAN (A) POWER SYSTEM.
A-c line power from the main fuse box in the house is run separately to the receiving equipment and to the transmitting equipment. Separate switches and fuse blocks then are available for the transmitters and for the auxiliary equipment. Since the fuses in the boxes at the operating room will be in series with those at the main fuse box, those in the operating room should have a lower rating than those at the main fuse box. Then it will always be possible to replace blown fuses without leaving the operating room. The fuse boxes can conveniently be located alongside one another on the wall of the operating room.

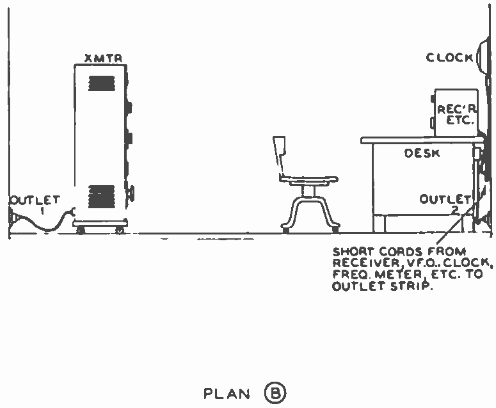


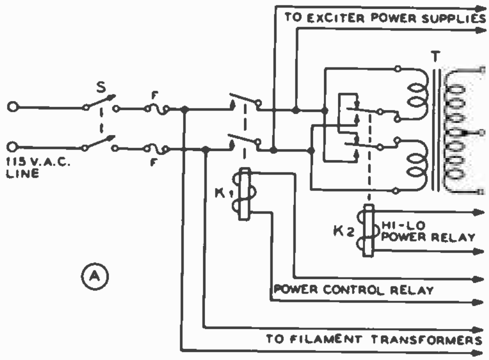
Figure 5.
THE PLAN (B) POWER SYSTEM.
This system is less convenient than the (A) system, but does not require extensive rewiring of the electrical system within the house to accommodate the arrangement. Thus it is better for a temporary or semi-permanent installation. In most cases it will be necessary to run an extra conduit from the main fuse box to the outlet from which the transmitter is powered, since the standard arrangement in most houses is to run all the outlets in one room (and sometimes all in the house) from a single pair of fuses and leads.

rear of the transmitter will have to be of an approved type. It is possible also that the BX cable will have to be permanently affixed to the transmitter with the connector at the fuse-box end. These details may be worked out in advance with the electrical inspector for your area.

The general aspects of Plan (B) are shown in figure 5. The basic difference between the two plans is that (A) represents a "permanent" installation even though a degree of mobility is allowed through the use of BX for power leads, while plan (B) is definitely a "temporary" type of installation as far as the electrical inspector is concerned. While it will be permissible in most areas to leave the transmitter cord plugged into the outlet even though it is turned off, the Fire Insurance Underwriters codes will make it necessary that the cord which runs to the group of outlets at the back of the operating desk be removed whenever the equipment is not actually in use.

Whether the general aspects of plans (A) or (B) are used it will be necessary to run a number of control wires, keying and audio leads, and an excitation cable from the operating desk to the transmitter. Control and keying wires can best be grouped into a multiple-wire rubber-covered cable between the desk and the transmitter. Such an arrangement gives a good appearance, and is particularly practical if cable connectors are used at each end. High-level audio at a moderate impedance level (600 ohms or below) may be run in the same control cable as the other leads. However, low-level audio can best be run in a small coaxial cable. Small coaxial cable such as RG-58/U or RG-59/U also is quite satisfactory and quite convenient for the signal from the v.f.o. to the r-f stages in the transmitter. Coaxial-cable connectors of the UG series are quite satisfactory for the terminations both for the v-f-o lead and for any low-level audio cables.

Checking an Outlet with a Heavy Load To make sure that an outlet will stand the full load of the entire transmitter, plug in an electric heater rated at about 50 per cent greater wattage than the power you



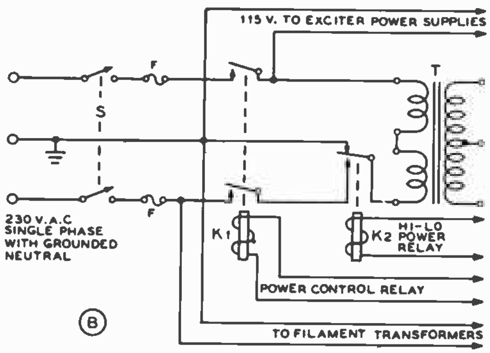


Figure 6.
FULL-VOLTAGE/HALF-VOLTAGE POWER CONTROL SYSTEMS.
The circuit at (A) is for use with a 115-volt a-c line. Transformer T is of the standard type having two 115-volt primaries; these primaries are connected in series for half-voltage output when the power control relay K_1 is energized but the hi-lo relay K_2 is not operated. When both relays are energized the full output voltage is obtained. At (B) is a circuit for use with a standard 230-volt residence line with grounded neutral. The two relays control the output of the power supplies the same as at (A).

expect to draw from the line. If the line voltage does not drop more than 5 volts (assuming a 117-volt line) under load and the wiring does not overheat, the wiring is adequate to supply the transmitter. About 600 watts total drain is the maximum that should be drawn from a 117-volt "lighting" outlet or circuit. For greater power, a separate pair of heavy conductors should be run right from the meter box. For a 1-kw. phone transmitter the total drain is so great that a 230-volt "split" system ordinarily will be required. Most of the newer homes are wired with this system, as are homes utilizing electricity for cooking and heating.

With a three-wire system, be sure there is no fuse in the neutral wire at the fuse box. A neutral fuse is not required if both "hot" legs are fused, and, should a neutral fuse blow, there is a chance that damage to the radio transmitter will result.

If you have a high power transmitter and do a lot of operating, it is a good idea to check on your local power rates if you are on a straight "lighting" rate. In some cities a lower rate can be obtained (but with a higher "minimum") if electrical equipment such as an electric heater drawing a specified amount of current is permanently wired in. It is not required that you use this equipment, merely that it be permanently wired into the electrical system. Naturally, however, there would be no saving unless you expect to occupy the same dwelling for a considerable length of time.

Outlet Strips The "outlet strips" which have been suggested for installation in the base board or for use on the rear of a desk are obtainable from the large electrical supply houses. If such a house is not in the vicinity it is probable that a local electrical contractor can order a suitable type of strip from one of the supply house catalogs. These strips are quite convenient in that they are available in varying lengths with provision for inserting a-c line plugs throughout their length. The a-c plugs from the various items of equipment on the operating desk then may be inserted in the outlet strip throughout its length. In many cases it will be desirable to reduce the equipment cord lengths so that they will plug neatly into the outlet strip without an excess to dangle behind the desk.

Contactors and Relays The use of power-control contactors and relays often will add considerably to the operating convenience of the station installation. The most practicable arrangement usually is to have a main a-c line switch on the front of the transmitter to apply power to the filament transformers and to the power control circuits. It also will be found quite convenient to have a single a-c line switch on the operating desk to energize or cut the power from the outlet strip on the rear of the operating desk. Through the use of such a switch it is not necessary to remember to switch off a large number of separate switches on each of the items of equipment on the operating desk. The alternative arrangement, and that which is approved by the Underwriters, is to remove the plugs from the wall both for the transmitter and for the operating-desk outlet strip when a period of operation has been completed.

While the insertion of plugs or operation of switches usually will be found best for applying the a-c line power to the equipment, the changing over between transmit and receive can best be accomplished through the use of relays. Such a system usually involves three relays, or three groups of relays. The relays and their functions are: (1) power control relay for the transmitter—applies 115-volt line to the primary of the high-voltage transformer and turns on the exciter; (2) control relay for the receiver—makes the receiver inoperative by any one of a number of methods when closed, also may apply power to the v.f.o. and to a keying or a phone monitor; and (3) the antenna changeover relay—connects the antenna to the transmitter when the transmitter is energized and to the receiver when the transmitter is not operating. Several circuits illustrating the application of relays to such control arrangements are discussed in the paragraphs to follow in this chapter.

Controlling Transmitter Power Output It is necessary, in order to comply with FCC regulations, that transmitter power output be limited to the minimum amount necessary to sustain communication. This requirement may be met in several ways. Many amateurs have two transmitters; one is capable of relatively high power output for use when calling, or when interference is severe, and the other is capable of considerably less power output. In many cases the lower powered transmitter acts as the exciter for the higher powered stage when full power output is required. But the majority of the amateurs using a high powered equipment have some provision for reducing the plate voltage on the high-level stages when reduced power output is desired.

One of the most common arrangements for obtaining two levels of power output involves the use of a plate transformer having a double primary for the high-voltage power supply. The majority of the high-power plate transformers of standard manufacture have just such a dual-primary arrangement. The two primaries are designed for use with either a 115-volt or 230-volt line. When such a transformer is to be operated from a 115-volt line, operation of both primaries in parallel will deliver full output from the plate supply. Then when the two primaries are connected in series and still operated from the 115-volt line the output voltage from the supply will be reduced approximately to one half. In the case of the normal class C amplifier, a reduction in plate voltage to one half will reduce the power input to the stage to one quarter.

If the transmitter is to be operated from a 230-volt line, the usual procedure is to operate the filaments from one side of the line, the low-voltage power supplies from the other side, and the primaries of the high-voltage transformer across the whole line for full power output. Then when reduced power output is required, the primary of the high-voltage plate transformer is operated from one side to center tap rather than across the whole line. This procedure places 115 volts across the 230-volt winding the same as in the case discussed in the previous paragraph. Figure 6 illustrates the two standard methods of power reduction with a plate transformer having a double primary; (A) shows the connections for use with a 115-volt line and (B) shows the arrangement for a 230-volt a-c power line to the transmitter.

The full-voltage/half-voltage methods for controlling the power input to the transmitter, as just discussed, are subject to the limitation that only two levels of power input (full power and quarter power) are obtainable. In many cases this will be found to be a limitation to flexibility. When tuning the transmitter, the antenna coupling network, or the antenna system itself it is desirable to be able to reduce the power input to the final stage to a relatively low value. And it is further convenient to be able to vary the power input continuously from this relatively low input up to the full power capabilities of the transmitter. The use of a variable-ratio auto-transformer in the circuit from the line to the primary of the plate transformer will allow a continuous variation in power input from zero to the full capability of the transmitter.

Variable-Ratio Auto-Transformers There are several types of variable-ratio auto-transformers available on the market. Of these, the most common

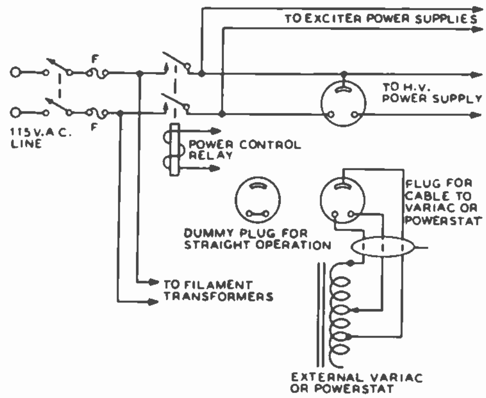


Figure 7.
CIRCUIT WITH VARIABLE-RATIO AUTO-TRANSFORMER.

When the dummy plug is inserted into the receptacle on the equipment, closing of the power control relay will apply full voltage to the primaries. With the cable from the Variac or Powerstat plugged into the socket the voltage output of the high-voltage power supply may be varied from zero to about 15 per cent above normal.

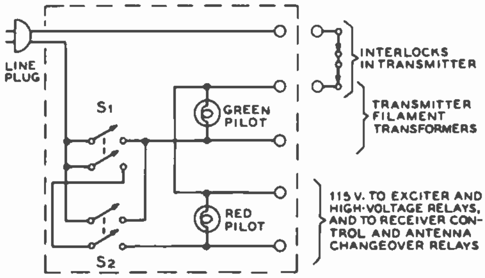


Figure 8.
PROTECTIVE CONTROL CIRCUIT.

With this circuit arrangement either switch may be closed first to light the heaters of all tubes and the filament pilot light. Then when the second switch is closed the high voltage will be applied to the transmitter and the red pilot will light. With a 30-second delay between the closing of the first switch and the closing of the second, the rectifier tubes will be adequately protected. Similarly, the opening of either switch will remove plate voltage from the rectifiers while the heaters remain lighted.

are the "Variac" manufactured by the General Radio Company, and the "Powerstat" manufactured by the Superior Electric Company. Both these types of variable-ratio transformers are excellently constructed and are available in a wide range of power capabilities. Each is capable of controlling the line voltage from zero to about 15 per cent above the nominal line voltage. Each manufacturer makes a single-phase unit capable of handling an output power of about 175 watts, one capable of about 750 to 800 watts, and a unit capable of about 1500 to 1800 watts. The maximum power-output capability of these units is available only at approximately the nominal line voltage, and must be reduced to a maximum current limitation when the output voltage is somewhat above or below the input line voltage. This, however, is not an important limitation for this type of application since the output voltage seldom will be raised above the line voltage, and when the output voltage is reduced below the line voltage the input to the transmitter is reduced accordingly.

One convenient arrangement for using a Variac or Powerstat in conjunction with the high-voltage transformer of a transmitter is illustrated in figure 7. In this circuit a heavy three-wire cable is run from a plug on the transmitter to the Variac or Powerstat. The Variac or Powerstat then is installed so that it is accessible from the operating desk so that the input power to the transmitter may be controlled during operation. If desired, the cable to the Variac or Powerstat may be unplugged from the transmitter and a dummy plus inserted in its place. With the dummy plug in place the transmitter will operate at normal plate voltage. This arrangement allows the transmitter to be wired in such a manner that an external Variac or Powerstat may be used if desired, even though the unit is not available at the time that the transmitter is constructed.

Notes on the Use of the Variac or Powerstat Plate voltage to the modulators may be controlled at the same time as the plate voltage to the final amplifier is varied if the modulator stage uses beam tetrode tubes; variation in the plate voltage on such tubes used as modulators causes only a moderate change in the standing plate current. Since the final amplifier plate voltage is being controlled simultaneously with the modulator plate voltage, the conditions of impedance match will not be seriously upset. In several high power transmitters using this system, and using beam-tetrode modulator tubes, it is possible to vary the plate input

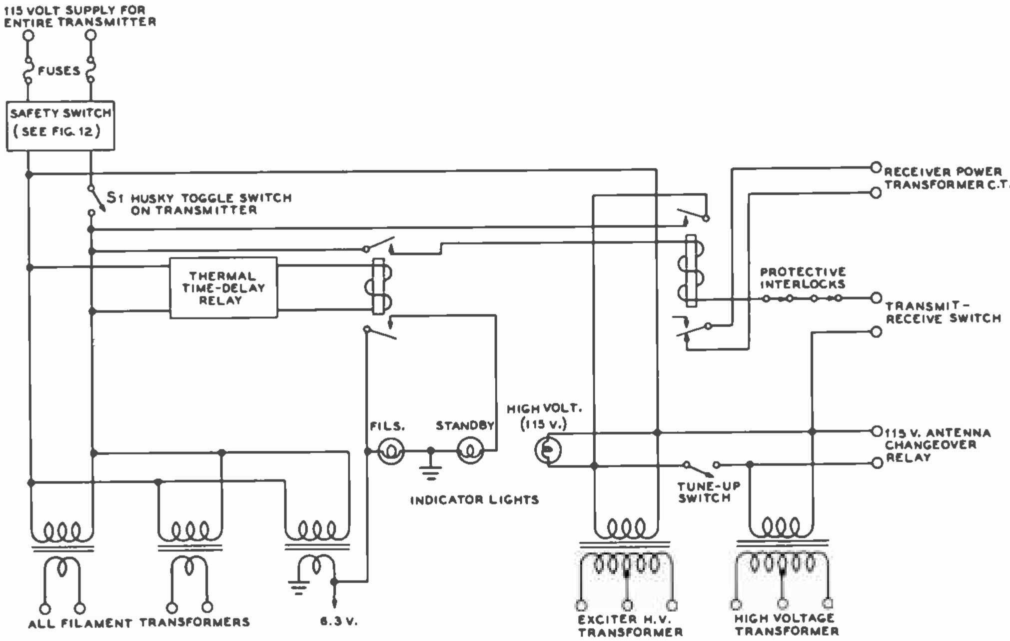


Figure 9.
TRANSMITTER CONTROL CIRCUIT.

Closing S_1 lights all filaments in the transmitter and starts the time-delay relay in its cycle. When the time-delay relay has operated closing the transmit-receive switch at the operating position will apply plate power to the transmitter and disable the receiver. A tune-up switch has been provided so that the exciter stages may be tuned without plate voltage on the final amplifier. The safety circuit of Figure 12 has been incorporated.

from about 50 watts to one kilowatt without a change other than a slight increase in audio distortion at the adjustment which gives the lowest power output from the transmitter.

With triode tubes as modulators it usually will be found necessary to vary the grid bias at the same time that the plate voltage is changed. This will allow the tubes to be operated at approximately the same relative point on their operating characteristic when the plate voltage is varied. When the modulator tubes are operated with zero bias at full plate voltage, it will usually be possible to reduce the modulator voltage along with the voltage on the modulated stage, with no apparent change in the voice quality. However, it will be necessary to reduce the audio gain at the same time that the plate voltage is reduced.

The manufacturers of the Variac and the Powerstat recommend that changes in the setting of the auto-transformers be made with the power *off*. when the load is inductive. The primary of the plate transformer in an amateur transmitter usually will present an inductive load to the line. Hence it is best that the transmitter be shut down for a moment before the setting of the auto-transformer is changed. If the setting is varied at a time when the transmitter is operating, it is possible that pitting of the surface of the winding and burning of the end of the brush will occur. If this should take place the unit must be disassembled and the contacting surface carefully cleaned.

10-4 Transmitter Control Methods

Almost everyone, when first getting a new transmitter on the air, has had the experience of having to throw several switches and pull or insert a few plugs when changing from receive to transmit. This is one extreme in the direction of how *not* to control a transmitter. At the other extreme we find systems where

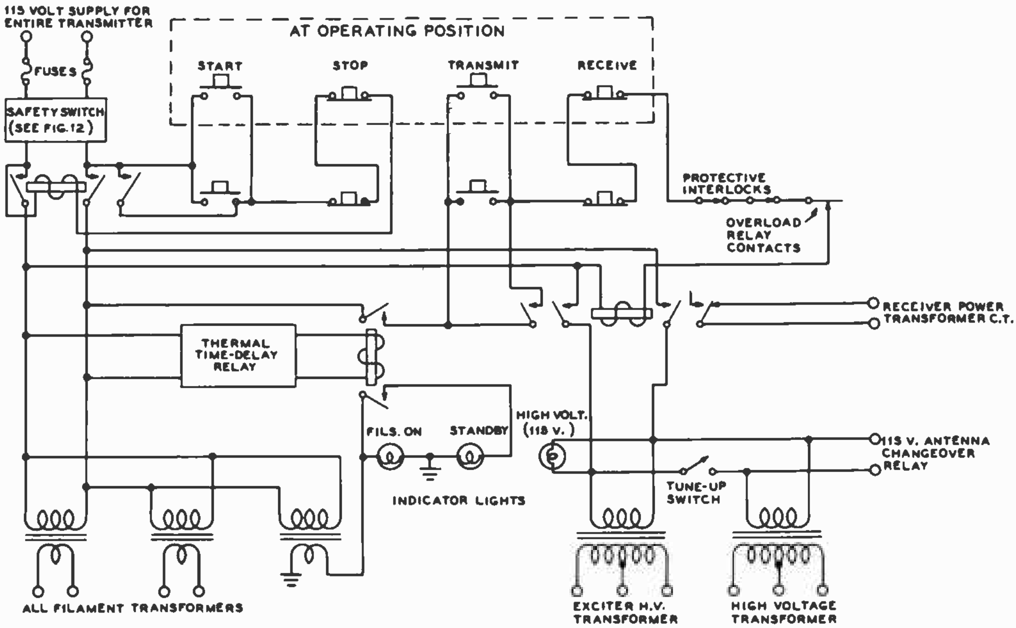


Figure 10.

PUSH-BUTTON TRANSMITTER-CONTROL CIRCUIT.

Pushing the START button either at the transmitter or at the operating position will light all filaments and start the time-delay relay in its cycle. When the cycle has been completed, a touch of the TRANSMIT button will put the transmitter on the air and disable the receiver. Pushing the RECEIVE button will disable the transmitter and restore the receiver. Pushing the STOP button will instantly drop the entire transmitter from the a-c line. If desired, a switch may be placed in series with the lead from the RECEIVE button to the protective interlocks; opening this switch will make it impossible for any person accidentally to put the transmitter on the air. Various other safety provisions, such as the protective-interlock arrangement described in the text and the circuit of figure 12 have been incorporated.

With the circuit arrangement shown for the overload-relay contacts, it is only necessary to use a simple normally-closed d-c relay with a variable shunt across the coil of the relay. When the current through the coil becomes great enough to open the normally-closed contacts the hold-circuit on the plate-voltage relay will be broken and the plate voltage will be removed. If the overload is only momentary, such as a modulation peak or a tank flashover, merely pushing the TRANSMIT button will again put the transmitter on the air. This simple circuit provision eliminates the requirement for expensive overload relays of the mechanically-latching type, but still gives excellent overload protection.

it is only necessary to speak into the microphone or touch the key to change both transmitter and receiver over to the transmit condition. Most amateur stations are intermediate between the two extremes in the control provisions and use some relatively simple system for transmitter control.

In figure 8 is shown an arrangement which protects mercury-vapor rectifiers against premature application of plate voltage without resorting to a time-delay relay. No matter which switch is thrown first, the filaments will be turned on first and off last. However, double-pole switches are required in place of the usual single-pole switches.

When assured time delay of the proper interval and greater operating convenience are desired, a group of inexpensive a-c relays may be incorporated into the circuit to give a control circuit such as is shown in figure 9. This arrangement uses a 115-volt thermal (or motor-operated) time-delay relay and a d-p-d-t 115-volt control relay. Note that the protective interlocks are connected in series with the coil of the relay which applies high voltage to the transmitter. A tune-up switch has been included so that the transmitter may be tuned up as far as the grid circuit of the final stage is concerned before application of high voltage to the final amplifier. Provisions for operating an antenna-changeover relay and for cutting the plate voltage to the receiver when the transmitter is operating have been included.

A circuit similar to that of figure 9 but incorporating push-button control of the transmitter is shown in figure 10. The circuit features a set of START-STOP and TRANSMIT-RECEIVE buttons at the transmitter and a separate set at the operating position. The control push buttons operate independently so that either set may be used to control the transmitter. It is only necessary to push the START button momentarily to light the transmitter filaments and start the time-delay relay in its cycle. When the standby light comes on it is only necessary to touch the TRANSMIT button to put the transmitter on the air and disable the receiver. Touching the RECEIVE button will turn off the transmitter and restore the receiver. After a period of operation it is only necessary to touch the STOP button at either the transmitter or the operating position to shut down the transmitter. This type of control arrangement is called an electrically-locking push-to-control system. Such systems are frequently used in industrial electronic control.

Automatic Control of Transmitter Plate Supplies A popular circuit for accomplishing the controlling of the transmitter plate supplies with the keying circuit was described in the RADIO HANDBOOK, Eleventh Edition, on page 439. This description concerned the conversion and the power supply circuits for an ART-13 transmitter. The essentials of this power-supply control arrangement are given in figure 11.

The circuit of figure 11 operates in the following manner: When the key is closed, the keying relay K_2 closes. One set of contacts on this relay acts to key the circuit in the transmitter which is doing the actual job of keying the output. Alternatively, this set of contacts may be connected to the keyer unit to be described for accomplishing screen-voltage keying of the keyed stage. The other set of contacts shorts the grid of the 6C5 tube to the cathode as it closes. This action causes the 6C5 to draw full plate current which immediately closes relay K_1. As K_1 closes it causes the relay or relays to close which turn on the high-voltage plate supplies to the transmitter. Then, with normal keying, relay K_1 will remain closed thus keeping the plate-

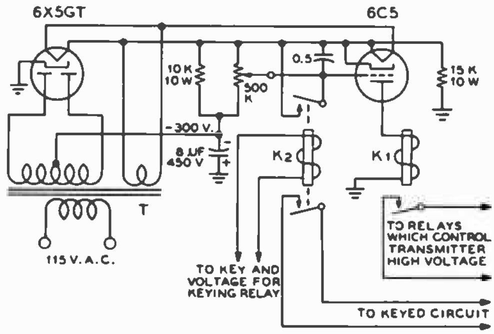


Figure 11.
AUTOMATIC-POWER-CONTROL CIRCUIT.
This circuit automatically turns on the high voltage power supplies when the key is first closed, then keeps the power supplies operating until a few seconds after the transmission of the last character.

K_1—2500-ohm s.p.s.t. relay
K_2—Double-circuit keying relay
T—Small power transformer such as Stancor P-6010 furnishing 650 v. c.t. at 40 ma., 6.3 v. at 2 a., and a 5-volt winding which is not used.

voltage supplies operative. Relay K_1 still acts, of course, to key the output of the transmitter.

When the transmission is completed, the 0.5-µfd. capacitor from grid to cathode on the 6C5 finally gets a chance to become charged from the tap on the 500,000-ohm potentiometer between negative high voltage and the cathode of the 6C5, thus cutting off the plate current to the 6C5. The time delay between the transmission of the last character and the opening of K_1 is controlled between a very short period and an interval of about ten seconds by the adjustment of the potentiometer. The 0.5-µfd. capacitor does not have time to build up sufficient charge during normal keying to cut off the tube since it is shorted out each time the key is closed.

10-5 Safety Precautions

The best way for an operator to avoid serious accidents from the high voltage supplies of a transmitter is for him to use his head, act only with deliberation, and not take unnecessary chances. However, no one is infallible, and chances of an accident are greatly lessened if certain factors are taken into consideration in the design of a transmitter, in order to

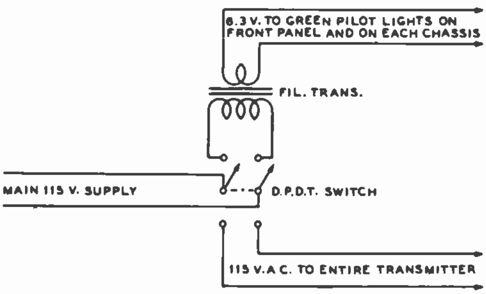


Figure 12.
COMBINED MAIN SWITCH AND SAFETY SIGNAL.
When shutting down the transmitter, throw the main switch to neutral. If work is to be done on the transmitter, throw the switch all the way to "pilot," thus turning on the green pilot lights on the panel and on each chassis, and insuring that no voltage can exist on the primary of any transformer, even by virtue of a short or accidental ground.

protect the operator in the event of a lapse of caution. If there are too many things one must "watch out for" or keep in mind there is a good chance that sooner or later there will be a mishap; and it only takes *one*. When designing or constructing a transmitter, the following safety considerations should be given attention.

Grounds For the utmost in protection, everything of metal on the front panel of a transmitter capable of being touched by the operator should be at ground potential. This includes dial set screws, meter "zero adjuster" screws, meter cases if of metal, meter jacks, *everything* of metal protruding through the front panel or capable of being touched or *nearly* touched by the operator. This applies whether or not the panel itself is of metal. Do not rely upon the insulation of meter cases or tuning knobs for protection.

The B negative or chassis of all plate power supplies should be connected together, and to an external ground such as a waterpipe. In the case of a bias supply, the B positive should be connected to the common ground.

Exposed Wires and Components It is not necessary to resort to rack and panel construction in order to provide complete enclosure of all components and wiring of the transmitter. Even with metal-chassis construction it is possible to arrange things so as to incorporate a protective shielding housing which will not interfere with ventilation yet will prevent contact with all wires and components carrying high voltage d.c. or a.c., in addition to offering shielding action.

If everything on the front panel is at ground potential (with respect to external ground) and all units are effectively housed with protective covers, then there is no danger except when the operator must reach into the interior part of the transmitter, as when changing coils, neutralizing, adjusting coupling, or shooting trouble. The latter procedure can be made safe by making it possible for the operator to be *absolutely certain* that all voltages have been turned off and that they cannot be turned on either by short circuit or accident. This can be done by incorporation of the following system of main primary switch and safety signal lights.

Combined Safety Signal and Switch The common method of using red pilot lights to show when a circuit is "on" is useless except from an ornamental standpoint. When the red pilot is not lit it *usually* means that the circuit is turned off, but it *can* mean that the circuit is on but the lamp is burned out or not making contact.

To enable you to grab the tank coils in your transmitter with absolute assurance that it is impossible for you to obtain a shock except from possible undischarged filter condensers (see following topic for elimination of this hazard), it is only necessary to incorporate a device similar to that of figure 12. It is placed near the point where the main 110-volt leads enter the room (preferably near the door) and in such a position as to be inaccessible to small children. Notice that this switch breaks *both* leads; switches that open just one lead do not afford complete protection, as it is sometimes possible to complete a primary circuit through a short or accidental ground. Breaking just one side of the line may be all right for turning the transmitter on and off, but when you are going to stick an arm inside the transmitter, *both* 110-volt leads should be broken.

When you are all through working your transmitter for the time being, simply throw the main switch to neutral. Then you can leave the transmitter and even go on a vacation with absolute peace of mind.

When you find it necessary to work on the transmitter or change coils, throw the switch so that the green pilots light up. These can be ordinary 6.3-volt pilot lamps behind green bezels or dipped in green lacquer. One should be placed on the front panel of the transmitter; others should be placed so as to be easily visible when changing coils or making adjustments requiring the operator to reach inside the transmitter.

For 100 per cent protection, just obey the following rule: *never work on the transmitter or reach inside any protective cover except when the green pilots are glowing.* To avoid confusion, no other green pilots should be used on the transmitter; if you want an indicator jewel to show when the filaments are lighted, use amber instead of green.

If the main switch is out of reach of small children, a conspicuous sign, such as "DO NOT TOUCH UNDER ANY CIRCUMSTANCES," placed on the switch cover will guard against the off chance that someone else would throw the switch unexpectedly. An alternative is to place the switch on the under side of the operating table out of sight. The latter is not so desirable when small children have access to the room.

Safety Bleeders Filter capacitors of good quality hold their charge for some time, and when the voltage is more than 1000 volts it is just about as dangerous to get across an undischarged 4-μfd. filter capacitor as it is to get across a high-voltage supply that is turned on. Most power supplies incorporate bleeders to improve regulation, but as these are generally wire-wound resistors, and as wire-wound resistors occasionally open up without apparent cause, it is desirable to incorporate an auxiliary safety bleeder across each heavy-duty bleeder. Carbon resistors will not stand much dissipation and sometimes change in value slightly with age. However, the chance of their opening up when run well within their dissipation rating is very small.

To make *sure* that all capacitors are bled, it is best to short each one with an insulated screwdriver. However, this is sometimes awkward and always inconvenient. One can be virtually sure by connecting auxiliary carbon bleeders across all wire-wound bleeders used on supplies of 1000 volts or more. For every 500 volts, connect in series a 500,000-ohm 1-watt carbon resistor. The drain will be negligible (1 ma.) and each resistor will have to dissipate only 0.5 watt. Under these conditions the resistors will last indefinitely with little chance of opening up. For a 1500-volt supply, connect three 500,000-ohm resistors in series. If the voltage exceeds an integral number of 500 volt divisions, assume it is the next higher integral value; for instance, assume 1800 volts as 2000 volts and use four resistors.

Do *not* attempt to use fewer resistors by using a higher value for the resistors; not over 500 volts should appear across any single 1-watt resistor.

In the event that the regular bleeder opens up, it will take several seconds for the auxiliary bleeder to drain the capacitors down to a safe voltage, because of the very high resistance. Hence, it is best to allow 10 or 15 seconds after turning off the plate supply before attempting to work on the transmitter.

"Hot" Adjustments Some amateurs contend that it is almost impossible to make certain adjustments, such as coupling and neutralizing, unless the transmitter is running. The best thing to do is to make all neutralizing and coupling devices adjustable from the front panel by means of flexible control shafts which are broken with insulated couplings to permit grounding of the panel bearing.

If your particular transmitter layout is such that this is impracticable and you refuse to throw the main switch to make an adjustment —throw the main switch—take a reading—throw the main switch—make an adjustment —and so on, then protect yourself by making use of long adjusting rods made from ½-inch dowel sticks which have been wiped with oil when perfectly free from moisture.

If you are addicted to the use of pickup loop and flashlight bulb as a resonance and neutralizing indicator, then fasten it to the end of a long dowel stick and use it in that manner.

Protective Interlocks With the increasing tendency toward construction of transmitters in enclosed steel cabinets a transmitter becomes a particularly lethal device unless adequate safety provisions have been incorporated. Even with a combined safety signal and switch as shown in figure 12 it is still conceivable that some person unfamiliar with the transmitter could come in contact with high voltage. It is therefore recommended that the transmitter, wherever possible, be built into a complete metal housing or cabinet and that *all* doors or access covers be provided with protective interlocks (all interlocks must be connected in *series*) to remove the high voltage whenever these doors or covers are opened. The term "high voltage" should mean any voltage above approximately 150 volts, although it is still possible to obtain a serious burn from a 150-volt circuit under certain circumstances. The 150-volt limit usually will mean that grid-bias packs as well as high-voltage packs should have their primary circuits opened when any interlock is opened.

10-6 Transmitter Keying

The carrier from a c-w telegraph transmitter must be broken into dots and dashes for the transmission of code characters. The carrier signal is of a constant amplitude while the key is closed, and is entirely removed when the key is open. When code characters are being transmitted, the carrier may be considered as being modulated by the keying. If the change from the no-output condition to full-output, or vice versa, occurs too rapidly, the rectangular pulses which form the keying characters contain high-frequency components which take up a wide frequency band as sidebands and are heard as clicks.

The two general methods of keying a c-w transmitter are those which control the excitation, and those which control the plate voltage which is applied to the final amplifier. *Excitation keying* can be of several forms, such as crystal-oscillator keying, buffer-stage keying, or blocked-grid keying. In such an arrangement, plate voltage is applied to the final amplifier at all times.

Key-Click Elimination Key-click elimination is accomplished by preventing a too-rapid make-and-break of power to the antenna circuit, rounding off the keying characters so as to limit the sidebands to a value which does not cause interference to adjacent channels. Too much lag will prevent fast keying, but fortunately key clicks can be practically eliminated without limiting the speed of manual (hand) keying. Some circuits which eliminate key clicks introduce too much time-lag and thereby add *tails* to the dots. These tails may cause the signals to be difficult to copy at high speeds.

Sparking Contacts Just as any electrical circuit producing sparks will cause interference to *nearby* receivers unless precautions are taken to prevent it, so will a sparking key or relay cause interference unless measures are taken to prevent it. The interference produced will have no correlation with the frequency upon which the transmitter is operating; the clicks produced are not keying sidebands, but rather are due to the sparking contacts and their associated wiring acting as a crude form of aperiodic spark transmitter.

Clicks due to key sparks can be minimized by limiting the amount of power handled by the key, and then putting an r-f bypass capacitor of 0.002 μfd. or so directly across the key terminals (on the key, not at the transmitter), and in stubborn cases a couple of r-f chokes in series with the key leads right at the key terminals.

A sparking relay, which usually will be called upon to handle considerably more power, can be prevented from causing trouble by housing it in a grounded metal can and by-passing to the can all leads to the relay at the point where they enter the can. If this does not suffice, inserting r-f chokes in series with the leads, right at the relay, often will prove satisfactory.

Clicks due to sparking contacts should not be confused with those due to keying sidebands. The former may be heard over most of the radio spectrum if not suppressed, but only for a short distance. Clicks due to keying sidebands are actually radiated by the transmitting antenna, and may be heard for a great distance, but under the worst conditions only over a band of frequencies a few per cent either side of the carrier frequency.

Primary Keying One simple form of clickless keying which is satisfactory for certain applications under some conditions is *primary keying*. The key or keying relay is placed in the primary winding of the a-c plate transformer feeding the final amplifier (and in some cases one or more of the preceding stages).

The inherent lag in the plate supply filter will "round off" the keying to the point where keying sidebands are insignificant. In fact, if a heavily filtered 60-cycle single-phase supply is used, there may be too much lag for anything but slow hand keying, and code characters will have objectionable "tails". How-

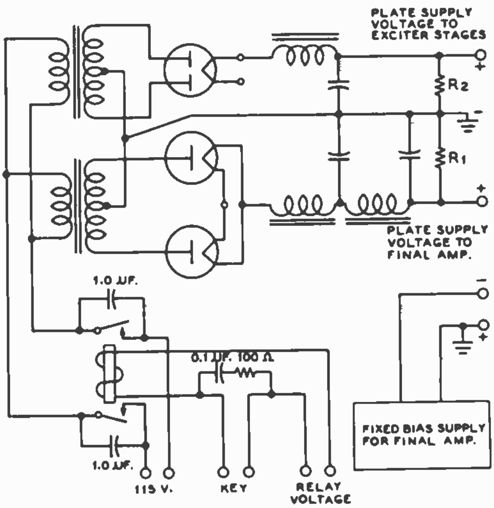


Figure 13.
IMPROVED PRIMARY-KEYING CIRCUIT.
If the system illustrated above is properly used it will be possible to obtain excellent primary keying without objectionable keying lag. The circuit characteristics are described in detail in the text.

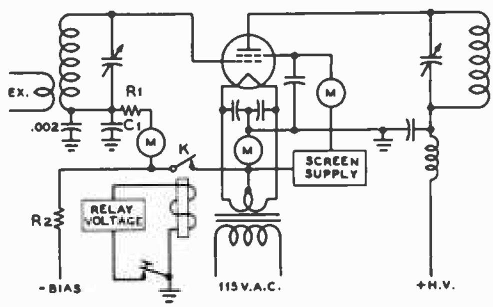


Figure 14.
GRID-BLOCK KEYING CIRCUIT.
The grid-bias supply voltage must be sufficient to cut off plate current to the amplifier stage in the presence of the excitation voltage. Resistor R_1 should be the normal grid-leak value for the amplifier tube. The values of R_1 and C_1 should be adjusted for correct keying action.

ever, if the plate supply filter is engineered as a multiple section low-pass filter working into its characteristic impedance and designed for about 40-cycle cutoff, it is possible to obtain nearly pure direct current and yet key through the filter cleanly at high speed.

When precautions are taken against spark radiation, this type of keying is an almost sure cure for clicks. The disadvantages are (1) a heavy relay is required, in order to avoid sticking contacts, and (2) a special filter is required to avoid keying tails.

Improved Primary-Keying Circuit Figure 13 illustrates an improved type of primary-keying arrangement which reduces greatly the keying lag which accompanies the usual primary-keying circuits. Upon first glance the circuit appears to be quite conventional: However, the difference lies in the fact that the plate supply for the exciter is keyed at the same time as the final plate supply, the final amplifier is biased by means of a power supply which provides sufficient voltage to bias the final amplifier tubes past cutoff at the operating plate voltage used, and a very high value of bleeder resistance is employed on the high-voltage plate supply. Adequate filter should be used on the high-voltage plate supply to insure that the transmitted signal will be adequately pure. However, the filter on the *exciter* plate supply should be reduced to the point where high-speed keying without tails is obtained. The resistor R_1 in the circuit diagram should be approximately one megohm for each 1000 volts of plate voltage. This means that this resistor can very well be the multiplier resistor associated with the high-voltage d-c voltmeter. Such a voltmeter is required anyway with all transmitters operating at more than 900 watts input. Resistor R_2 can be a conventional bleeder resistor of 15,000 to 50,000 ohms depending upon the voltage.

A word should be mentioned in regard to safety when using only a very high value of bleeder resistor on a high-voltage plate supply. Some provision should be included in the transmitter-control system to apply a comparatively *low* (20,000 to 50,000 ohms) value of bleeder resistor across the high-voltage plate supply except during the period when actually keying the transmitter.

Blocked-Grid Keying The grid bias in an r-f amplifier may be increased in the key-up position to a value sufficient to reduce the output from the amplifier to zero. Figure 14 illustrates such a circuit applied to a beam-tetrode amplifier stage. Also, through the use of the circuit as diagrammed, the milliammeter placed in the cathode return

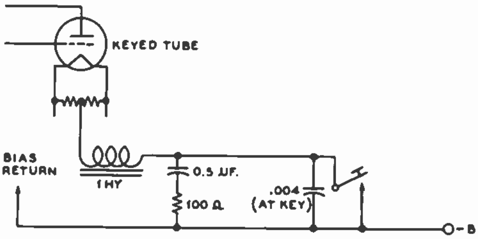


Figure 15.
CENTER-TAP KEYING WITH CLICK FILTER.

The constants shown above are suggested as starting values; considerable variation in these values can be expected for optimum keying of amplifiers of different operating conditions. It is suggested that a keying relay be substituted for the key in the circuit above wherever practicable.

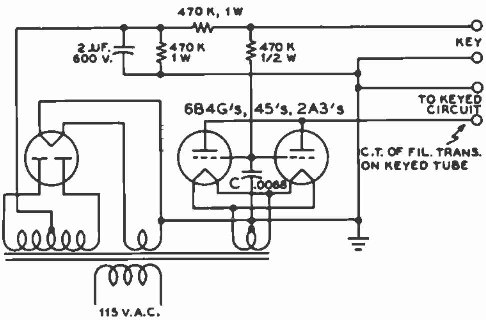


Figure 16.
VACUUM-TUBE CENTER-TAP KEYING.

Some variation in the capacitance of C may be necessary to obtain smooth keying without clicks or tails at the keying speed normally used.

of the amplifier will indicate only the plate current of the stage—*minus* the small amount of current which passes through R_2 and the keying relay to the cathode milliammeter and ground. In a circuit such as this, a paper capacitor of 0.1-µfd. or so is placed across the r-f by-pass for the grid tank to ground. This additional capacitor, C_1 in figure 14, is used to control the time constant of the keying circuit in conjunction with resistors R_1 and R_2. The time constant for the keying "make" is established by C_1 and R_1, in which case R_1 also acts as the grid leak for the stage. The time constant for the keying "break" is determined by C_1 and R_1 plus R_2 in series. Hence the time constant for the make in such a circuit will be shorter than the time constant for the keying break.

Oscillator Keying, Break-in A stable and quick-acting crystal oscillator may be keyed in the plate, cathode or screen-grid circuit for break-in operation.

Assuming that the crystal oscillator itself is capable of being keyed without clicks, it is still possible to transmit serious keying sideband clicks if the oscillator is followed by several heavily driven amplifier stages. A heavily-excited Class C amplifier or multiplier acts like a "clipper" stage, tending to square up a rounded excitation impulse, and the cumulative effect of several such stages cascaded is sufficient to square up the "softened" characters out of the oscillator to the point where bad clicks result.

Parasitics with Oscillator Keying When keying in the crystal stage, or, for that matter, any stage ahead of the final amplifier, the stages following the keyed one must be absolutely stable so that parasitic or output frequency oscillation will not occur when the excitation is rising on the beginning of each keying impulse. This type of oscillation gives rise to extremely offensive clicks which cannot be eliminated by any type of filter; in fact, a filter designed to slow up the rate at which signal comes to full strength may only make them worse.

Center-Tap Keying The lead from the center-tap connection to the filament of an r-f amplifier or oscillator tube can be opened and closed for keying a circuit. Such a keying system opens the plate-voltage circuit and at the same time opens the grid-bias return lead. For this reason, the grid circuit is blocked at the same time that the plate circuit is opened, so that excessive sparking does not occur at the key contacts. Unfortunately, this method of keying applies the power too suddenly to the tube, producing a serious key click. This click often can be eliminated with the key-click filter shown in figure 15.

Vacuum Tube C.T. Keying A variation on the center-tap keying circuit of figure 15 producing virtually no clicks is one in which the key or relay is replaced by one or more low resistance triodes in parallel, as in figure 16. These tubes act as a very high resistance when sufficient blocking bias is applied to them, and as a very low resistance when the bias is removed. The desired amount of lag or "cushioning effect" can be obtained by employing suitable resistance and capacitance values in the grid of the keyer tube(s). Because very little spark is produced at the key, due to the small amount of power in the key circuit, sparking clicks are easily suppressed.

One type 45 tube should be used for every 50 ma. of plate current. Type 6B4G or 2A3 tubes may also be used; allow one 6B4G tube for every 80 ma. of plate current.

Because of the series resistance of the keyer tubes, the plate voltage at the keyed tube will be from 30 to 60 volts less than the power supply voltage. This voltage appears as cathode bias on the keyed tube, assuming the bias return is made to ground, and should be taken into consideration when providing bias.

10-7 Break-In Operation

To the experienced c-w operator the most satisfactory method of communication between stations is obtained through the use of automatic break-in operation. This type of operation is not obtained easily, nor can it be purchased ready-made at the nearest distributor of amateur components. The components for accomplishing automatic break-in operation are not necessarily expensive, nor are the electrical circuits unduly complex. But this type of operation is a challenge to the operator, since careful planning and skillful tailoring of components and circuits to the equipment of the station will be required before smooth break-in operation with a medium power transmitter can be obtained.

Keyer Circuit Requirements In the first place it may be established that the majority of new design transmitters, and many of those of older design as well, use a medium power beam tetrode tube either as the output stage or as the exciter for the output stage of a high power transmitter. Thus the transmitter usually will end up with a tube such as a 2E26, 807, 814, 813, 4-65A, 4E27/257B, 4-125A or similar type, or one of these tubes will be used as the stage just ahead of the output stage.

Second, it may be established that it is undesirable to key further down in the transmitter chain than the output stage or the stage just ahead of the final amplifier. If a low-level stage, which is followed by a series of class C amplifiers, is keyed, serious transients will be generated in the output of the transmitter even though the keyed stage is being turned on and off very smoothly. This condition arises as a result of "pulse sharpening," which has been discussed previously.

Third, the output from the stage should be completely cut off when the key is up, and the time constant of the rise and decay of the keying wave should be easily controllable.

Fourth, it should be possible to make the rise period and the decay period of the keying wave approximately equal. This type of keying envelope is the only one tolerable for commercial work, and is equally desirable for obtaining clean cut and easily readable signals in amateur work.

Fifth, it is desirable that the keying circuit be usable without a keying relay, even when a high-power stage is being keyed.

Sixth, for the sake of simplicity and safety, it should be possible to ground the frame of the key, and yet the circuit should be such that placing the fingers across the key will not result in electrical shock. In other words, the keying circuit should be inherently safe.

Last, it should be possible to use the circuit in conjunction with a break-in system, so that no signal from the exciter stages of the transmitter will be audible in the receiver with the key up.

All these requirements have been met in the keying circuit to be described. The circuit itself has been installed in one of the exciters described in Chapter Twenty-One, and several keyer units such as illustrated in figures 17, 18, and 19 have been constructed and installed as an added portion of existing transmitters.

Screen-Voltage Keyer Unit The unit to be described may be used as a screen-voltage keyer for any of the tubes previously mentioned. In addition, the circuit principles may be used in designing a keyer unit with larger current capability for screen-grid keying of higher powered beam-tetrode tubes.

Screen-voltage keying was chosen as the

most logical keying method for the following reasons: (1) by keying the screen-grid voltage of an amplifier tube it is possible to control a relatively large power input to the plate circuit with a much smaller amount of voltage and current capability in the keyer. (2) As mentioned before, either the output stage or the exciter for the final almost invariably uses a medium power beam-tetrode tube. Grid-block keying could have been used except that: the keyed transmitter-output waveform obtainable with grid-block keying is a function of the excitation conditions of the stage and may vary widely on different bands in a particular transmitter or on different transmitters; the grid-block keying voltage required would vary widely with the manner in which screen voltage is fed to the stage; and the screen voltage could soar to high values with key up and series screen feed. Hence (3), screen-voltage keying offers the advantage of uniformity of keying characteristics in transmitters with widely differing designs since the screen voltage for the tube not only is controlled by the keyer unit but actually *is supplied to the keyed tube by the keyer unit.*

Circuit of the Keyer The keyer itself requires only two tubes: a dual triode (of which one half is used in the break-in circuit and may be eliminated if desired) d-c voltage amplifier, and a cathode-follower output stage. The keyer requires a positive supply of 375 to 400 volts and a negative supply of 275 to 300 volts. The current drain from the positive supply is only slightly greater than the screen current to the keyed stage, and the drain from the negative supply is less than 5 ma. The unit illustrated includes a simple dual-voltage power supply for the positive and negative voltage; the exciter described in Chapter Twenty-One which uses the keyer circuit does not have additional power supplies for the keyer alone but uses the normal bias and plate voltage supplies included within the equipment.

The first condition established in the circuit design of the keyer was the fact that the output stage should be a cathode follower. The cathode follower has excellent operating characteristics as a series screen-voltage control tube since it acts effectively as a series resistor, from the plate-voltage supply, whose resistance may be varied over a very wide range. With key down the drop across the tube is less than 100 volts, but with the key up the tube drop is over 400 volts. Note that the drop across the tube with the key up is greater than the plate supply voltage. This condition exists since the cathode of the tube is connected through a relatively large resistor to the negative 275-volt line.

Operation of the Keyer In order for the cathode follower to operate in the desired manner, the grid potential on the cathode-follower tube should vary from a value of about −100 volts with the key up all the way to a value only slightly less than the plate supply voltage with the key down. It would be possible to obtain this voltage variation with a keying relay and two RC circuits. But it is much more convenient if the keying relay may be eliminated and replaced by a vacuum tube operating as a d-c voltage amplifier. The first half of the 12AU7 tube operates in such a manner in the circuit of figure 19. Also, through the use of the d-c amplifier stage, one side of the key may be grounded, and the voltage appearing across the

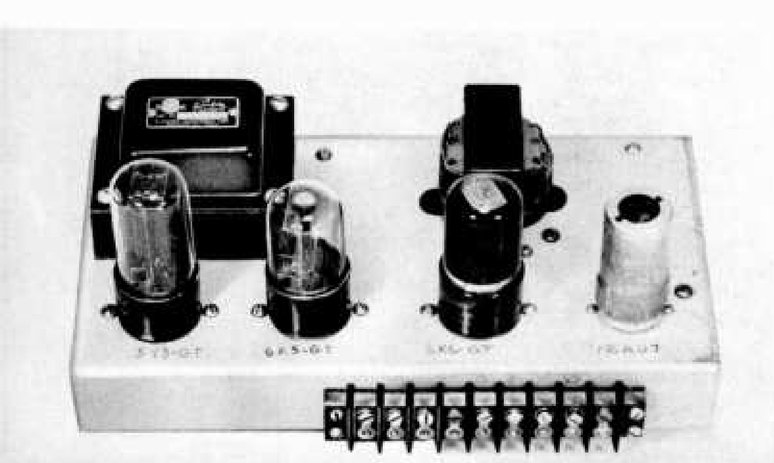


Figure 17.
TOP VIEW OF THE KEYER UNIT.

key with key up is fed through such a high value of resistance that no shock is felt when the fingers are placed across the key.

The operation of the half-12AU7 d-c amplifier is as follows: With the key up the voltage appearing at the ungrounded key contact is about −110 volts. But the 1-megohm and 470,000-ohm resistors which control the grid bias on the 12AU7 would tend to set the grid potential at about −85 volts. However, this would place the grid at a positive potential with respect to the cathode; this condition cannot take place due to the high values of feed resistance in the grid lead. Hence the grid voltage is pulled up to the cathode voltage by grid current in the half-12AU7. Since grid current is flowing, the grid is actually slightly positive with respect to the cathode potential.

With the grid of the d-c amplifier drawing a slight amount of grid current, the tube is in a highly conductive condition as far as cathode-to-plate current is concerned. Since plate current is being fed through a relatively high resistor (470,000 ohms) and since the tube internal resistance is relatively low, the plate of the half-12AU7 is positive only about 20 volts with respect to cathode. Hence the plate of the half-12AU7 operates at a plate potential with respect to ground of about −90 volts.

The plate of the d-c amplifier is directly connected to the grid of the cathode-follower stage. Since the control grid of the 6K6-GT thus is at a potential of about −90 volts and since the cathode current is very low, the tube is essentially cut off. Thus the key-up cathode voltage drops to about −35 volts. This is the voltage which is fed to the screen grid of the tube in the r-f stage being keyed.

When the key is closed the following sequence of events takes place: The cathode of the half-12AU7 is grounded. Since the grid of this tube now is at a potential of about −85 volts, the current flow through the tube is completely cut off. Hence the 0.025-μfd. capacitor from plate to ground is charged exponentially by the 470,000-ohm series resistor from the plate-supply voltage. The grid voltage of the 6K6-GT thus rises smoothly, followed in turn by the cathode voltage. The actual cathode voltage rises from the key-up potential of about −35 volts to a potential of 275 to 350 volts, depending upon the screen current of the r-f tube being keyed. The 10,000-ohm resistor shown in series with the output lead is there simply to limit the maximum screen voltage and screen current to the r-f tube. This resistor may be eliminated for the larger tube types, or it may be increased in value for keying smaller screen-grid tubes.

When the key is lifted again, the 0.025-μfd. capacitor is discharged by the current through the half-12AU7 and the resistor from its cathode to the −275-volt line. Thus the cathode voltage of the 6K6-GT cathode follower falls smoothly to the key-up potential of about −35 volts.

Time Constant Changes The time constant of the keying wave is determined by the resistors in the plate and cathode circuits of the half-12AU7 and by the value of the capacitor from the plate of the 12AU7 to ground. Sharper keying may be obtained by decreasing the value of this capacitor, and softer keying may be had by increasing the value above 0.025 μfd. The value given has proven quite satisfactory for normal service with a straight key and 0.0068 μfd. has given good results with "bug" keying.

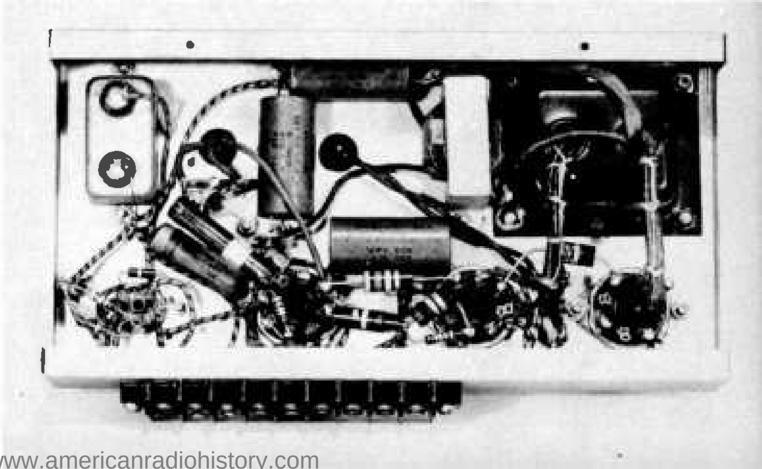


Figure 18.
UNDERCHASSIS PHOTO OF THE KEYER UNIT.

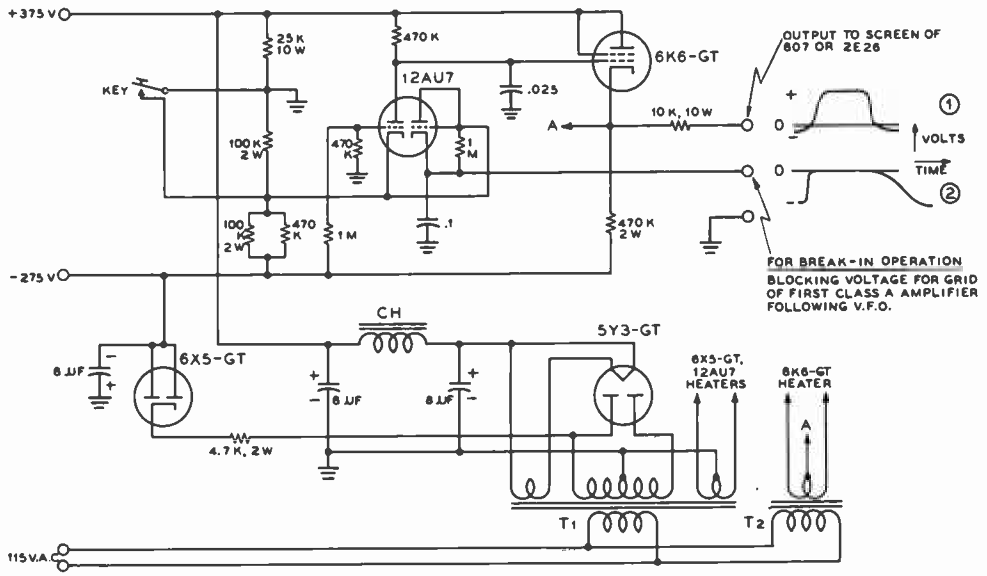


Figure 19.
SCHEMATIC DIAGRAM OF THE KEYER UNIT.

CH—Small 40 or 50 ma. filter choke (Stancor C-1706)
T_1—650 c.t. 40 ma., 5 v. 3 a., 6.3 v. 2 a. (Stancor P-6010)
T_2—Small 6.3-volt filament transformer (Stancor P-6134)

Plate-Current Cut-Off on the Keyed Stage Experience with screen-grid keying circuits has shown that merely dropping the screen potential to zero will not completely cut off plate current in the keyed stage with most beam-tetrode tubes. However, when the screen potential is dropped to about −35 volts with the key up it was found that plate current is completely cut off in the case of all the tube types tested. It is for that reason that provision was included in the circuit illustrated for dropping the screen voltage to a negative value with the key up.

The Exciter Blocking Circuit Also included within the keyer unit is a circuit for supplying a blocking voltage to the low-level stages of the exciter unit. The normal procedure is to feed the exciter-blocking voltage, when break-in operation is desired, to the grid return of the class A amplifier stage which follows the variable-frequency oscillator. This exciter-blocking circuit is included as a portion of the 70E-8/2E26 exciter described in Chapter Twenty-One.

The blocking circuit requires only a diode, one resistor and one capacitor. In the unit illustrated the diode consists of the second half of the 12AU7 tube with the grid and plate strapped together. The waveform of the exciter-blocking voltage is illustrated at (2) in figure 19. With key up both cathode and plate of the diode are at the potential of the ungrounded contact of the key—approximately −110 volts.

When the key is closed, the plate of the diode is grounded. A heavy current flows through the diode in discharging the 0.1-µfd. capacitor rapidly. This rapid discharge to zero voltage produces a sharp front on the exciter-blocking wave. Hence the exciter stages come into operation quickly, but the transient generated by this sharp waveform dies out before energy begins to be delivered by the keyed stage. In other words the keying waveform lags behind the exciter *un*-blocking waveform on the "make" of each character.

When the key is lifted the plate of the diode immediately drops to the key potential of −110 volts, and the keying wave (waveform (1) of figure 19 cuts off the output from the transmitter. But the blocking-voltage waveform falls more slowly since the 0.1-µfd. capacitor must be charged through the 1-meg-ohm resistor. Hence, on the "break" of each character the keying waveform is *ahead* of the blocking waveform. Thus, as far as the output stage of the transmitter is concerned the exciter is operating by the time it is turned on and is still operating when it goes off. But listening in the receiver one can hear the exciter stop operating a fraction of a second after the last character has been sent. Thus for continuous rapid keying the exciter is heard steadily in the receiver. But with slow sending, or with a slight break in the regularity of rapid sending, signals may be heard in the receiver on the operating frequency of the transmitter.

Using the Keyer in a Full Break-In System Through the use of the keyer unit just described it is possible to key the transmitter without clicks or chirps, and still be able to receive signals when the key is up. But there are three other functions which may or may not be included within the overall keying system in the complete break-in hookup. These are: (1) shorting of the receiver input whenever the key is closed; (2) automatically blocking the output of the receiver when the key is closed; and (3) automatic control of the transmitter high-voltage supplies.

These three additional possible features of a complete keying system for break-in will be discussed in turn. First, it is usually necessary to short the antenna input of the receiver when a high-power transmitter is in use and when a *separate* antenna is employed for receiving. If the receiver input is not shorted when the key is closed there is a possibility of damage to the receiver as a result of the amount of r-f being picked up by the receiving antenna. Also likely to occur is blocking of the receiver for an interval of several seconds after the transmission of the last character.

Receiver blocking with transmitter keying is desirable to eliminate the strong keying thump in the phones or loudspeaker. The blocking action may be accomplished by means of an external unit, or it may be accomplished by biasing off the receiver at an appropriate point in the circuit. The Collins 75A receivers have included within the detector circuit an arrangement for accomplishing the biasing-off action.

Automatic control of the high-voltage supplies of the transmitter is a desirable feature of a break-in system, especially in the case of a high-power transmitter. Through the use of such a system (illustrated in figure 11) it is necessary only to touch the key for turning on all the power supplies in the transmitter. The supplies then stay turned on as long as the operator is sending steadily. But when he pauses for a short period (the period usually is adjustable from a fraction of a second to 10 or 15 seconds) the power supplies automatically shut down.

Shorting the Receiver Antenna Terminals The BC-312 and BC-342 receivers already have built into them a relay for shorting the antenna input of the receiver as the transmitter is keyed. This relay, which is designed to operate on 12 volts d.c., may be wired into the keying circuit so that it closes whenever the key is closed. The coil of this relay may be picked up at terminal J of the large plug on the front of the receiver, and on one side of the SEND-RECEIVE toggle switch at the bottom of the front panel.

Complete Break-In Systems Shown in figures 20 and 21 are two alternative circuit arrangements for accomplishing break-in operation. The circuits are basically similar, but offer different features and are for use with different types of v-f-o arrangements. The circuit at figure 20 is best suited for use with a home-constructed v.f.o. where it is convenient to get to the grid return of the first stage following the oscillator tube. In this case the exciter blocking voltage is fed to this first r-f stage, while one of the higher level stages in the transmitter is screen keyed by the keyer unit diagrammed in figure 19.

The significant addition included in this circuit is the use of a "Millisec" relay (manufactured by Stevens-Arnold, Inc., 22 Elkins Street, South Boston, Mass.) to accomplish a number of functions in addition to keying the transmitter. The relay itself is operated directly from the key with the aid of a small "A" battery which furnishes 6 volts. The relay drain is about 40 ma. so the life of the battery should be very long. With the key

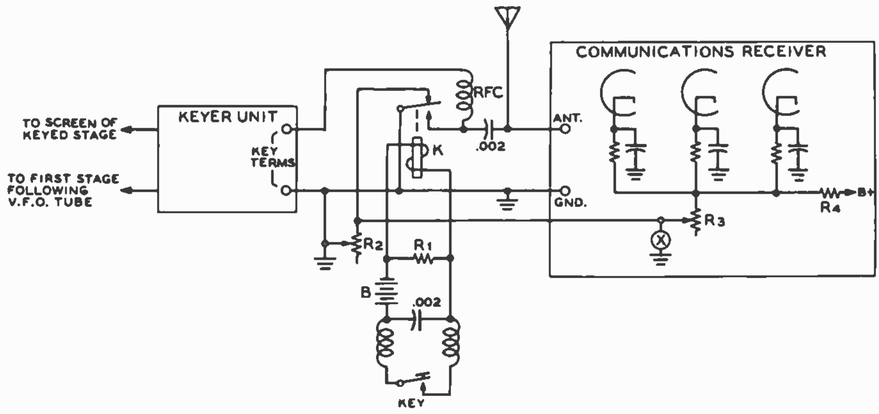


Figure 20.
SCHEMATIC OF BREAK-IN SYSTEM.

R_1—330-ohm ½-watt resistor
R_2—100,000-ohm potentiometer
R_3—Manual gain control inside receiver
R_4—Resistor to positive voltage inside receiver
K—6-volt type 172 "Millisec" relay
B—6-volt dry-cell battery
RFC—R-f choke suitable for band in use — Note that r-f chokes also are connected in series with the key terminals
(X)—Point where ground return of manual gain control inside receiver is broken for connection of lead to external circuit

open, the relay is open, and the receiver operates normally. When the key is closed a number of events occur in sequence: First, the moving contact of the relay opens the "muting" circuit of the receiver. This circuit is connected into the communications receiver merely by lifting the ground return of the r-f/i-f gain control of the set, and then bringing out the lead which originally went to ground from this point on the potentiometer. The resistor R_2, which can be mounted externally to the receiver, is then adjusted to a point which will allow the receiver gain to drop to a very low value when the "muting" circuit is in operation.

Then when the moving contact of the "Millisec" relay closes to the bottom contact (at a time about one-thousandth second after the key is closed) the receiving antenna is effectively shorted to ground. At the same time the key terminals of the keyer unit are closed. Thus it can be seen that the receiver already has been muted and its antenna shorted to ground before the emission of a signal from the transmitter. But when the key is lifted the receiver is placed into operation at about the same time that the signal from the transmitter is dying out. Hence, the use of this circuit may result in the presence of fairly strong thumps in the phones each time that the key is raised. However, by proper adjustment of the value of R_2, and sometimes by the addition of a relatively large capacitor from ground to the junction of R_2 and R_1, it usually will be possible to delay the action of the muting circuit to a point where serious keying thumps will not be heard.

The circuit of figure 21 is best suited to use with a v-f-o unit of the manufactured type where it is not practicable or desirable to cut into the v-f-o circuit. In this arrangement the key is connected to the key terminals of the keyer unit and the antenna shorting relay is operated by a vacuum tube connected to the "BREAK-IN" output signal from the keyer unit. One additional triode tube of the 6C5 or 6J5 type is required for the operation of this circuit as compared to those of figure 19 and figure 20.

The action of the circuit of figure 21 is as follows: When the key is closed the potential fed to the grid of the added 6C5 is immediately brought up to ground potential. The tube conducts quickly and closes the "Millisec" relay K. When this relay closes the receiver is muted and the antenna shorted as before. But the added circuit on the shorting relay in this case serves to close the normal keying circuit

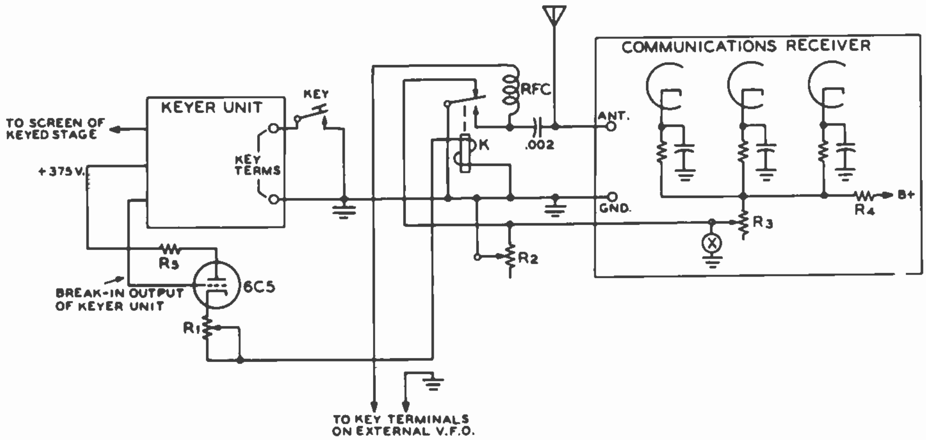


Figure 21.
ALTERNATIVE BREAK-IN SYSTEM.

R_1—10,000-ohm wire-wound potentiometer
R_2, R_3, R_4—Same as in figure 20
R_5—3300-ohm 2-watt resistor
RFC—R-f choke suitable for band in use
K—18-volt type 172 "Millisec" relay

of the external v.f.o. All this action takes place very quickly, and before the screen voltage output of the keyer unit is sufficiently high so that the keyed tube is ready to conduct. But the screen voltage output of the keyer unit does rise smoothly to the normal value so that full output from the transmitter is delivered.

When the key is lifted, the keying voltage to the keyed stage drops off normally. But due to the delay in the drop off of the voltage fed to the external 6C5, the "Millisec" relay remains closed for a moment. With rapid keying this relay will remain closed, but it will open between words if so adjusted.

The circuit of figure 21 is somewhat more smooth in operation than that of figure 20, but is more complex and hence more difficult to get into operation. The components associated with the relay K must be varied until the timing between the operation of this relay and the operation of the screen-voltage keying circuit is proper for smooth and clickless keying without objectionable thumps in the receiver.

Control of the Transmitter Plate Supplies The operation of the circuits of figures 19, 20, and 21 is based on the assumption that the transmitter plate supplies will be turned on and off manually. A key switch on the operating desk may be closed to turn on the supplies whenever transmission is started, and the plate supplies then may be turned off when a transmission is completed. But for rapid back-and-forth break-in, using the systems mentioned, it will be necessary to leave the transmitter plate supplies on substantially all the time. This is a reasonable method of operation in certain cases, especially with a transmitter of moderate power. But in the case of a high-power transmitter the bleeder power alone on the high-voltage supplies becomes an appreciable power drain. Hence it becomes desirable not only that the plate supplies be turned on as soon as transmission is begun, but that these supplies turn themselves off automatically as soon as a certain time interval has elapsed after the transmission of the last character.

One such circuit for accomplishing automatic control of the transmitter plate supplies was illustrated in figure 11 of this chapter. An alternative circuit is given in figure 22. This circuit is identically the same as that of figure 21 except that an additional 12AU7 tube and a sensitive relay have been added. The operation of the 12AU7, which is connected as a diode feeding a triode, is the same as the operation of the circuit which operates the "Millisec" relay in figure 21. However, the time constant has been increased considerably

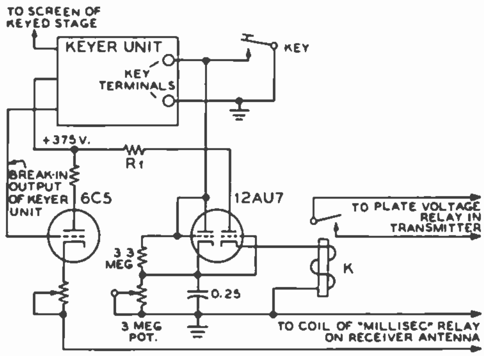


Figure 22.

ALTERNATIVE AUTOMATIC-POWER-CONTROL CIRCUIT.

This circuit is a modification of the circuit of figure 21, with the automatic-power-control circuit added. All circuit details not given are the same as in figure 21.

R_1—4700-ohm 2-watt resistor. Some variation in the value of this resistor will be required to obtain proper timing of the action of the power-control relay, K. Action of this relay may be speeded up somewhat, if required, by placing a 20-µfd. 250-volt electrolytic capacitor across resistor R_1.

K—2500-ohm s.p.s.t. relay.

so that the transmitter plate supplies may remain turned on during a normal pause in break-in operation of 5 to 10 seconds.

In the case of any of these circuits for automatically controlling the transmitter plate supplies, there is a certain amount of delay from the first closing of the key until the voltages being delivered by the supplies are high enough for normal operation of the transmitter. This means that if one merely starts keying without waiting for the plate supplies to become fully operative, the first dot or a portion of the first dash will not be transmitted. A very simple method for overcoming this condition is available. Simply tap the key for a very short dot an instant before you begin keying. This short dot due to the quick tap on the key will not be transmitted; but an instant later you may begin keying normally since all the control relays will have been operated by the first dot. This quick tap before you begin keying soon will become automatic in your operating. The timing is not at all difficult, since the control relays will be heard to close following the short tap, giving the signal that the transmitter is ready. Similarly, the transmitter control relays will be heard to drop out after a period of transmission. It is then obvious and more or less automatic to give the key a tap for closing the relays before actual keying of the transmitter is begun.

Receiving Antennas for Break-In One of the problems associated with break-in operation is the matter of the receiving antenna. For break-in operation on the lower frequency bands, where directive antenna arrays are not normally used, the normal procedure is to use a separate antenna for receiving. The receiving antenna usually is somewhat smaller than the transmitting antenna and may consist of a straight piece of wire located as far as possible from the transmitting antenna. Signal strengths on the lower frequencies usually are great enough so that a smaller and less efficient antenna than that used for transmitting may be used for receiving.

But for operation on the 14-Mc. band and above, it usually becomes desirable to use the transmitting antenna also for receiving, even for break-in operation. This becomes almost a necessity when a directive antenna array is used for transmitting. One satisfactory system for accomplishing the changeover is to tie the operation of the antenna changeover relay to the operation of the "Millisec" relay at the antenna posts of the receiver. A procedure such as this will involve a lot of clatter at the antenna changeover relay unless the time constants of the circuits accomplishing operation of the changeover relay are carefully controlled. But in any event it is necessary that the antenna changeover relay close quickly when the key is closed so that no sparking will take place at the changeover relay as a result of the power emitted by the transmitter. The opening of the changeover relay should be timed so that it will remain closed for a word, but will open between words and between sentences. Solenoid-type changeover relays are not rapid enough in their action, and are too noisy, for use with break-in; however, the simpler clapper-type of relay has proven to be rapid enough.

An alternative type of control for the changeover relay, which has proven satisfactory for the dx type of break-in operation or for traffic break-in where a delay in breaking of a few seconds may be tolerated on occasion, is to tie the changeover relay into the operation of the high-voltage plate supplies. This

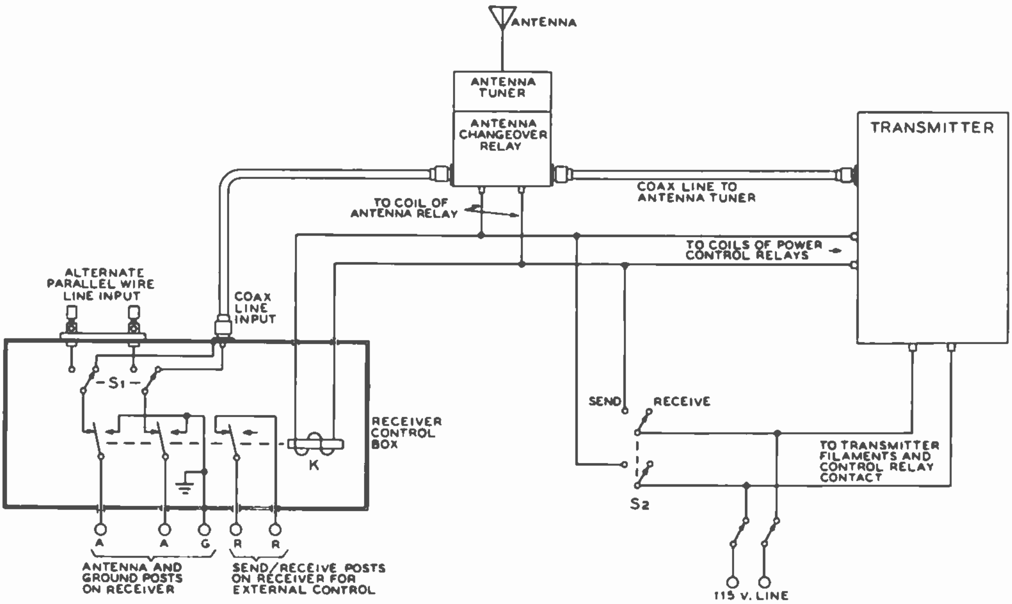


Figure 23.
SEND-RECEIVE CIRCUIT FOR PHONE OPERATION.
Switch S_1 changes the receiver from two-wire line to coaxial input. The relay K is a standard ceramic insulated antenna-changeover relay with a third set of contacts on the center of the relay structure for receiver control.

would require a control circuit such as figure 11 or figure 22, with the changeover relay coil connected across the primary of one of the high-voltage transformers.

When using the above system of break-in with a strong received signal, a breaking signal usually can be heard on the leads from the receiver to the changeover relay, even though the antenna is connected to the transmitter. When operating with a relatively weak received signal, it will be necessary to pause periodically to listen for a breaking signal, thus allowing the plate supplies to shut down and the changeover relay to connect the antenna to the receiver. It has been found desirable in certain cases to connect a switch in series with a power supply control circuit such as figure 11 or figure 22, so that the power supplies may *manually* be shut down briefly and the antenna changed over for a short listening period.

Antenna and Transmitter Control for Phone Operation

Control systems for the antenna, for the transmitter, and for the receiver at the station using mostly phone operation or using c.w. without break-in, are considerably simpler than those for break-in c-w operation. In the main, the conditions of "transmit" and "receive" are established in nearly all cases by the operation of a single switch which may have either two or three positions. When a third position is used it usually is for the position where both the transmitter and the receiver are inoperative.

A more or less standard control circuit for transmitter, receiver, and antenna changeover relay is diagrammed in figure 23. A minimum of three relays are required for this system. These relays are: the antenna changeover relay, the main power control relay in the transmitter, and the receiver control relay. When the control switch S_2 is placed in the transmit position, all three relays close more or less simultaneously. The antenna changeover relay disconnects the antenna from the receiver and connects it to the transmitter; the receiver control relay disables the receiver and shorts the antenna input posts; and the transmitter power control relay or relays turn on the exciter and apply plate voltage to the transmitter.

The Receiver Control Box may best be placed at the rear of the receiver, and may conveniently be grounded to the receiver chassis. An alternate switch has been shown as a portion of the receiver control box for changing the connections inside the box for operation of the receiver from a two-wire line instead of operation from coaxial-cable input.

In many installations it may be found that when the SEND-RECEIVE switch is changed to the RECEIVE position, the receiver will be blocked for an instant by the signal emitted from the transmitter. This condition is the result of the energy stored in the filter capacitors of the transmiter. These capaciors frequently will hold enough charge so that the transmitter will continue to emit a signal for a moment after the switch is placed in the receive position. Since the receiver input is opened and the receiver turned on at the same time that the transmitter is turned off, this last bit of signal from the transmitter may block the receiver.

If the blocking condition is serious enough to cause trouble, the transmitter-receiver-antenna control circuit may be changed over to that illustrated in figure 24. In this arrangement a three-position switch is used for the SEND-RECEIVE control. In one position the transmitter is turned on and the antenna changeover relay is in the transmit position. In the center position both the transmitter and the receiver are inoperative. And in the third position the receiver is turned on. If the control switch is consciously stopped in the center position when going from TRANSMIT to RECEIVE, the blocking condition in the receiver can be eliminated. The receiver input remains shorted, and the receiver is in the inoperative condition until the control switch is placed in the RECEIVE position. In fact, there are switches available on the market which have a stop in the center position of such a nature that it is necessary to press the switch lever twice to go from one extreme position to the other extreme. The switch is pressed once to go from the SEND to the neutral position, then the force is released, then the switch is pressed once more to go from the neutral to the RECEIVE position. The use of a switch of this type will insure that an instant's time delay is allowed, even by a guest operator, in going from SEND to RECEIVE.

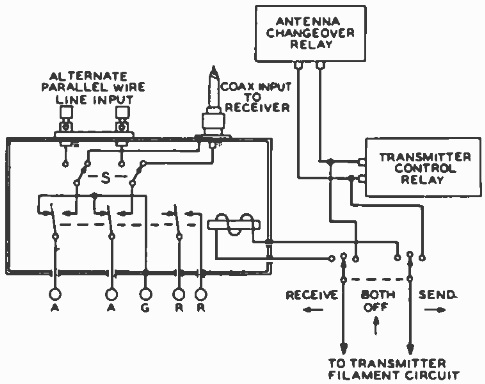


Figure 24.
SEND-RECEIVE CIRCUIT WITH CENTER OFF POSITION.
This circuit provides an off position on the transmit-receive switch for disabling both the transmitter and the receiver. The balance of the circuit is the same as that of figure 23.

CHAPTER ELEVEN

Transmitter Adjustment and Loading

While there are as many different tuning procedures as there are types of transmitters, there are certain general rules which should be followed regardless of the type of transmitter. Also, there are certain initial checks that should be made on a new transmitter regardless of the type. A sequence of ten such checks is given in the following section.

11-1 Initial Transmitter Tune-Up

In making the initial adjustments upon a new transmitter it is recommended that an orderly procedure be followed in checking out first the simpler circuits and then proceeding to the more complex circuits so that it will be known, as soon as difficulties are encountered, that all the simpler circuits are operating properly. It is suggested that the following steps be followed, wherein they apply, in checking a transmitter for proper operation the first time it is tuned up, or for that matter, the first time it is operated after a long period of inactivity.

1. Check Filament and Heater Voltages Apply line voltage to the transmitter and make sure that all the plate supplies and bias power supplies are completely disconnected from the line. Check the filament or heater voltage of each tube in the transmitter with a voltmeter whose accuracy has been checked. When checking a-c voltages it is best to use an instrument of the iron-vane type since rectifier-type instruments often drift far out of calibration. All filament and heater voltages should be checked directly at the *socket* terminals.

It is best to adjust filament-transformer primary taps and drop resistors (if used) so that the filament voltage actually at the socket is about 5 per cent *above* the rated value. This procedure is recommended since the line voltage to a transmitter will usually drop from 3 to 8 per cent when plate power is applied — unless, of course, a special low-drop line has been run in to operate the transmitter.

Heater-cathode tubes such as the 6L6, 807, 2E26 and similar types can be operated over a plus or minus 10 per cent range in heater voltage above and below the rated value. However, it is best to operate tubes of this type at least within 5 per cent of the rated voltage. Directly-heated or filament tubes such as the great majority of transmitting tubes and rectifiers can be operated 5 per cent above or below rated voltage but it is best to operate these types within a few per cent of the rated value. This applies particularly to low-voltage high-current tubes such as the Eimac and HK types and the RCA 806. This limitation in voltage tolerance applies also to rectifier tubes such as the 816, 866A/866, and 872A/872.

2. Check Control Circuits, Plate and Bias Supplies All control circuits, time-delay relays, interlocks, and overload provisions should be checked to make sure that all switches control the circuits they were intended to control, and that the transmitter is safe to operate with the control provisions that have been made.

No-load voltages of the various plate and bias supplies should be checked with a high-resistance voltmeter to make sure that these voltages are within the expected range, and are not excessive for the capacitors being used. No-load voltages on plate supplies having a capacitor-input filter can be expected to be from 40 to 50 per cent high. No-load voltage output of a supply having a choke-input filter, however, should be not more than about 10 or 15 per cent above the expected operating value. If the voltage output of a choke-input power supply *does* soar, it means that the maximum inductance of the input swinging choke is not sufficiently high or that the value of bleeder resistance used is too great. If the voltage output of a choke-input supply is excessive with no load the input inductance must be increased to the critical value of R/1000 henries, where R is the value of the bleeder resistor. Alternatively, the bleeder resistor may be reduced to a value of resistance such that the above expression is satisfied with the input choke in use. The inductance of a swinging choke can often be increased by removing some of the material in the air gap of the core. This subject is covered in Chapter Twenty-five, *Power Supplies*.

The easiest way to make a first check on the presence of a bleeder of correct value on a power supply is to remove completely the power plug going to the supply, short one of the filter capacitors with an insulated screwdriver, remove all output leads from the power supply, remove the screwdriver, and check for resistance with an ohmmeter across one of the filter capacitors. The resistance value noted should be the value of the bleeder resistor across the power supply.

Another method of checking bleeder operation is to connect a high-resistance voltmeter to the supply, apply plate voltage for a moment, and then note the rate at which the voltage decreases to zero. If the voltage falls very gradually after the plate transformer has been removed from the line, *beware*, and wait for all the charge to drain from the capacitors, as indicated by the voltmeter indication dropping to zero, before working on the supply. Then short the filter capacitors with a screwdriver, and apply a clip lead across one of the filter capacitors before working on the rewiring. *Do not* attempt to drain the charge on the filter capacitors of a high-voltage supply with a defective bleeder with a screwdriver — the resulting outrush of current will certainly damage the screwdriver, the procedure is very dangerous, and the filter capacitors may be damaged.

3. Check First R-F Stage in Transmitter The first r-f stage in a transmitter will normally be either a crystal oscillator stage or the first tuned stage following the v-f-o. The operation of a crystal oscillator depends to a great extent upon the activity of the crystal, and the activity varies widely with different crystals. The oscillator should be tuned for the greatest output or lowest plate current which will provide strong, stable oscillations. An attempt to adjust the oscillator for every last milliwatt of output will result in the crystal's not starting "cleanly" each time the plate voltage is applied or the key is pressed. A receiver or monitor will be required for this check, during which a check also should be made on the frequency.

The first time the crystal oscillator is operated a check also should be made upon the r-f crystal current (unless the oscillator is run at very low screen and plate voltages) to make sure that it is not excessive at any setting of the plate tuning capacitor.

If the first r-f stage in the transmitter is fed from a v.f.o. the stage should be excited from the v.f.o. and the plate tank tuned to resonance with the aid of a dial lamp and a loop of wire, or by the grid current indication on the next stage. A wavemeter should be used to make sure that the stage is being tuned to the correct frequency. The plate tank of the stage should then be tuned through its full range while listening on a receiver to make sure that the stage does not self-oscillate at any tank capacitor setting.

4. Tune Successive Stages to Resonance Each stage following the crystal oscillator or v-f-o amplifier should then be tuned to resonance with the aid of a wavemeter and the grid and plate milliammeters on the stages. If there are any neutralized stages in the chain these should be neutralized in accordance with the procedure given in Chapter Seven, *Generation of R-F Energy*. Make sure that adequate grid current is obtainable on the final amplifier stage on each of the bands on which the exciter has been designed to operate. The operating currents and voltages on grids, plates, and screens of each of the exciter stages should be measured to make sure that they are within the rated values for the tube types concerned. If they are not, alter the resistor values in the various feed circuits until voltages and currents are within ratings.

5. Neutralize Final Amplifier The recommended procedure for amplifier neutralization has been covered in Chapter Seven. Triode amplifier stages will always require neutralization unless the grounded-grid or cathode-follower circuit has been used. Even then the stage *may* require neutralization. Beam-tetrode and pentode stages will frequently require neutralization, although neutralization is less frequently required with pentodes than with tetrodes. A very small value of neutralizing capacitance will be required when neutralizing tetrodes or pentodes.

6. Operate Final Amplifier into Dummy Load Apply reduced plate voltage to the final amplifier and couple the stage to a dummy load. Reduced plate voltage may be obtained from the final plate supply by connecting the primary of the transformer for half voltage if the taps are available, it may be obtained by running the final stage from one of the buffer plate supplies, or reduced voltage may be obtained by connecting a lamp or screw-base heater element in series with the primary of the plate transformer.

The dummy load may consist of Ohmite Dummy Antenna Resistors, Sprague non-inductive resistors of appropriate resistance and wattage, or it may consist simply of a number of ordinary 115-volt lamps.

When using ordinary 115-volt lamps it has been found wise to use somewhat more lamp wattage than the expected output of the transmitter. If this is not done, difficulty may be encountered in the case of high-power transmitters with dielectric breakdown in the base or the stem of the lamp. A bank of nine 100-watt lamps connected three in series and three groups in parallel has been found to operate satisfactorily on a 1-kilowatt transmitter. But a single 1000-watt lamp will break down in the base after a short period of operation at the higher frequencies.

The coupling to the dummy load should be adjusted so that the plate current to the stage bears the same ratio to the desired plate current as does the test plate voltage to the plate voltage at which the stage is to be operated (assuming that screen voltage is varied along with plate voltage if tetrodes or pentodes are being used). When this adjustment has been made, full plate voltage may be applied, *carefully*, to the final amplifier stage. When all stages seem to be operating correctly, the transmitter may be checked for parasitic oscillations.

7. Checking for Parasitic Oscillations It is an unusual transmitter which harbors no parasitic oscillations when first constructed and tested. Hence it is always wise to follow a definite procedure in checking a new transmitter for parasitic oscillations.

Parasitic oscillations (as distinguished from self-oscillation on the tuned frequency of the amplifier) ordinarily occur in two types: low-frequency parasitics from 20 to 200 or 300 kc., and high-frequency parasitics from 40 to 200 Mc. Low-frequency parasitics can easily be detected by the fact that they will modulate the carrier frequency of the transmitter producing strong sidebands, usually of rough tone, on either side of the carrier frequency and spaced from it and from each other by the parasitic oscillation frequency. Thus if the transmitter carrier is on 14.1 Mc. and the parasitic oscillation frequency is 100 kc. the spurious sidebands will be heard at 13.9, 14.0, 14.2, 14.3, 14.4 Mc. and so on on either side of the carrier.

High-frequency or v-h-f parasitics on the other hand usually cause a roughening of the carrier signal of the transmitter but must be tuned in on a v-h-f receiver to be heard. A systematic procedure for determining the presence or absence of parasitic oscillations is given in the following paragraphs.

a. Tune a communications receiver about 20 kc. or 30 kc. to one side or the other of the carrier frequency of the transmitter.

b. Apply plate voltage to the stage being checked and detune first the plate tank and then the grid tank as far either side of resonance as can be done without exceeding the plate dissipation rating of the tube or tubes in the stage. It is wise to have a resistor in series with the primary of the plate transformer of a high-power amplifier stage so that the plate voltage will drop when the plate current increases.

c. If there are no sudden jumps in either the grid current or the plate current of the stage, and if no spurious signals can be heard with any tuning adjustment of the stage on the receiver when tuned from the carrier to several hundred kilocycles either side, it can be assumed that there are no low-frequency parasitics present. Also, if the plate and grid currents behave in an orderly manner it is probable that v-h-f parasitics are not present. However, it is still wise to listen on a v-h-f receiver covering the range from 28 to perhaps 150 Mc. (if such a receiver is available) to make sure that parasitics in this frequency range are not present. Parasitic oscillations can almost always be detected on a v-h-f receiver by the fact that their tone will be rough and unstable as compared to the clean and stable tone that normal harmonics of the carrier frequency will have.

If parasitic oscillations are found they should be eliminated by the procedure discussed in Section 11-2 of this chapter. If parasitic oscillations are not found, or after they have been eliminated, it is then possible to apply full power to the transmitter and check its modulation or keying while operating into the dummy load.

8. Check Modulation or Keying The transmitter should now be operated at full power and modulated or keyed in the manner normally to be employed. Again it is convenient to check for key clicks or spurious sidebands under modulation by means of a communications receiver tuned either to one side or the other of the carrier frequency of the transmitter. Further check for parasitic oscillations under modulation or keying should be made.

9. Checking on Other Bands After the transmitter has been checked out completely with the energy fed into a dummy antenna on one band, the dial settings for all the tuning controls should be noted and the transmitter shifted to another frequency in another band on which it is desired to operate. The procedure given before should be followed and when satisfactory operation has been obtained on that band the dial settings should again be noted and operation shifted to another band. After it has been determined that satisfactory operation can be obtained on all the bands in which it is desired to operate the transmitter it is well to give the equipment a heat run of moderate duration.

10. Making a Heat Run on the Equipment It is always wise to make a short life test or heat run on a new transmitter to make sure that the operating conditions of the transmitter will remain stable over a period of time and to determine whether or not any components are experiencing excessive heating as a result of their normal operation. For the first test the transmitter should be run for perhaps 10 minutes into a dummy load — with key down if the rig is a c-w transmitter and with about 60 per cent sine-wave modulation if the transmitter is to be used with amplitude modulation. After the first 10-minute test the transmitter should be shut down, the connection to the 115-volt line completely severed, the filter capacitors shorted with a screwdriver, and the various components felt cautiously with the hand to determine whether or not excessive heating has taken place. The bleeder resistors should be quite warm (if they are not it is possible that they are open) but the other components except for occasional resistors should not be too hot to touch.

If the equipment passes the 10-minute run without trouble the equipment should be operated for about 30 minutes and the check again made. After this test the equipment

HARMONIC FREQUENCY CHART

All Frequencies in Megacycles

ELEVEN-METER BAND

6.74	×	4	=	26.960
6.75	×	4	=	27.000
6.775	×	4	=	27.100
6.8	×	4	=	27.200
6.8075	×	4	=	27.230

TEN-METER BAND

3.500	×	8	=	28.0
3.53125	×	8	=	28.25
3.5625	×	8	=	28.5
3.59395	×	8	=	28.75
3.625	×	8	=	29.0
3.65625	×	8	=	29.25
3.6875	×	8	=	29.5
3.7125	×	8	=	29.7
7.0	×	4	=	28.0
7.0625	×	4	=	28.25
7.125	×	4	=	28.5
7.1879	×	4	=	28.75
7.25	×	4	=	29.0
7.3125	×	4	=	29.25
7.375	×	4	=	29.5
7.425	×	4	=	29.7

SIX-METER BAND

3.125	×	16	=	50.0
3.15625	×	16	=	50.5
3.1875	×	16	=	51.0
3.21875	×	16	=	51.5
3.250	×	16	=	52.0
3.28125	×	16	=	52.5
3.3125	×	16	=	53.0
3.34375	×	16	=	53.5
3.375	×	16	=	54.0
6.250	×	8	=	50.0
6.3125	×	8	=	50.5
6.375	×	8	=	51.0
6.4375	×	8	=	51.5
6.50	×	8	=	52.0
6.5625	×	8	=	52.5
6.625	×	8	=	53.0
6.6875	×	8	=	53.5
6.750	×	8	=	54.0
8.33333	×	6	=	50.0
8.41666	×	6	=	50.5
8.50	×	6	=	51.0
8.58333	×	6	=	51.5
8.70	×	6	=	52.0
8.75	×	6	=	52.5
8.8333	×	6	=	53.0
8.9166	×	6	=	53.5
9.0	×	6	=	54.0

TWO-METER BAND

9.0	×	16	=	144
9.0625	×	16	=	145
9.125	×	16	=	146
9.1875	×	16	=	147
9.250	×	16	=	148
8.0	×	18	=	144
8.0555	×	18	=	145
8.1111	×	18	=	146
8.1666	×	18	=	147
8.2222	×	18	=	148
6.0	×	24	=	144
6.04166	×	24	=	145
6.08333	×	24	=	146
6.125	×	24	=	147
6.16666	×	24	=	148

should be left on for about 1 hour. Many components designed for amateur transmitters (particularly inexpensive plate transformers and chokes) are designed for not more than about 1 hour continuous duty. This design limitation is usually satisfactory for amateur operation since an amateur transmitter is almost invariably standing by a somewhat greater percentage of the time than it is transmitting. Most components of the size usually employed in amateur transmitters will reach their ultimate temperature after a continuous run of approximately one hour. However, if any components show signs of dangerous heating (but not actually excessive) it is probably best to operate the transmitter again for an additional half-hour period to determine whether the component will remain in operation. It is better to find that a component is inadequate while testing a transmitter than to have it break down while the equipment is being used in an emergency.

Transformers and chokes are usually operating satisfactorily insofar as heating is concerned if they are a little too hot to hold the hand in contact with after such a period of operation. Filter capacitors should remain as cool as the ambient temperature of the air within the transmitter enclosure. Mica capacitors may run warm but certainly not far above body temperature.

11-2 Elimination of Parasitic Oscillations

Parasitic oscillations in transmitter stages are of such varying types, frequencies, and amplitudes that it is very difficult to set forth a definite procedure to be followed in attempting to eliminate them. However, a number of general suggestions will be made, and the applicability to the particular case at hand will have to be determined by the person who has tackled the task of parasitic elimination.

It may be said in general that the presence of low-frequency parasitics indicates that somewhere in the oscillating circuit there is an impedance which is high at a frequency in the upper audio or low r-f range. This impedance may include one or more r-f chokes of the conventional variety, power supply chokes, modulation components, or the high impedance may be presented simply by an RC circuit such as might be found in the screen-feed circuit of a beam-tetrode amplifier stage. The presence of low-frequency parasitics is easily determined by the method discussed in paragraph 7 of Section 11-1 earlier in this chapter.

The most usual source of low-frequency parasitics is the presence of an r-f choke in both the grid circuit and plate circuit of an amplifier. Hence, if such parasitics are encountered it is best to replace first the grid r-f choke by a resistor (or by a tuned circuit if the stage is shunt fed) and then the plate r-f choke by a resistor, checking in each case on both sides of the carrier for some distance to determine whether or not the parasitic has been eliminated. If this expedient does not eliminate the condition, and the stage under investigation uses a beam-tetrode tube, negative resistance can exist in the screen circuit of such tubes. Try larger and smaller screen by-pass capacitors to determine whether or not they have any effect. If the condition is coming from the screen circuit an audio choke with a resistor across it in series with the screen feed lead will often eliminate the trouble.

Low-frequency parasitic oscillations can ofter take place in the audio system of an AM transmitter, and their presence will not be known until the transmitter is checked on a receiver. It is easy to determine whether or not the oscillations are coming from the modulator simply by switching off the modulator tubes. If the oscillations are coming from the modulator, the stage in which they are being generated can be determined by removing tubes successively, starting with the first speech amplification stage, until the oscillation stops. When the stage has been found, remedial steps can be taken on that stage.

If the stage causing the oscillation is a low-level speech stage it is possible that the trouble is coming from r-f or power-supply feedback, or it may be coming about as a result of inductive coupling between two transformers. If the oscillation is taking place in a high-level audio stage, it is possible that inductive or capacitive coupling is taking place back to one of the low-level speech stages. It is also possible, in certain cases, that parasitic push-pull oscillation can take place in a Class B or Class AB modulator as a result of the grid-to-plate capacitance within the tubes and in the stage wiring. This condition is more likely to occur if capacitors have been placed across the secondary of the driver transformer and across the primary of the modulation transformer to act in the reduction of the amplitude of the higher audio frequencies. Relocation of wiring or actual neutralization of the audio stage in the manner used for r-f stages may be required.

In general, however, low-frequency parasitics are comparatively easy to find and easy to eliminate since their frequency is most frequently far removed from the carrier frequency of the transmitter and from any possible modulation frequency.

Elimination of V-H-F Parasitic Oscillations V-h-f parasitic oscillations are often difficult to locate and quite difficult to eliminate since their frequency often is only moderately above the desired frequency of operation. But it may be said that v-h-f parasitics always may be eliminated if the operating frequency is appreciably below the upper frequency limit for the tubes used in the stage. However, the elimination of a persistent parasitic oscillation on a frequency only moderately higher than the desired operating frequency will involve a sacrifice in either the power output or the power sensitivity of the stage, or in both.

Beam-tetrode stages, particularly those using 807 type tubes, will almost invariably have one or more v-h-f parasitic oscillations unless adequate precautions have been taken in advance. Many of the units described in the constructional section of this edition had parasitic oscillations when first constructed. But these oscillations were eliminated in each case; hence, the expedients used in these equipments should be studied.

Parasitic Oscillations with Triodes Triode stages are less subject to parasitic oscillations primarily because of the much lower power sensitivity of such tubes as compared to heam tetrodes. But such oscillations can and do take place. Usually, however, it is not necessary to incorporate losser resistors as normally is the case with beam tetrodes, unless the triodes are operated quite near to their upper frequency limit, or the tubes are characterized by a relatively high transconductance. Triode v-h-f parasitic oscillations normally may be eliminated by adjustment of the lengths and effective inductance of the leads to the elements of the tubes.

In the case of triodes, v-h-f parasitic oscillations often come about as a result of inductance in the neutralizing leads. This is particularly true in the case of push-pull amplifiers. The cure for this effect will usually be found in reducing the length of the neutralizing leads and increasing their diameter. Both the reduction in length and increase in diameter will reduce the inductance of the leads and tend to raise the parasitic oscillation frequency until it is out of the range at which the tubes will oscillate. The use of straightforward circuit design with short leads will assist in forestalling this trouble at the outset. Butterfly-type tank capacitors with the neutralizing capacitors built into the unit (such as the B&W type) are effective in this regard.

V-h-f parasitic oscillations may take place as a result of inadequate by-passing or long by-pass leads in the filament, grid-return and plate-return circuits. Such oscillations also can take place when long leads exist between the grids and the grid tuning capacitor or between the plates and the plate tuning capacitor. The grid and plate leads should be kept short, but the leads from the tuning capacitors to the tank coils can be of any reasonable length insofar as parasitic oscillations are concerned. In an amplifier where oscillations have been traced to the grid or plate leads, their elimination can often be effected by making the grid leads much longer than the plate leads or vice versa. Sometimes parasitic oscillations can be eliminated by using iron or nichrome wire for the grid or plate leads, or for the neutralizing leads. But in any event it will always be found best to make the neutralizing leads as short and of as heavy conductor as is practicable.

Small v-h-f tank circuits, consisting of a few turns of heavy wire tuned by an APC capacitor, connected in series with the grid leads of an amplifier sometimes will effect a cure. This expedient is not desirable, however, since such circuits may do an adequate job when the amplifier is operated only over a comparatively narrow frequency range.

In cases where it has been found that increased length in the grid leads for an amplifier is required, this increased length can often be wound into the form of a small coil and still obtain the desired effect. Winding these small coils of iron or nichrome wire may sometimes be of assistance.

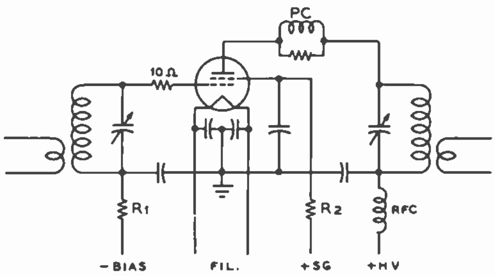


Figure 1.
BEAM-TETRODE PARASITIC SUPPRESSION.

Showing the basic circuit for parasitic suppression in beam-tetrode r-f amplifier stages. The grid and plate parasitic suppressors are discussed in the text. Resistor R_1 is the conventional grid leak for the stage; R_2 is the current-limiting resistor for the screen circuit.

Parasitics with Beam Tetrodes Where beam-tetrode tubes are used in the stage which has been found to be generating the parasitc oscillation, all the foregoing suggestions (except those specifically related to neutralizing-lead inductance) apply in general. However, there are certain additional considerations involved in elimination of parasitics from beam-tetrode amplifier stages. These considerations involve the facts that a beam-tetrode amplifier stage has greater power sensitivity than an equivalent triode amplifier, such a stage has a certain amount of screen-lead inductance which may give rise to trouble, and such stages have a small amount of feedback capacitance.

Beam-tetrode stages often will require the inclusion of a neutralizing circuit to eliminate oscillation on the operating frequency. Such neutralizing circuits are discussed in detail in Chapter Seven. However, oscillation on the operating frequency normally is not called a parasitic oscillation, and different measures are required to eliminate the condition.

Basically, parasitic oscillations in beam-tetrode amplifier stages fall into two classes: cathode-grid-screen oscillations, and cathode-screen-plate oscillations. Both these types of oscillation *can* be eliminated through the use of a resistor in the lead between the screen by-pass capacitor and the screen lead. But the inclusion of such a resistor will result in a large energy loss and loss in power sensitivity at the operating frequency, and probably will result in cathode-grid-plate oscillations. So the inclusion of a resistor in series with the screen lead of the tube is *not* recommended.

However, the grid-screen oscillations may be eliminated through the use of a resistor in series with the grid of the tube. And the screen-plate oscillations may be eliminated through the use of a resistor in series with the plate lead of the tube. These measures will *always* work. But, the resistor in series with the grid will require an increase in driving power on the higher frequency bands, so a compromise value must be chosen. Normally, a resistor between 10 and 22 ohms (make sure that this resistor is of the carbon type such as an Ohmite or Allen-Bradley) will effectively suppress parasitic oscillations of the grid-screen variety in all the common types of beam-tetrode tubes. The driving power will be increased for frequencies above about 15 Mc., but not to an extent which will require an increase over the normal reserve capability included in the driver for a beam-tetrode stage. Also, the dissipation in this resistor due to loss of driving power will be appreciable at frequencies above perhaps 25 Mc. So the grid losser resistor should be checked for excessive heating from excitation energy.

Screen-plate oscillations always may be eliminated through the use of a resistor in series with the plate lead. However, the power loss in this resistor will become excessive at a frequency near the upper frequency limit of the tube. The power loss in the medium-frequency range may be reduced greatly through paralleling the plate losser resistor with a small inductor. This is standard practice with commercially manufactured parasitic suppressors.

The parasitic suppressor for the plate circuit of a small tube such as an 807, 2E22, 2E26 or similar type normally may consist of a 47-ohm *carbon* resistor of the 2-watt size with 6 turns of No. 18 or No. 20 tinned or bare wire wound around the resistor. This type of parasitic suppressor normally will run cool on the frequency ranges below 15 Mc., and will be suitable for operation up to about 30 Mc. without excessive heating. However, for operation on frequencies of 50 Mc. and above, special tailoring of the value of the resistor and the size of the coil wound around it will be required in order to attain satisfactory parasitic suppression without excessive power loss in the parasitic suppressor.

Parasitic suppressors for larger tubes such as the 4-65A, 4E27 series, 813, 4-125A, 4-250A and such types normally will be required to have greater dissipation than can be handled by receiver-type carbon resistors. Commercial transmitters normally use large carbon or "Globar" resistors; when these are available they are ideal. But Sprague *non-inductive* "Koolohm" resistors have been found quite satisfactory, and are generally available. The resistance value usually will fall between 5 and 50 ohms, and the shunting coil may have 3 to 6 turns with a diameter about twice that of the resistor which it surrounds.

11-3 Adjustment of Class B, Class C, and FM Amplifiers

The excitation, tuning, and loading of Class C, FM, and Class B amplifiers of the type where linearity is not important, is essentially the same. The adjustment of Class B linear amplifiers is discussed in Section 11-4.

Plate Circuit Tuning After an amplifier is completely neutralized, reduced plate voltage should be applied before any load is coupled to the amplifier. This reduction in plate voltage should be at least 50 per cent of normal value, because the plate current will rise to excessive values when the plate tuning capacitor is not adjusted to the point of resonance as indicated by the greatest dip in reading of the d-c plate current milliammeter.

With no load, the r-f voltage may be several times as high as when operating under conditions of full load; this may result in capacitor flashover if normal d-c voltage is applied. The no-load plate current at resonance should dip to roughly 15 per cent of normal value. If the plate circuit losses are excessive, or if parasitic oscillations are taking place, the no-load plate current will be higher.

Loading The load (antenna or succeeding r-f stage) then can be coupled to the amplifier under test. The coupling can be increased until the plate current at resonance (greatest dip in plate current meter reading) approaches the normal value for which the tube is rated. The value at reduced plate voltage should be proportionately less in order to prevent excessive plate current when normal plate voltage is applied. Full plate voltage should not be applied to an amplifier unless the r-f load also is connected; otherwise the tuning capacitors may arc or flash over, thereby causing an abnormally high plate current which may damage the tube. The tuned circuit impedance is lowered when the amplifier is loaded, as are the r-f voltages across the plate and neutralizing capacitors.

Grid Excitation Excessive grid excitation is just as injurious to a vacuum tube as abnormal plate current or low filament voltage. Too much grid driving power will overheat the grid wires in the tube, and will cause a release of gas in certain types of tubes. An excess of grid drive will not appreciably increase the power output and can increase the efficiency only slightly. The grid current in the tube should not exceed the values listed for the type of tube being used, and care also should be exercised to have the bias voltage low enough to prevent flashover in the stem of the vacuum tube.

Grid excitation usually refers to the actual r-f power input to the grid circuit of the vacuum tube, part of which is used to drive the tube, and part of which is lost in the grid-bias supply. There is no way to avoid wasting a portion of the excitation power in the bias supply. The loss is the same with battery bias as with grid-leak bias or power-supply bias.

It is natural that the grid current to an amplifier stage should fall off somewhat with application of plate voltage, the drop in grid current becoming greater as the loading and plate current on the stage are increased. If the excitation is adjusted for maximum permissible grid current with the tubes loaded, this value will be exceeded when the plate voltage or load is removed, particularly when no grid-leak bias is employed. However, under these conditions, the grid impedance drops to such a low value that the high value of grid current represents but little increase in power, and there is little likelihood that the tube will be damaged unless the grid current increases to more than twice its rated maximum for normal (loaded) operation.

Tuning Under Load Amplifier stages always should be tuned for maximum output. This does not mean that the coupling must be adjusted until the stage will deliver the maximum power of which it is capable, but that the tank tuning capacitor always should be adjusted to the setting which permits maximum output. If the stage is not heavily loaded, this will correspond closely to minimum plate current. However, if the two do not correspond exactly, the stage should be tuned for maximum output rather than minimum plate current. If the difference is appreciable, especially in that amplifier which feeds the antenna, the amplifier should be redesigned to utilize a higher value of tank capacitance.

Screen-grid tubes never should be operated with full screen voltage when the plate voltage is removed, as the screen dissipation will become excessive and the tube may be permanently damaged. Neither should screen-grid and beam-tetrode amplifier stages be operated with plate voltage applied and without load since the screen current also will become very high under such conditions unless some provision has been made, such as a resistor in the screen series circuit, to limit the screen current.

When all stages are operating properly, the filament voltage on the final amplifier tubes should be checked with the transmitter operating to make sure that it is neither excessive nor deficient, one being about as bad as the other. Unless the line voltage varies at least several volts throughout the day, filament meters are not required on all stages of a multi-stage transmitter. An initial check when the transmitter is put into operation for the first time is sufficient; after that a single filament meter permanently wired across the filament or filaments of the final amplifier stage will be sufficient. If the filament voltage reads high on that stage, it can be assumed to be high on all stages if the filament voltages were adjusted correctly in the first place. Filament voltage always should be measured *directly at the tube socket.*

11-4 Adjustment of Class B Linear Amplifiers

The establishment of operating conditions and the adjustment of a Class B linear amplifiers, or of any type of efficiency-modulated amplifier, are somewhat more critical than for a similar Class C amplifier for one basic reason: the linear amplifier or efficiency-modulated amplifier must be linear as an *amplifier* with respect to the parameter which is to be varied to effect modulation. Hence, the operating value of bias, the amount of excitation, and the plate circuit loading must be approximately at the correct values before linear amplification or modulation may be attained. A Class C modulated amplifier must be linear only with respect to plate-voltage variation.

Establishing Operating Conditions Satisfactory operating conditions for use of many tube types as Class B linear amplifiers are included in the recommended operating conditions published by tube manufacturers. The RCA HB-3 Tube Handbook gives a comprehensive group of operating conditions for such RCA tubes as are recommended for use as Class B linear amplifiers. However, there are many tube types which would be suitable for use as a Class B linear amplifier for which recommended operating conditions are not published. Operating conditions for such tube types may be determined through use of the procedures given in Sections 5-12 and 5-13 of Chapter Five. Also, Class AB_1 or AB_2 *audio* amplifier ratings normally are suitable for use with the same tubes as a linear r-f amplifier.

Basically, a Class B linear amplifier is operated at a grid-bias voltage such that with excitation removed the plate dissipation of the tube will be perhaps 10 to 20 per cent of the rated dissipation of the tube. Loading of the tube will be approximately twice as heavy

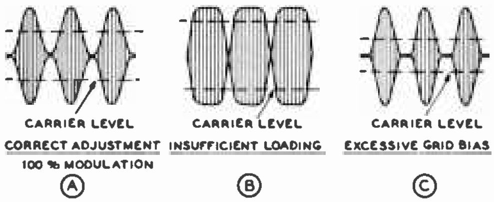


Figure 2.
LINEAR-AMPLIFIER OUTPUT WAVEFORMS.
Showing the carrier envelopes obtained with proper and improper adjustment of a Class B linear amplifier.

as with the same tube operating as a Class C amplifier. This means that the effective load impedance which the tube sees will be about one-half the value which would be used if the tube were to operate as a Class C amplifier. Hence, in order that the loaded Q of the tank circuit in the plate circuit of the tube shall be 15 or higher, approximately *twice* as much tuning capacitance for a specified operating frequency will be required — as compared to the same tube with the same carrier input operating as a Class C amplifier. Thus figure 20 in Chapter Seven may be used in the case of a Class B linear amplifier for a conventional AM signal, but the tank capacitance value obtained from figure 20 should be multiplied by 2. However, the peak voltage rating required of the tank capacitor will be one-half that which would be used for a plate-modulated Class C amplifier.

In the case of a Class B linear amplifier which is to be used without carrier for the amplification of single-sideband signals, the correct value of load impedance may be determined through use of the calculation procedures given in Chapter Five, and then the tank circuit capacitance calculated with the aid of the equations given in Chapter Seven. However, an approximation to the correct value of load impedance for a single-sideband Class B linear amplifier may be determined as follows: (1) Assume a peak power output value for the tube. This value will be determined by the operating conditions chosen, and usually will lie between three and five times the rated plate dissipation of the tube, assuming that the tube will be operated at a relatively high plate voltage and that the maximum capabilities of the tube are to be used. (2) Assume a peak value of plate voltage swing at maximum output from the stage. This value will normally be about 0.9 times the operating voltage on the stage. (3) Determine the approximate R_L for the tube from:

$$R_L = (0.9\ E_{bb})^2/2\ W_{o\ pk}$$

The above is only an approximation, but it is of sufficient accuracy to be of value in establishing the correct value of tank circuit capacitance for a specified nominal value of circuit loaded Q. The reactance required in the tank capacitor or the tank coil at the operating frequency then is equal to: R_L/Q_L, where Q_L is the value of loaded Q desired in the tank circuit.

Adjustment of the Linear Amplifier An oscilloscope is practically a necessity in adjusting a Class B linear amplifier, or other type of efficiency-modulated amplifier, for proper operating conditions. The output signal from the amplifier to be adjusted is coupled to the vertical deflection plates of the oscilloscope. Then the horizontal sweep oscillator on the oscilloscope is adjusted to a frequency which is appropriate for the audio modulation frequency which will be used to adjust the amplifier.

If the amplifier is to be used as a Class B linear amplifier of a modulated signal, 100 per cent modulation is applied to the signal which is being fed to the grid of the linear amplifier. The audio modulating frequency normally will be in the vicinity of 400 cycles for the first checks. Then the excitation coupling, output-circuit loading, and grid-circuit swamping are successively adjusted until the plate power input to the linear amplifier is at the correct value, and the output modulated wave is a faithful replica of the signal being fed to the grid circuit. Watch particularly for flattening of the tops of the modulated wave, and for bright spots at the center of the carrier which denote negative-peak clipping. Flattening of the tops may be the result of insufficient antenna coupling or of insufficient swamping of the grid circuit of the amplifier. Negative-peak clipping may be the result of excessive grid bias or excessive excitation signal. After a satisfactory set of adjustments has been obtained, removal of the modulation

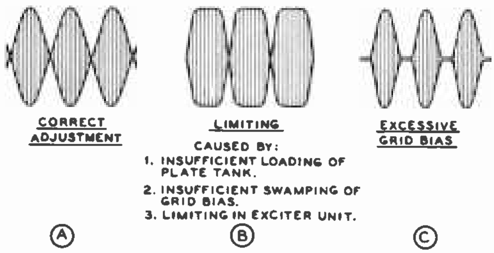


Figure 3.
SSB LINEAR-AMPLIFIER OUTPUT WAVEFORMS.

With improper adjustment the output envelope from an ssb linear amplifier appears similar to the analogous conditions obtained with a conventional AM linear amplifier. But when operating conditions are correct, with a two-tone modulating signal or with carrier and one sideband of full amplitude, the envelope has the appearance shown at (A) above.

from the input signal should not cause an appreciable change in the plate power input to the linear amplifier.

The adjustment of a linear amplifier of a single-sideband signal may be done in several ways. One satisfactory way, when the linear amplifier is to be operated relatively conservatively, is to insert a carrier signal in the single-sideband exciter of approximately one-half the peak output voltage capability of the exciter. Then an audio modulating tone of about 400 cycles is applied to the ssb exciter, and the excitation and loading adjustments to the linear amplifier are made in a similar manner to the procedure just discussed for a conventional AM signal.

Another method of adjustment of a linear amplifier for single-sideband signals is to apply two tones to the exciter, say 1000 cycles and 600 cycles, so that the difference frequency will be about 400 cycles. Then loading, excitation, and bias to the linear amplifier are adjusted until the waveform output of the exciter is undistorted in passing through the linear amplifier. The envelope of the amplified ssb signal should appear as in figure 3.

Due to the large signal excursions encountered in a Class B linear amplifier of single-sideband signals, it is important that the bias voltage and plate voltage have good regulation, and in the case of a tetrode am-

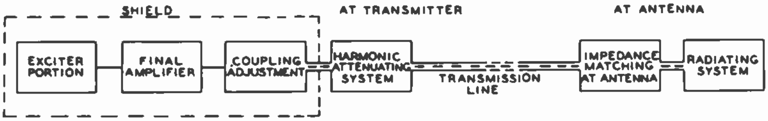


Figure 4.

ANTENNA COUPLING SYSTEM.

The harmonic suppressing antenna coupling system illustrated above is for use when the antenna transmission line has a low standing-wave ratio, and when the characteristic impedance of the antenna transmission line is the same as the nominal impedance of the low-pass harmonic-attenuating filter.

plifier the screen voltage must have extremely good regulation. Also, to obtain best linearity when using tetrode tubes as ssb linear amplifiers, it has been found best to operate with a grid bias somewhat less than that which would be asked for straight Class B operation. The tube manual values specified for Class AB_1 or Class AB_2 operation as an *audio* amplifier have proven most satisfactory for linear operation of tetrode tubes as single-sideband amplifiers.

11-5 Coupling to the Antenna System

When coupling an antenna feed system to a transmitter the most important considerations are as follows: (1) means should be provided for varying the load on the amplifier; (2) the two tubes in a push-pull amplifier should be equally loaded; (3) the load presented to the final amplifier should be resistive (non-reactive) in character; and (4) means should be provided to reduce harmonic coupling between the final amplifier plate tank circuit and the antenna or antenna transmission line to an *extremely low* value.

The Transmitter-Loading Problem The problem of coupling the power output of a high-frequency or v-h-f transmitter to the radiating portion of the antenna system has been materially complicated by the virtual necessity for eliminating interference to TV reception. However, the TVI-elimination portion of the problem may *always* be accomplished by adequate shielding of the transmitter, by filtering of the control and power leads which enter the transmitter enclosure, and by the inclusion of a harmonic-attenuating filter between the output of the transmitter and the antenna system. Methods for eliminating TVI are discussed in detail in Chapter Seventeen.

Although TVI may be eliminated through inclusion of a filter between the output of a shielded transmitter and the antenna system, the fact that such a filter must be included in the link between transmitter and antenna makes it necessary that the transmitter-loading problem be re-evaluated in terms of the necessity for inclusion of such a filter.

Harmonic-attenuating filters are discussed in detail in Chapter Seventeen, but there is one characteristic of such filters that is pertinent to this discussion: harmonic-attenuating filters must be operated at an impedance level which is close to their design value; hence, they must operate *into* a resistive termination substantially equal to the characteristic impedance of the filter. If such filters are operated into an impedance which is not resistive and approximately equal to their characteristic impedance: (1) the capacitors used in the filter sections will be subjected to high peak voltages and may be damaged, (2) the harmonic-attenuating properties of the filter will be decreased, and (3) the impedance at the input end of the filter will be different from that seen by the filter at the load end (except in the case of the half-wave type of filter). Hence it is important that the filter be included in the transmitter-to-antenna circuit at a point where the impedance is close to the nominal value of the filter, and at a point where this impedance is likely to remain fairly constant with variations in frequency.

Block Diagrams of Transmitter-to-Antenna Coupling Systems There are two basic arrangements which include all the provisions required in

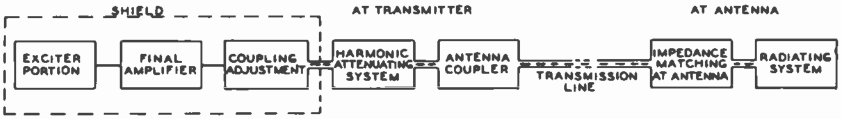


Figure 5.
ANTENNA COUPLING SYSTEM.
The antenna coupling system illustrated above is for use when the antenna transmission line does not have the same characteristic impedance as the TVI filter, and when the standing-wave ratio on the antenna transmission line may or may not be low.

the transmitter-to-antenna coupling system, and which permit the harmonic-attenuating filter to be placed at a position in the coupling system where it can be operated at an impedance level close to its nominal value. These arrangements are illustrated in block-diagram form in figures 4 and 5.

The arrangement of figure 4 is recommended for use with a single-band antenna system, such as a dipole or a rotatable array, wherein an impedance matching system is included within or adjacent to the antenna. The feed line coming down from the antenna system should have a characteristic impedance equal to the nominal impedance of the harmonic filter, and the impedance matching at the antenna should be such that the standing-wave ratio on the antenna feed line is less than 2 to 1 over the range of frequency to be fed to the antenna. Such an arrangement may be used with open-wire line, ribbon or tubular line, or with coaxial cable. The use of coaxial cable is to be recommended, but in any event the impedance of the antenna transmission line should be the same as the nominal impedance of the harmonic filter. The arrangement of figure 4 is more or less standard for commercially manufactured equipment for amateur and commercial use in the h-f and v-h-f range.

The arrangement of figure 5 merely adds an antenna coupler between the output of the harmonic attenuating filter and the antenna transmission line. The antenna coupler will have some harmonic-attenuating action, but its main function is to transform the impedance at the station end of the antenna transmission line to the nominal value of the harmonic filter. Hence the arrangement of figure 5 is more general than the figure 4 system, since the inclusion of the antenna coupler allows the system to feed an antenna transmission line of any reasonable impedance value, and also without regard to the standing-wave ratio which might exist on the antenna transmission line. Antenna couplers are discussed in a following section.

Output Coupling Adjustment It will be noticed by reference both to figure 4 and figure 5 that a box labeled "Coupling Adjustment" is included in the block diagram. Such an element is necessary in the complete system to afford an adjustment in the value of load impedance presented to the tubes in the final amplifier stage of the transmitter. The impedance at the input terminal of the harmonic filter is established by the antenna, through its matching system and the antenna coupler if used. At any event the impedance at the input terminal of the harmonic filter should be very close to the nominal impedance of the filter. Then the "Coupling Adjustment" provides a means for transforming this impedance value to the correct operating value of load impedance which should be presented to the final amplifier stage.

There are two common ways for accomplishing the antenna coupling adjustment, as illustrated in figures 6 and 7. Figure 6 shows the variable-link arrangement most commonly used in home-constructed equipment, while the pi-network coupling arrangement commonly used in commercial equipment is illustrated in figure 7. Either method may be used, and each has its advantages.

Variable-Link Coupling The variable-link method illustrated in figure 6 has the advantage that standard manufactured components may be used with no changes. However, for greatest bandwidth of operation of the coupling circuit, the reactance of the link coil, L, and the reactance of the

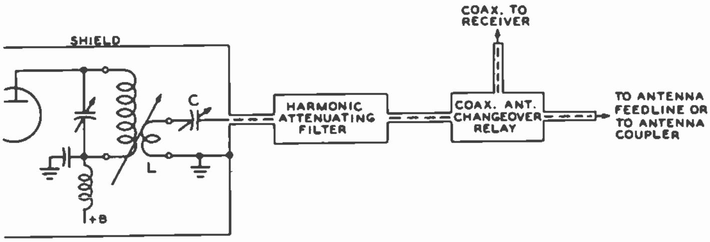


Figure 6.
TUNED-LINK OUTPUT CIRCUIT.
Capacitor C should be adjusted so as to tune out the inductive reactance of the coupling link, L. Loading of the amplifier then is varied by physically varying the coupling between the plate tank of the final amplifier and the antenna coupling link.

link tuning capacitor, C, should both be between 3 and 4 times the nominal load impedance of the harmonic filter. This is to say that the inductive reactance of the coupling link L should be tuned out or resonated by capacitor C, and the operating Q of the L-C link circuit should be between 3 and 4. If the link coil is not variable with respect to the tank coil of the final amplifier, capacitor C *may* be used as a loading control; however, this system is not recommended since its use will require adjustment of C whenever a frequency change is made at the transmitter. If L and C are made resonant at the center of a band, with a link circuit Q of 3 to 4, and coupling adjustment is made by physical adjustment of L with respect to the final amplifier tank coil, it usually will be possible to operate over an entire amateur band without change in the coupling system. Capacitor C normally may have a low voltage rating, even with a high power transmitter, due to the low Q and low impedance of the coupling circuit.

Pi-Network Coupling The pi-network coupling system offers two advantages: (1) a *mechanical* coupling variation is not required to vary the loading of the final amplifier, and (2) the pi network (if used with an operating Q of about 15) offers within itself a harmonic attenuation of 40 db or more, in addition to the harmonic attenuation provided by the additional harmonic attenuating filter. Some commercial equipments (such as the Collins amateur transmitters) incorporate an L network in addition to the pi network, for accomplishing the impedance transformation in two steps and to provide additional harmonic attenuation. The design of L and pi matching networks is discussed in detail in Section 7-10 of Chapter Seven.

Tuning the Pi-Section Coupler Tuning of a pi-network coupling circuit such as illustrated in figure 7 is accomplished in the following manner: First remove the connection between the output of the amplifier and the harmonic filter (load). Tune C_2 to a capacitance which is large for the band in use, adding suitable additional capacitance by switch S if operation is to be on one of the lower frequency bands. Apply reduced plate voltage to the stage and dip to resonance with C_1. It may be necessary to vary the inductance in coil L, but in any event resonance should be reached with a setting of C_1 which is approximately correct for the desired value of operating Q of the pi network (refer to Section 7-10 for a discussion of operating Q of a pi network).

Then couple the load to the amplifier (through the harmonic filter), apply reduced plate voltage again and dip to resonance with C_1. If the plate current dip with load is too low (taking into consideration the reduced plate voltage), *decrease* the capacitance of C_2 and again dip to resonance, repeating the procedure until the correct value of plate current is obtained with full plate voltage on the stage. There should be a relatively small change required in the setting of C_1 (from the original setting of C_1 without load) if the operating Q of the network is correct and if a large value of impedance transformation is being employed — as would be the case when transforming from the plate impedance of a single-ended output stage down to the 50-ohm im-

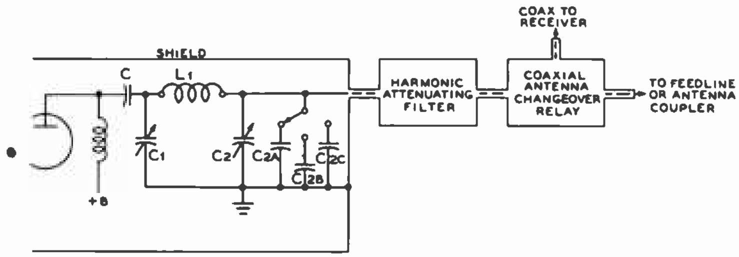


Figure 7.
PI-NETWORK ANTENNA COUPLER.
The design of pi-network output circuits is discussed in Chapter Seven. The additional output-end shunting capacitors selected by switch S are for use on the lower frequency ranges. Inductor L_1 may be selected by a tap switch, it may be continuously variable, or plug-in inductors may be used.

pedance of the usual harmonic filter and its subsequent load.

In a pi network of this type the harmonic attenuation of the section will be adequate when the correct value of C_1 and L are being used, and when the resonant dip in C_1 is sharp. If the dip in C_1 is broad, or if the plate current persists in being too high with C_2 at maximum setting, it means that a *greater* value of capacitance is required at C_2—assuming that the values of C_1 and L are correct.

11-6 Antenna Couplers

As stated in the previous section, an antenna coupler is not required when the impedance of the antenna transmission line is the same as the nominal impedance of the harmonic filter, assuming that the antenna feed line is being operated with a low standing-wave ratio. (Matching the antenna to the feed line is discussed in detail in Chapter Thirteen.) However, there are many cases where it is desirable to feed a multi-band antenna from the output of the harmonic filter, where a tuned line is being used to feed the antenna, or where a long wire without a separate feed line is to be fed from the output of the harmonic filter. In such cases an antenna coupler is required.

Some harmonic attenuation will be provided by the antenna coupler, particularly if it is well shielded. In certain cases when a pi network is being used at the output of the transmitter, the addition of a shielded antenna coupler will provide sufficient harmonic attenuation. But in all normal cases it will be necessary to include a harmonic filter between the output of the transmitter and the antenna coupler. When an adequate harmonic filter is being used, it will not be necessary in normal cases to shield the antenna coupler, except from the standpoint of safety or convenience.

Function of an Antenna Coupler The function of the antenna coupler is, basically, to transform the impedance of the antenna system being used to the correct value of *resistive* impedance for the harmonic filter, and hence for the transmitter. Thus the antenna coupler may be used to resonate the feeders or the radiating portion of the antenna system, in addition to its function of impedance transformation.

It is important to remember that there is *nothing* that can be done at the antenna coupler which will eliminate standing waves on the antenna transmission line. Standing waves are the result of reflection from the antenna, and the coupler can do nothing about this condition. However, the antenna coupler can resonate the feed line (by introducing a conjugate impedance) in addition to providing an impedance transformation. Thus, a resistive impedance of the correct value can be presented to the harmonic filter, as in figure 5, regardless of any reasonable standing-wave ratio on the antenna transmission line.

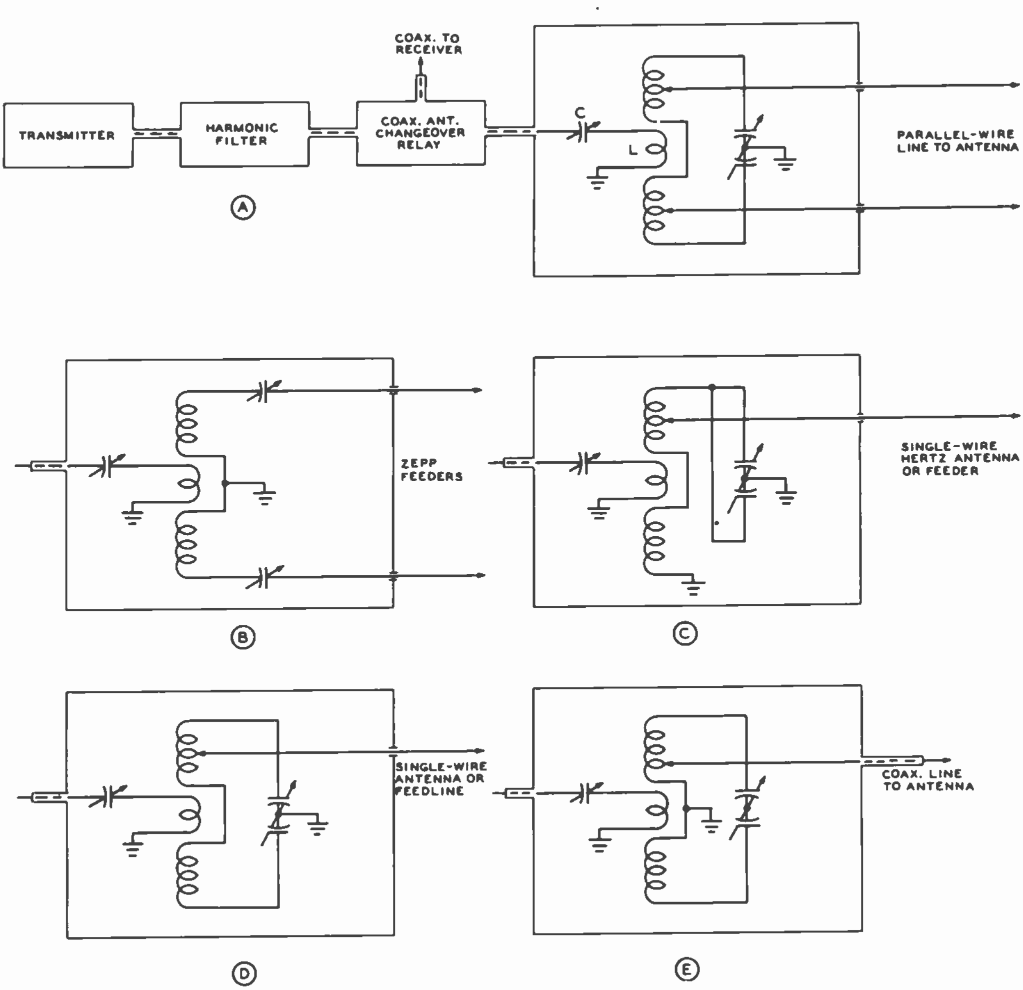


Figure 8.

ALTERNATIVE ANTENNA-COUPLER CIRCUITS.

Plug-in coils, one or two variable capacitors of the split-stator variety, and a system of switches or plugs and jacks may be used in the antenna coupler to accomplish the feeding of different types of antennas and antenna transmission lines from the coaxial input line from the transmitter or from the antenna changeover relay. Link L should be resonated with capacitor C at the operating frequency of the transmitter so that the harmonic filter will operate into a resistive load impedance of the correct nominal value.

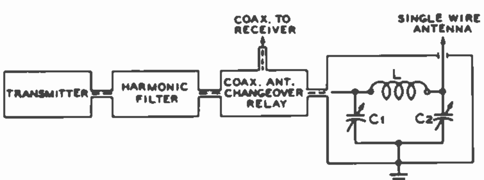


Figure 9.

PI-NETWORK ANTENNA COUPLER.

An arrangement such as illustrated above is convenient for feeding an end-fed Hertz antenna, or a random length of wire for portable or emergency operation, from the nominal value of impedance of the harmonic filter.

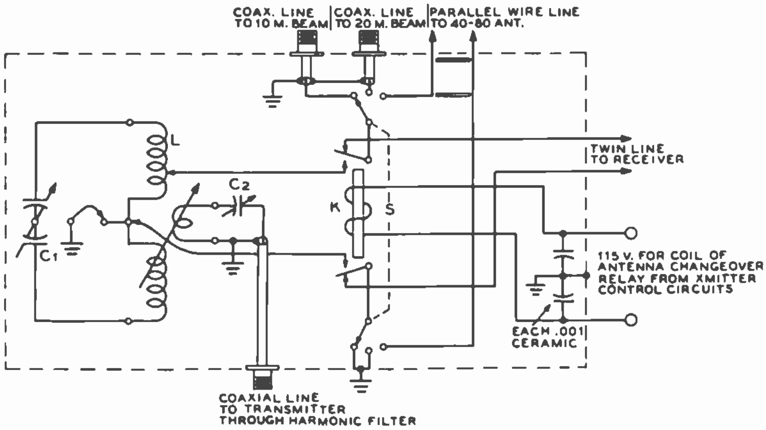


Figure 10.

CIRCUIT OF A PRACTICAL ANTENNA COUPLER.

This arrangement includes the antenna changeover relay, an antenna selector switch, and a universal-type antenna coupler. The tuned circuit of the antenna coupler is not in the circuit in the receive position of the changeover relay. Capacitor C_2 serves to tune out the inductive reactance of the coupling link.

Types of Antenna Couplers All usual types of antenna couplers fall into two classifications: (1) inductively coupled resonant systems as exemplified by those shown in figure 8, and (2) conductively coupled pi-network systems such as shown in figure 9. The inductively-coupled system is much more commonly used, since it is convenient for feeding a balanced line from the coaxial output of the usual harmonic filter. The pi-network system is most useful for feeding a length of wire from the output of a transmitter.

Several general methods for using the inductively-coupled resonant type of antenna coupler are illustrated in figure 8. The coupling between the link coil L and the main tuned circuit need not be variable; in fact it is preferable that the correct link size and placement be determined for the

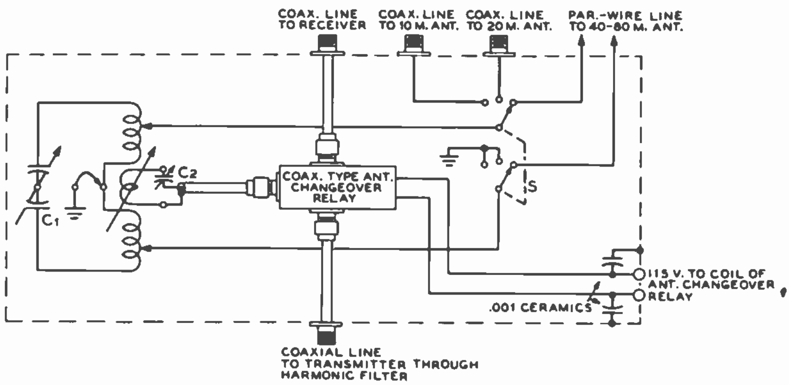


Figure 11.

COAXIAL-TYPE ANTENNA COUPLER.

This arrangement differs from that of figure 10 in that a coaxial changeover relay is used, coaxial line is used from the coupler to the receiver, and the tuned circuit of the coupler is in the circuit for receiving as well as transmitting.

tank coil which will be used for each band, and then that the link be made a portion of the plug-in coil. Capacitor C then can be adjusted to a pre-determined value for each band such that it will resonate with the link coil for that band. The reactance of the link coil (and hence the reactance of the capacitor setting which will resonate the coil) should be about 3 or 4 times the impedance of the transmission line between the antenna coupler and the harmonic filter, so that the link coupling circuit will have an operating Q of 3 or 4. The use of capacitor C to resonate with the inductance of the link coil L will make it easier to provide a low standing-wave ratio to the output of the harmonic filter, simply by adjustment of the antenna-coupler tank circuit to resonance. If this capacitor is not included, the system still will operate satisfactorily, but the tank circuit will have to be detuned slightly from resonance so as to cancel the inductive reactance of the coupling link and thus provide a resistive load to the output of the harmonic filter. Variations in the loading of the final amplifier should be made by the coupling adjustment at the final amplifier, not at the antenna coupler.

The pi-network type of antenna coupler, as shown in figure 9, is useful for certain applications, but is primarily useful in feeding a single-wire antenna from a low-impedance transmission line. In such an application the operating Q of the pi network may be somewhat lower than that of a pi network in the plate circuit of the final amplifier of a transmitter, as shown in figure 7. An operating Q of 3 or 4 in such an application will be found to be adequate, since harmonic attenuation has been accomplished ahead of the antenna coupler. However, the circuit will be easier to tune, although it will not have so great a bandwidth, if the operating Q is made higher.

Three complete antenna couplers are illustrated in figures 10, 11, and 12. The circuit of figure 10 is for use with twin-line feed from the antenna coupler to the receiver, and does not feed the received signal through

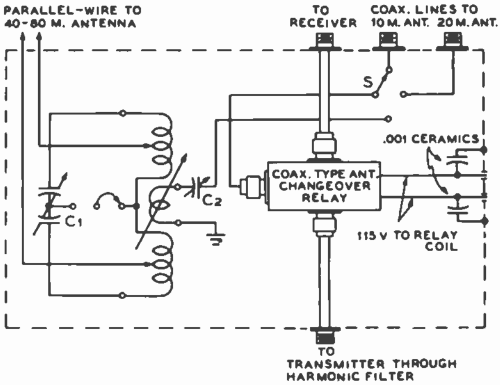


Figure 12.
ALTERNATIVE COAXIAL ANTENNA COUPLER.

This circuit is recommended not only as being most desirable when coaxial lines with low s.w.r. are being used to feed antenna systems such as rotatable beams, but when it also is desired to feed through open-wire line to some sort of multi-band antenna for the lower frequency ranges. The tuned circuit of the antenna coupler is operative only when using the open-wire feed, and then it is in operation both for transmit and receive.

the antenna tuning network. This is desirable in that it allows the receiver to be tuned over a wide range of frequencies without adjustment of the antenna coupler. On the other hand, best impedance matching between the antenna transmission line and the input of the *receiver* is not provided.

The alternative arrangement shown in figure 11 utilizes a coaxial feed line to the receiver, and feeds the received signal through the antenna tuning network, when the network is being used. An alternative arrangement would be to use the antenna coupling tank circuit only when feeding the coaxial output of the transmitter to the open-wire feed line (or similar multi-band antenna) of the 40-80 meter antenna. The coaxial lines to the 10-meter beam and to the 20-meter beam would be fed directly from the output of the coaxial antenna changeover relay through switch S. Such an arrangement is illustrated in figure 12.

CHAPTER TWELVE

Radiation, Propagation, and Transmission Lines

Radio waves are electromagnetic waves similar in nature but much lower in frequency than light waves or heat waves. Such waves represent electric energy traveling through space. Radio waves travel in free space with the velocity of light and can be reflected and refracted much the same as light waves.

12-1 Radiation from an Antenna

Alternating current passing through a conductor creates an alternating electromagnetic field around that conductor. Energy is alternately stored in the field, and then returned to the conductor. As the frequency is raised, more and more of the energy does not return to the conductor, but instead is radiated off into space in the form of electromagnetic waves, called radio waves. Radiation from a wire, or wires, is materially increased whenever there is a sudden *change* in the *electrical constants* of the line. These sudden changes produce reflection, which places *standing waves* on the line.

When a wire in space is fed radio frequency energy having a wavelength of approximately 2.1 times the length of the wire in meters, the wire *resonates* as a *half-wave dipole* antenna at that wavelength or frequency. The greatest possible change in the electrical constants of a line is that which occurs at the open end of a wire. Therefore, a dipole has a great mismatch at each end, producing a high degree of reflection. We say that the dipole is terminated in an infinite impedance (open circuit).

A returning wave which has been reflected meets the next incident wave, and the voltage and current at any point along the antenna are the vector sum of the two waves. At the ends of the dipole, the voltages add, while the currents of the two waves cancel, thus producing *high voltage* and *low current* at the *ends* of the dipole or half-wave section of wire. In the same manner, it is found that the currents add while the voltages cancel at the center of the dipole. Thus, at the *center* there is *high current* but *low voltage*.

Inspection of figure 1 will show that the current in a dipole decreases sinusoidally towards either end, while the voltage similarly increases. The voltages at the two ends of the antenna are 180° out of phase, which means that the polarities are opposite, one being plus while the other is minus at any instant. A curve representing either the voltage or current on a dipole represents a *standing wave* on the wire.

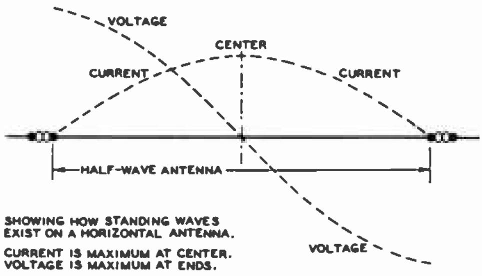


Figure 1.
STANDING WAVES ON A RESONANT ANTENNA.

Radiation from Sources other than Antennas Radiation can and does take place from sources other than antennas. Undesired radiation can take place from open-wire transmission lines, both from single-wire lines and from lines comprised of more than one wire. In addition, radiation can be made to take place in a very efficient manner from electromagnetic horns, from plastic lenses or from electromagnetic lenses made up of spaced conducting planes, from slots cut in a piece of metal, from dielectric wires, or from the open end of a wave guide.

Directivity of Radiation The radiation from any physically practicable radiating system is directive to a certain degree. The degree of directivity can be enhanced or altered when desirable through the combination of radiating elements in a prescribed manner, through the use of reflecting planes or curved surfaces, or through the use of such systems as mentioned in the preceding paragraph. The construction of directive antenna arrays is covered in detail in the chapters which follow.

Polarization Like light waves, radio waves can have a definite polarization. In fact, while light waves ordinarily have to be reflected or passed through a polarizing medium before they have a definite polarization, a radio wave leaving a simple radiator will have a definite polarization, the polarization being indicated by the orientation of the electric-field component of the wave. This, in turn, is determined by the orientation of the radiator itself, as the magnetic-field component is always at right angles to a linear radiator, and the electric-field component is always in the same plane as the radiator. Thus we see that an antenna that is vertical with respect to the earth will transmit a vertically polarized wave, as the electrostatic lines of force will be vertical. Likewise, a simple horizontal antenna will radiate horizontally polarized waves.

Because the orientation of a simple linear radiator is the same as the polarization of the waves emitted by it, the radiator itself is referred to as being either vertically or horizontally polarized. Thus, we say that a horizontal antenna is horizontally polarized.

Figure 2A illustrates the fact that the polarization of the electric field of the radiation from a vertical dipole is vertical. Figure 2B, on the other hand, shows that the polarization of electric-field radiation from a vertical *slot* radiator is horizontal. This fact has been utilized in certain commercial FM antennas where it is desired to have horizontally polarized radiation but where it is more convenient to use an array of vertically stacked slot arrays. If the metallic sheet is bent into a cylinder with the slot on one side, substantially omnidirectional horizontal coverage is obtained with horizontally-polarized radiation when the cylinder with the slot in one side is oriented vertically. An arrangement of this type is shown in figure 2C. Several such cylinders may be stacked vertically to reduce high-angle radiation and to concentrate the radiated energy at the useful low radiation angles.

In any event the polarization of radiation from a radiating system is parallel to the electric field as it is set up inside or in the vicinity of the radiating system.

12-2 General Characteristics of Antennas

All antennas have certain general characteristics to be enumerated. It is the result of differences in these general characteristics which makes one type of antenna system most suitable for one type of application and another type best for a different application. Six of the more important characteristics are: (1)

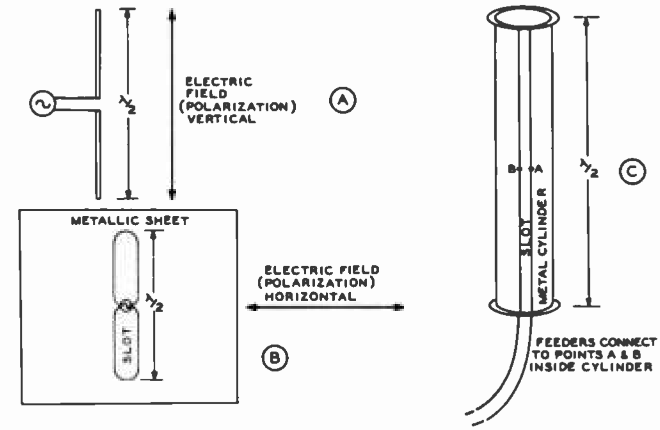


Figure 2.
ANTENNA POLARIZATION.
The polarization (electric field) of the radiation from a resonant dipole such as shown at (A) above is parallel to the length of the radiator. In the case of a resonant slot cut in a sheet of metal and used as a radiator, the polarization (of the electric field) is perpendicular to the length of the slot. In both cases, however, the polarization of the radiated field is parallel to the potential gradient of the radiator; in the case of the dipole the electric lines of force are from end to end, while in the case of the slot the field is across the sides of the slot. The metallic sheet containing the slot may be formed into a cylinder to make up the radiator shown at (C). With this type of radiator the radiated field will be horizontally polarized even though the radiator is mounted vertically.

polarization, (2) radiation resistance, (3) horizontal directivity, (4) vertical directivity, (5) bandwidth, and (6) effective power gain.

The *polarization* of an antenna or radiating system is the direction of the electric field vector and has been defined in Section 12-1.

The *radiation resistance* of an antenna system is normally referred to the feed point in an antenna fed at a current loop, or it is referred to a current loop in an antenna system fed at another point. The radiation resistance is that value of resistance which, if inserted in series with the antenna at a current loop, would dissipate the same energy as is actually radiated by the antenna if the antenna current at the feed point were to remain the same.

The *horizontal* and *vertical directivity* can best be expressed as a *directive pattern* which is a graph showing the relative radiated field intensity against *azimuth* angle for horizontal directivity and field intensity against *elevation* angle for vertical directivity.

The *bandwidth* of an antenna is a measure of its ability to operate within specified limits over a range of frequencies. Bandwidth can be expressed either "operating frequency plus-or-minus a specified per cent of operating frequency" or "operating frequency plus-or-minus a specified number of megacycles" for a certain standing-wave-ratio limit on the transmission line feeding the antenna system.

The *effective power gain* or *directive gain* of an antenna is the ratio between the power required in the specified antenna and the power required in a reference antenna (usually a half-wave dipole) to attain the same field strength in the favored direction of the antenna under measurement. Directive gain may be expressed either as an actual power ratio, or as is more common, the power ratio may be expressed in decibels.

Physical Length of a Half-Wave Antenna If the cross section of the conductor which makes up the antenna is kept very small with respect to the antenna length, an electrical half wave is a fixed percentage shorter than a physical half-wavelength. This percentage is approximately 5 per cent. Therefore, most linear half-wave antennas are close to 95 per cent of a half wave long physically. Thus, a half-wave antenna resonant at exactly 80 meters would be one-half of 0.95 times 80 meters in length. Another way of saying the same thing is that a wire resonates at a wavelength of about 2.1 times its length in meters. If the diameter of the conductor begins to be an appreciable fraction of a wavelength, as when tubing is used as a v-h-f radiator, the factor becomes slightly less than 0.95. For the use of wire and not tubing on frequencies below 30 Mc., however, the figure of 0.95 may be taken as accurate. This assumes a radiator removed from surrounding objects, and with *no bends*.

Simple conversion into feet can be obtained by using the factor 1.56. To find the physical length of a half-wave 80-meter antenna, we multiply 80 times 1.56, and get 124.8 feet for the length of the radiator.

It is more common to use frequency than wavelength when indicating a specific spot in the radio spectrum. For this reason, the relationship between wavelength and frequency must be kept in mind. As the velocity of radio waves through space is constant at the speed of light, it will be seen that the more waves that pass a point per second (higher frequency), the closer together the peaks of those waves must be (shorter wavelength). Therefore, the higher the frequency the lower the wavelength.

A radio wave in space can be compared to a wave in water. The wave, in either case, has peaks and troughs. One peak and one trough constitute a *full wave,* or *one wavelength.*

Frequency describes the number of wave cycles or peaks passing a point per second. Wavelength describes the distance the wave travels through space during one cycle or oscillation of the antenna current; it is the distance in meters between adjacent peaks or adjacent troughs of a wave train.

As a radio wave travels 300,000,000 meters a second (speed of light), a frequency of 1 cycle per second corresponds to a wavelength of 300,000,000 meters. So, if the frequency is multiplied by a million, the wavelength must be divided by a million, in order to maintain their correct ratio.

A frequency of 1,000,000 cycles per second (1,000 kc.) equals a wavelength of 300 meters. Multiplying frequency by 10 and dividing wavelength by 10, we find: a frequency of 10,000 kc. equals a wavelength of 30 meters. Multiplying and dividing by 10 again, we get: a frequency of 100,000 kc. equals 3 meters wavelength. Therefore, to change wavelength to frequency (in kilocycles), simply divide 300,000 by the wavelength in meters (λ).

$$F_{kc} = \frac{300{,}000}{\lambda}$$

$$\lambda = \frac{300{,}000}{F_{kc}}$$

Now that we have a simple conversion formula for converting wavelength to frequency and vice versa, we can combine it with our wavelength versus antenna length formula, and we have the following:

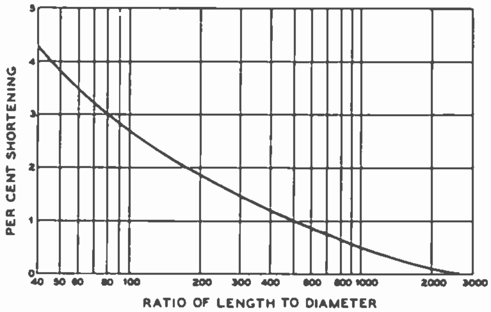


Figure 3.

CHART SHOWING SHORTENING OF A RESONANT ELEMENT IN TERMS OF RATIO OF LENGTH TO DIAMETER.

The use of this chart is based on the basic formula where radiator length in feet is equal to 468/frequency in Mc. This formula applies on frequencies below perhaps 30 Mc. when the radiator is made from wire. On higher frequencies, or on 14 and 28 Mc. when the radiator is made of large-diameter tubing, the radiator is shortened from the value obtained with the above formula by an amount determined by the ratio of length to diameter of the radiator. The amount of this shortening is obtainable from the chart shown above.

Length of a half-wave radiator made from wire (no. 14 to no. 10):

3.5-Mc. to 30-Mc. bands

$$\text{Length in feet} = \frac{468}{\text{Freq. in Mc.}}$$

50-Mc. band

$$\text{Length in feet} = \frac{460}{\text{Freq. in Mc.}}$$

$$\text{Length in inches} = \frac{5600}{\text{Freq. in Mc.}}$$ 466,666 fr

144-Mc. band

$$\text{Length in inches} = \frac{5500}{\text{Freq. in Mc.}}$$ 458 1/3

Length-to-Diameter Ratio When a half-wave radiator is constructed from tubing or rod whose diameter is an appreciable fraction of the length of the radiator, the resonant length of a half-wave antenna will be shortened. The amount of shortening can be determined with the aid of the chart of figure 3. In this chart the amount of additional shortening over the values given in the previous paragraph is plotted against the ratio of the length to the diameter of the half-wave radiator.

The length of a wave in free space is somewhat longer than the length of an antenna for the same frequency. The actual free-space wavelength is given by the following expressions:

$$\text{Wavelength} = \frac{492}{\text{Freq. in Mc.}} \text{ in feet}$$

$$\text{Wavelength} = \frac{5905}{\text{Freq. in Mc.}} \text{ in inches}$$

Harmonic Resonance A wire in space can resonate at more than one frequency. The *lowest* frequency at which it resonates is called its *fundamental* frequency, and at that frequency it is approximately a half wavelength long. A wire can have two, three, four, five, or more standing waves on it, and thus it resonates at approximately the integral harmonics of its fundamental frequency. However, the higher harmonics are not exactly integral multiples of the lowest resonant frequency as a result of *end effects.*

A harmonic operated antenna is somewhat longer than the corresponding integral number of dipoles, and for this reason, the dipole length formula cannot be used simply by multiplying by the corresponding harmonic. The intermediate half wave sections do not have "end effects." Also, the current distribution is disturbed by the fact that power can reach some of the half wave sections only by flowing through other sections, the latter then acting not only as radiators, but also as transmission lines. For the latter reason, the resonant length will be dependent to an extent upon the method of feed, as there will be less attenuation of the current along the antenna if it is fed at or near the center than if fed towards or at one end. Thus, the antenna would have to be somewhat longer if fed near one end than if fed near the center. The difference would be small, however, unless the antenna were many wavelengths long.

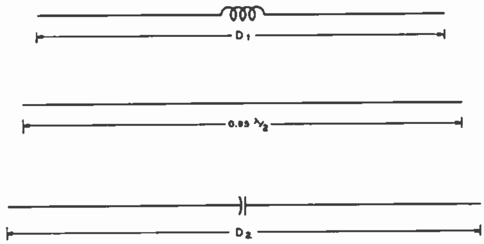


Figure 4.
EFFECT OF SERIES INDUCTANCE AND CAPACITANCE ON THE LENGTH OF A HALF-WAVE RADIATOR.

The top antenna has been electrically lengthened by placing a coil in series with the center. In other words, an antenna with a lumped inductance in its center can be made shorter for a given frequency than a plain wire radiator. The bottom antenna has been capacitively shortened electrically. In other words, an antenna with a capacitor in series with it must be made longer for a given frequency since its effective electrical length as compared to plain wire is shorter.

Under conditions of severe current attenuation, it is possible for some of the nodes, or loops, actually to be slightly greater than a physical half wavelength apart. Practice has shown that the most practical method of resonating a harmonically operated antenna accurately is by cut and try, or by using a feed system in which both feed line and antenna are resonated at the station end as an integral system.

A dipole or half-wave antenna is said to operate on its fundamental or first harmonic. A full wave antenna, 1 wavelength long, operates on its second harmonic. An antenna with five half-wavelengths on it would be operating on its fifth harmonic. Observe that the fifth harmonic antenna is *2½ wavelengths* long, *not* 5 wavelengths.

Antenna Resonance Most types of antennas operate most efficiently when tuned or *resonated* to the frequency of operation. This consideration of course does not apply to the rhombic antenna and to the parasitic elements of arrays employing parasitically excited elements. However, in practically every other case it will be found that increased efficiency results when the entire antenna system is resonant, whether it be a simple dipole or an elaborate array. The radiation efficiency of a resonant wire is many times that of a wire which is not resonant.

If an antenna is slightly too long, it can be resonated by series insertion of a variable capacitor at a high current point. If it is slightly too short, it can be resonated by means of a variable inductance. These two methods, illustrated schematically in figure 4, are generally employed when part of the antenna is brought into the operating room.

With an antenna array, or an antenna fed by means of a transmission line, it is more common to cut the elements to exact resonant length by "cut and try" procedure. Exact antenna resonance is more important when the antenna system has low radiation resistance; an antenna with low radiation resistance has higher Q (tunes sharper) than an antenna with high radiation resistance. The higher Q does not indicate greater efficiency; it simply indicates a sharper resonance curve.

12-3 Radiation Resistance and Feed-Point Impedance

In many ways, a half-wave antenna is like a tuned tank circuit. The main difference lies in the fact that the elements of inductance, capacitance, and resistance are *lumped* in the tank circuit, and are *distributed* throughout the length of an antenna. The center of a half-wave radiator is effectively at ground potential as far as r-f voltage is concerned, although the current is highest at that point.

When the antenna is resonant, and it always should be for best results, the impedance at the center is substantially resistive, and is termed the radiation resistance. Radiation resistance is a fictitious term; it is that value of resistance (referred to the current loop) which would dissipate the same amount of power as is being radiated by the antenna, when fed with the current flowing at the current loop.

The radiation resistance depends on the antenna length and its proximity to nearby objects which either absorb or re-radiate power, such as the ground, other wires, etc.

The Marconi Antenna Before going too far with the discussion of radiation resistance, an explanation of the Marconi (grounded quarter wave) antenna is in order. The Marconi antenna is a special type of Hertz antenna in which the earth acts as the "other half" of the dipole. In other words, the current flows into the earth instead of into a similar quarter-wave section. Thus, the current loop of a Marconi antenna is at the *base* rather than in the *center*. In either case, it is a quarter wavelength from the end (or ends).

A half-wave dipole far from ground and other reflecting objects has a radiation resistance at the center of about 73 ohms. A Marconi antenna is simply one-half of a dipole. For that reason, the radiation resistance is roughly half the 73-ohm impedance of the dipole or 36.5 ohms. The radiation resistance of a Marconi antenna such as a mobile whip will be lowered by the proximity of the automobile body.

Antenna Impedance Because the power throughout the antenna is the same, the *impedance* of a resonant antenna at any point along its length merely expresses the ratio betwen voltage and current at that point. Thus, the lowest impedance occurs where the current is highest, namely, at the center of a dipole, or a quarter wave from the end of a Marconi. The impedance rises uniformly toward each end, where it is about 2000 ohms for a dipole remote from ground, and about twice as high for a vertical Marconi.

If a vertical half-wave antenna is set up so that its lower end is at the ground level, the effect of the ground reflection is to increase the radiation resistance to approximately 100 ohms. When a horizontal half-wave antenna is used, the radiation resistance (and, of course, the amount of energy radiated for a given antenna current) depends on the height of the antenna above ground, since the height determines the phase and amplitude of the wave reflected from the ground back to the antenna. Thus the resultant current in the antenna for a given power is a function of antenna height.

Along a half-wave antenna, the *impedance* varies from a minimum at the center to a maximum at the ends. The impedance is that property which expresses the *ratio* between the antenna current at any point along the wire for the value of r-f voltage at that point.

The curves of figure 5 indicate the theoreti-

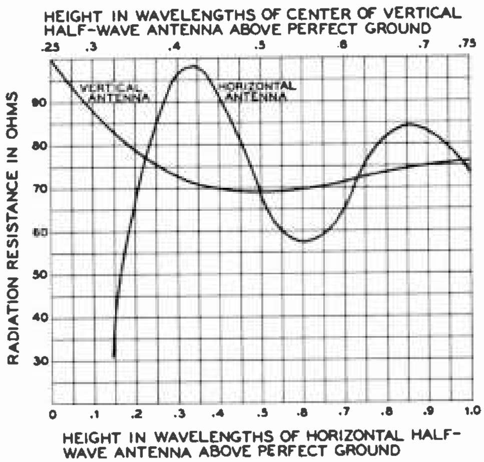


Figure 5.

EFFECT OF HEIGHT ON THE RADIATION RESISTANCE OF A DIPOLE SUSPENDED ABOVE PERFECT GROUND.

cal center-point radiation resistance of a half-wave antenna for various heights above perfect ground. These values are of importance in matching untuned radio-frequency feeders to the antenna, in order to obtain a good impedance match and an absence of standing waves on the feeders.

Above *average* ground, the actual radiation resistance of a dipole will vary from the exact value of figure 5, since the latter assumes a hypothetical, perfect ground having no loss and perfect reflection. Fortunately, the curves for the radiation resistance over most types of earth will correspond rather closely with those of the chart, except that the radiation resistance for a horizontal dipole does not fall off as rapidly as is indicated for heights below an eighth wavelength. However, with the antenna so close to the ground and the soil in a strong field, much of the radiation resistance is actually represented by ground loss; this means that a good portion of the antenna power is being dissipated in the earth, which, unlike the hypothetical perfect ground, has resistance. In this case, an appreciable portion of the "radiation resistance" actually is loss resistance. The type of soil also has an effect upon the radiation *pattern*, especially in the vertical plane, as will be seen later.

The radiation resistance of an antenna generally increases with length, although this increase varies up and down about a constantly increasing average. The peaks and dips are caused by the reactance of the antenna, when its length does not allow it to resonate at the operating frequency.

Antenna Efficiency Antennas have a certain loss resistance as well as a radiation resistance. The loss resistance defines the power lost in the antenna due to ohmic resistance of the wire, ground resistance (in the case of a Marconi), corona discharge, and insulator losses.

The approximate effective radiation efficiency (expressed as a decimal) is equal to: $N_r = R_a/(R_a+R_L)$ where R_a is equal to the radiation resistance and R_L is equal to the effective loss resistance of the antenna. The loss resistance will be of the order of 0.25 ohm for large-diameter tubing conductors such as are most commonly used in multi-element parasitic arrays, and will be of the order of 0.5 to 2.0 ohms for arrays of normal construction using copper wire.

When the radiation resistance of an antenna or array is very low, the current at a voltage node will be quite high for a given power. Likewise, the voltage at a current node will be very high. Even with a heavy conductor and excellent insulation, the losses due to the high voltage and current will be appreciable if the radiation resistance is sufficiently low.

Usually, it is not considered desirable to use an antenna or array with a radiation resistance of less than approximately 5 ohms unless there is sufficient directivity, compactness, or other advantage to offset the losses resulting from the low radiation resistance.

Ground Resistance The radiation resistance of a Marconi antenna, especially, should be kept as high as possible. This will reduce the antenna current for a given power, thus minimizing loss resulting from the series resistance offered by the earth connection. The radiation resistance can be kept high by making the Marconi radiator somewhat longer than a quarter wave, and shortening it by series capacitance to an electrical quarter wave. This reduces the current flowing in the earth connection. It also should be removed from ground as much as possible (ver-

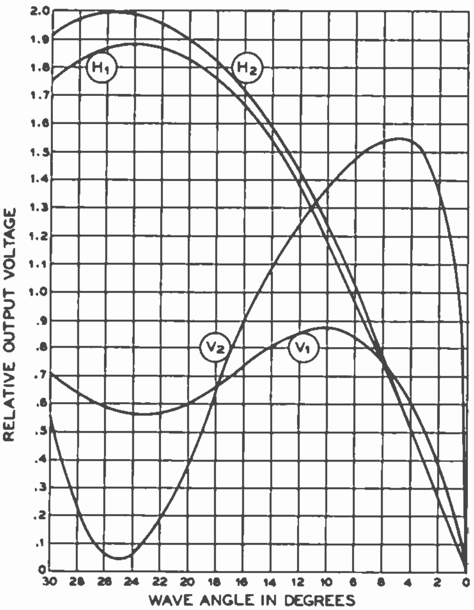


Figure 6.

VERTICAL-PLANE DIRECTIONAL CHARACTERISTICS OF HORIZONTAL AND VERTICAL DOUBLETS ELEVATED 0.6 WAVELENGTH AND ABOVE TWO TYPES OF GROUND.

H_1 represents a horizontal doublet over typical farmland. H_2 over salt water. V_1 is a vertical pattern of radiation from a vertical doublet over typical farmland, V_2 over salt water. A salt water ground is the closest approach to an extensive ideally perfect ground that will be met in actual practice.

tical being ideal). Methods of minimizing the resistance of the earth connection will be found in the discussion of the Marconi antenna in Chapter Thirteen.

12-4 Horizontal Directivity

When choosing and orienting an antenna system, the radiation patterns of the various common types of antennas should be given careful consideration. The directional characteristics are of still greater importance when a directive antenna array is used.

Horizontal directivity is always desirable on any frequency for point-to-point work. However, it is not always attainable with reasonable antenna dimensions on the lower frequencies. Further, when it is attainable, as on the frequencies above perhaps 7 Mc., with reasonable antenna dimensions, operating convenience is greatly furthered if the maximum lobe of the horizontal directivity is controllable. It is for this reason that rotatable antenna arrays have come into such common usage.

Considerable horizontal directivity can be used to advantage when: (1) only point-to-point work is necessary, (2) several arrays are available so that directivity may be changed by selecting or reversing antennas, (3) a single rotatable array is in use. Signals follow the great-circle path, or within 2 or 3 degrees of that path under all normal propagation conditions. However, under turbulent ionosphere conditions, or when unusual propagation conditions exist, the deviation from the great-circle path for greatest signal intensity may be as great as 90°. Making the array rotatable overcomes these difficulties, but arrays having extremely high horizontal directivity become too cumbersome to be rotated, except perhaps when designed for operation on frequencies above 50 Mc.

12-5 Vertical Directivity

Vertical directivity is of the greatest importance in obtaining satisfactory communication above 14 Mc. whether or not horizontal directivity is used. This is true simply because *only* the energy radiated between certain definite *elevation* angles is useful for communication. Energy radiated at other elevation angles is lost and performs no useful function.

Optimum Angle of Radiation The optimum angle of radiation for propagation of signals between two points is dependent upon a number of variables. Among these significant variables are: (1) height of the ionosphere layer which is providing the reflection, (2) distance betwen the two stations, (3) number of hops for propagation between the two stations. For communication on the 14-Mc. band it is often possible for different modes of propagation to provide signals between two points. This means, of

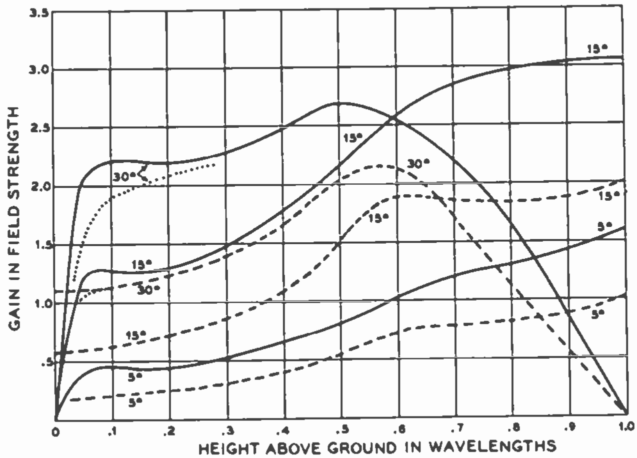


Figure 7.
EFFECT OF HEIGHT UPON ANTENNA GAIN.

Showing the effect of height above ground upon the gain of a horizontal single-section flat-top beam with one-eighth wave spacing (solid curves) and on the gain of a horizontal half-wave dipole antenna (dashed curves) at vertical angles of 5°, 15° and 30°. The gain is referred to a half wave antenna in free space. Perfectly conducting ground is assumed. The short dotted curves show the effect of 0.5-ohm loss resistance on the effective gain of the dipole and the flat-top beam.

course, that more than one angle of radiation can be used. If *no* elevation directivity is being used under this condition of propagation, selective fading will take place because of interference between the waves arriving over the different paths.

On the 28-Mc. band it is by far the most common condition that only one mode of propagation will be possible between two points at any one time. This explains, of course, the reason why rapid fading in general and selective fading in particular are almost absent from signals heard on the 28-Mc. band (except for fading caused by local effects).

Measurements have shown that the angles useful for communication on the 14-Mc. band are from 3° to about 30°—angles above about 15° being useful only for local work. On the 28-Mc. band measurements have shown that the useful angles range from about 3° to 18°—angles above about 12° being useful only for local (less than 3000 miles) work. These figures assume normal propagation by virtue of the F_2 layer.

Angle of Radiation of Typical Antennas and Arrays

It now becomes of interest to determine the amount of radiation available at these useful lower angles of radiation from commonly used antennas and antenna arrays. Figure 6 shows relative output voltage plotted against elevation angle (wave angle) in degrees above the horizontal, for horizontal and vertical doublets elevated 0.6 wavelength above two types of ground. It is obvious by inspection of the curves that a horizontal dipole mounted at this height above ground (20 feet on the 28-Mc. band) is radiating only a small amount of energy at angles useful for communication on the 28-Mc. band. Most of the energy is being radiated uselessly upward. The vertical antenna above a good reflecting surface appears much better in this respect—and this fact has been proven many times by actual installations.

It might immediately be thought that the amount of radiation from a horizontal or vertical dipole could be increased by raising the antenna higher above the ground. This is true to an extent in the case of the horizontal dipole; the low-angle radiation does increase *slowly* after a height of 0.6 wavelength is reached but at the expense of greatly increased high-angle radiation and the formation of a number of nulls in the elevation pattern. No signal can be transmitted or received at the elevation angles where these nulls have been formed. Tests have shown that a center height of 0.6 wavelength for a vertical dipole (0.35 wavelength to the bottom end) is about optimum for this type of array.

Figure 7 shows the relative gain in field strength for different elevation angles of radiation for a horizontal dipole at different heights above ground. The effect of placing a hori-

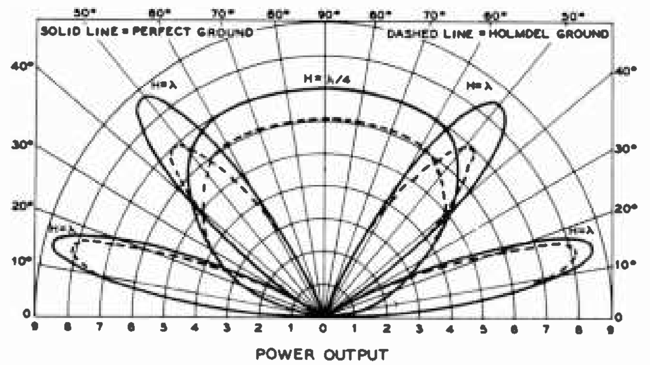


Figure 8.

VERTICAL RADIATION PATTERNS.

Showing the vertical radiation patterns for half-wave antennas (or colinear half-wave or extended half-wave antennas) at different heights above average ground and perfect ground. Note that such antennas one-quarter wave above ground concentrate most radiation at the very high angles which are useful for communication only on the lower frequency bands. Antennas one-half wave above ground are not shown, but the elevation pattern shows one lobe on each side at an angle of 30° above horizontal.

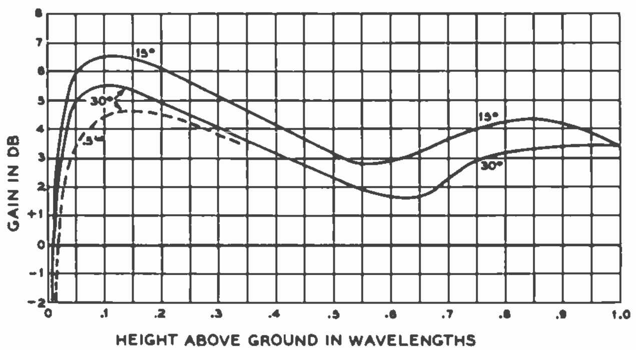


Figure 9.

HEIGHT AGAINST GAIN FOR A FLAT-TOP BEAM.

Showing the effect of height above ground on the gain of a single-section flat-top beam with one-eighth wave spacing over a horizontal half-wave antenna at the same height above ground for vertical angles of 15° and 30°. Multi-section flat-top beams will show approximately the same gain over a colinear half-wave antenna of the same overall length for the same height above ground. The effect of 0.5-ohm loss resistance in the flat-top beam is shown by the dashed curve.

zontal dipole still higher above ground is illustrated by figure 8 showing the vertical radiation pattern of a dipole elevated one wavelength above ground. It is easily seen by reference to figure 8 (and figure 10 which shows the radiation from a dipole at ¾ wave height) that a large percentage of the total radiation from the dipole is being radiated at relatively high angles which are useless for communication on the 14-Mc. and 28-Mc. bands. Thus we see that in order to obtain a worthwhile increase in the ratio of low-angle radiation to high-angle radiation it is necessary to place the antenna high above ground, *and in addition* it is necessary to use additional means for suppressing high-angle radiation.

High-angle radiation can be suppressed, and this radiation can be added to that going out at low angles, only through the use of some sort of *directive* antenna system. There are three general types of antenna arrays composed of dipole elements commonly used which concentrate radiation at the lower more effective angles for high-frequency communication. These types are: (1) the close-spaced out-of-phase system as exemplified by the "flat-top beam" or W8JK array, (2) the wide-spaced in-phase system as exemplified by the "lazy-H" and similar arrays, and (3) the close-spaced parasitic system as exemplified by the "three-element rotary" and similar arrays using varying numbers of elements and different spacings.

A comparison between the radiation from a dipole, a "flat-top beam" and a pair of dipoles stacked one above the other (half of a "lazy H"), in each case with the top of the antenna at a height of ¾ wavelength is shown in figure 11. The improvement in the amplitude of low-angle radiation at the expense of the useless high-angle radiation with these simple arrays as contrasted to the dipole is quite marked.

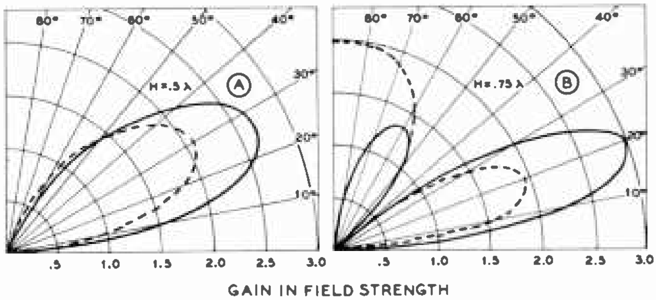


Figure 10.
VERTICAL RADIATION PATTERNS.

Showing vertical-plane radiation patterns of a horizontal single-section flat-top beam with one-eighth wave spacing (solid curves) and a horizontal half-wave antenna (dashed curves) when both are 0.5 wavelength (A) and 0.75 wavelength (B) above ground.

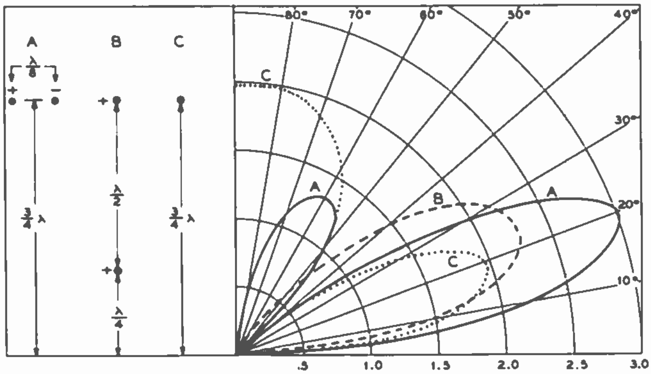


Figure 11.
COMPARATIVE VERTICAL RADIATION PATTERNS.

Showing the vertical radiation patterns of a horizontal single-section flat-top beam (A), an array of two stacked horizontal in-phase half-wave elements—half of a "Lazy H"—(B), and a horizontal dipole (C). In each case the top of the antenna system is 0.75 wavelength above ground, as shown to the left of the curves.

Effect of Average Ground on Antenna Radiation Discussions of antenna radiation usually are based upon the perfect-ground assumption, in order to cover the subject in the most simple manner. When the earth is less than a perfect conductor, it becomes a dielectric or, perhaps in an extreme case, a "leaky insulator."

The resulting change in the vertical pattern of a horizontal antenna is shown in figure 6. The ground constants, in this case, are for flat farmland, which probably is similar to mid-western farmland. The ocean is the closest practical approach to a theoretically perfect ground. It will be noted that there is only a slight loss in power due to the imperfect ground as compared to the ocean horizontal.

The effect of the earth on the radiation pattern of a vertical dipole is much greater. Radiation from a half-wavelength vertical wire is severely reduced by deficiencies of the ground.

12-6 Bandwidth

The bandwidth of an antenna or an antenna array is a function primarily of the radiation resistance and of the shape of the conductors which make up the antenna system. For arrays of essentially similar construction the bandwidth (or the deviation in frequency which the system can handle without mismatch) is increased with increasing radiation resistance, and the bandwidth is increased with the use of conductors of larger diameter (smaller ratio of length to diameter). This is to say that if an array of any type is constructed of large diameter tubing or spaced wires, its bandwidth will be greater than that of a similar array constructed of single wires.

The radiation resistance of antenna arrays of the types mentioned in the previous paragraphs may be increased through the use of wider spacing between elements. With increased radiation resistance in such arrays the "radiation efficiency" increases since the ohmic losses within the conductors become a smaller percentage of the radiation resistance, and the bandwidth is increased proportionately.

12-7 Propagation of Radio Waves

The preceding sections have discussed the manner in which an electromagnetic-wave or radio-wave field may be set up by a radiating system. However, for this field to be useful for communication it must be propagated to some distant point where it may be received, or where it may be reflected so that it may be recived at some other point. Radio waves may be propagated to a remote point by either or both of two general methods. Propagation may take place as a result of the *ground wave,* or as a result of the *sky wave* or *ionospheric wave.*

The Ground Wave The term "ground wave" actually includes several different types of waves which usually are called: (1) the *surface wave,* (2) the *direct wave,* and (3) the *ground-reflected wave.* The latter two waves combine at the receiving antenna to form the *resultant wave* or the *space wave.* The distinguishing characteristic of the components of the ground wave is that all travel along or over the surface of the earth, so that they are affected by the conductivity and terrain of the earth's surface.

The Ionospheric Wave or Sky Wave Intense bombardment of the upper regions of the atmosphere by radiations from the sun results in the formation of ionized layers. These ionized layers, which form the *ionosphere,* have the capability of reflecting or refracting radio waves which impinge upon them. A radio wave which has been propagated as a result of one or more reflections from the ionosphere is known as an "ionospheric wave" or a "sky wave." Such waves make possible long distance radio communication. Propagation of radio signals by ionospheric waves is discussed in detail in Section 12-9.

12-8 Ground-Wave Communication

As stated in the preceding paragraph, the term "ground wave" applies both to the "sur-

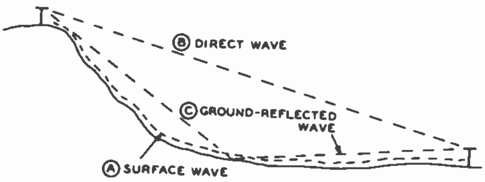


Figure 12.
GROUND-WAVE SIGNAL PROPAGATION.
The illustration above shows the three components of the ground wave: (A), the surface wave; (B), the direct wave; and (C), the ground-reflected wave. The direct wave and the ground-reflected wave combine at the receiving antenna to make up the space wave.

face wave" and to the "space wave" (the resultant wave from the combination of the direct wave and the ground-reflected wave) or to a combination of the two. The three waves which may combine to make up the ground wave are illustrated in figure 12.

The Surface Wave The surface wave is that wave which we normally receive from a standard broadcast station. It travels directly along the ground and terminates on the earth's surface. Since the earth is a relatively poor conductor, the surface wave is attenuated quite rapidly. The surface wave is attenuated less rapidly as it passes over sea water, and the attenuation decreases for a specific distance as the frequency is decreased. The rate of attenuation with distance becomes so large as the frequency is increased above about 3 Mc. that the surface wave becomes of little value for communication.

The Space Wave The resultant wave or space wave is illustrated in figure 12 by the combination of (B) and (C). It is this wave path, which consists of the combination of the direct wave and the ground-reflected wave at the receiving antenna, which is the *normal* path of signal propagation for line-of-sight or near line-of-sight communication or FM and TV reception on frequencies above about 40 Mc.

Below line-of-sight over plane earth or water, when the signal source is effectively at the horizon, the ground-reflected wave does not exist, so that the "direct" wave is the only component which goes to make up the space wave. But both the signal source and the receiving antenna are elevated with respect to the intervening terrain, the ground-reflected wave is present and adds vectorially to the direct wave at the receiving antenna. The vectorial addition of the two waves, which travel over different path lengths since one of the waves has been reflected from the ground, results in an interference pattern. The interference between the two waves brings about a cyclic variation in signal strength as the receiving antenna is raised above the ground. This effect is illustrated in figure 13. From this figure it can be seen that best space-wave reception of a v-h-f signal often will be obtained with the receiving antenna quite close to the ground. This subject, along with other aspects of v-h-f signal propagation and reception, are discussed in considerable detail in a recently published book on fringe-area TV reception.*

The distance from an elevated point to the geometrical horizon is given by the approximate equation: $d = 1.22\sqrt{H}$ where the distance d is in miles and the antenna height H is in feet. This equation must be applied separately to the transmitting and receiving antennas and the results added. However, refraction and diffraction of the signal around the spherical earth cause a smaller reduction in field strength than would occur in the absence of such bending, so that the average radio horizon is somewhat beyond the geometrical horizon. The equation $d = 1.4\sqrt{H}$ is sometimes used for determining the radio horizon.

Tropospheric Propagation Propagation by signal bending in the lower atmosphere, called tropospheric propagation, can result in the reception of signals over a much greater distance than would be the case if the lower atmosphere were homogeneous. In a homogeneous or well-mixed lower atmosphere, called a *normal* or *standard* atmosphere, there is a gradual and uniform decrease in index of refraction with height. This effect is due to the combined effects of a decrease in temperature, pressure, and water-vapor content with height.

This gradual decrease in refractive index

*"Better TV Reception," by W. W. Smith and R. L. Dawley, published by Editors and Engineers, Ltd., Santa Barbara, Calif.

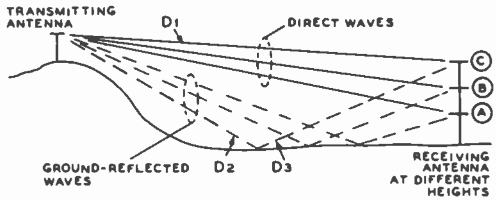


Figure 13.

WAVE INTERFERENCE WITH HEIGHT.

When the source of a horizontally-polarized space-wave signal is above the horizon, the received signal at a distant location will go through a cyclic variation as the antenna height is progressively raised. This is due to the difference in total path length between the direct wave and the ground-reflected wave, and to the fact that this path length difference changes with antenna height. When the path length difference is such that the two waves arrive at the receiving antenna with a phase difference of 360° or some multiple of 360°, the two waves will appear to be in phase as far as the antenna is concerned and maximum signal will be obtained. On the other hand, when the antenna height is such that the path length difference for the two waves causes the waves to arrive with a phase difference of an odd multiple of 180° the two waves will substantially cancel, and a null will be obtained at that antenna height. The difference between D_1 and D_2 plus D_3 is the path-length difference. Note also that there is an additional 180° phase shift in the ground-reflected wave at the point where it is reflected from the ground. It is this latter phase shift which causes the space-wave field intensity of a horizontally polarized wave to be zero with the receiving antenna at ground level.

with height causes waves radiated at very low angles with respect to the horizontal to be bent downward slightly in a curved path. The result of this effect is that such waves will be propagated beyond the "true" or "geometrical" horizon. In a so-called standard atmosphere the effect of the curved path is the same as though the radius of the earth were increased by approximately one third. This condition extends the horizon by approximately 30 per cent for normal propagation, and the extended horizon is known as the "radio path horizon," mentioned before.

Conditions Leading to Tropospheric Stratification When the temperature, pressure, or water-vapor content of the atmosphere does not change smoothly with rising altitude, the discontinuity or stratification will result in the reflection or refraction of incident v-h-f signals. Ordinarily this condition is more prevalent at night and in the summer. In certain areas, such as along the west coast of North America, it is frequent enough to be considered normal. Signal strength decreases slowly with distance and, if the favorable condition in the lower atmosphere covers sufficient area, the range is limited only by the transmitter power, antenna gain, receiver sensitivity, and signal-to-noise ratio. There is no skip distance. Usually, transmission due to this condition is accompanied by slow fading, although fading can be violent at a point where direct waves of about the same strength are also received.

Bending in the troposphere, which refers to the region from the earth's surface up to about 10 kilometers, is more likely to occur on days when there are stratus clouds than on clear, cool days with a deep blue sky. The temperature or humidity discontinuities may be broken up by vertical convection currents over land in the daytime but are more likely to continue during the day over water. This condition is in some degree predictable from weather information several days in advance. It does not depend on the sunspot cycle. Like direct communication, best results require similar antenna polarization or orientation at both the transmitting and receiving ends, whereas in transmission via a reflection in the ionosphere (that part of the atmosphere between about 50 and 500 kilometers high) it makes little difference whether antennas are similarly polarized.

Duct Formation When bending conditions are particularly favorable they may give rise to the formation of a *duct* which can propagate waves with very little attenuation over great distances in a manner similar to the propagation of waves through a wave guide. *Guided propagation* through a duct in the atmosphere can give quite remarkable transmission conditions. However, such ducts usually are formed only on an over-water path. The depth of the duct over the water's surface may be only 20 to 50 feet, or it may be 1000 feet deep or more. Ducts exhibit a low-frequency cutoff characteristic similar to a wave guide. The cutoff frequency is determined by depth of the duct and by the strength of the discontinuity in refractive in-

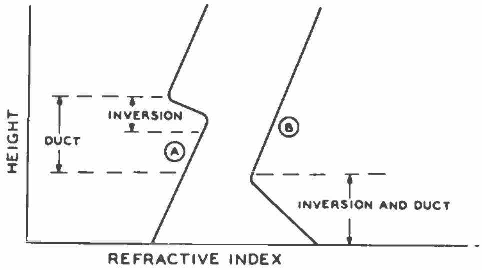


Figure 14.
ILLUSTRATING DUCT TYPES.
Showing two types of variation in refractive index with height which will give rise to the formation of a duct. An elevated duct is shown at (A), and a ground-based duct is shown at (B). Such ducts can propagate ground-wave signals far beyond their normal range.

dex at the upper surface of the duct. The lowest frequency that can be propagated by such a duct seldom goes below 50 Mc., and usually will be greater than 100 Mc. even along the Pacific Coast.

Stratospheric Reflection Communication by virtue of stratospheric reflection can be brought about during magnetic storms, aurora borealis displays, and during meteor showers. Dx communication during extensive meteor showers is characterized by frequent bursts of great signal strength followed by a rapid decline in strength of the received signal. The motion of the meteor forms an ionized trail of considerable extent which can bring about effective reflection of signals. However, the ionized region persists only for a matter of seconds so that a shower of meteors is necessary before communication becomes possible.

The type of communication which is possible during visible displays of the aurora borealis and during magnetic storms has been called *aurora-type dx.* These conditions reach a maximum somewhat after the sunspot cycle peak, possibly because the spots on the sun are nearer to its equator (and more directly in line with the earth) in the latter part of the cycle. Ionospheric storms generally accompany magnetic storms. The normal layers of the ionosphere may be churned or broken up, making radio transmission over long distances difficult or impossible on high frequencies. Unusual conditions in the ionosphere sometimes modulate v-h-f waves so that a definite tone or noise modulation is noticed even on transmitters located only a few miles away.

A peculiarity of this type of auroral propagation of v-h-f signals in the northern hemisphere is that directional antennas usually must be pointed in a northerly direction for best results for transmission or reception, regardless of the direction of the other station being contacted. Distances out to 700 or 800 miles have been covered during magnetic storms, using 30 and 50 Mc. transmitters, with little evidence of any silent zone between the stations communicating with each other. Generally, voice-modulated transmissions are difficult or impossible due to the tone or noise modulation on the signal. Most of the communication of this type has taken place by c.w. or by tone modulated waves with a keyed carrier.

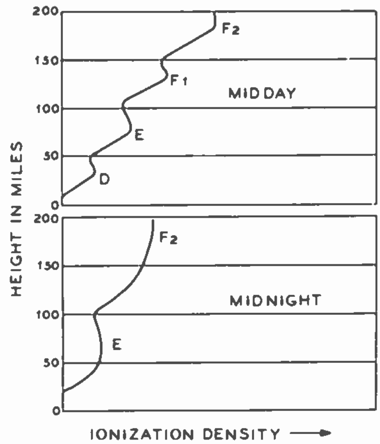


Figure 15.
IONIZATION DENSITY IN THE IONOSPHERE.

Showing typical ionization density of the ionosphere in mid-summer. Note that the F_1 and D layers disappear at night, and that the density of the E layer falls to such a low value that it is ineffective.

12-9 Ionospheric Propagation

Propagation of radio waves for communication on frequencies between perhaps 3 and 30 Mc. is normally carried out by virtue of *ionospheric reflection* or *refraction*. Under conditions of abnormally high ionization in the ionosphere, communication has been known to have taken place by ionospheric reflection on frequencies higher than 50 Mc.

The ionosphere consists of layers of ionized gas located above the stratosphere, and extending up to possibly 300 miles above the earth. Thus we see that high-frequency radio waves may travel over short distances in a direct line from the transmitter to the reeciver, or they can be radiated upward into the ionosphere to be bent downward in an indirect ray, returning to earth at considerable distance from the transmitter. The wave reaching a receiver via the ionosphere route is termed a *sky wave*. The wave reaching a rece'ver by traveling in a direct line from the transmitting antenna to the receiving antenna is commonly called a *ground wave* and has been discussed in Section 12-8.

The amount of bending at the ionosphere which the sky wave can undergo depends upon its frequency, and the amount of *ionization* in the ionosphere, which is in turn dependent upon radiation from the sun. The sun increases the density of the ionosphere layers, and lowers their effective height. For this reason, the ionosphere acts very differently at different times of day, and at different times of the year.

The higher the frequency of a radio wave, the farther it penetrates the ionosphere, and the less it tends to be bent back toward the earth. The lower the frequency, the more easily the waves are bent, and the less they penetrate the ionosphere. 160-meter and 80-meter signals will usually be bent back to earth even when sent straight up, and may be considered as being *reflected* rather than *refracted*. As the frequency is raised beyond about 5,000 kc. (dependent upon the critical frequency of the ionosphere at the moment), it is found that waves transmitted at angles higher than a certain critical angle *never return to earth*. Thus, on the higher frequencies, it is necessary to confine radiation to low angles, since the high angle waves simply penetrate the ionosphere and are lost.

The F_2 Layer The higher of the two major reflection regions of the ionosphere is called the F_2 layer. This layer has a virtual height of approximately 175 miles at night, and in the daytime it splits up into two layers, the upper one being called the F_2 layer and the lower being called the F_1 layer. The height of the F_2 layer during daylight hours is normally about 250 miles on the average and the F_1 layer often has a height of as low as 140 miles. It is the F_2 layer which supports all nighttime dx communication and nearly all daytime dx propagation.

The E Layer Below the F_1 layer is another layer, called the *E* layer, which is of importance in daytime communication over moderate distances in the frequency range between 3 and 8 Mc. This layer has an almost constant height at about 70 miles. Since the re-combination time of the ions at this height is rather short, the *E* layer disappears almost completely a short time after local sunset.

The D Layer Below the *E* layer at a height of about 35 miles is an *absorbing* layer, called the *D* layer, which exists in the middle of the day in the summertime. The layer also exists during midday in the winter time during periods of high solar activity, but the layer disappears completely at night. It is this layer which causes high absorption of signals in the medium and high-frequency range during the middle of the day.

Critical Frequency The critical frequency of an ionospheric layer is the highest frequency which will be reflected when the wave strikes the layer at vertical incidence. The critical frequency of the most highly ionized layer of the ionosphere may be as low as 2 Mc. at night and as high as 12 to 13 Mc. in the middle of the day. The critical frequency is directly of interest in that a "skip-distance" zone will exist on all frequencies *greater* than the highest critical frequency at that time. The critical frequency is a measure of the density of ionization of the reflecting layers. The higher the critical frequency the greater the density of ionization.

Maximum Usable Frequency The *maximum usable frequency* or *m.u.f.* is of great importance in long-distance communication since this frequency is the highest that can be used for communication between any two specified areas. The m.u.f. is the highest frequency at which a wave projected into space in a certain direction will be returned to earth in a specified region by ionospheric reflection. The m.u.f. is highest at noon or in the early afternoon and is highest in periods of greatest sunspot activity, often going to frequencies higher than 50 Mc.

The m.u.f. often drops to frequencies below 10 Mc. in the early morning hours. The high m.u.f. in the middle of the day is brought about by reflection from the F_2 layer. M.u.f. data is published periodically in the magazines devoted to amateur work, and the m.u.f. can be calculated with the aid of *Basic Radio Propagation Predictions,* CRPL-D, published monthly by the Government Printing Office, Washington, D.C.

Absorption and Optimum Working Frequency The optimum working frequency for any particular direction and distance is usually about 15 per cent less than the m.u.f. for contact with that particular location. The absorption by the ionosphere becomes greater and greater as the operating frequency is progressively lowered below the m.u.f. It is this condition which causes signals to increase tremendously in strength on the 14-Mc. and 28-Mc. bands just before the signals drop completely out. At the time when the signals are greatest in amplitude the operating frequency is equal to the m.u.f. Then as the signals drop out the m.u.f. has become lower than the operating frequency.

Skip Distance The shortest distance from a transmitting location at which signals reflected from the ionosphere can be returned to the earth is called the *skip distance.* As was mentioned above under *Critical Frequency* there is no skip distance for a frequency below the critical frequency of the most highly ionized layer of the ionosphere at the time of transmission. However, the skip distance is always present on the 14-Mc. band and is almost always present on the 3.5-Mc. and 7-Mc. bands at night. The actual measure of the skip distance is the distance between the point where the ground wave falls to zero and the point where the sky wave begins to return to earth. This distance may vary from 40 to 50 miles on the 3.5-Mc. band to thousands of miles on the 28-Mc. band.

The Sporadic-E Layer Occasional patches of extremely high ionization density appear at intervals throughout the year at a height approximately equal to that of the *E* layer. These patches, called the *sporadic-E* layer may be very small or may be up to several hundred miles in extent. The critical frequency of the *sporadic-E* layer may be greater than twice that of the normal ionosphere layers which exist at the same time.

It is this *sporadic-E* condition which provides "short-skip" contacts from 400 to perhaps 1200 miles on the 28-Mc. band in the evening. It is also the *sporadic-E* condition which provides the more common type of "band opening" experienced on the 50-Mc. band when very loud signals are received from stations from 400 to 1200 miles distant.

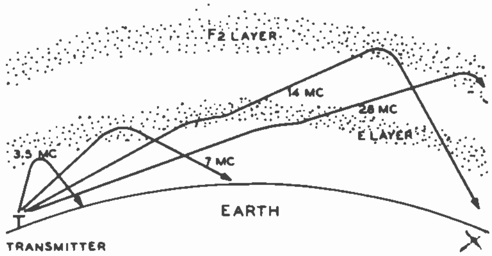


Figure 16.
IONOSPHERE-REFLECTION WAVE PATHS.

Showing typical ionosphere-reflection wave paths during daylight hours when ionization density is such that frequencies as high as 28 Mc. will be returned to earth. The distance between ground-wave range and that range where the ionosphere-reflected wave of a specific frequency first will be returned to earth is called the skip distance.

Cycles in Ionosphere Activity The ionization density of the ionosphere is determined by the amount of radiation (probably ultra violet) which is being received from the sun. Consequently, ionosphere activity is a function of the amount of radiation of the proper character being emitted by the sun and is also a function of the relative aspect of the regions in the vicinity of the location under discussion to the sun. There are four main cycles in ionosphere activity. These cycles are: the daily cycle which is brought about by the rotation of the earth, the 27-day cycle which is caused by the rotation of the sun, the seasonal cycle which is caused by the movement of the earth in its orbit, and the 11-year cycle which is a cycle in sunspot activity. The effects of these cycles are superimposed insofar as ionosphere activity is concerned. Also, the cycles are subject to short term variations as a result of magnetic storms and similar terrestrial disturbances.

Fading The lower the angle of radiation of the wave, with respect to the horizon, the farther away will the wave return to earth, and the greater the skip distance. The wave can be reflected back up into the ionosphere by the earth, and then be reflected back down again, causing a second skip distance area. The drawing of figure 16 shows the multiple reflections possible. When the receiver receives signals which have traveled over more than one path between transmitter and receiver, the signal impulses will not all arrive at the same instant, as they do not all travel the same distance. When two or more signals arrive in the same phase at the receiving antenna, the resulting signal in the receiver will be quite strong. On the other hand, if the signals arrive 180° out of phase, so they tend to cancel each other, the received signal will drop,—perhaps to zero if perfect cancellation occurs. This explains why high-frequency signals are subject to fading.

Fading can be greatly reduced on the high frequencies by using a transmitting antenna with sharp vertical directivity, thus cutting down the number of possible paths of signal arrival. A receiving antenna with similar characteristics (sharp vertical directivity) will further reduce fading. It is desirable, when using antennas with sharp vertical directivity, to use the lowest vertical angle consistent with good signal strength for the frequency used.

Transmission Lines

For many reasons it is desirable to place an antenna or radiating system as high and in the clear as is physically possible, utilizing some form of nonradiating transmission line to carry energy with as little loss as possible from the transmitter to the radiating antenna, and conversely from the antenna to the receiver.

There are many different types of transmis-

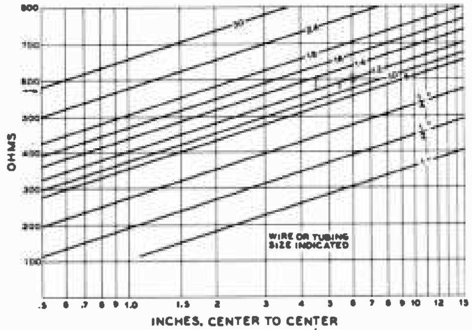


Figure 17.
CHARACTERISTIC IMPEDANCE OF TYPICAL TWO-WIRE OPEN LINES.

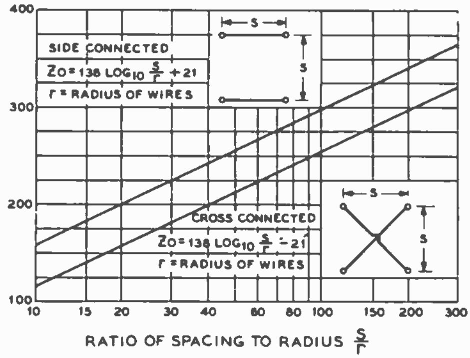


Figure 18.
CHARACTERISTIC IMPEDANCE OF FOUR-WIRE CROSS-CONNECTED AND SIDE-CONNECTED OPEN LINES.

sion lines and, generally speaking, practically any type of transmission line or feeder system may be used with any type of antenna. However, mechanical or electrical considerations often make one type of transmission line better adapted for use to feed a particular type of antenna than any other type.

Transmission lines for carrying r-f energy are of two general types: *non-resonant* and *resonant*. A non-resonant transmission line is one on which a successful effort has been made to eliminate reflections from the termination (the antenna in the transmitting case and the receiver for a receiving antenna) and hence one on which standing waves do not exist or are relatively small in magnitude. A resonant line, on the other hand, is a transmission line on which standing waves of appreciable magnitude do appear, either through inability to match the characteristic impedance of the line to the termination or through intentional design.

The principal types of transmission line in use or available at this time include the open-wire line (two-wire and four-wire types), two-wire solid-dielectric line ("Twin-Lead" and similar ribbon or tubular types), two-wire polyethylene-filled shielded line, coaxial line of the solid-dielectric, beaded, stub-supported, or pressurized type, rectangular and cylindrical wave guide, and the single-wire feeder operated against ground. The significant characteristics of the more popular types of transmission line available at this time are given in the chart of figure 19.

12-10 Non-Resonant Transmission Lines

A non-resonant or untuned transmission line is a line with negligible standing waves. Hence, a non-resonant line is a line carrying r-f power only in one direction — from the source of energy to the load.

Physically, the line itself should be *identical throughout its length*. There will be a smooth distribution of voltage and current throughout its length, both tapering off very slightly towards the load end of the line as a result of line losses. The attenuation (loss) in certain types of untuned lines can be kept very low for line lengths up to several thousand feet. In other types, particularly where the dielectric is not air (such as in the twisted-pair line), the losses may become excessive at the higher frequencies, unless the line is relatively short.

Transmission-Line Impedance All transmission lines have distributed inductance, capacitance and resistance. Neglecting the resistance, as it is of minor importance in short lines, it is found that the *inductance and capacitance per unit length* determine the characteristic or surge impedance of the line. Thus, the surge impedance depends upon the nature and spacing of the conductors, and the dielectric separating them.

CHARACTERISTICS OF COMMON TRANSMISSION LINES

	ATTENUATION db/100 FEET VSWR = 1.0			VELOCITY FACTOR V	μμFD PER FT.	REMARKS
	30 MC	100 MC	300 MC			
OPEN WIRE LINE, № 12 COPPER.	0.15	0.3	0.8	0.98-0.99	—	BASED UPON 4" SPACING BELOW 50 MC.; 2" SPACING ABOVE 50 MC. RADIATION LOSSES INCLUDED. CLEAN, LOW LOSS CERAMIC INSULATION ASSUMED. RADIATION HIGH ABOVE 150 MC.
RIBBON LINE, REC TYPE, 300 OHMS. (7/28 CONDUCTORS)	0.86	2.2	5.3	0.82ψ	6ψ	FOR CLEAN, DRY LINE. WET WEATHER PERFORMANCE RATHER POOR. BEST LINE IS SLIGHTLY CONVEX. AVOID LINE THAT HAS CONCAVE DIELECTRIC. SUITABLE FOR LOW POWER TRANSMITTING APPLICATIONS. LOSSES INCREASE AS LINE WEATHERS. HANDLES 400 WATTS AT 30 MC. IF VSWR IS LOW.
TUBULAR "TWIN-LEAD" REC. TYPE, 300 OHMS, 5/16" O.D. (AMPHENOL TYPE 14-271)	—	—	—	—	—	CHARACTERISTICS SIMILAR TO RECEIVING TYPE RIBBON LINE EXCEPT FOR MUCH BETTER WET WEATHER PERFORMANCE.
RIBBON LINE, TRANS. TYPE, 300 OHMS.	—	—	—	—	—	CHARACTERISTICS VARY SOMEWHAT WITH MANUFACTURER, BUT APPROXIMATE THOSE OF RECEIVING TYPE RIBBON EXCEPT FOR GREATER POWER HANDLING CAPABILITY AND SLIGHTLY BETTER WET WEATHER PERFORMANCE.
TUBULAR "TWIN-LEAD" TRANS. TYPE, 7/16 O.D. (AMPHENOL 14-076)	0.85	2.3	5.4	0.79	6.1	FOR USE WHERE RECEIVING TYPE TUBULAR "TWIN-LEAD" DOES NOT HAVE SUFFICIENT POWER HANDLING CAPABILITY. WILL HANDLE 1 KW AT 30 MC. IF VSWR IS LOW.
RIBBON LINE, RECEIVE. TYPE, 150 OHMS.	1.1	2.7	6.0	0.77ψ	10ψ	USEFUL FOR QUARTER WAVE MATCHING SECTIONS. NO LONGER WIDELY USED AS A LINE.
RIBBON LINE, RECEIVE. TYPE, 75 OHMS.	2.0	5.0	11.0	0.68ψ	19ψ	USEFUL MAINLY IN THE H-F RANGE BECAUSE OF EXCESSIVE LOSSES AT V-H-F AND U-H-F. LESS AFFECTED BY WEATHER THAN 300 OHM RIBBON.
RIBBON LINE, TRANS. TYPE, 75 OHMS.	1.5	3.9	8.0	0.71ψ	18ψ	VERY SATISFACTORY FOR TRANSMITTING APPLICATIONS BELOW 30 MC. AT POWERS UP TO 1 KW. NOT SIGNIFICANTLY AFFECTED BY WET WEATHER.
RG-8/U COAX (52 OHMS)	1.0	2.1	4.2	0.66	29.5	WILL HANDLE 2 KW AT 30 MC. IF VSWR IS LOW. 0.4" O.D. 7/21 CONDUCTOR.
RG-11/U COAX (75 OHMS)	0.94	1.9	3.8	0.66	20.5	WILL HANDLE 1.4 KW AT 30 MC. IF VSWR IS LOW. 0.4" O.D. 7/26 CONDUCTOR.
RG-17/U COAX (52 OHMS)	0.38	0.85	1.6	0.66	29.5	WILL HANDLE 7.8 KW. AT 30 MC. IF VSWR IS LOW. 0.87" O.D. 0.19" DIA. CONDUCTOR
RG-58/U COAX (53 OHMS)	1.95	4.1	8.0	0.66	28.5	WILL HANDLE 430 WATTS AT 30 MC. IF VSWR IS LOW. 0.2" O.D. № 20 CONDUCTOR.
RG-59/U COAX (73 OHMS)	1.9	3.6	7.0	0.66	21	WILL HANDLE 680 WATTS AT 30 MC. IF VSWR IS LOW. 0.24" O.D. № 22 CONDUCTOR.
TV-59 COAX (72 OHMS)	2.0	4.0	7.0	0.66	22	"COMMERCIAL" VERSION OF RG-59/U FOR LESS EXACTING APPLICATIONS. LESS EXPENSIVE.
RG-22/U SHIELDED PAIR (95 OHMS)	1.7	3.0	5.5	0.66	16	FOR SHIELDED, BALANCED-TO-GROUND APPLICATIONS. VERY LOW NOISE PICK UP. 0.4" O.D.
K-111 SHIELDED PAIR (300 OHMS)	2.0	3.5	6.1	—	4	DESIGNED FOR TV LEAD-IN IN NOISY LOCATIONS. LOSSES HIGHER THAN REGULAR 300 OHM RIBBON, BUT DO NOT INCREASE AS MUCH FROM WEATHERING

ψ APPROXIMATE. EXACT FIGURE VARIES SLIGHTLY WITH MANUFACTURER.

FIGURE 19.

Speaking in electrical terms, the characteristic impedance of a transmission line is simply the ratio of the voltage across the line to the current which is flowing, the same as is the case with a simple resistor: $Z_0 = E/I$. Also, in a substantially loss-less line (one whose attenuation per wavelength is small) the energy stored in the line will be equally divided between the capacitive field and the inductive field which serve to propagate the energy along the line. Hence the characteristic impedance of a line may be expressed as:

$$Z_0 = \sqrt{L/C}$$

Two-Wire Open Line A two-wire transmission system is easy to construct. Its surge impedance can be calculated quite easily, and when properly adjusted and balanced to ground, with a conductor spacing which is negligible in terms of the wavelength of the signal carried, undesirable feeder radiation is minimized; the current flow in the adjacent wires is in opposite directions, and the magnetic fields of the two wires are in opposition to each other. When a two-wire line is terminated with the equivalent of a pure resistance equal to the characteristic impedance of the line, the line becomes a non-resonant line.

Expressed in physical terms, the characteristic impedance of a two-wire open line is equal to:

$$Z_0 = 276 \log_{10} \frac{2S}{d}$$

Where:

S is the exact distance between wire centers in some convenient unit of measurement, and

d is the diameter of the wire measured in the same units as the wire spacing, S.

Since $\frac{2S}{d}$ expresses a ratio only, the units of measurement may be centimeters, millimeters, or inches. This makes no difference in the answer, so long as the substituted values for S and d are in the *same units.*

The equation is accurate *so long as the wire spacing is relatively large as compared to the wire diameter.*

Surge impedance values of less than 200 ohms are seldom used in the open-type two-wire line, and, even at this rather high value of Z_0, the wire spacing S is uncomfortably close, being only 5.3 times the wire diameter d.

Figure 17 gives in graphical form the surge impedance of practicable two-wire lines. The chart is self-explanatory, and is sufficiently accurate for practical purposes.

Four-Wire Open-Wire Lines Under certain conditions it is desirable to have the power-handling capacity of an open-wire line and yet have a line whose characteristic impedance is of the order of 200 ohms rather than being in the vicinity of 500 ohms for practicable two-wire lines. The four-wire open line has this lowered characteristic impedance and has been used in many applications as a quarter-wave matching transformer (Q section). Design data for four-wire lines are given in figure 18. The four wires of such a transmission line are spaced on the corners of a square (or equally spaced on the circumference of a circle) and *opposite* wires are connected together at each end of the line. Make sure that the same wires are paralleled at each end of the transmission line section.

The spacers for the line can be constructed from crossed strips of polystyrene cemented with polystyrene cement or of crossed strips of lucite cemented with chloroform. Plastic iced-tea coasters of suitable diameter can also be used as spacers. Experience has indicated that the spacers should be placed every 2 or 3 feet along the line and that the line should be operated under some tension to prevent twisting.

The Double Two-Wire Line It sometimes is of importance to have an open-wire line with a characteristic impedance higher than the practicable value for a four-wire open line, but lower than the practicable range of values for a standard two-wire line. Such a line can be constructed in the same manner as a four-wire cross-connected line, except that the *adjacent* wires at one end are connected together, and the same wires are paralleled at the other end. Such a line will have a characteristic impedance 42 ohms higher than a four-wire cross-connected line with the same conductor diameter and spacing. The equation for the characteristic impedance of such a line is:

$$Z_0 = 138 \log_{10} \frac{S}{r} + 21 \text{ ohms},$$

while the equation for the impedance of a four-wire cross-connected line is:

$$Z_0 = 138 \log_{10} \frac{S}{r} - 21 \text{ ohms}.$$

In each case S represents the spacing between adjacent conductors while r represents the radius of the wires. These expressions, similar to the equation for the characteristic impedance of a two-wire line, are accurate only for cases where the wire spacing is much greater than the wire diameter.

Ribbon and Tubular Transmission Line Instead of using spacer insulators placed periodically along the transmission line it is possible to mold the line conductors into a ribbon or tube of flexible low-loss dielectric material. Such line, with polyethylene dielectric, is used in enormous quantities as the lead-in transmission line for FM and TV receivers. The line is available from several manufacturers in the ribbon and tubular configuration, with characteristic impedance values from 75 to 300 ohms. Receiving types, and transmitting types for power levels up to one kilowatt in the h-f range, are listed with their pertinent characteristics, in the table of figure 19.

Coaxial Line Several types of coaxial cable have come into wide use for feeding power to an antenna system. A cross-sectional end view of a coaxial cable (sometimes called concentric cable or line) is shown in figure 20.

As in the parallel-wire line, the power lost in a properly terminated coaxial line is the sum of the effective resistance losses along the length of the cable and the dielectric losses between the two conductors.

Of the two losses, the effective resistance loss is the greater; since it is largely due to the skin effect, the line loss (all other conditions the same) will increase directly as the square root of the frequency.

Figure 20 shows that, instead of having two conductors running side by side, one of the conductors is placed *inside* of the other. Since the outside conductor completely shields the inner one, no radiation takes place. The conductors may both be tubes, one within the other; the line may consist of a solid wire within a tube, or it may consist of a stranded or solid inner conductor with the outer conductor made up of one or two wraps of copper shielding braid.

In the type of cable most popular for military and non-commercial use the inner conductor consists of a heavy stranded wire, the outer conductor consists of a braid of copper wire, and the inner conductor is supported within the outer by means of a semi-solid dielectric of exceedingly low loss characteristics called polyethylene. The Army-Navy designation on one size of this cable suitable for power levels up to one kilowatt at frequencies as high as 30 Mc. is AN/RG-8/U. The outside diameter of this type of cable is approximately one-half inch. The characteristic impedance of this cable type is 52 ohms, but other similar types of greater and smaller power-handling capacity are available in impedances of 52, 75, and 95 ohms.

When using solid dielectric coaxial cable it is necessary that precautions be taken to insure that moisture cannot enter the line. If the better grade of connectors manufactured for the line are employed as terminations, this condition is automatically satisfied. If connectors are not used, it is necessary that some type of moisture-proof sealing compound be

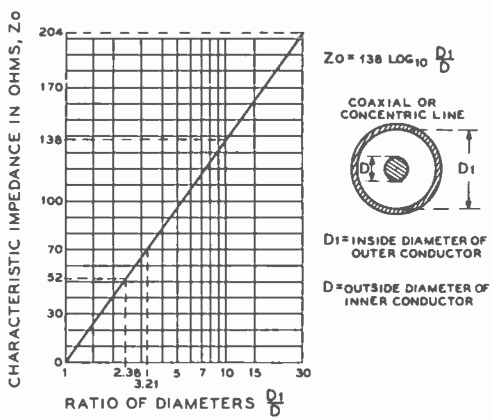


Figure 20.

CHARACTERISTIC IMPEDANCE OF AIR-FILLED COAXIAL LINES.

If the filling of the line is a dielectric material other than air, the characteristic impedance of the line will be reduced by a factor proportional to the square-root of the dielectric constant of the material used as a dielectric within the line.

applied to the end of the cable where it will be exposed to the weather.

Nearby metallic objects cause no loss, and coaxial cable may be run up air ducts or elevator shafts, inside walls, or through metal conduit. Insulation troubles can be forgotten. The coaxial cable may be buried in the ground or suspended above ground.

Standing Waves Standing waves on a transmission line *always* are the result of the reflection of energy. The only significant reflection which takes place in a normal installation is that at the load end of the line. But reflection can take place from discontinuities in the line, such as caused by insulators, bends, or metallic objects adjacent to an unshielded line.

When a uniform transmission line is terminated in an impedance equal to its surge impedance, reflection of energy does not occur, and no standing waves are present. When the load termination is exactly the same as the line impedance, it simply means that the load takes energy from the line just as fast as the line delivers it, no slower and no faster.

Thus, for proper operation of an untuned line (with standing waves eliminated), some form of impedance-matching arrangement must be used between the transmission line and the antenna, so that the radiation resistance of the antenna is reflected back into the line as a nonreactive impedance equal to the line impedance.

The termination at the antenna end is the only critical characteristic about the untuned line fed by a transmitter. It is the reflection from the antenna end which starts waves moving back toward the transmitter end. When waves moving in both directions along a conductor meet, standing waves are set up.

Semi-Resonant Parallel-Wire Lines A well-constructed open-wire line has acceptably low losses when its length is less than about two wavelengths even when the voltage standing-wave ratio is as high as 10 to 1. A transmission line constructed of ribbon or tubular line, however, should have the standing-wave ratio kept down to not more than about 3 to 1 both to reduce power loss and because the energy dissipation on the line will be localized, causing overheating of the line at the points of maximum current.

Because moderate standing waves can be tolerated on open-wire lines without much loss, a standing-wave ratio of 2/1 or 3/1 is considered acceptable with this type of line, *even when used in an "untuned" system.* Strictly speaking, a line is untuned, or nonresonant, only when it is perfectly "flat," with a standing-wave ratio of 1 (no standing waves). However, some mismatch can be tolerated with open-wire untuned lines, so long as the reactance is not objectionable, or is eliminated by cutting the line to approximately resonant length.

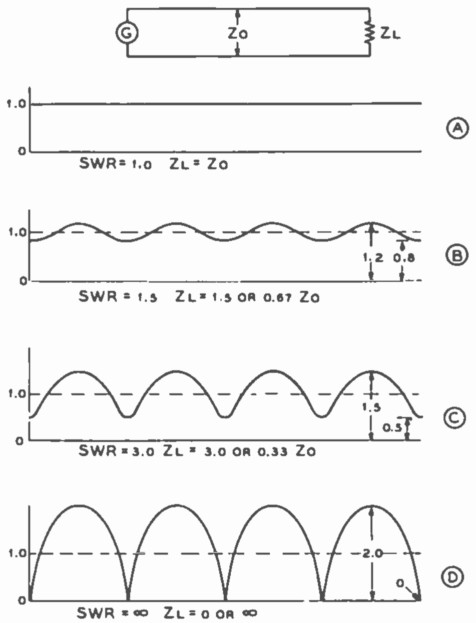


Figure 21.
STANDING WAVES ON A TRANSMISSION LINE.

As shown at (A), the voltage and current are constant on a transmission line which is terminated in its characteristic impedance, assuming that losses are small enough so that they may be neglected. (B) shows the variation in current or in voltage on a line terminated in a load with a reflection coefficient of 0.2 so that a standing-wave ratio of 1.5 to 1 is set up. At (C) the reflection coefficient has been increased to 0.5, with the formation of a 3 to 1 standing-wave ratio on the line. At (D) the line has been terminated in a load which has a reflection coefficient of 1.0 (short, open circuit, or a pure reactance) so that all the energy is reflected with the formation of an infinite standing-wave ratio.

12-11 Tuned or Resonant Lines

If a transmission line is terminated in its *characteristic surge impedance,* there will be no reflection at the end of the line, and the current and voltage distribution will be uniform along the line. If the end of the line is either open-circuited or short-circuited, the reflection at the end of the line will be 100 per cent, and *standing waves* of very great amplitude will appear on the line. There will still be practically no radiation from the line if it is closely spaced, but voltage nodes will be found along the line, spaced a half wavelength. Likewise, voltage loops will be found every half wavelength, the voltage loops corresponding to current nodes.

If the line is terminated in some value of resistance other than the characteristic surge impedance, there will be some reflection, the amount being determined by the amount of mismatch. With reflection, there will be standing waves (excursions of current and voltage) along the line, though not to the same extent as with an open-circuited or short-circuited line. The current and voltage loops will occur at the same *points* along the line as with the open- or short-circuited line, and as the terminating impedance is made to approach the characteristic impedance of the line, the current and voltage along the line will become more uniform. The foregoing assumes, of course, a purely resistive (non-reactive) load. If the load is reactive, standing waves also will be formed. But with a reactive load the nodes will occur at different locations from the node locations encountered with wrong-value resistive termination.

A well built 500- to 600-ohm transmission line may be used as a resonant feeder for lengths up to several hundred feet with very low loss, so long as the amplitude of the standing waves (ratio of maximum to minimum voltage along the line) is *not too great.* The amplitude, in turn, depends upon the mismatch at the line termination. A line of no. 12 wire, spaced 6 inches with good ceramic or plastic spreaders, has a surge impedance of approximately 600 ohms, and makes an excellent tuned feeder for feeding anything between 60 and 6000 ohms (at frequencies below 30 Mc.). If used to feed a load of higher or lower impedance than this, the standing waves become great enough in amplitude that some loss will occur unless the feeder is kept short. At frequencies above 30 Mc., the spacing becomes an appreciable fraction of a wavelength, and radiation from the line no longer is negligible. Hence, coaxial line or close-spaced parallel-wire line is recommended for v-h-f work.

If a transmission line is not perfectly matched, it should be made *resonant,* even though the amplitude of the standing waves (voltage variation) is not particularly great. This prevents reactance from being coupled into the final amplifier. A feed system having moderate standing waves may be made to present a non-reactive load to the amplifier either by tuning or by pruning the feeders to approximate resonance.

Usually it is preferable with tuned feeders to have a current loop (voltage minimum) at the transmitter end of the line. This means that when voltage-feeding an antenna, the tuned feeders should be made an odd number of quarter wavelengths long, and when current-feeding an antenna, the feeders should be made an even number of quarter wavelengths long. Actually, the feeders are made about 10 per cent of a quarter wave longer than the calculated value (the value given in the tables) when they are to be series tuned to resonance by means of a capacitor, instead of being trimmed and pruned to resonance.

When tuned feeders are used to feed an antenna on more than one band, it is necessary to compromise and make provision for both series and parallel tuning, inasmuch as it is impossible to cut a feeder to a length that will be optimum for several bands. If a voltage loop appears at the transmitter end of the line on certain bands, parallel tuning of the feeders will be required in order to get a transfer of energy. It is impossible to transfer energy by inductive coupling unless current is flowing. This is effected at a voltage loop by the presence of the resonant tank circuit formed by parallel tuning of the antenna coil.

CHAPTER THIRTEEN

Antennas and Antenna Matching

Antennas for the lower frequency portion of the h-f spectrum (perhaps from 1.8 to 7.0 Mc.), and temporary or limited use antennas for the upper portion of the h-f range, usually are of a relatively simple type in which directivity is not a prime consideration. Also, it often is desirable, in amateur work, that a single antenna system be capable of operation at least on the 3.5-Mc. and 7.0-Mc. range, and preferably on other frequency ranges. Consequently, the first portion of this chapter will be devoted to a discussion of such antenna systems. The latter portion of the chapter is devoted to the general problem of matching the antenna transmission line to antenna systems of the fixed type. Matching the antenna tranmission line to the rotatable directive array is discussed in Chapter Sixteen.

13-1 End-Fed Half-Wave Horizontal Antennas

The half-wave horizontal dipole is the most common and the most practical antenna for the 3.5-Mc. and 7-Mc. amateur bands. The form of the dipole and the manner in which it is fed are capable of a large number of variations. Figure 2 shows a number of practicable forms of the simple dipole antenna along with methods of feed.

Usually a high-frequency doublet is mounted as high and as much in the clear as possible, for obvious reasons. However, it is sometimes justifiable to bring part of the radiating system directly to the transmitter, feeding the antenna without benefit of a transmission line. This is permissible when (1) there is insufficient room to erect a 75- or 80-meter horizontal dipole and feed line, (2) when a long wire is also to be operated on one of the higher frequency bands on a harmonic. In either case, it is usually possible to get the main portion of the antenna in the clear because of its length. This means that the power lost by bringing the antenna directly to the transmitter is relatively small.

Even so, it is not best practice to bring the high-voltage end of an antenna into the operating room because of the increased difficulty in eliminating BCI and TVI. For this reason one should dispense with a feed line in conjunction with a Hertz antenna only as a last resort.

End-Fed Antennas The end-fed antenna has no form of transmission line to couple it to the transmitter, but brings the radiating portion of the antenna right down to the transmitter, where some form of coupling system is used to transfer energy to the antenna.

This antenna always is voltage-fed, and always consists of an *even* number of quarter-wavelengths. Figure 1 shows two common methods of feeding the Fuchs antenna or "end-fed Hertz." Some harmonic-attenuating provision (in addition to the usual low-pass TVI filter) must be included in the coupling system, as an end-fed antenna itself offers no discrimination against harmonics, either odd or even.

The end-fed Hertz antenna has rather high losses unless at least three-quarters of the radiator can be placed outside the operating room and in the clear. As there is high r-f voltage at the point where the antenna enters the operating room, the insulation at that point should be several times as effective as the insulation commonly used with low-voltage feeder systems. This antenna can be operated on all of its higher harmonics with good efficiency, and can be operated at half frequency against ground as a quarter-wave Marconi.

As the frequency of an antenna is raised slightly when it is bent anywhere except at a voltage or current loop, an end-fed Hertz antenna usually is a few per cent longer than a straight half-wave doublet for the same frequency, because, ordinarily, it is impracticable to bring a wire in to the transmitter without making several bends.

The Zepp Antenna System The zeppelin or "zepp" antenna system, illustrated in figure 2A is very convenient when it is desired to operate a single radiating wire on a number of harmonically related frequencies.

The zepp antenna system is easy to tune, and can be used on several bands by merely retuning the feeders. The overall efficiency of the zepp antenna system is not quite as high for long feeder lengths as for some of the antenna systems which employ non-resonant transmission lines, but where space is limited and where operation on more than one band is desired, the zepp has some decided advantages.

As the radiating portion of the zepp antenna system must always be some multiple of a half wave long, there is always high voltage present at the point where the live zepp feeder attaches to the end of the radiating portion of the antenna. Thus, this type of zepp antenna system is *voltage fed.*

Stub-Fed Zepp-Type Radiator Figure 2C shows a modification of the zepp-type antenna system to allow the use of a non-resonant transmission line between the radiating portion of the antenna and the transmitter. The "zepp" portion of the antenna is resonated as a quarter-wave stub and the non-resonant feeders are connected to the stub at a point where standing waves on the feeder are minimized. The procedure for making these adjustments is described in detail in Section 13-9. This type of antenna system is quite satisfactory when it is necessary physically to end feed the antenna, but where it is necessary also to use non-resonant feeder between the transmitter and the radiating system.

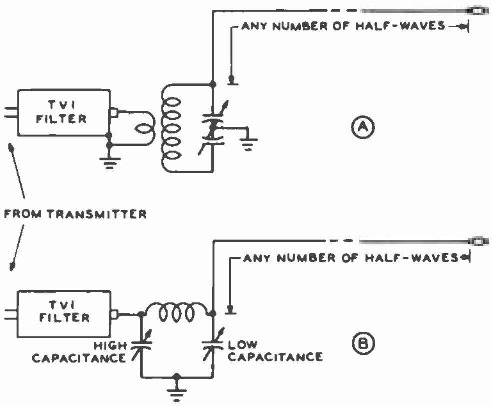


Figure 1.
THE END-FED HERTZ ANTENNA.
Showing the manner in which an end-fed Hertz antenna may be fed through a low-impedance line and low-pass filter by using a resonant tank circuit as at (A), or through the use of a reverse-connected pi network as at (B).

13-2 Center-Fed Half-Wave Horizontal Antennas

The center feeding of a half-wave antenna system is usually to be desired over an end-fed system since the center-fed system is inherently balanced to ground and is therefore less likely to be troubled by feeder radiation. A number of center-fed systems are illustrated in figure 2.

The Tuned Doublet The current-fed doublet with spaced feeders, sometimes called a center-fed zepp, is an inherently balanced system if the two legs of the radiator are electrically equal. This fact holds true re-

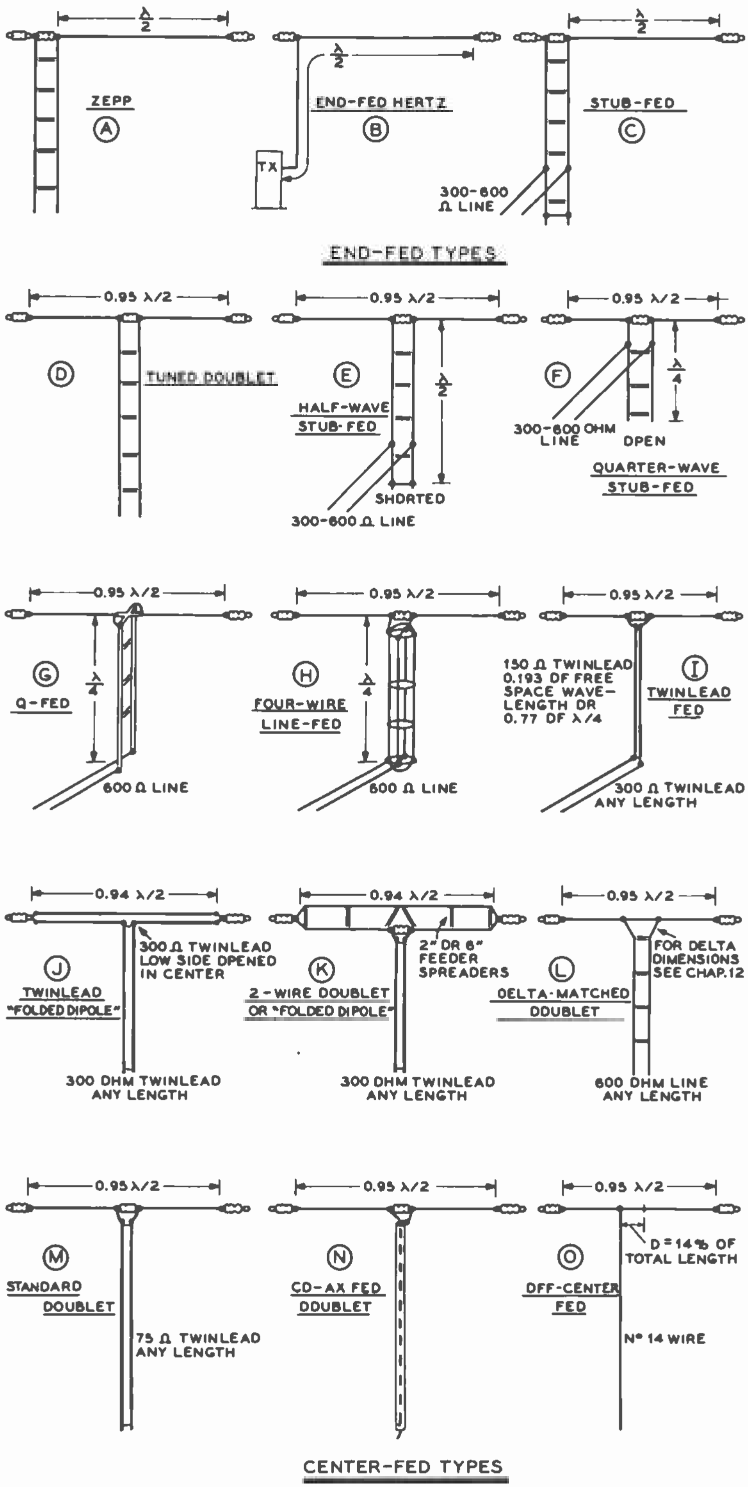


Figure 2.
ALTERNATIVE METHODS OF FEEDING A HALF-WAVE DIPOLE.

gardless of the frequency, or of the harmonic, on which the system is operated. The system can successfully be operated over a wide range of frequencies if the system as a whole (both tuned feeders and the center-fed flat top) can be resonated to the operating frequency. It is usually possible to tune such an antenna system to resonance with the aid of a tapped coil and a tuning capacitor that can optionally be placed either in series with the antenna coil or in parallel with it. A series tuning capacitor can be placed in series with one feeder leg without unbalancing the system.

The tuned-doublet antenna is shown in figure 2D. The antenna is a current-fed system when the radiating wire is a half wave long electrically, or when the system is operated on its odd harmonics, but becomes a voltage-fed radiator when operated on its even harmonics.

The antenna has a different radiation pattern when operated on its harmonics, as would be expected. The arrangement used on the second harmonic is better known as the Franklin colinear array and is described in Chapter Fourteen. The pattern is similar to a half-wave dipole except that it is sharper in the broadside direction. On higher harmonics of operation there will be multiple lobes of radiation from the system.

Figures 2E and 2F show alternative arrangements for using an untuned transmission line between the transmitter and the tuned-doublet radiator. In figure 2E a half-wave shorted line is used to resonate the radiating system, while in figure 2F a quarter-wave open line is utilized. The adjustment of quarter-wave and half-wave stubs is discussed in Section 13-9.

Doublets with Quarter-Wave Transformers The average value of feed impedance for a center-fed half-wave doublet is 75 ohms. The actual value varies with height and is shown in figure 5 of Chapter Twelve. Alternative methods of matching this rather low value of impedance to a medium-impedance transmission line are shown in (G), (H), and (I) of figure 2. Each of these three systems uses a quarter-wave transformer to accomplish the impedance transformation. The only difference between the three systems lies in the type of transmission line used in the quarter-wave transformer. (G) shows the "Johnson Q" system whereby a line made up of ½-inch dural tubing is used for the low-impedance linear transformer. A line made up in this manner is frequently called a set of "Q bars." Illustration (H) shows the use of a four-wire line as the linear transformer, and (I) shows the use of a piece of 150-ohm Twin-Lead electrically ¼-wave in length as the transformer between the center of the dipole and a piece of 300-ohm Twin-Lead. In any case the impedance of the quarter-wave transformer will be of the order of 150 to 200 ohms. The use of sections of transmission line as linear transformers is discussed in detail in Section 13-10.

Multi-Wire Doublets An alternative method for increasing the feed-point impedance of a dipole so that a medium-impedance transmission line may be used is shown in figures 2J and 2K. This system utilizes more than one wire in parallel for the radiating element, but only one of the wires is broken for attachment of the feeder. The theory of this type of antenna has been discussed in Chapter Twelve, but the most common arrangement uses two wires in the flat top of the antenna so that an impedance multiplication of four is obtained.

The antenna shown in figure 2J is the so-called Twin-Lead "folded dipole" which is a commonly used antenna system on the medium-frequency amateur bands. In this arrangement both the antenna and the transmission line to the transmitter are constructed of 300-ohm Twin-Lead. The flat top of the antenna is made slightly less than the conventional length ($462/F_{Mc}$ instead of $468/F_{Mc}$ for a single-wire flat top) and the two ends of the Twin-Lead are joined together at each end. The center of one of the conductors of the Twin-Lead flat top is broken and the two ends of the Twin-Lead feeder are spliced into the flat top leads. As a protection against moisture pieces of flat polyethylene taken from another piece of 300-ohm Twin-Lead may be molded over the joint between conductors with the aid of an electric iron or soldering iron.

Better bandwidth characteristics can be obtained with a folded dipole made of ribbon line if the two conductors of the ribbon line are shorted a distance of 0.82 (the velocity factor of ribbon line) of a free-space quarter wavelength from the center or feed point. This procedure is illustrated in figure 3A. An alternative arrangement for a Twin-Lead folded dipole (which, incidentally, requires less Twin-Lead) is illustrated in figure 3B. This type of

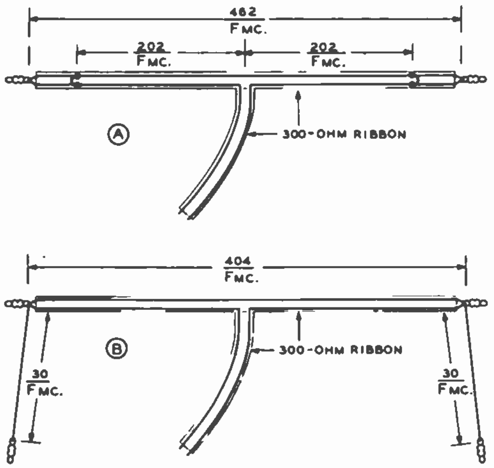


Figure 3.
FOLDED DIPOLE WITH SHORTING STRAPS.

The impedance match and bandwidth characteristics of a folded dipole may be improved by shorting the two wires of the ribbon a distance out from the center equal to the velocity factor of the ribbon times the half-length of the dipole as shown at (A). An alternative arrangement with bent down ends for space conservation is illustrated at (B).

half-wave antenna system is convenient for use on the 3.5-Mc. band when the 116 to 132 foot distance required for a full half-wave is not quite available in a straight line, since the single-wire end pieces may be bent away or downward from the direction of the main section of the antenna.

Figure 2K shows the basic type of 2-wire doublet or "folded dipole" wherein the radiating section of the system is made up of standard antenna wire spaced by means of feeder spreaders. The feeder again is made of 300-ohm Twin-Lead since the feed-point impedance is approximately 300 ohms, the same as that of the Twin-Lead folded dipole.

The folded-dipole type of antenna has the broadest response characteristic (greatest bandwidth) of any of the conventional half-wave antenna systems constructed of small wires or conductors. Hence such an antenna may be operated over the greatest frequency range without serious standing waves of any common half-wave antenna type.

The increased bandwidth of the multi-wire doublet type of radiator, and the fact that the feed-point resistance is increased several times over the radiation resistance of the element, have both contributed to the frequent use of the multi-wire radiator as the driven element in a parasitic antenna array.

Delta-Matched Doublet and Standard Doublet These two types of radiating elements are shown in figure 2L and figure 2M. The delta-matched doublet is described in detail in Section 13-8 and is illustrated in figure 15 of this chapter. The standard doublet, shown in figure 2M, is fed in the center by means of 75-ohm Twin-Lead, either the transmitting or the receiving type, or it may be fed by means of twisted-pair feeder or by means of parallel-wire lamp-cord. Any of these types of feed line will give an approximate match to the center impedance of the dipole, but the 75-ohm Twin-Lead is far to be preferred over the other types of low-impedance feeder due to the much lower losses of the polyethylene-dielectric transmission line.

The coaxial-cable-fed doublet shown in figure 2N is a variation on the system shown in figure 2M. Either 52-ohm coaxial cable or 75-ohm coaxial cable may be used to feed the center of the dipole, although the 75-ohm type will give a somewhat better impedance match at normal antenna heights. Due to the asymmetry of the coaxial feed system difficulty may be encountered with waves traveling on the outside of the coaxial cable. For this reason the use of Twin-Lead is normally to be preferred over the use of coaxial cable for feeding the center of a half-wave dipole.

Off-Center Fed Doublet The system shown in figure 2(O) is sometimes used to feed a half-wave dipole, especially when it is desired to use the same antenna on a number of harmonically-related frequencies. The feeder wire (no. 14 enamelled wire should be used) is tapped a distance of 14 per cent of the total length of the antenna either side of center. The feeder wire, operating against ground for the return current, has an impedance of approximately 600 ohms. The system works well over highly conducting ground, but will introduce rather high losses when the antenna is located above rocky or poorly conducting soil. The off-center fed antenna has a further disadvantage that it is highly responsive to harmonics fed to it from the transmitter.

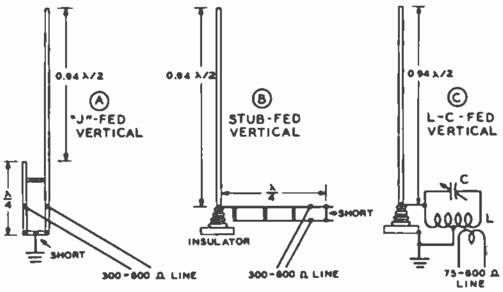


Figure 4.

HALF-WAVE VERTICAL ANTENNA SHOWING ALTERNATIVE METHODS OF FEED.

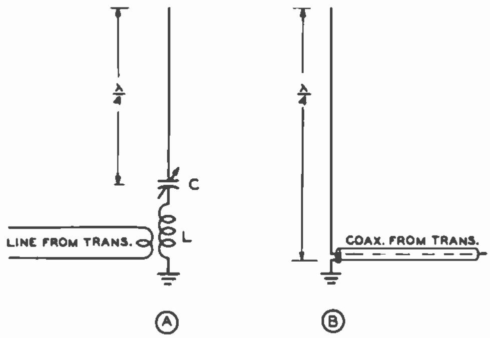


Figure 5.

FEEDING A QUARTER-WAVE MARCONI ANTENNA.

When an open-wire line is to be used, it may be link coupled to a series-resonant circuit between the bottom end of the Marconi and ground, as at (A). Alternatively, a reasonably good impedance match may be obtained between 52-ohm coaxial line and the bottom of a resonant quarter-wave antenna, as illustrated at (B) above.

The effectiveness of the antenna system in radiating harmonics is of course an advantage when operation of the antenna on a number of frequency bands is desired. But again it is quite necessary to use a harmonic filter to insure that only the desired frequency is fed from the transmitter to the antenna.

13-3 The Half-Wave Vertical Antenna

The half-wave vertical antenna with its bottom end from 0.1 to 0.2 wavelength above ground is an effective transmitting antenna for low-angle radiation, where ground conditions in the vicinity of the antenna are good. Such an antenna is not good for short-range *sky*-wave communication, such as is the normal usage of the 3.5-Mc. amateur band, but is excellent for short-range *ground*-wave communication such as on the standard broadcast band and on the amateur 1.8-Mc. band. The vertical antenna normally will cause greater BCI than an equivalent horizontal antenna, due to the much greater ground-wave field intensity. Also, the vertical antenna is poor for receiving under conditions where man-made interference is severe, since such interference is predominantly vertically polarized.

Three ways of feeding a half-wave vertical antenna from an untuned transmisison line are illustrated in figure 4. The J-fed system shown in figure 4A is obviously not practicable except on the higher frequencies where the extra length for the stub may easily be obtained. However, in the normal case the ground-plane vertical antenna is to be recommended over the J-fed system for v-h-f work.

13-4 The Marconi Antenna

A grounded quarter-wave Marconi antenna, widely used on frequencies below 3 Mc., is sometimes used on the 3.5-Mc. band, and is also used in v-h-f mobile services where a compact antenna is required. The Marconi type antenna allows the use of half the length of wire that would be required for a half-wave Hertz radiator. The ground acts as a mirror, in effect, and takes the place of the additional quarter-wave of wire that would be required to reach resonance if the end of the wire were not returned to ground.

The fundamental practical form of the Marconi antenna system is shown in figure 5A. Other Marconi antennas differ from this type primarily in regard to the method of feeding the energy to the radiator. The feed method shown in figure 5B can often be used to advantage, particularly in mobile work.

Variations on the basic Marconi antenna are shown in the illustrations of figure 6. Figures 6B and 6C show the "L"-type and "T"-type Marconi antennas. These arrangements have been more or less superseded by the top-loaded forms of the Marconi antenna shown in figures

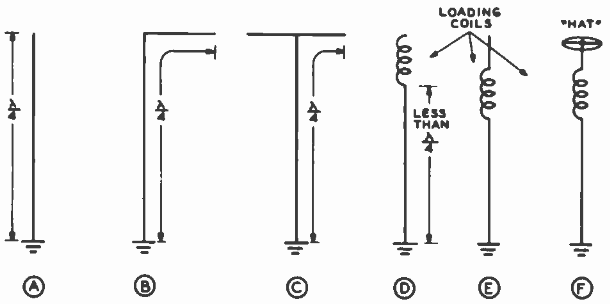


Figure 6.
LOADING THE MARCONI ANTENNA.
The various loading systems are discussed in the accompanying text.

6D, 6E, and 6F. In each of these latter three figures an antenna somewhat less than one quarter wave in length has been *loaded* to increase its effective length by the insertion of a *loading coil* at or near the top of the radiator. The arrangement shown at figure 6D gives the least loading but is the most practical mechanically. The system shown at figure 6E gives an intermediate amount of loading, while that shown at figure 6F, utilizing a "hat" just above the loading coil, gives the greatest amount of loading. The object of all the top-loading methods shown is to produce an increase in the effective length of the radiator, and thus to raise the point of maximum current in the radiator as far as possible above ground. Raising the maximum-current point in the radiator above ground has two desirable results: The percentage of low-angle radiation is increased and the amount of ground current at the base of the radiator is reduced, thus reducing the ground losses.

To estimate whether a loading coil will probably be required, it is necessary only to note if the length of the antenna wire and ground lead is over a quarter wavelength; if so, no loading coil is needed, provided the series tuning capacitor has a high maximum capacitance.

Amateurs primarily interested in the higher frequency bands, but who like to work 80 meters occasionally, can usually manage to resonate one of their antennas as a Marconi by working the whole system, feeders and all, against a water pipe ground, and resorting to a loading coil if necessary. A high-frequency-rotary, zepp, doublet, or single-wire-fed antenna will make quite a good 80-meter Marconi if high and in the clear, with a rather long feed line to act as a radiator on 80 meters. Where two-wire feeders are used, the feeders should be tied together for Marconi operation.

Importance of Ground Connection

With a quarter-wave antenna and a ground, the antenna current generally is measured with a meter placed in the antenna circuit close to the ground connection. Now, if this current flows through a resistor, or if the ground itself presents some resistance, there will be a power loss in the form of heat. Improving the ground connection, therefore, provides a definite means of reducing this power loss, and thus increasing the radiated power.

The best possible ground consists of as many wires as possible, each at least a quarter wave long, buried just below the surface of the earth, and extending out from a common point in the form of radials. Copper wire of any size larger than no. 16 is satisfactory, though the larger sizes will take longer to disintegrate. In fact, the radials need not even be buried; they may be supported just above the earth, and insulated from it. This arrangement is called a *counterpoise,* and operates by virtue of its high capacitance to ground.

If the antenna is physically shorter than a quarter wavelength, the antenna current is higher, due to lower radiation resistance. Consequently, the power lost in resistive soil is greater. The importance of a good ground with short, inductive-loaded Marconi radiators is, therefore, quite obvious. With a good ground system, even very short (one-eighth wavelength) antennas can be expected to give a high percentage of the efficiency of a quarter-wave antenna used with the same ground system. This is especially true when the short radiator is *top loaded* with a high Q (low loss) coil.

Water-Pipe Grounds

Water pipe, because of its comparatively large surface and cross section, has a relatively low r-f resistance. If it is possible to attach to a junction of several water pipes (where they branch in several directions and run for some distance under ground), a satisfactory ground connection will be obtained. If one of the pipes attaches to a lawn or garden sprinkler system in the immediate vicinity of the antenna, the effectiveness of the system will approach that of buried copper radials.

The main objection to water-pipe grounds is the possibility of high resistance joints in the pipe, due to the "dope" put on the coupling threads. By attaching the ground wire to a junction with three or more legs, the possibility of requiring the main portion of the r-f current to flow through a high resistance connection is greatly reduced.

The presence of water in the pipe adds nothing to the conductivity; therefore it does not relieve the problem of high resistance joints. Bonding the joints is the best insurance, but this is, of course, impracticable where the pipe is buried. Bonding together with copper wire the various water faucets above the surface of the ground will improve the effectiveness of a water-pipe ground system hampered by high-resistance pipe couplings.

Marconi Dimensions A Marconi antenna is an odd number of electrical quarter waves long (usually only one quarter wave in length), and is always resonated to the operating frequency. The correct loading of the final amplifier is accomplished by varying the coupling, *rather than by detuning the antenna from resonance.*

Physically, a quarter-wave Marconi may be made anything from one-eighth to three-eighths wavelength overall, meaning the total length of the antenna wire and ground lead from the end of the antenna to the point where the ground lead attaches to the junction of the radials or counterpoise wires, or the water pipe enters the ground. The longer the antenna is made physically, the lower will be the current flowing in the ground connection, and the greater will be the overall radiation efficiency. However, when the antenna length exceeds three-eighths wavelength, the antenna becomes difficult to resonate by means of a series capacitor, and it begins to take shape as an end-fed Hertz, requiring a method of feed such as a pi network.

A radiator physically much shorter than a quarter wavelength can be lengthened electrically by means of a series loading coil, and used as a quarter-wave Marconi. However, if the wire is made shorter than approximately one-eighth wavelength, the radiation resistance will be quite low. This is a special problem in mobile work below about 20-Mc., and is discussed in detail in Chapter Nineteen.

13-5 Space-Conserving Antennas

In many cases it is desired to undertake a considerable amount of operation on the 80-meter or 40-meter band, but sufficient space is simply not available for the installation of a half-wave radiator for the desired frequency of operation. This is a common experience of apartment dwellers. The shortened Marconi antenna operated against a good ground *can* be used under certain conditions, but the shortened Marconi is notorious for the production of broadcast interference, and a good ground connection is usually completely unobtainable in an apartment house.

Essentially, the problem in producing an antenna for lower frequency operation in restricted space is to erect a short radiator which is balanced with respect to ground and which is therefore independent of ground for its operation. Several antenna types meeting this set of conditions are shown in figure 7. Figure 7A shows a conventional center-fed doublet with bent-down ends. This type of antenna can be fed with 75-ohm Twin-Lead in the center, or it may be fed with a resonant line for operation on several bands. The overall length of the radiating wire will be a few per cent greater than the normal length for such an antenna since the wire is bent at a position intermediate between a current loop and a voltage loop. The actual length will have to be determined by the cut-and-try process due to the increased effect of interfering objects on the effective electrical length of an antenna of this type.

Figure 7B shows a method for using a two-wire doublet on one half of its normal operating frequency. It is recommended that spaced open conductor be used both for the radiating portion of the "folded dipole" and for the feed line. The reason for this recommendation lies in the fact that the two wires of the flat

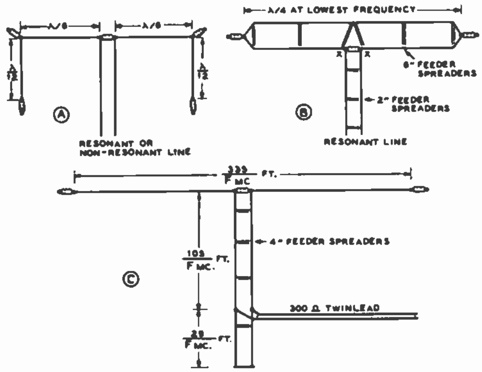


Figure 7.

THREE EFFECTIVE SPACE CONSERVING ANTENNAS.

The arrangements shown at (A) and (B) are satisfactory where resonant feed line can be used. However, non-resonant 75-ohm feed line may be used in the arrangement at (A) when the dimensions in wavelengths are as shown. In the arrangement shown at (B) low standing waves will be obtained on the feed line when the overall length of the antenna is a half wave. The arrangement shown at (C) may be tuned for any reasonable length of flat top to give a minimum of standing waves on the transmission line.

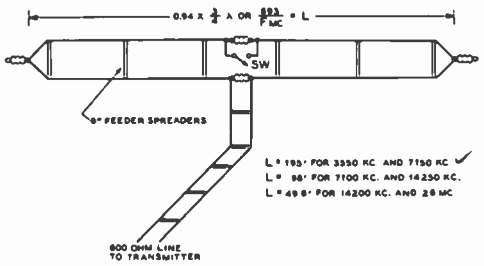


Figure 8.

THE THREE-QUARTER WAVE FOLDED DOUBLET.

This antenna arrangement will give very satisfactory operation with a 600-ohm feed line for operation with the switch open on the fundamental frequency and with the switch closed on twice frequency.

top are *not* at the same potential throughout their length when the antenna is operated on one-half frequency. Twin-Lead may be used for the feed line if operation on the frequency where the flat top is one-half wave in length is most common, and operation on one-half frequency is infrequent. However, if the antenna is to be used primarily on one-half frequency as shown it should be fed by means of an open-wire line. If it is desired to feed the antenna with a non-resonant line, a quarter-wave stub may be connected to the antenna at the points X, X in figure 7B. The stub should be tuned and the transmission line connected to it in the normal manner.

The antenna system shown in figure 7C may be used when not quite enough length is available for a full half-wave radiator. The dimensions in terms of frequency are given on the drawing. An antenna of this type is 93 feet long for operation on 3600 kc. and 86 feet long for operation on 3900 kc. This type of antenna has the additional advantage that it may be operated on the 7-Mc. and 14-Mc. bands, when the flat top has been cut for the 3.5-Mc. band, simply by changing the position of the shorting bar and the feeder line on the stub.

A sacrifice which must be made when using a shortened radiating system, as for example the types shown in figure 7, is in the bandwidth of the radiating system. The frequency range which may be covered by a shortened antenna system is approximately in proportion to the amount of shortening which has been employed. For example, the antenna system shown in figure 7C may be operated over the range from 3800 kc. to 4000 ks. without serious standing waves on the feed line. If the antenna had been made full length it would be possible to cover about half again as much frequency range for the same amount of mismatch on the extremes of the frequency range.

13-6 Multi-Band Antennas

The availability of a multi-band antenna is a great operating convenience to an amateur station. In most cases it will be found best to install an antenna which is optimum for the band which is used for the majority of the available operating time, and then to have an additional multi-band antenna which may be pressed into service for operation on another band when propagation conditions on the most frequently used band are not suitable. Most amateurs use, or plan to install, at least one directive array for one of the higher-frequency bands, but find that an additional antenna which may be used on the 3.5-Mc. and 7.0-

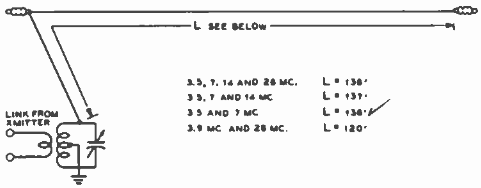


Figure 9.
RECOMMENDED LENGTHS FOR THE END-FED HERTZ.

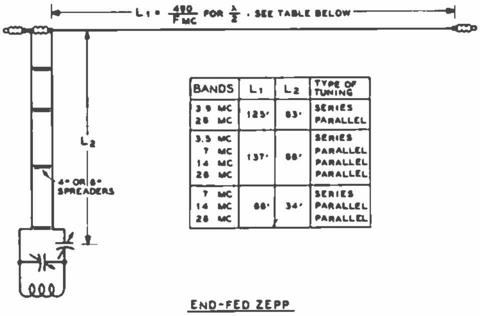


FIGURE 10.

Mc. band or even up through the 28-Mc. band is almost indispensable.

The choice of a multi-band antenna depends upon a number of factors such as the amount of space available, the band which is to be used for the majority of operating with the antenna, the radiation efficiency which is desired, and the type of antenna tuning network to be used at the transmitter. A number of recommended types for use under differing conditions are illustrated in figures 8 through 11.

The ¾-Wave Folded Doublet Figure 8 shows an antenna type which will be found to be very effective when a moderate amount of space is available, when most of the operating will be done on one band with occasional operation on the second harmonic. The system is quite satisfactory for use with high-power transmitters since a 600-ohm non-resonant line is used from the antenna to the transmitter and since the antenna system is balanced with respect to ground. With operation on the fundamental frequency of the antenna where the flat top is ¾ wave long the switch SW is left open. The system affords a very close match between the 600-ohm line and the feed point of the antenna. Kraus has reported a standing-wave ratio of approximately 1.2 to 1 over the 14-Mc. band when the antenna was located approximately one-half wave above ground (*Radio*, June 1939).

For operation on the second harmonic the switch SW is *closed*. The antenna is still an effective radiator on the second harmonic but the pattern of radiation will be different from that on the fundamental and the standing-wave ratio on the feed line will be greater. The flat top of the antenna must be made of open wire rather than ribbon or tubular line.

The End-Fed Hertz The end-fed Hertz antenna shown in figure 9 is not as effective a radiating system as many other antenna types, but it is particularly convenient when it is desired to install an antenna in a hurry for a test, or for field-day work. The flat top of the radiator should be as high and in the clear as possible. In any event at least three quarters of the total wire length should be in the clear. Dimensions for optimum operation on various amateur bands are given in addition in figure 9. Additional information on the end-fed Hertz is given in Section 12-1 and figure 1.

The End-Fed Zepp The end-fed Zepp has long been a favorite for multi-band operation. It is shown in figure 10 along with recommended dimensions for operation on various amateur band groups. Since this antenna type is an unbalanced radiating system, its use is not recommended with high-power transmitters where interference to broadcast listeners is likely to be encountered. The r-f voltages encountered at the end of zepp feeders and at points an electrical half wave from the end are likely to be quite high. Hence the feeders should be supported an adequate distance from surrounding objects and sufficiently in the clear so that a chance encounter between a passerby and the feeder is unlikely.

The coupling coil at the transmitter end of the feeder system should be link coupled to the output of the low-pass TVI filter in order to reduce harmonic radiation. Transmitter-to-feed line coupling methods have been discussed in detail in Chapter Eleven.

The Center-Fed Multi-Band Antenna Several types of center-fed antenna systems are shown in figure 11. If the feed line is made up in the conventional manner of no. 12 or no. 14 wire spaced 4 to 6 inches the antenna system is sometimes called a center-fed zepp. With this type of feeder the impedance at the transmitter end of the feeder varies from about 70 ohms to approximately 5000 ohms, the same as is encountered in an end-fed zepp antenna. This great impedance ratio requires provision for either series or parallel tuning of the feeders at the transmitter, and involves quite high r-f voltages at various points along the feed line.

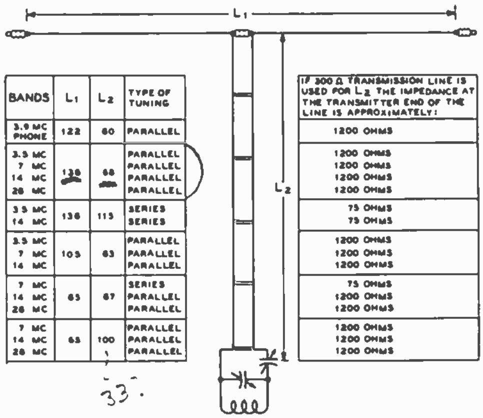
If the feed line between the transmitter and the antenna is made to have a characteristic impedance of approximately 300 ohms the excursions in end-of-feeder impedance are greatly reduced. In fact the impedance then varies from approximately 75 ohms to 1200 ohms. With this much lowered impedance variation it is usually possible to use series tuning on all bands, or merely to couple the antenna directly to the output tank circuit or the harmonic reduction circuit without any separate feeder tuning provision.

There are four practicable types of transmission line which can give an impedance of approximately 300 ohms. The first is, obviously, 300-ohm Twin-Lead. Twin-Lead of the receiving type *may* be used as a resonant feed line in this case, but its use is not recommended with power levels greater than perhaps 100 watts, and it should not be used when lowest loss in the transmission line is desired. Although Twin-Lead is quite satisfactory for a non-resonant line, it has not been designed for application as a resonant line. The second type is a two-wire air-spaced line with large conductors and close spacing. The ratio of conductor spacing to conductor radius for a characteristic impedance of 300 ohms is 12.22. In other words, no. 10 wire should be spaced 0.6", no. 8 wire should be spaced 0.79", 3/16" copper tubing should be spaced about 1.14", and ¼-inch copper tubing should be spaced 1.53/". Johnson no. 132 2" feeder spreaders have a notch cut in the side for use with ¼-inch copper tubing at 1½" spacing for making up a 300-ohm line.

The third type of transmission line which may be used to obtain an impedance of 300 ohms is the standard cross-connected four-wire air-spaced line. In this case to obtain, from the equation:

$$Z_o = 138 \log_{10} \frac{S}{r} - 21$$

an impedance of 300 ohms requires a ratio of conductor spacing to conductor radius of 211.2. This high ratio requires the use of rather small wires or rather large spacings. Number 18 wires would be spaced 4" on the sides of the square, or equally about a 5.65" diameter circle.



BANDS	L1	L2	TYPE OF TUNING	IF 300 Ω TRANSMISSION LINE IS USED FOR L2 THE IMPEDANCE AT THE TRANSMITTER END OF THE LINE IS APPROXIMATELY:
3.9 MC PHONE	122	60	PARALLEL	1200 OHMS
3.5 MC 7 MC 14 MC 28 MC	136	68	PARALLEL PARALLEL PARALLEL PARALLEL	1200 OHMS 1200 OHMS 1200 OHMS 1200 OHMS
3.5 MC 14 MC	136	115	SERIES SERIES	75 OHMS 75 OHMS
3.5 MC 7 MC 14 MC	105	63	PARALLEL PARALLEL PARALLEL	1200 OHMS 1200 OHMS 1200 OHMS
7 MC 14 MC 28 MC	65	67	SERIES PARALLEL PARALLEL	75 OHMS 1200 OHMS 1200 OHMS
7 MC 14 MC 28 MC	65	100	PARALLEL PARALLEL PARALLEL	1200 OHMS 1200 OHMS 1200 OHMS

CENTER-FED ANTENNA

FIGURE 11.

Probably the most satisfactory method for obtaining a low-loss 300-ohm line of simple construction is to use a four-wire *side-connected* transmission line. This type of transmission line is constructed the same as a conventional cross-connected four-wire line but adjacent pairs of wire are strapped together at the ends (make sure that the same pair is strapped at each end). This type of transmission line has not been very commonly used in the past but it lends itself particularly to the construction of transmission lines with impedances in the range between 260 ohms (above which the four-wire cross-connected line becomes difficult mechanically) and 400 ohms (below which the two-wire line becomes mechanically difficult to construct).

The equation for the characteristic impedance of such a line where the four wires are placed on the corners of a square is:

$$Z_o = 138 \log_{10} \frac{S}{r} + 21$$

where S is the spacing between adjacent wires and r is the radius of the wires. Inspection of the equation shows that the form of the equation is the same as that for the cross-connected four-wire line except that a plus sign has been substituted for the minus sign between the last two terms. This means that if we change the end-strap connections on a four-wire cross-connected line to make it an adjacent-connected line, the impedance of the line will be raised by 42 ohms. The actual ratio of spacing to radius for 300 ohms impedance using this type of line is 105.5 It will be noticed that this ratio is exactly half that required for the same value of impedance using a cross-connected line; this fact will be found true in all cases when comparing the two methods of connection for a four-wire line—the ratio of spacings for a cross-connected line is exactly twice that for an adjacent-connected line having the same impedance.

Folded Flat-Top Dual-Band Antenna As has been mentioned earlier, there is an increasing tendency among amateur operators to utilize rotary or fixed arrays for the 14-Mc. band and those higher in frequency. In order to afford complete coverage of the amateur bands it is then desirable to have an additional system which will operate with equal effectiveness on the 3.5-Mc. and 7-Mc. bands, but this low-frequency antenna system will not be required to operate on any bands higher in frequency than the 7-Mc. band. The antenna system shown in figure 12 has been developed to fill this need.

This system consists essentially of an open-line folded dipole for the 7-Mc. band with a special feed system which allows the antenna to be fed with minimum standing waves on the feed line on both the 7-Mc. and 3.5-Mc. bands. The feed-point impedance of a folded dipole on its fundamental frequency is approximately 300 ohms. Hence the 300-ohm Twin-Lead shown in figure 12 can be connected directly into the center of the system for operation only on the 7-Mc. band and standing waves on the feeder will be very small. However, it is possible to insert an electrical half-wave of transmission line of any

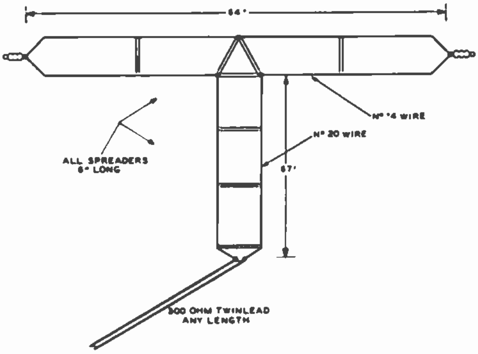


Figure 12.
FOLDED-TOP DUAL-BAND ANTENNA.

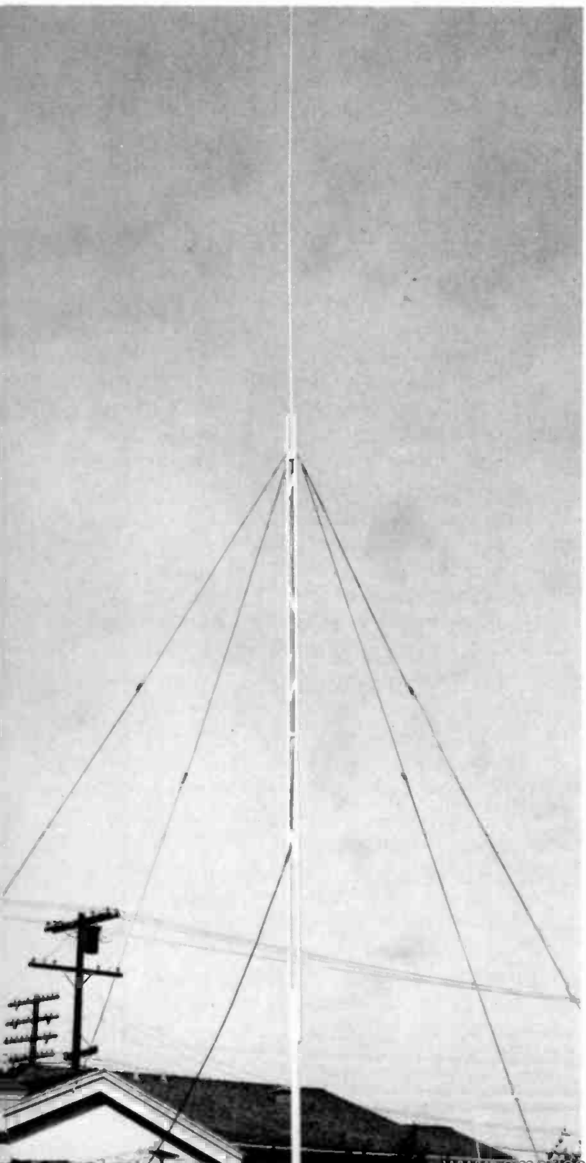
characteristic impedance into a feeder system such as this and the impedance at the far end of the line will be exactly the same value of impedance which the half-wave line sees at its termination. Hence this has been done in the antenna system shown in figure 12; an electrical half wave of line has been inserted between the feed point of the antenna and the 300-ohm transmission line to the transmitter.

The characteristic impedance of this additional half-wave section of transmission line has been made about 715 ohms (no. 20 wire spaced 6 inches), but since it is an electrical half wave long at 7 Mc. and operates into a load of 300 ohms at the antenna the 300-ohm Twin-Lead at the bottom of the half-wave section still sees an impedance of 300 ohms. The additional half-wave section of transmission line introduces a negligible amount of loss since the current flowing in the section of line is the same which would flow in a 300-ohm line at each end of the half-wave section, and at all other points it is *less* than the current which would flow in a 300-ohm line since the effective impedance is *greater* than 300 ohms in the center of the half-wave section. This means that the loss is less than it would be in an equivalent length of 300-ohm Twin-Lead since this type of manufactured transmission line is made up of conductors which are equivalent to no. 20 wire.

So we see that the added section of 715-ohm line has substantially no effect on the operation of the antenna system on the 7-Mc. band. However, when the flat top of the antenna is operated on the 3.5-Mc. band the

Figure 13.
THE "DROOPING GROUND PLANE" VERTICAL ANTENNA.

This vertically polarized antenna uses the four guy wires as a "drooping ground plane" for the vertical radiator. Note the use of two sets of insulators in the guy wires. One set is spaced 8½ feet from the metal collar about the mast; the guys between these insulators and the mast are used on the 28-Mc. band. The second set is spaced 8½ feet from the first set; when jumpers are placed across the first set of insulators the total length of the ground plane then becomes about 17 feet for operation on the 14-Mc. band. Additional sets of insulators may be placed at appropriate points in the guy wires for operation in conjunction with various vertical radiator rods on other amateur bands.



feed-point impedance of the flat top is approximately 3500 ohms. Since the section of 715-ohm transmission line is an electrical *quarter-wave* in length on the 3.5-Mc. band, this section of line will have the effect of transforming the approximately 3500 ohms feed-point impedance of the antenna down to an impedance of about 150 ohms which will result in a 2:1 standing-wave ratio on the 300-ohm Twin-Lead transmission line from the transmitter to the antenna system.

The antenna system of figure 12 operates with very low standing waves over the entire 7-Mc. band, and it will operate with moderate standing waves from 3500 to 3800 kc. in the 3.5-Mc. band and with sufficiently low standing-wave ratio so that it is quite usable over the entire 3.5-Mc. band.

This antenna system, as well as all other types of multi-band antenna systems, must be used in conjunction with some type of harmonic-reducing antenna tuning network even though the system does present a convenient impedance value on both bands.

"Drooping Ground Plane" Antenna.

Although the ground-plane vertical is not the ideal antenna by any standards, it still is a type of antenna system which can give excellent performance in terms of the amount of space required. It radiates a moderate amount of power at the extremely low angles of radiation which characterize dx propagation to certain portions of the world. Further, it is easy to install, particularly so when constructed as shown in figures 13 and 14, and is easily altered for operation on another frequency band.

The capability of easy alteration may be of advantage in the event that the 21-Mc. band is opened with little prior notice. The array shown in figure 13 may be lowered in just a few moments by loosening one of the guy wires, the vertical rod may be lengthened or shortened by adding or removing a section, and the placement of the insulators in the guy wires may be altered. However, if it is desired to make a quick change from a lower frequency to a higher frequency band, experience has shown that the guy-wire "ground plane" may be left at the full length for the lower frequency band. Thus when making a quick change from 20 to 10 the guy-wire ground plane may be left at its full 17-foot length. The standing-wave ratio on the coaxial feed line is increased somewhat, but the antenna system still operates satisfactorily.

Several other advantages of the ground-plane vertical as a standby antenna may be cited: Such an antenna will give a much improved range of reliable ground-wave communication with 10 and 11 meter mobile stations, as compared with the usual horizontally polarized rotary array. The vertical is a useful accessory to the rotary beam on the same band for obtaining a quick check on the direction from which the most satisfactory signals are arriving. A further application of the ground plane is as an additional antenna system for operation in diversity with the existing horizontally polarized antenna system; the fading characteristics of a signal received on a vertical antenna almost invariably will be sufficiently different from the fading of the same signal on a horizontal antenna so that diversity reception will allow the signal to be read solid a greater percentage of the time than on either of the antennas alone.

Construction of the Antenna System Select a straight 16-foot length of 1" by 4" which is free of knots and other imperfections. Have the piece ripped down the center to make a matched pair of nominal 1 by 2's. If there is a slight curvature the pieces may be assembled with the curves opposing so that the net result is a straight pole. If more height is desired, a 2 by 2 will fit nicely between the two legs at the bottom. The sections should be overlapped about one foot and bolts run through the three pieces. This has been done in the case of the antenna illustrated in figure 13.

Because the block under the insulator is small (about the same size as the insulator base) and has a hole through the center, it will have a tendency to split unless some close-grained wood such as oak is used. Undersize holes should be drilled before inserting the wood screws which hold the insulator base to the block. Two countersunk wood screws on each side are used to hold the block between the 1 by 2 uprights. The metal-strip clamp shown in figure 14 serves to make the whole assembly quite strong at this point.

A quarter wave of tubing at 14 Mc. has

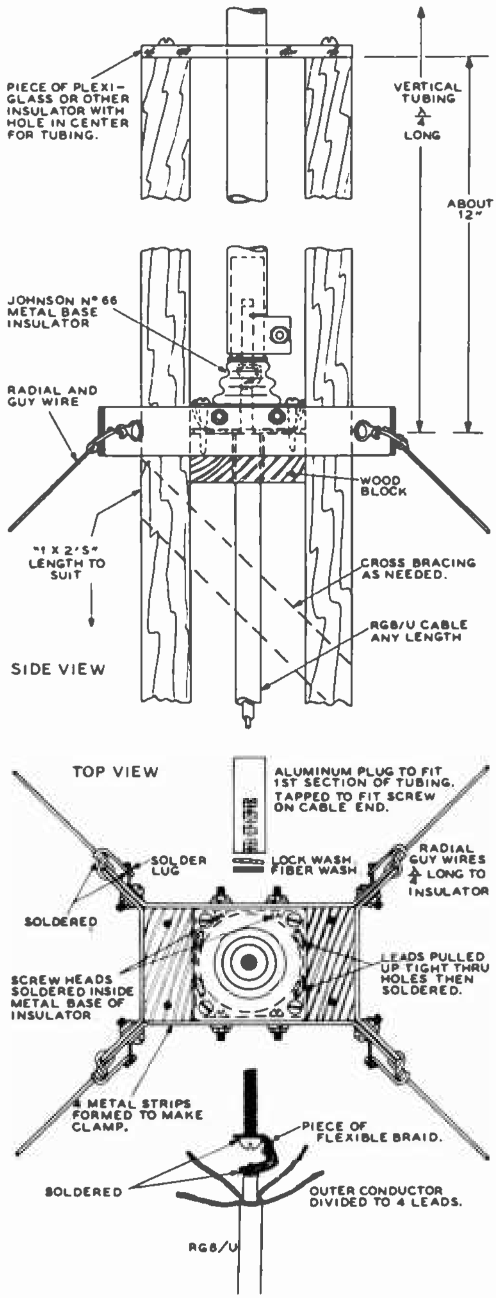


Figure 14.

CONSTRUCTIONAL DETAILS OF THE "DROOPING GROUND PLANE" VERTICAL.

considerable wind resistance, which may cause small tubing to bend at the top of the pole. Hence it will be wise to run the piece of tubing inserted into the first section clear to the bottom. This procedure will in effect double the wall thickness at the bottom and serve to strengthen the structure. Of course a standard whip antenna, either of the sectional or drawn tapered type may be used for the self-supporting radiator portion of the antenna system in place of the sections of aluminum tubing.

When the antenna is to be used on several bands, install "egg" insulators at the appropriate positions in the guy-wire "radials." These then may be jumpered for the lower frequency bands, and only the first section used on the highest frequency band. Similarly, the first section of aluminum tubing above the supporting insulator should be the correct length for the highest frequency band. Then the added sections for the lower frequency bands should be pre-cut to length so that when they are inserted fully to the base of the bottom section the total length will be correct for the desired band.

13-7 Dummy Antennas

In order to test a radio transmitter it is necessary that the full power output of the transmitter be delivered to some type of dissipative load. The radio law states that it is not permissible to test transmitter operation with the antenna connected except for very brief periods. This means therefore that for any type of extensive testing some sort of a dummy antenna must be provided.

The cheapest form of dummy antenna will consist of a 115-volt lamp or a group of 115-volt lamps coupled to the output of the transmitter.

If a lamp or lamps are chosen of such value that they light up approximately to normal brilliance at normal transmitter input, the approximate output may be determined by comparing the brilliance of the lamps with similar lamps connected to the 115-volt line. It is difficult to obtain an accurate measurement of the output by measuring the r-f current through the lamps and applying Ohm's Law because the resistance of the filament within the lamp cannot be determined accurately. The resistance of a light bulb varies considerably with the amount of current passing through it and with the frequency of the current.

It will be found best when testing a high-power transmitter to use a number of medium wattage lamps (100 to 200 watts) in series parallel rather than coupling the entire output of the transmitter to a single high wattage lamp. If the full energy output of the transmitter is coupled into a single lamp at frequencies in the amateur range, it is quite likely that dielectric breakdown within the stem of the lamp will be the result.

Another type of dummy load that may be used with transmitters having power output from 500 watts to many kilowatts is simply a tank of water. To control the rate of temperature rise in the water the tank (which may be constructed of water-proofed plywood with the seams sealed with tar) should contain approximately five gallons of water for each kilowatt to be dissipated. Pieces of no. 10 copper wire are inserted into the water several inches and spaced as far as it is possible to space them within the dimensions of the tank. Such a load presents a fairly low capacitive reactance in addition to its resistive component. The capacitive reactance may be tuned out with the antenna coupling network of the transmitter. The resistive component of the impedance may be varied from perhaps 100 to 600 ohms by varying the spacing and the depth of insertion of the wire electrodes.

For relatively accurate measurements of the r-f output of the transmitter dummy antenna resistors having a resistance that is substantially constant with varying dissipation are offered by Ohmite in various power ratings. These resistors are available in various resistances between 50 and 600 ohms and can be considered purely resistive and substantially constant in value at frequencies below 15 Mc. It will be noted that the stock resistance values of the dummy antenna loads correspond to the surge impedances of the most common transmission lines.

The Sprague Mfg. Company also makes available small non-inductive resistors in their Koolohm series which are suitable for making power measurements on medium-power transmitters. This type of non-inductive resistor is available in ratings from 5 watts to 120 watts and in resistance values over a rather complete range.

13-8 Matching Non-Resonant Lines to the Antenna

Present practice in regard to the use of transmission lines for feeding antenna systems on the amateur bands is about equally divided between three types of transmission line: (1) Ribbon or tubular molded 300-ohm line is widely used up to moderate power levels (the "transmitting" type is usable up to the kilowatt level). (2) Open-wire 400 to 600 ohm line is most commonly used when the antenna is some distance from the transmitter, due to the low attenuation of this type of line. (3) Coaxial line (usually RG-8/U with a 52-ohm characteristic impedance) is widely used in v-h-f work and also on the lower frequencies where the feed line must run underground or through the walls of a building. Coaxial line also is of assistance in TVI reduction since the r-f field is entirely enclosed within the line. Molded 75-ohm line is sometimes used to feed a doublet antenna, but the doublet has been largely superseded by the folded-dipole antenna fed by 300-ohm ribbon or tubular line when an antenna for a single band is required.

Standing Waves As was discussed in Chapter Twelve, standing waves on the antenna transmission line, in the transmitting case, are a result of reflection from the point where the feed line joins the antenna system. The magnitude of the standing waves is determined by the degree of mismatch between the characteristic impedance of the transmission line and the input impedance of the antenna system. When the feed-point impedance of the antenna is resistive and of the same value as the characteristic impedance of the feed line, standing waves will not exist on the feeder. It may be well to repeat at this time that there is no adjustment which can be made at the transmitter end of the feed line which will change the magnitude of the standing waves on the antenna transmission line.

Delta-Matched Antenna System The delta type matched-impedance antenna system is shown in figure 15. The impedance of the transmission line is transformed gradually into a higher value by the fanned-out Y portion of the feeders, and the Y portion is tapped on the antenna at points

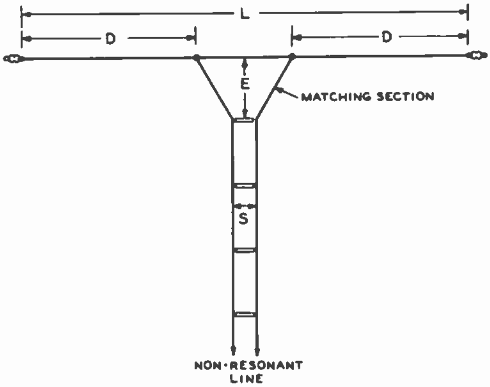


Figure 15.
THE DELTA-MATCHED DIPOLE ANTENNA.
The dimensions for the portions of the antenna are given in the text.

where the antenna impedance is a compromise between the impedance at the ends of the Y and the impedance of the unfanned portion of the line.

The constants of the system are rather critical, and the antenna must resonate at the operating frequency in order to minimize standing waves on the line. Some slight readjustment of the taps on the antenna is desirable, if appreciable standing waves persist in appearing on the line.

The constants are determined by the following formulas:

$$L_{feet} = \frac{467.4}{F_{megacycles}}$$

$$D_{feet} = \frac{175}{F_{megacycles}}$$

$$E_{feet} = \frac{147.6}{F_{megacycles}}$$

where L is antenna length; D is the distance *in from each end* at which the Y taps on; E is the height of the Y section.

As these constants are corrected only for a 600-ohm transmission line, the spacing S of the line must be approximately 75 times the diameter of the wire used in the transmission line. For no. 14 B & S wire, the spacing will be slightly less than 5 inches. For no. 12 B &

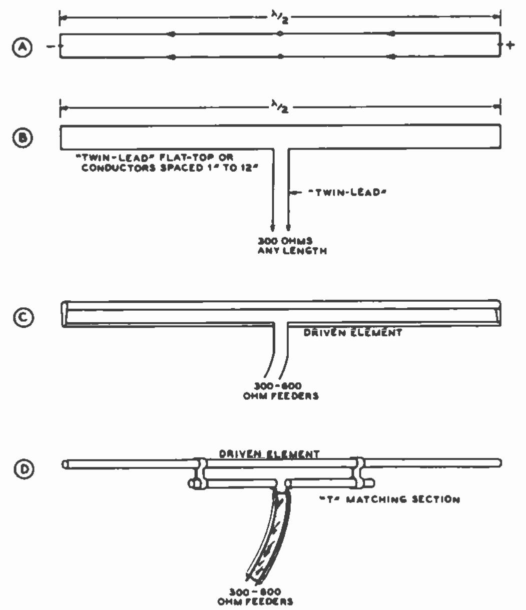


Figure 16.

FOLDED-ELEMENT MATCHING SYSTEMS.

Drawing (A) above shows a half-wave made up of two parallel wires. If one of the wires is broken as in (B) and the feeder connected, the feed-point impedance is multiplied by four; such an antenna is commonly called a "folded doublet." The feed-point impedance for a simple half-wave doublet fed in this manner is approximately 300 ohms, depending upon antenna height. Drawing (C) shows how the feed-point impedance can be multiplied by a factor greater than four by making the half of the element that is broken smaller in diameter than the unbroken half. An extension of the principles of (B) and (C) is the arrangement shown at (D) where the section into which the feeders are connected is considerably shorter than the driven element. This system is most convenient when the driven element is too long (such as for a 28-Mc. or 14-Mc. array) for a convenient mechanical arrangement of the system shown at (C).

S, the spacing should be 6 inches. This system should never be used on either its even or odd harmonics, as entirely different constants are required when more than a single half wavelength appears on the radiating portion of the system.

Using Delta Match to Parasitic Arrays The delta match is one of the two most popular systems for feeding a parasitic array such as the three-element and four-element rotaries. The other commonly-used system is the T-match which is discussed in the following section. Experience has shown that the adjustment for accurate match between the transmission line and the driven element of the array must be made by the cut-and-try process for minimum standing waves. An adjustment which has given low standing waves on a 480-ohm line (no. 12 wire spaced 2 inches) is to have the line tapped on 24 inches each side of the center of the array, with the drop to the first spreader (dimension E in figure 15) of about 40 inches. These dimensions are for a 28-Mc. three-element parasitic array with 0.15 wavelength spacing between driven element and either of the other two elements. These dimensions can be used as a starting point, and the exact adjustment determined by cut-and-try from these dimensions. The dimensions should of course be doubled for a 14-Mc. array.

Multi-Wire Doublets When a doublet antenna or the driven element in an array consists of more than one wire or tubing conductor the radiation resistance of the antenna or array is increased slightly as a result of the increase in the effective diameter of the element. Further, if we split just one wire of such a radiator, as shown in figure 16, the effective *feed-point* resistance of the antenna or array will be increased by a factor of N^2 where N is equal to the number of conductors, all in parallel, of the same diameter in the array. Thus if there are two conductors of the same diameter in the driven element or the antenna the feed-point resistance will be multiplied by 2^2 or 4. If the antenna has a radiation resistance of 75 ohms its feed-point resistance will be 300 ohms—this is the case of the conventional *folded-dipole* as shown in figure 16B.

If three wires are used in the driven radiator the feed-point resistance is increased by a factor of 9; if four wires are used the impedance is increased by a factor of 16, and so on. In certain cases when feeding a parasitic array it is desirable to have an impedance step up different from the value of 4:1 obtained with two elements of the same diameter and 9:1 with three elements of the same diameter. Intermediate values of impedance step up may be obtained by using two elements of different diameter for the complete driven element as shown in figure 16C. If the conductor that is broken for the feeder is of *smaller* diameter than the other conductor of the radiator, the impedance step up will be *greater* than 4:1. On the other hand if the *larger* of the two elements is broken for the feeder the impedance step up will be *less* than 4:1.

As an example of the use of the system shown in figure 16C, the following parasitic array was tested: frequency, 50-Mc. band; 0.2 wavelength spacing between reflector and driven element, between driven element and first director, and between first and second director. With the driven element and all parasitic elements constructed from 1-inch dural tubing, and with a full-length piece of quarter-inch copper tubing below the driven element as shown in figure 16C, an excellent match with low standing waves was obtained with 300-ohm Twin-Lead feeder connected in the center of the piece of ¼-inch copper tubing. The spacing between the outsides of the ¼-inch section and the 1-inch section of the driven element was 1 inch. Should this array be scaled up or down for use on another band the dimensions for the diameters and spacing for the two pieces which make up the driven element should not be changed.

The "T" Match A method of matching a low-impedance transmission line to the driven element of a parasitic array is the "T" match illustrated in figure 16D. This method is an adaptation of the multi-wire doublet principle which is more practicable for lower-frequency parasitic arrays such as those for use on the 14-Mc. and 28-Mc. bands. In the system a section of tubing of the same diameter as the driven element is spaced about 1 inch from the driven element by means of clamps which hold the "T" section mechanically and which make electrical connection to the driven element. The length of the "T" section is normally between 24" and 40" each side of the center of the driven dipole for the case of feeding a 3- or 4-element array from 300- to 600-ohm transmission line when the array is operating on the 28-Mc. band. These dimensions either side of center are doubled for the 14-Mc. band. It is well to allow for an adjustment out to 4 feet either side of center on the "T" section for a 28-Mc. antenna.

One particular array constructed for the 28-Mc. band ended up with the following dimensions for minimum standing waves of a 300-ohm Twin-Lead feeder: three-element array with 0.2 spacing between elements; tubing diameter for all elements, 1½-inch dural; spacing betwen "T" section and driven element, 2"; gap between inside faces of "T" section where feeder is attached, ½ inch; length of "T" section either side of center 40".

13-9 Matching Stubs

By hanging a resonant section of transmission line (called a matching stub) from either a voltage or current loop and attaching parallel-wire non-resonant feeders to the resonant stub at a suitable voltage (impedance) point, standing waves on the line may be virtually eliminated. The stub is made to serve as an auto-transformer. Stubs are particularly adapted to matching an open line to certain directional arrays, as will be described later.

Voltage Feed When the stub attaches to the antenna at a voltage loop, the stub should be a quarter wavelength long electrically, and be shorted at the bottom end. The stub can be resonated by sliding the shorting bar up and down before the non-resonant feeders are attached to the stub, the antenna being shock-excited from a separate radiator during the process. Slight errors in the length of the radiator can be compensated for by adjustment of the stub if both sides of the stub are connected to the radiator in a symmetrical manner. Where only one side of the stub connects to the radiating system, as in the Zepp and in certain antenna arrays, the radiator length must be exactly right in order to prevent excessive unbalance in the untuned line.

If only one leg of a stub is used to voltage-feed a radiator, it is impossible to secure a perfect balance in the transmission line due to a slight inherent unbalance in the stub itself when one side is left floating. This unbalance should not be aggravated by a radiator of improper length.

Current Feed When a stub is used to current-feed a radiator, the stub should either be left *open* at the bottom end instead of shorted, or else made a *half wave* long. The open stub should be resonated in the same manner as the shorted stub before attaching the transmission line; however, in this case,

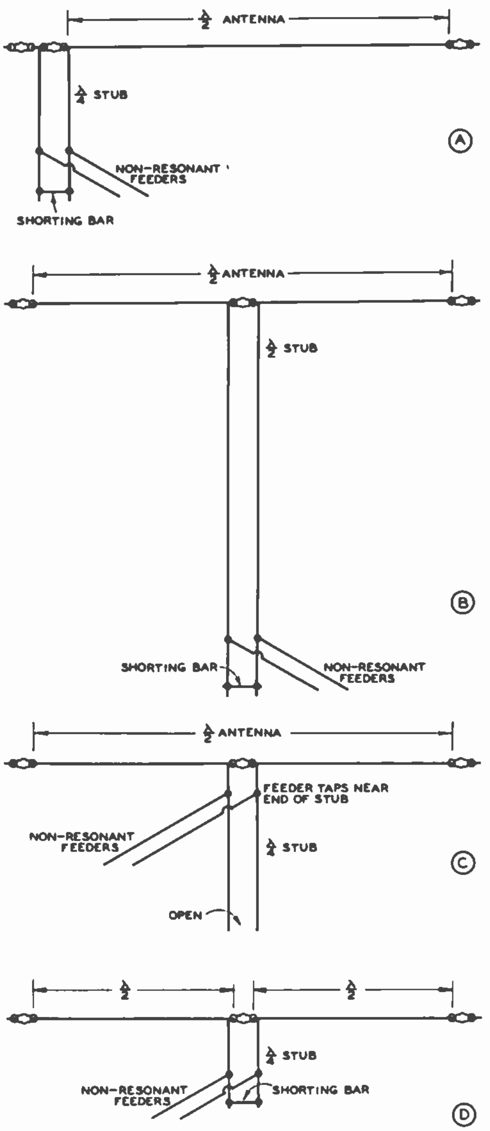


Figure 17.

MATCHING-STUB APPLICATIONS.

An end-fed half-wave antenna with a quarter-wave shorted stub is shown at (A). (B) shows the use of a half-wave shorted stub to feed a relatively low impedance point such as the center of the driven element of a parasitic array, or the center of a half-wave dipole. The use of an open-ended quarter-wave stub to feed a low impedance is illustrated at (C). (D) shows the conventional use of a shorted quarter-wave stub to voltage feed two half-wave antennas with a 180° phase difference.

it is necessary to prune the stub to resonance, as there is no shorting bar.

Sometimes it is handy to have a stub hang from the radiator to a point that can be reached from the ground, in order to facilitate adjustment of the position of the transmission-line attachment. For this reason, a quarter-wave stub is sometimes made three-quarters wavelength long at the higher frequencies, in order to bring the bottom nearer the ground. Operation with any *odd* number of quarter waves is the same as for a quarter-wave stub.

Any number of *half waves* can be added to either a quarter-wave stub or a half-wave stub without disturbing the operation, though losses and frequency sensitivity will be lowest if the shortest usable stub is employed.

Stub Length (Electrical)	Current-Fed Radiator	Voltage-Fed Radiator
¼-¾-1¼-etc. wavelengths	Open	Shorted
½-1-1½-2-etc. wavelengths	Shorted	Open

Shorted-Stub Tuning Procedure When the antenna requires a shorted stub (odd number of quarter waves if the antenna is voltage-fed, or an even number if the radiator is current-fed), the tuning procedure is as follows:

Shock-excite the radiator (or one of the half-wave sections, if harmonically operated) by means of a makeshift doublet located as far distant as will give a useful indication, with the temporary antenna connected to the transmitter by means of any convenient type of transmission line.

With the feeders and shorting bar disconnected from the stub, slide along an r-f milliammeter or low-current dial light at about where you calculate the shorting bar should be, and find the point of maximum current (in other words, use the meter or lamp as a shorting bar).

It is best to start with reduced power to the transmitter, until you see how much of an indication you may expect; otherwise, the meter or lamp may be blown on the initial trial. The lamp or meter leads should be no longer than necessary to reach across the stub.

After finding the point of maximum current, remove the lamp or meter and connect a piece of wire across the stub at that point.

Starting at a point about a quarter of a quarter wave (2 feet on 10 meters) from the shorting bar, connect the feeders to the stub. Then, move the feeders up and down the stub until the standing waves on the line are at a minimum. The makeshift doublet should, of course, be disconnected and the regular feeders connected to the transmitter during this process. Slight readjustment of the shorting bar to lengthen the stub will usually result in further improvement.

The standing wave indicator may be either a voltage device, such as a neon bulb, or a current device, such as an r-f milliammeter connected to a pickup coil, or a bridge-type standing-wave indicator may be used. A high degree of accuracy is not required.

The following rule will indicate in which direction the feeders should be moved in an attempt to minimize standing waves: If the current increases on the transmission line as the indicator is moved away from the point of attachment to the stub, the feeders are attached too far from the shorting bar, and must be moved closer to the shorting bar; if the current decreases, the feeders must be attached farther from the shorting bar.

Open-Ended Stub Tuning Procedure If the antenna requires an open stub (even number of quarter waves if the antenna is voltage-fed, odd number of quarter waves if it is current-fed), the tuning procedure is as follows:

Shock-excite the radiator as described for tuning a shorted-stub system, feeders disconnected from the stub, and stub cut slightly longer than the calculated value. Place a field strength meter (the standing wave indicator can be very easily converted into one by addition of a tuned tank) close enough to one end of the radiator to get a reading, and as far as possible from the makeshift exciting antenna. Now, start folding and clipping the stub wires back on themselves a few inches at a time, effectively shortening their length, until you find the peak as registered on the field meter.

Next, attach the feeders to the stub as described for the shorted-stub system, but, for the initial trial connection, the feeders will attach at a distance more nearly three-quarters of a quarter wave from the end of the stub instead of a quarter of a quarter wave, as is the case for a shorted stub. After attaching the feeders, move them along the stub as necessary to minimize standing waves on the line. If sliding the feeders along the stub a few inches makes the standing waves worse, it means the correct connecting point is in the other direction.

After the optimum point on the stub is found for the feeder attachment, the length of the stub can be "touched up" for a final adjustment to minimize standing waves. This is advisable because the attachment of the feeder usually will detune the stub slightly.

Important Note on Stub Adjustment When a stub is used to match a line to an impedance *of the same order of impedance as that of the surge impedance of the stub and line* (assuming the stub and line use the same wire size and spacing), it will be found that attaching the feeders to the stub introduces a large amount of reactance. The length of the stub then must be altered considerably to restore resonance.

Unfortunately, alteration of the stub length requires that the position of attachment of the feeders be readjusted. Consequently, the adjustments entail considerable juggling of both stub length and point of feeder attachment, in order to minimize both reactance and standing waves. Hence, it is recommended that a quarter-

Frequency in Kilocycles	Quarter-wave matching section or stub	Half-wave radiator
3500	70'3"	133'7"
3600	68'5"	129'10"
3700	67'6"	126'4"
3800	64'10"	123'
3900	63'1"	119'10"
3950	62'3"	118'4"
4000	61'6"	116'10"
7000	35'1"	66'9"
7150	34'5"	65'4"
7300	33'8"	64'
14,000	17'7"	33'5"
14,200	17'4"	32'11"
14,400	17'1"	32'6"
28,000	8'9"	16'8"
28,500	8'7"	16'5"
29,000	8'6"	16'1"
29,500	8'4"	15'10"

DIMENSIONS FOR HALF-WAVE RADIATOR AND QUARTER-WAVE MATCHING STUB OR Q SECTION.

wave transformer (Section 13-10) be used instead of a stub for impedance matching when the main feed-line impedance is within a ratio of perhaps 4 to 1 to the feed-line impedance.

If a *shorted* stub is used to feed an impedance of *more* than 4 times that of the surge impedance of the stub and line, this effect will be small, and it is not absolutely necessary that the stub length be readjusted after the feeders are attached. Likewise, the length of an *open* stub need not be altered after attachment of the feeders, if the stub feeds an impedance of *less* than ¼ that of the surge impedance of the stub and line.

When not sure of the exact order of impedance into which the stub works, it is always advisable to try "touching up" the stub length after the feeders are attached.

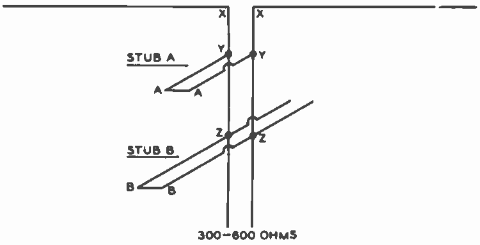


Figure 18.
TWO-FREQUENCY STUB-MATCHED ANTENNA SYSTEM.
Any antenna which has a radiating system capable of efficient operation on two widely separated frequencies may be matched to an open wire transmission line on both frequencies by use of two "reactance stubs" as shown here. Operation and adjustment are explained in the text.

Two-Frequency Stub Matching It is practicable to use matching stubs to match an untuned line to an antenna or array on two frequencies. The frequencies need not be harmonically related if the antenna itself is capable of good efficiency on both frequencies. However, the frequencies should be in a ratio not exceeding 4/1 nor less than 1.3/1.

The arrangement is illustrated in figure 18. The system is tuned up on the lowest frequency for minimum standing waves by means of adjusting the length and point of attachment of stub "A," stub "B" not yet being connected. After the standing waves are reduced to a negligible value, the transmitter is changed to the higher frequency. Stub "B," which is a quarter wave long on the *lower* frequency, then is attached experimentally, and the point of attachment varied until standing waves are at a minimum on the higher frequency. Because stub "B" is exactly a quarter wave long on the lower frequency, its attachment will have virtually no effect upon the operation of the antenna system at the lower frequency.

It should be kept in mind that stub "A" is tuned by varying the distances XY and AY; the stub does not "hang over" as does stub "B." The overall length of stub "B" is not altered; only the distances XZ and BZ are varied when adjusting for minimum standing waves on the higher frequency. It is possible that the position of the two stubs will be reversed from that shown in figure 18. This depends upon the particular antenna being fed, and the characteristic impedance of the feed line.

Standing Wave Indicators Many simple devices can be used for detecting the presence and approximate ratio of standing waves on a feed line. A 1-turn pickup loop, about 4 or 5 inches in diameter, may be attached to a current indicator, such as a small Mazda bulb or an r-f thermogalvanometer, to indicate current excursions along the line. The device should be attached to the end of a wood stick at least a foot long in order to minimize body capacitance. The loop is moved along the line in inductive relation to the feed line, care being taken to see that the loop always is in *exactly the same inductive relation* to the line. It should be kept in mind that this type of indicator is a *current* indicator.

A small neon bulb also may be used to indicate standing waves on a feed line. In this case, the indicator works on *voltage*, and it should be kept in mind that the *voltage* on the line normally is highest where the current is lowest. This type of indicator is operated by touching various parts of the bulb to *one* feeder wire until an indication of medium brilliancy is obtained. The bulb is then slid along the wire, *in exactly the same position and point of contact with the wire.* If the enamel insulation is not intact on all portions of the wire and the wire is exposed in spots, deceptive "bumps" will be noticed. The wire should be either uniformly insulated or uniformly bare throughout its length; otherwise, it will be necessary to place a thickness of insulating material over the exposed metal parts of the neon bulb, the bulb then working by virtue of capacitance to the wire, rather than direct contact.

If it is desired to measure the exact rather than relative standing wave ratio and an r-f meter is not available, a low range d-c milliammeter may be used instead, if a suitable rectifier is placed in series with the d-c meter. A 0-1 ma. d-c milliammeter in series with a 1N34 crystal rectifier is commonly used. As noted before, this type of indicator is a *current* indicator.

If a considerable amount of antenna and feeder work is planned or in progress, a direct-reading standing-wave indicator such as is described in Chapter Twenty-Six may be constructed, or a manufactured instrument such as the "Micro-Match" for making such measurements may be purchased.

13-10 Linear R. F. Transformers

A resonant quarter-wave line has the unusual property of acting much as a transformer. Let us take, for example, a section consisting of no. 12 wire spaced 6 inches, which happens to have a surge impedance of 600 ohms. Let the far end be terminated with a pure resistance, and let the near end be fed with radio-frequency energy at the frequency for which the line is a quarter wavelength long. If an impedance measuring set is used to measure the impedance at the near end while the impedance at the far end is varied, an interesting relationship betwen the 600-ohm characteristic surge impedance of this particular quarter-wave matching line, and the impedance at the ends will be discovered.

When the impedance at the far end of the line is the same as the characteristic surge impedance of the line itself (600 ohms), the impedance measured at the near end of the quarter-wave line will also be found to be 600 ohms.

Under these conditions, the line would not have any standing waves on it, since it is terminated in its characteristic impedance. Now, let the resistance at the far end of the line be doubled, or changed to 1200 ohms. The impedance measured at the near end of the line will be found to have been cut in half, to 300 ohms. If the resistance at the far end is made half the original value of 600 ohms or 300 ohms, the impedance at the near end doubles the original value of 600 ohms, and becomes 1200 ohms. As one resistance goes up, the other goes down proportionately.

It always will be found that the characteristic surge impedance of the quarter-wave matching line is the geometric mean betwen the impedance at both ends. This relationship is shown by the following formula:

$$Z_{MS} = \sqrt{Z_A Z_L}$$

where

Z_{MS} = Impedance of matching section.

Z_A = Antenna resistance.

Z_L = Line impedance.

Quarter-Wave Matching Transformers The impedance inverting characteristic of a quarter-wave section of transmission line is widely used by making such a section of line act as a *quarter-wave transformer*. The "Johnson Q" feed system is a widely known application of the quarter-wave transformer to the feeding of a dipole antenna and array consisting of two dipoles. However, the quarter-wave transformer may be used in a wide number of applications wherever a transformer is required to match two impedances whose geometric mean is somewhere between perhaps 25 and 750 ohms when transmission line sections can be used. Paralleled coaxial lines may be used to obtain the lowest impedance mentioned, and open-wire lines composed of small conductors spaced a moderate distance may be used to obtain the higher impedance. A short list of impedances which may be matched by quarter-wave sections of transmission line having specified impedances is given below.

Load or Ant. Impedance	300	480	600	Feed-Line Impedance
20	77	98	110	Quarter-Wave Transformer Impedance
30	95	120	134	
50	110	139	155	
75	150	190	212	
100	173	220	245	

The impedance values from 20 to 50 ohms are obtained in the center of the driven element of a *wide-spaced* (0.2 wave-length or so) parasitic array or at the bottom of the stub of a "flat-top beam." 75 ohms represents

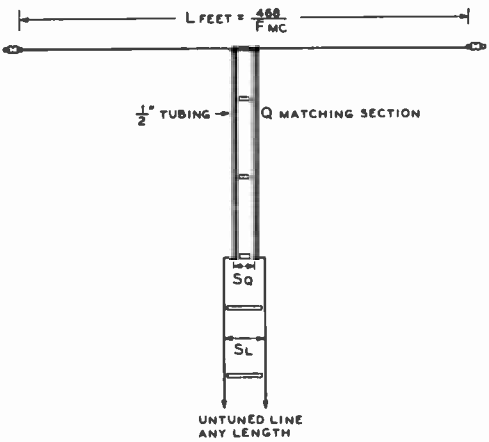


Figure 19.
HALF-WAVE RADIATOR FED BY "Q BARS."

The Q matching section is simply a quarter-wave transformer whose impedance is equal to the geometric mean between the impedance at the center of the antenna and the impedance of the transmission line to be used to feed the bottom of the transformer. The transformer may be made up of parallel tubing, a four-wire line, or any other type of transmission line which has the correct value of impedance.

the average center impedance of a half-wave doublet, and 100 ohms represents the approximate center impedance of one half wave of a full-wave antenna. Impedance values of 75 and 150 ohms can of course be obtained in Amphenol Twin-Lead, 100 and 200 ohms can be obtained, though less readily, in ribbon line of other manufacture. Impedance values from 175 to 275 ohms can readily be obtained either from a four-wire line or from large-diameter dural or aluminum tubing spaced closely together ("Q Bars").

Johnson-Q Feed System The standard form of Johnson-Q feed to a doublet is shown in figure 19. An impedance match is obtained by utilizing a matching section, the surge impedance of which is the geometric mean betwen the transmission line surge impedance and the radiation resistance of the radiator. A sufficiently good match usually can be obtained by either designing or adjusting the matching section for a dipole to have a surge impedance that is the geometric mean betwen the line impedance and 72 ohms, the latter being the theoretical radiation resistance of a half-wave doublet either infinitely high or a half wave above a perfect ground.

Though the radiation resistance may depart somewhat from 72 ohms under actual conditions, satisfactory results will be obtained with this assumed value, so long as the dipole radiator is more than a quarter wave above effective earth, and reasonably in the clear. The small degree of standing waves introduced by a slight mismatch will not increase the line losses appreciably, and any *small* amount of reactance present can be tuned out at the transmitter termination with no bad effects. If the reactance is objectionable, it may be minimized by making the untuned line an integral number of quarter waves long.

A Q-matched system can be adjusted precisely, if desired, by constructing a matching section to the calculated dimensions with provision for varying the spacing of the Q section conductors slightly, after the untuned line has been checked for standing waves.

The Q section usually will require about 200 ohms surge impedance when used to match a half-wave doublet, actually varying from about 150 to 250 ohms with different installations. This impedance is difficult to obtain with a two-wire line, as very close spacing would be required. For this reason, either a four-wire line or a line consisting of two half-inch aluminum tubes is ordinarily used. The four-wire section has the advantage of lightness and cheapness, and can be used where the approximate radiation resistance is known with certainty, thus making it possible to design the matching section for a certain value of surge impedance with some assurance that it will turn out to be sufficiently accurate.

PARALLEL TUBING SURGE IMPEDANCE FOR MATCHING SECTIONS.

Center to Center Spacing in Inches	Impedance in Ohms for ½" Diameters	Impedance in Ohms for ¼" Diameters
1	170	250
1.25	188	277
1.5	207	298
1.75	225	318
2	248	335

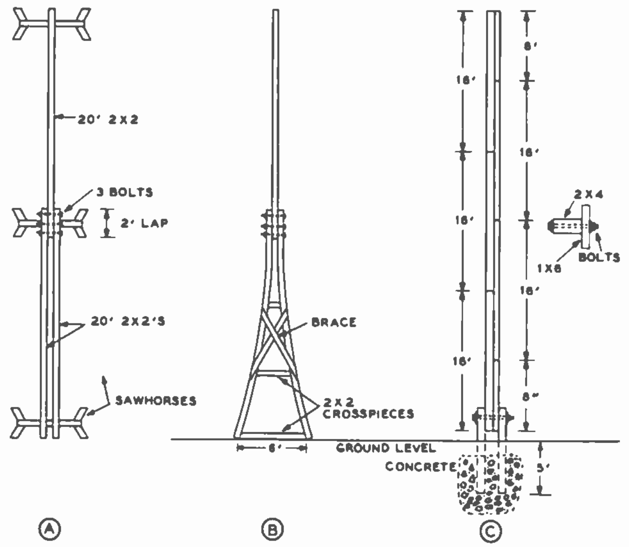


Figure 20.
TWO SIMPLE WOOD MASTS.
Shown at (A) is the method of assembly, and at (B) is the completed structure, of the conventional "A-frame" antenna mast. At (C) is shown a structure which is heavier but more stable than the A-frame for heights above about 40 feet.

Matching to Rotory Beam Antennas A considerable amount of additional information, with detailed tuning and adjustment procedure, on the problem of matching the antenna input impedance to that of the transmission line is given in Chapter Sixteen, *Rotatable Antenna Arrays.*

13-11 Antenna Construction

The foregoing portion of this chapter has been concerned primarily with the *electrical* characteristics and considerations of antennas. Some of the physical aspects and mechanical problems incident to the actual erection of antennas and arrays will be discussed in the following section.

Up to 60 feet, there is little point in using mast-type antenna supports unless guy wires either must be eliminated or kept to a minimum. While a little more difficult to erect, because of their floppy nature, fabricated wood poles of the type to be described will be just as satisfactory as more rigid types, *provided* many guy wires are used.

Rather expensive when purchased through the regular channels, 40- and 50-foot telephone poles sometimes can be obtained quite reasonably. In the latter case, they are hard to beat, inasmuch as they require no guying if set in the ground six feet (standard depth), and the resultant pull in any lateral direction is not in excess of a hundred pounds or so.

For heights of 80 to 100 feet, either three-sided or four-sided lattice type masts are most practicable. They can be made self-supporting, but a few guys will enable one to use a smaller cross section without danger from high winds. The torque exerted on the base of a high self-supporting mast is terrific during a strong wind.

The "A-Frame" Mast Figures 20A and 20B show the standard method of construction of the "A-frame" type of mast. This type of mast is quite frequently used since there is only a moderate amount of work involved in the construction of the assembly and since the material cost is relatively small. The three pieces of selected 2 by 2 are first set up on three sawhorses or boxes and the holes drilled for the three ¼-inch bolts through the center of the assembly. Then the base legs are spread out to about 6 feet and the bottom braces installed. Then the upper braces and the cross pieces are installed and the assembly given several coats of good-quality paint as a protection against weathering.

Figure 20C shows another common type of mast which is made up of sections of 2 by 4 placed end-to-end with stiffening sections of

1 by 6 bolted to the edge of the 2 by 4 sections. Both types of masts will require a set of top guys and another set of guys about one-third of the way down from the top. Two guys spaced about 90 to 100 degrees and pulling against the load of the antenna will normally be adequate for the top guys. Three guys are usually used at the lower level, with one directly behind the load of the antenna and two more spaced 120 degrees from the rear guy.

The raising of the mast is made much easier if a "gin pole" about 20 feet high is installed about 30 or 40 feet to the rear of the direction in which the antenna is to be raised. A line from a pulley on the top of the gin pole is then run to the top of the pole to be raised. The gin pole comes into play when the center of the mast has been raised 10 to 20 feet above the ground and an additional elevated pull is required to keep the top of the mast coming up as the center is raised further above ground.

Using TV Masts Steel tubing masts of the telescoping variety are widely available at a moderate price for use in supporting television antenna arrays. These masts usually consist of several 10-foot lengths of electrical metal tubing (EMT) of sizes such that the sections will telescope. The 30-foot and 40-foot lengths are well suited as masts for supporting antennas and arrays of the types used on the amateur bands. The masts are constructed in such a manner that the bottom 10-foot length may be guyed permanently before the other sections are raised. Then the upper sections may be extended, beginning with the top-mast section, until the mast is at full length (provided a strong wind is not blowing) following which all the guys may be anchored. It is important that there be no load on the top of the mast when the "vertical" raising method is to be employed.

Guy Wires Guy wires should never be pulled taut; a *small* amount of slack is desirable. Galvanized wire, somewhat heavier than seems sufficient for the job, should be used. The heavier wire is a little harder to handle, but costs only a little more and takes longer to rust through. Care should be taken to make sure that no kinks exist when the pole or tower is ready for erection, as the wire will be greatly weakened at such points if a kink is pulled tight, even if it is later straightened.

If "dead men" are used for the guy wire terminations, the wire or rod reaching from the dead men to the surface should be of non-rusting material, such as brass, or given a heavy coating of asphalt or other protective substance to prevent destructive action by the damp soil. Galvanized iron wire will last only a short time when buried in moist soil.

Only strain-type (compression) insulators should be used for guy wires. Regular ones might be sufficiently strong for the job, but it is not worth taking chances, and egg-type strain halyard insulators are no more expensive.

Only a brass or bronze pulley should be used for the halyard, as a nice high pole with a rusted pulley is truly a sad affair. The bearing of the pulley should be given a few drops of heavy machine oil before the pole or tower is raised. The halyard itself should be of good material, preferably water-proofed. Hemp rope of good quality is better than window sash cord from several standpoints, and is less expensive. Soaking it thoroughly in engine oil of medium viscosity, and then wiping it off with a rag, will not only extend its life but minimize shrinkage in wet weather. Because of the difficulty in replacing a broken halyard it is a good idea to replace it periodically, without waiting for it to show excessive wear or deterioration.

It is an excellent idea to tie both ends of the halyard line together in the manner of a flag-pole line. Then the antenna is tied onto the place where the two ends of the halyard are joined. This procedure of making the halyard into a loop prevents losing the "up" end of the halyard should the antenna break near the end, and it also prevents losing the halyard completely should the end of the halyard carelessly be allowed to go free and be pulled through the pulley at the top of the mast by the antenna load. A somewhat longer piece of line is required but the insurance is well worth the cost of the additional length of rope.

Trees as Supports Often a tall tree can be called upon to support one end of an antenna, but one should not attempt to attach anything to the top, as the swaying of the top of the tree during a heavy wind will complicate matters.

If a tree is utilized for support, provision should be made for keeping the antenna taut without submitting it to the possibility of being severed during a heavy wind. This can be done by the simple expedient of using a pulley and halyard, with weights attached to the lower end of the halyard to keep the antenna taut. Only enough weight to avoid excessive sag in the antenna should be tied to the halyard, as the continual swaying of the tree submits the pulley and halyard to considerable wear.

Galvanized iron pipe, or steel-tube conduit, is often used as a vertical radiator, and is quite satisfactory for the purpose. However, when used for supporting antennas, it should be remembered that the grounded supporting poles will distort the field pattern of a vertically polarized antenna unless spaced some distance from the radiating portion.

Painting The life of a wood mast or pole can be increased several hundred per cent by protecting it from the elements with a coat or two of paint. And, of course, the appearance is greatly enhanced. The wood should first be given a primer coat of flat white outside house paint, which can be thinned down a bit to advantage with second-grade linseed oil. For the second coat, which should not be applied until the first is thoroughly dry, *aluminum paint* is not only the best from a preservative standpoint, but looks very well. This type of paint, when purchased in quantities, is considerably cheaper than might be gathered from the price asked for quarter-pint cans.

Portions of posts or poles below the surface of the soil can be protected from termites and moisture by painting with creosote. While not so strong initially, redwood will deteriorate much more slowly when buried than will the white woods, such as pine.

Antenna Wire The antenna or array itself presents no especial problem. A few considerations should be borne in mind, however. For instance, soft-drawn copper should not be used, as even a short span will stretch several per cent after whipping around in the wind a few weeks, thus affecting the resonant frequency. Enameled-copper wire, as ordinarily available at radio stores, is usually soft drawn, but by tying one end to some object such as a telephone pole and the other to the frame of an auto, a few husky tugs can be given and the wire, after stretching a bit, is equivalent to hard drawn.

Where a long span of wire is required, or where heavy insulators in the center of the span result in considerable tension, copper-clad steel wire is somewhat better than hard-drawn copper. It is a bit more expensive, though the cost is far from prohibitive. The use of such wire, in conjunction with strain insulators, is advisable, where the antenna would endanger persons or property should it break.

For transmission lines and tuning stubs steel-core or hard-drawn wire will prove awkward to handle, and soft-drawn copper should, therefore, be used. If the line is long, the strain can be eased by supporting it at several points.

The use of copper tubing for antennas (except at v.h.f.) is not only expensive but unjustifiable. Though it was a fad at one time, there is no excuse for using anything larger than no. 10 copper or copper-clad wire for any power up to 1 kilowatt. In fact, no. 12 will do the trick just as well, and passes the underwriter's rules if copper-clad steel is used. For powers of less than 100 watts, the underwriter's rules permit no. 14 wire of solid copper. This size is practically as efficient as larger wire, but will not stand the pull that no. 12 or no. 10 will, and the underwriter's rules call for the latter for powers in excess of 100 watts, if solid copper conductor is used.

More important from an electrical standpoint than the actual size of wire used is the soldering of joints, especially at current loops in an antenna of low radiation resistance. In fact, it is good practice to solder *all* joints, thus insuring quiet operation when the antenna is used for receiving.

Insulation A question that often arises is that of insulation. It depends, of course, upon the r-f voltage at the point at which the insulator is placed. The r-f voltage, in turn, depends upon the distance from a current node, and the radiation resistance of the antenna. Radiators having low radiation resistance have very high voltage at the voltage loops; consequently, better than usual insulation is advisable at those points.

Open-wire lines operated as non-resonant lines have little voltage across them; hence the most inexpensive ceramic types are sufficiently good electrically. With tuned lines, the voltage depends upon the amplitude of the standing waves. If they are very great, the voltage will reach high values at the voltage loops, and the best spacers available are none too good. At the current loops the voltage is quite low, and almost anything will suffice.

When insulators are subject to very high r-f voltages, they should be cleaned occasionally if in the vicinity of sea water or smoke. Salt scum and soot are not readily dislodged by rain, and when the coating becomes heavy enough, the efficiency of the insulators is greatly impaired.

If a very pretentious installation is to be made, it is wise to check up on both underwriter's rules and local ordinances which might be applicable. If you live anywhere near an airport, and are contemplating a tall pole, it is best to investigate possible regulations and ordinances pertaining to towers in the district, before starting construction.

CHAPTER FOURTEEN

High Frequency Directive Antenna Arrays

It is becoming of increasing importance in most types of radio communication to be capable of concentrating the radiated signal from the transmitter in a certain desired direction and to be able to discriminate at the receiver against reception from directions other than the desired one. Such capabilities involve the use of directive antenna arrays.

Few simple antennas, except the single vertical element, radiate energy equally well in all azimuth (horizontal or compass) directions. All horizontal antennas, except those specifically designed to give an omnidirectional azimuth radiation pattern such as the turnstile, have some directive properties. These properties depend upon the length of the antenna in wavelengths, the height above ground, and the slope of the radiator.

The various forms of the half-wave horizontal antenna produce maximum radiation at right angles to the wire, but the directional effect is not great. Nearby objects also minimize the directivity of a dipole radiator, so that it hardly seems worth while to go to the trouble to rotate a simple half-wave dipole in an attempt to improve transmission and reception in any direction.

The half-wave doublet, folded dipole, zepp, single-wire-fed, matched impedance, and Johnson Q antennas all have practically the same radiation pattern *when properly built and adjusted*. They all are dipoles, and the feeder system, if it does not radiate in itself, will have no effect on the radiation pattern.

Directive Antennas When a multiplicity of radiating elements is located and phased so as to reinforce the radiation in certain desired directions and to neutralize radiation in other directions, a directive antenna array is formed.

The function of a directive antenna when used for *transmitting* is to give an increase in signal strength in some direction at the expense of radiation in other directions. For *reception*, one might find useful an antenna giving little or no gain in the direction from which it is desired to receive signals if the antenna is able to *discriminate against interfering* signals and static arriving from other directions. A good directive transmitting antenna, however, can also be used to good advantage for reception.

If radiation can be confined to a narrow beam, the signal intensity can be increased a great many times in the desired direction of transmission. This is equivalent to increasing the power output of the transmitter. On the higher frequencies, it is more economical to use a directive antenna than to increase transmitter power, if more than a few watts of power is being used.

Directive antennas for the high-frequency range have been designed and used commer-

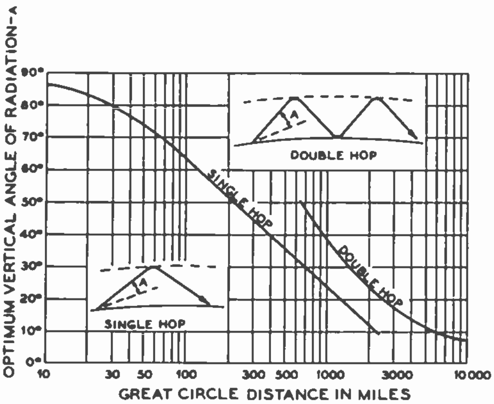


Figure 1.

OPTIMUM ANGLE OF RADIATION WITH RESPECT TO DISTANCES.

Shown above is a plot of the optimum angle of radiation for one-hop and two-hop communication. An operating frequency close to the optimum working frequency for the communication distance is assumed.

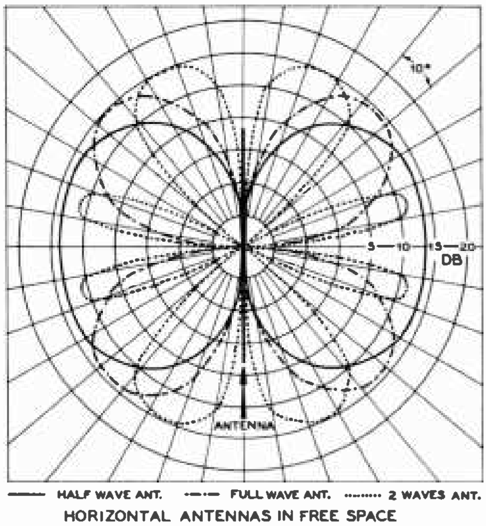


Figure 2.

FREE-SPACE FIELD PATTERNS OF LONG-WIRE ANTENNAS.

The presence of the earth distorts the field pattern in such a manner that the azimuth pattern becomes a function of the elevation angle.

cially with gains as high as 23 db over a simple dipole radiator. Gains as high as 35 db are common in direct-ray microwave communication and radar systems. A gain of 23 db represents a power gain of 200 times and a gain of 35 db represents a power gain of almost 3500 times. However, an antenna with a gain of only 15 to 20 db is so sharp in its radiation pattern that it is usable to full advantage only for point-to-point work.

The increase in radiated power in the desired direction is obtained at the expense of radiation in the undesired directions. Power gains of 3 to 12 db seem to be most practicable for amateur communication, since the width of a beam with this order of power gain is wide enough to sweep a fairly large area. Gains of 3 to 12 db represent effective transmitter power increases from 2 to 16 times.

Horizontal Pattern vs. Vertical Angle There is a certain optimum vertical angle of radiation for sky-wave communication, this angle being dependent upon distance, frequency, time of day, etc. Energy radiated at an angle much lower than this optimum angle is largely lost, while radiation at angles much higher than this optimum angle oftentimes is not nearly so effective.

For this reason, the horizontal directivity pattern as measured on the ground is of no import when dealing with frequencies and distances dependent upon sky-wave propagation. It is the horizontal directivity (or gain or discrimination) *measured at the most useful vertical angles of radiation* that is of consequence. The horizontal radiation pattern, as measured on the ground, is considerably different from the pattern obtained at a vertical angle of 15°, and still more different from a pattern obtained at a vertical angle of 30°. In general, the energy which is radiated at angles higher than approximately 30° above the earth is effective at any frequency only for local work.

For operation at frequencies in the vicinity of 14 Mc., the most effective angle of radiation is usually about 15° above the horizon, from *any* kind of antenna. The most effective angles for 10-meter operation are those in the vicinity of 10°. Figure 1 is a chart giving the optimum vertical angle of radiation for sky-wave propagation in terms of the great-circle distance between the transmitting and receiving antennas.

Types of Directive Arrays There is an enormous variety of directive antenna arrays that can give a substan-

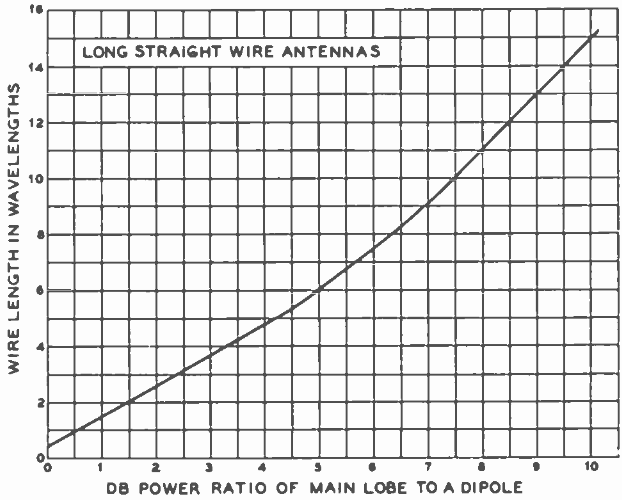


Figure 3.
DIRECTIVE GAIN OF LONG-WIRE ANTENNAS.

tial power gain in the desired direction of transmission or reception. However, some are more effective than others requiring the same space. In general it may be stated that long-wire antennas of various types, such as the single long wire, the V beam, and the rhombic, are less effective for a given space than arrays composed of resonant elements, but the long-wire arrays have the significant advantage that they may be used over a relatively large frequency range while resonant arrays are usable only over a quite narrow frequency band.

14-1 Long Wire Radiators

Harmonically operated long wires radiate better in certain directions than others, but cannot be considered as having appreciable directivity unless several wavelengths long. The current in adjoining half-wave elements flows in opposite directions at any instant, and thus, the radiation from the various elements adds in certain directions and cancels in others.

A half-wave doublet in free space has a "doughnut" of radiation surrounding it. A full wave has 2; 3 half waves 3; and so on. When the radiator is made more than 4 half wavelengths long, the *end* lobes (cones of radiation) begin to show noticeable power gain over a half-wave doublet, while the broadside lobes get smaller and smaller in amplitude, even though numerous.

The horizontal radiation pattern of such antennas depends upon the vertical angle of radiation being considered. If the wire is more than 4 wavelengths long, the maximum radiation at vertical angles of 15° to 20° (useful for dx) is in line with the wire, being slightly greater a few degrees either side of the wire than directly off the ends. The directivity of the main lobes of radiation is not particularly sharp, and the minor lobes fill in between the main lobes to permit working stations in nearly all directions, though the power radiated broadside to the radiator will not be great if the radiator is more than a few wavelengths long. The directive gain of long-wire antennas, in terms of the wire length in wavelengths is given in figure 3.

To maintain the out-of-phase condition in adjoining half-wave elements throughout the length of the radiator, it is necessary that a harmonic antenna be fed either at *one end* or at a *current* loop. If fed at a voltage loop, the adjacent sections will be fed *in* phase, and a different radiation pattern will result.

The directivity of a long wire does not increase very much as the length is increased beyond about 15 wavelengths. This is due to the fact that all long-wire antennas are adversely affected by the r-f resistance of the wire, and because the current amplitude begins to become unequal at different current loops, as a result of attenuation along the wire caused by radiation and losses. As the length is increased, the tuning of the antenna becomes quite broad. In fact, a long wire about

LONG-ANTENNA DESIGN TABLE.

Approximate Length in Feet — End-Fed Antennas

Frequency in Mc.	1λ	1½λ	2λ	2½λ	3λ	3½λ	4λ	4½λ
30	32	48	65	81	97	104	130	146
29	33	50	67	84	101	118	135	152
28	34	52	69	87	104	122	140	157
14.4	66½	100	134	169	203	237	271	305
14.2	67½	102	137	171	206	240	275	310
14.0	68½	103½	139	174	209	244	279	314
7.3	136	206	276	346	416	486	555	625
7.15	136½	207	277	347	417	487	557	627
7.0	137	207½	277½	348	418	488	558	628
4.0	240	362	485	618	730	853	977	1100
3.9	246	372	498	625	750	877	1000	1130
3.8	252	381	511	640	770	900	1030	1160
3.7	259	392	525	658	790	923	1060	1190
3.6	266	403	540	676	812	950	1090	1220
3.5	274	414	555	696	835	977	1120	
2.0	480	725	972	1230	1475			
1.9	504	763	1020	1280				
1.8	532	805	1080					

15 waves long is practically aperiodic, and works almost equally well over a wide range of frequencies.

One of the most practical methods of feeding a long-wire antenna is to bring one end of it into the radio room for direct connection to a tuned antenna circuit which is link-coupled through a harmonic-attenuating filter to the transmitter. The antenna can be tuned effectively to resonance for operation on any harmonic by means of the tuned circuit which is connected to the end of the antenna. A ground is sometimes connected to the center of the tuned coil.

If desired, the antenna can be opened and current-fed at a point of maximum *current* by means of low-impedance ribbon line, or by a quarter-wave matching section and open line.

14-2 The V Antenna

If two long-wire antennas are built in the form of a V, it is possible to make two of the maximum lobes of one leg shoot in the same direction as two of the maximum lobes of the other leg of the V. The resulting antenna is bidirectional (two opposite directions) for the main lobes of radiation. Each side of the V can be made any odd or even number of quarter wavelengths, depending on the method of feeding the apex of the V. The complete system must be a multiple of half waves. If each leg is an even number of quarter waves long, the antenna must be voltage-fed at the apex; if an odd number of quarter waves long, current feed must be used.

By choosing the proper apex angle, figure 4 and figure 5, the lobes of radiation from the two long-wire antennas aid each other to form a bidirectional beam. Each wire by itself would have a radiation pattern similar to that for a long wire. The reaction of one upon the

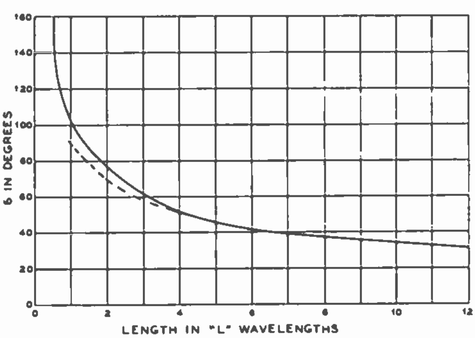


Figure 4.

INCLUDED ANGLE FOR A "V" BEAM.

Showing the included angle between the legs of a V beam for various leg lengths. For optimum alignment of the radiation lobe at the correct vertical angle with leg lengths less than three wavelengths, the optimum included angle is shown by the dashed curve.

V-ANTENNA DESIGN TABLE.

Frequency in Kilocycles	L = λ δ = 90°	L = 2λ δ = 70°	L = 4λ δ = 52°	L = 8λ δ = 39°
28000	34'8"	69'8"	140'	280'
28500	34'1"	68'6"	137'6"	275'
29000	33'6"	67'3"	135'	271'
29500	33'	66'2"	133'	266'
14050	69'	139'	279'	558'
14150	68'6"	138'	277'	555'
14250	68'2"	137'	275'	552'
14350	67'7"	136'	273'	548'
7020	138'2"	278'	558'	1120'
7100	136'8"	275'	552'	1106'
7200	134'10"	271'	545'	1090'
7280	133'4"	268'	538'	1078'

other removes two of the four main lobes, and increases the other two in such a way as to form two lobes of still greater magnitude.

The correct wire lengths and the degree of the angle δ are listed in the *V-Antenna Design Table* for various frequencies in the 10-, 20- and 40-meter amateur bands. Apex angles for all side lengths are given in figure 4. The gain of a "V" beam in terms of the side length when optimum apex angle is used is given in figure 6.

The legs of a very long V antenna are usually so arranged that the included angle is twice the angle of the major lobe from a single wire if used alone. This arrangement concentrates the radiation of each wire along the bisector of the angle, and permits part of the other lobes to cancel each other.

With legs shorter than 3 wavelengths, the best directivity and gain are obtained with a somewhat smaller angle than that determined by the lobes. Optimum directivity for a one-wave V is obtained when the angle is 90° rather than 108°, as determined by the ground pattern alone.

If very long wires are used in the V, the angle between the wires is almost unchanged when the length of the wires in wavelengths is altered. However, an error of a few degrees causes a much larger loss in directivity and gain in the case of the longer V than in the shorter one.

The vertical angle at which the wave is best transmitted or received from a horizontal V antenna depends largely upon the included angle. The sides of the V antenna should be at least a half wavelength above ground; com-

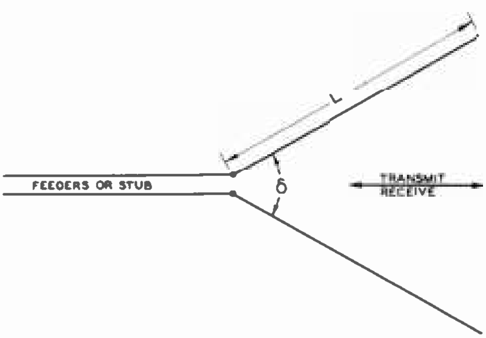


Figure 5.
TYPICAL "V" BEAM ANTENNA.

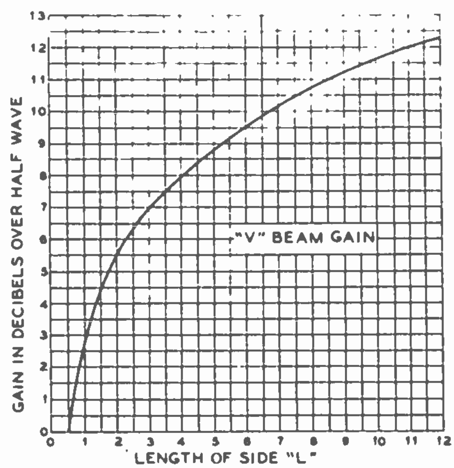


Figure 6.
DIRECTIVE GAIN OF A "V" BEAM.

This curve shows the approximate directive gain of a V beam with respect to a half-wave antenna located the same distance above ground, in terms of the side length L.

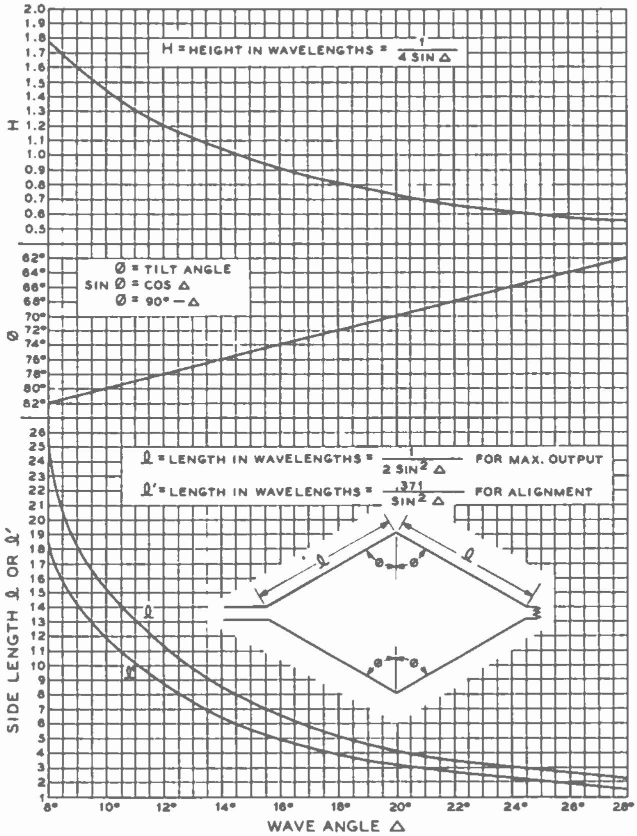


Figure 7.
RHOMBIC ANTENNA DESIGN TABLE.

Design data is given in terms of the wave angle (vertical angle of transmission and reception) of the antenna. The lengths l are for the "maximum output" design; the shorter lengths l' are for the "alignment" method which gives approximately 1.5 db less gain with a considerable reduction in the space required for the antenna. The values of side length, tilt angle, and height for a given wave angle are obtained by drawing a vertical line upward from the desired wave angle.

mercial practice dictates a height of approximately a full wavelength above ground.

14-3 The Rhombic Antenna

The terminated *rhombic* or *diamond* is probably the most effective directional antenna that is practical for amateur communication. This antenna is non-resonant, with the result that it can be used on three amateur bands, such as 10, 20, and 40 meters. When the antenna is non-resonant, i.e., properly terminated, the system is unidirectional, and the wire dimensions are not critical.

Rhombic Termination When the free end is terminated with a resistance of a value between 700 and 800 ohms the backwave is eliminated, the forward gain is increased, and the antenna can be used on several bands without changes. The terminating resistance should be capable of dissipating one-third the power output of the transmitter, and should have very little reactance. For medium or low power transmitters, the non-inductive *plaque* resistors will serve as a satisfactory termination. Several manufacturers offer special resistors suitable for terminating a rhombic antenna. The terminating device should, for technical reasons, present a small amount of inductive reactance at the point of termination.

A compromise terminating device commonly used consists of a terminated 250-foot or longer length of line, made of resistance wire which *does not have too much resistance per unit length.* If the latter qualification is not met, the reactance of the line will be excessive. A 250-foot line consisting of no. 25 nichrome wire, spaced 6 inches and terminated with 800 ohms, will serve satisfactorily. Because of the attenuation of the line, the lumped resistance

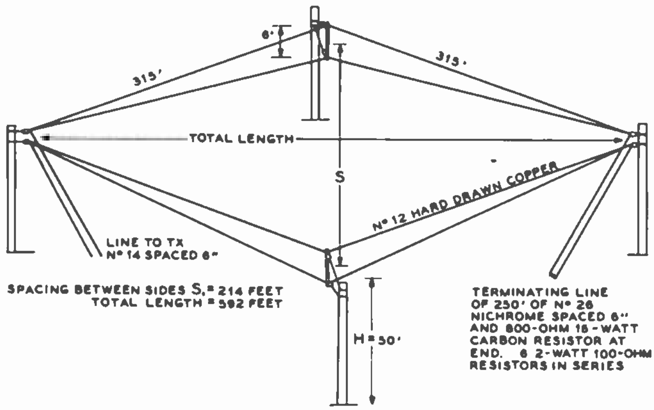


Figure 8.

TYPICAL RHOMBIC ANTENNA DESIGN.

The antenna system illustrated above may be used over the frequency range from 7 to 29 Mc. without change. The directivity of the system may be reversed by the system discussed in the text.

at the end of the line need dissipate but a few watts even when high power is used. A half-dozen 5000-ohm 2-watt carbon resistors in parallel will serve for all except very high power. The attenuating line may be folded back on itself to take up less room.

The determination of the best value of terminating resistor may be made while receiving, if the input impedance of the receiver is approximately 800 ohms. The value of resistor which gives the best directivity on reception will not give the most gain when transmitting, but there will be little difference between the two conditions.

The input resistance of the rhombic which is reflected into the transmission line that feeds it is always somewhat less than the terminating resistance, and is around 700 to 750 ohms when the terminating resistor is 800 ohms.

The antenna should be fed with a non-resonant line having a characteristic impedance of 650 to 700 ohms. The four corners of the rhombic should be at least one-half wavelength above ground for the lowest frequency of operation. For three-band operation the proper tilt angle ϕ for the center band should be observed.

The rhombic antenna transmits a horizontally-polarized wave at a relatively low angle above the horizon. The angle of radiation (wave angle) decreases as the height above ground is increased in the same manner as with a dipole antenna. The rhombic should not be tilted in any plane. In other words, the poles should all be of the same height and the plane of the antenna should be parallel with the ground.

A considerable amount of directivity is lost when the terminating resistor is left off the end and the system is operated as a resonant antenna. If it is desired to reverse the direction of the antenna it is much better practice to run transmission lines to both ends of the antenna, and then run the terminating line to the operating position. Then with the aid of two d-p-d-t switches it will be possible to connect either feeder to the antenna changeover switch and the other feeder to the terminating line, thus reversing the direction of the array and maintaining the same termination for either direction of operation.

Figure 7 gives curves for optimum-design rhombic antennas by both the maximum-output method and the alignment method. The alignment method is about 1.5 db down from the maximum output method but requires only about 0.74 as much leg length. The height and tilt angle is the same in either case. Figure 8 gives construction data for a recommended rhombic antenna for the 7.0 through 29.7 Mc. bands. This antenna will give about 11 db gain in the 14.0-Mc. band. The approximate gain of a rhombic antenna over a dipole, both above normal soil, is given in figure 9.

14-4 Stacked-Dipole Arrays

The characteristics of a half-wave dipole already have been described. When another

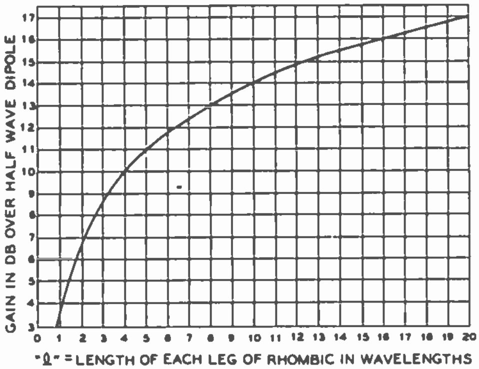


Figure 9.

RHOMBIC ANTENNA GAIN.

Showing the theoretical gain of a rhombic antenna, in terms of the side length, over a half-wave antenna mounted at the same height above the same type of soil.

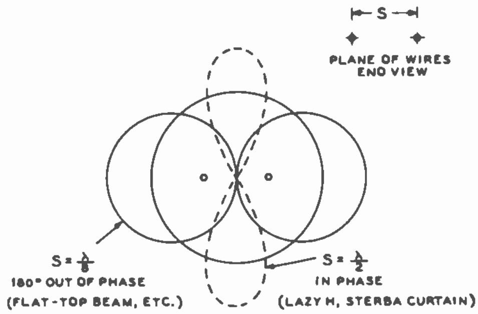


Figure 10.

RADIATION PATTERNS OF A PAIR OF DIPOLES OPERATING WITH IN-PHASE EXCITATION, AND WITH EXCITATION 180° OUT OF PHASE.

If the dipoles are oriented horizontally most of the directivity will be in the vertical plane; if they are oriented vertically most of the directivity will be in the horizontal plane.

dipole is placed in the vicinity and excited either directly or parasitically, the resultant radiation pattern will depend upon the spacing and phase differential, as well as the relative magnitude of the currents. With spacings less than 0.65 wavelength, the radiation is mainly broadside to the two wires (bidirectional) when there is no phase difference, and *through* the wires (end fire) when the wires are 180° out of phase. With phase differences between 0° and 180° (45°, 90°, and 135° for instance), the pattern is unsymmetrical, the radiation being *greater in one direction* than in the opposite direction.

With spacings of more than 0.8 wavelength, more than two main lobes appear for all phasing combinations; hence, such spacings are seldom used.

With the dipoles driven so as to be in phase, the most effective spacing is between 0.5 and 0.7 wavelength. The latter provides greater gain, but minor lobes are present which do not appear at 0.5-wavelength spacing. The radiation is broadside to the plane of the wires, and the gain is slightly greater than can be obtained from two dipoles out of phase. The gain falls off rapidly for spacings less than 0.375 wavelength, and there is little point in using spacing of 0.25 wavelength or less with in-phase dipoles, except where it is desirable to increase the radiation resistance. (See *Multi-Wire Doublet.*)

When the dipoles are fed 180° *out of phase,* the directivity is through the plane of the wires, and is greatest with *close spacing,* though there is but little difference in the pattern after the spacing is made less than 0.125 wavelength. The radiation resistance becomes so low for spacings of less than 0.1 wavelength that such spacings are not practicable.

In the three foregoing examples, most of the directivity provided is in a plane at a right angle to the length of the wires, though when out of phase, the directivity is in a line *through* the wires, and when in phase, the directivity is *broadside* to them. Thus, if the wires are oriented vertically, mostly horizontal directivity will be provided. If the wires are oriented horizontally, most of the directivity obtained will be *vertical* directivity.

To increase the sharpness of the directivity in all planes that include one of the wires, additional identical elements are added *in the line of the wires,* and fed so as to be *in phase.* The familiar H array is one array utilizing both types of directivity in the manner prescribed. The two-section Kraus flat-top beam is another.

These two antennas in their various forms are directional in a horizontal plane, in addition to being low-angle radiators, and are perhaps the most practicable of the *bidirectional* stacked-dipole arrays for amateur use. More phased elements can be used to provide greater directivity in planes including one of the radiating elements. The H then becomes a Sterba-curtain array.

For unidirectional work the most practicable stacked-dipole arrays for amateur-band use are parasitically-excited systems using relatively close spacing between the reflectors and the directors. Antennas of this type are described in detail in Chapter Sixteen. The next most practicable unidirectional array is an H or a Sterba curtain with a similar system placed approximately one-quarter wave behind. The added array may be directly fed as shown in figure 14, or it may be parasitically excited. The use of a reflector system in conjunction with any type of stacked-dipole broadside array will increase the gain by 3 db.

Colinear Arrays The simple colinear antenna array is a very effective radiating system for the 3.5-Mc. and 7.0-Mc. bands, but its use is not recommended on higher frequencies since such arrays do not possess any vertical directivity. The elevation radiation pattern for such an array is essentially the same as for a half-wave dipole. This consideration applies whether the elements are of normal length or are extended.

The colinear antenna consists of two or more radiating sections from 0.5 to 0.65 wavelengths long, with the current in phase in each section. The necessary phase reversal between sections is obtained through the use of resonant tuning stubs as illustrated in figure 11. The gain of a colinear array using half-wave elements in decibels is approximately equal to the number of elements in the array. The exact figures are as follows:

Number of Elements	2	3	4	5	6
Gain in Decibels	1.8	3.3	4.5	5.3	6.2

As additional in-phase colinear elements are added to a doublet, the radiation resistance goes up much faster than when additional half waves are added out of phase (harmonic operated antenna).

For a colinear array of from 2 to 6 elements, the terminal radiation resistance in ohms at any current loop is approximately 100 times the number of elements.

It should be borne in mind that the *gain* from a colinear antenna depends upon the *sharpness* of the horizontal directivity since no vertical directivity is provided. An array with several colinear elements will give considerable gain, but will have a sharp horizontal radiation pattern.

Double Extended Zepp The gain of a conventional two-element Franklin colinear antenna can be increased to a value approaching that obtained from a three-element Franklin, simply by making the two radiating elements 230° long instead of 180° long. The phasing stub is shortened correspondingly to maintain the whole array in resonance. Thus, instead of having 0.5-wavelength elements and 0.25-wavelength stub, the elements are made 0.64 wavelength long and the stub is approximately 0.18 wavelength long.

The correct radiator dimensions for a 230° double zepp can be obtained from the *Colinear Antenna Design Chart* simply by multiplying the L_1 values by 1.29. The length for L_3 must be determined experimentally for best results. It will be between 0.15 and 0.2 wavelength.

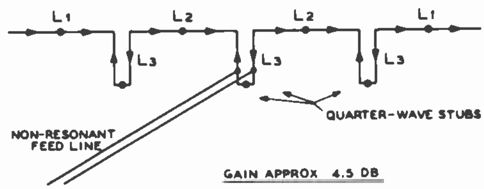


Figure 11.
THE FRANKLIN OR COLINEAR ANTENNA ARRAY.

An antenna of this type, regardless of the number of elements, attains all of its directivity through sharpening of the horizontal or azimuth radiation pattern; no vertical directivity is provided. Hence a long antenna of this type has an extremely sharp azimuth pattern, but no vertical directivity.

COLINEAR ANTENNA DESIGN CHART

FREQUENCY IN MC.	L_1	L_2	L_3
14.4	33'4"	34'3"	17'1"
14.2	33'8"	34'7"	17'3"
14.0	34'1"	35'	17'6"
7.3	65'10"	67'6"	33'9"
7.15	67'	68'8"	34'4"
7.0	68'5"	70'2"	35'1"
4.0	120'	123'	61'6"
3.9	123'	126'	63'
3.6	133'	136'5"	68'2"

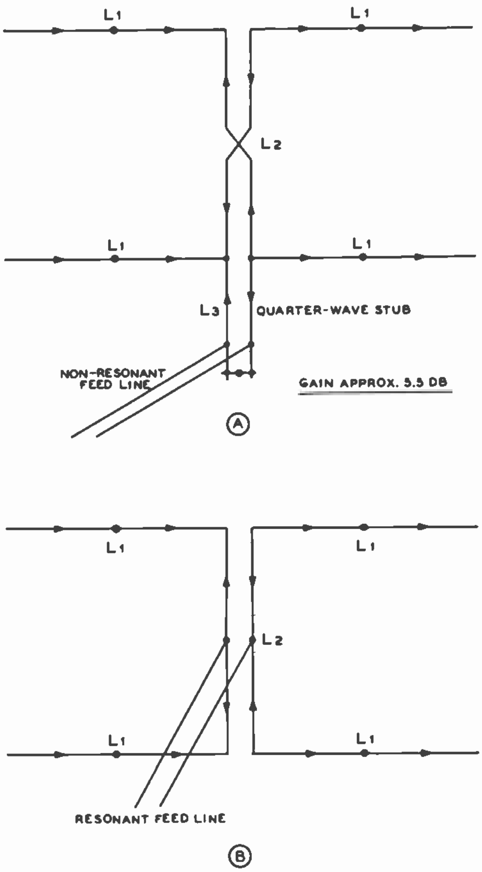


Figure 12.
THE "LAZY H" ANTENNA SYSTEM.
Stacking the colinear pairs gives both horizontal and vertical directivity. As shown the array will give about 5.5 db gain. Note that the array may be fed either at the center of the phasing section or at the bottom; if fed at the bottom the phasing section must be twisted through 180°.

The *vertical* directivity of a colinear antenna having 230° elements is the same as for one having 180° elements. There is little advantage in using extended sections when the total length of the array is to be greater than about 1.5 wavelength overall since the gain of a colinear antenna is proportional to the overall length, whether the individual radiating elements are ¼ wave, ½ wave or ¾ wave in length.

14-5 Broadside Arrays

Colinear elements may be stacked above or below another string of colinear elements to produce what is commonly called a *broadside* array. Such an array, when horizontal elements are used, possesses vertical directivity in proportion to the number of broadsided (vertically stacked) sections which have been used. Since broadside arrays do have good vertical directivity their use is recommended on the 14-Mc. band and on those higher in frequency. One of the most popular of simple broadside arrays is the "Lazy H" array of figure 12. Horizontal colinear elements stacked two above two make up this antenna system which is highly recommended for work on frequencies above perhaps 14-Mc. when moderate gain without too much directivity is desired. It has high radiation resistance and a gain of approximately 5.5 db. The high radiation resistance results in low voltages and a broad resonance curve, which permits use of inexpensive insulators and enables the array to be used over a fairly wide range in frequency. For dimensions, see the stacked dipole design table.

The Sterba Curtain Vertical stacking may be applied to strings of colinear elements longer than two half waves. In such arrays, the end quarter wave of each string of radiators usually is bent in to meet a similar bent quarter wave from the opposite end radiator. This provides better balance and better coupling between the upper and lower elements when the array is current-fed. Arrays of this type are shown in figure 13, and are commonly known as Sterba curtains.

STACKED DIPOLE DESIGN TABLE

FREQUENCY IN MC.	L_1	L_2	L_3
7.0	68'2"	70'	35'
7.3	65'10"	67'6"	33'9"
14.0	34'1"	35'	17'6"
14.2	33'8"	34'7"	17'3"
14.4	33'4"	34'2"	17'
21.0	22'9"	23'3"	11'8"
21.5	22'3"	22'9"	11'5"
27.3	17'7"	17'10"	8'11"
28.0	17'	17'7"	8'9"
29.0	16'6"	17'	8'6"
50.0	9'7"	9'10"	4'11"
52.0	9'3"	9'5"	4'8"
54.0	8'10"	9'1"	4'6"
144.0	39.8"	40.5"	20.3"
146.0	39"	40"	20"
148.0	38.4"	39.5"	19.8"

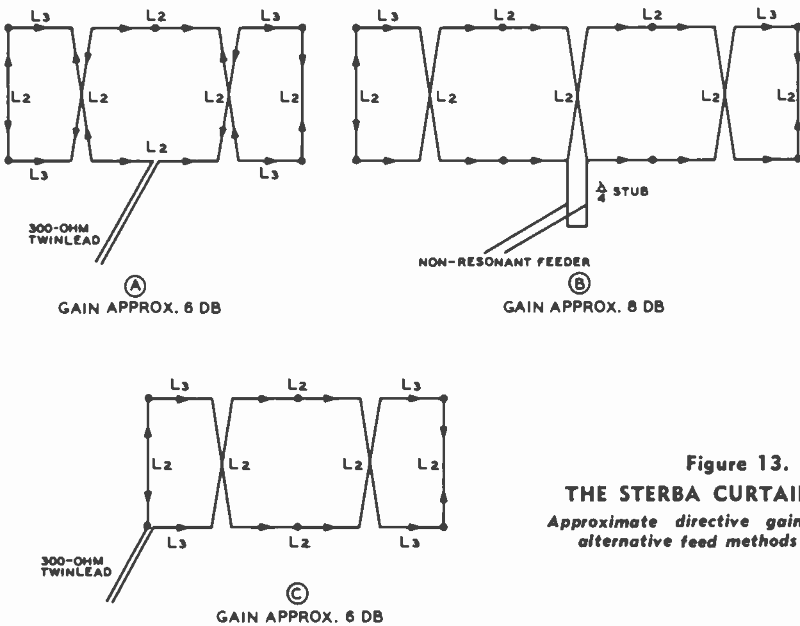


Figure 13.
THE STERBA CURTAIN ARRAY.
Approximate directive gains along with alternative feed methods are shown.

Correct length for the elements and stubs can be determined for any stacked dipole array from the *Stacked-Dipole Design Table.*

In the sketches of figure 13 the arrowheads represent the direction of current flow at any given instant. The dots on the radiators represent points of maximum current. All arrows should point in the same direction in each portion of the radiating sections of an antenna in order to provide a field in phase for broadside radiation. This condition is satisfied for the arrays illustrated in figure 13. Figures 13A and 13C show alternative methods of feeding a short Sterba curtain, while an alternative method of feed is shown in the higher gain antenna of figure 13B.

In the case of each of the arrays of figure 13, and also the "Lazy H" of figure 12, the array may be made unidirectional and the gain increased by 3 db if an exactly similar array is constructed and placed approximately ¼ wave behind the driven array. A screen or mesh of wires slightly greater in area than the antenna array may be used instead of an additional array as a reflector to obtain a unidirectional system. The spacing between the reflecting wires may vary from 0.05 to 0.1 wavelength with the spacing between the reflecting wires the smallest directly behind the driven elements. The wires in the untuned reflecting system should be parallel to the radiating elements of the array, and the spacing of the complete reflector system should be approximately 0.2 to 0.25 wavelength behind the driven elements.

On frequencies below perhaps 100 Mc. it normally will be impracticable to use a wire-screen reflector behind an antenna array such as a Sterba curtain or a "Lazy H." Parasitic elements may be used as reflectors or directors, but parasitic elements have the disadvantage that their operation is selective with respect to relatively small changes in frequency. Nevertheless, parasitic reflectors for such arrays are quite widely used. One quite satisfactory arrangement for adding 3 db to the gain of a "Lazy H" while at the same time making the array unidirectional is illustrated in figure 14. Four driven reflectors are used in conjunction with the four driven elements. The complete system exhibits about 10 db gain, has an excellent front-to-back ratio, and is quite non-selective as to frequency.

14-6 End-Fire Directivity

By spacing two half-wave dipoles, or colinear arrays, at a distance of from 0.1 to 0.25 wave-

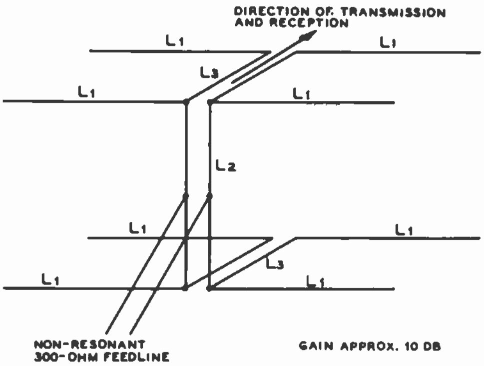


Figure 14.
THE UNIDIRECTIONAL "LAZY H" ARRAY.
Through the use of an additional set of elements directly fed from the main elements, the lazy H may be made unidirectional and will provide a directive gain of about 10 db in the favored direction.

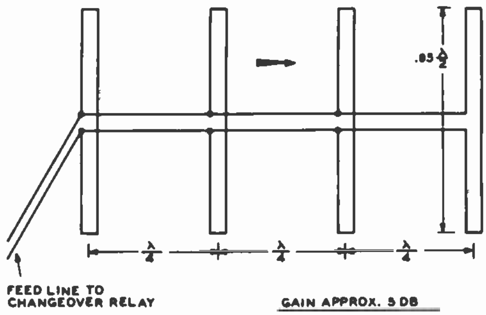


Figure 15.
A UNIDIRECTIONAL END-FIRE ARRAY WITH FOLDED ELEMENTS.
Such an array may conveniently be made rotatable, and will show substantially no change in its radiation pattern over a moderate frequency range.

length and driving the two 180° out of phase, directivity is obtained *through the two wires* at right angles to them. Hence, this type of bidirectional array is called *end fire.* A better idea of end-fire directivity can be obtained by referring to figure 10.

Remember that *end-fire* refers to the radiation with *respect to the two wires* in the array rather than with respect to the array as a whole.

The vertical directivity of an end-fire bidirectional array which is oriented horizontally can be increased by placing a similar end-fire array a half wave below it, and excited in the same phase. Such an array is a combination broadside and end-fire affair.

Unidirectional End-Fire Arrays A simple unidirectional end-fire array is illustrated in figure 15. If such an array is made two wavelengths long its gain will be approximately 8.5 db, if it is made three wavelengths long its gain will be about 10 db. Such arrays are convenient when it is desired to construct a high-gain array to radiate in a line between two poles. End-fire arrays of this type have one characteristic which is similar to the rhombic; they are effective in concentrating radiation both in the elevation and azimuth planes. Hence such arrays are good low-angle radiators.

Kraus Flat-Top Beam A very effective bidirectional end-fire array is the Kraus or 8JK *Flat-Top Beam.* Essentially, this antenna consists of two close-spaced dipoles or colinear arrays. Because of the close spacing, it is possible to obtain the proper phase relationships in multi-section flat tops by crossing the wires at the voltage loops, rather than by resorting to phasing stubs. This greatly simplifies the array. (See figure 16.) Any number of sections may be used, though the one- and two-section arrangements are the most popular. Little extra gain is obtained by using more than four sections, and trouble from phase shift may appear.

A center-fed single-section flat-top beam cut according to the table, can be used quite successfully on its second harmonic, the pattern being similar except that it is a little sharper. The single-section array can also be used on its fourth harmonic with some success, though there then will be four cloverleaf lobes, much the same as with a full-wave antenna.

If a flat-top beam is to be used on more than one band, tuned feeders are necessary.

The radiation resistance of a flat-top beam is rather low, especially when only one section is used. This means that the voltage will be high at the voltage loops. For this reason,

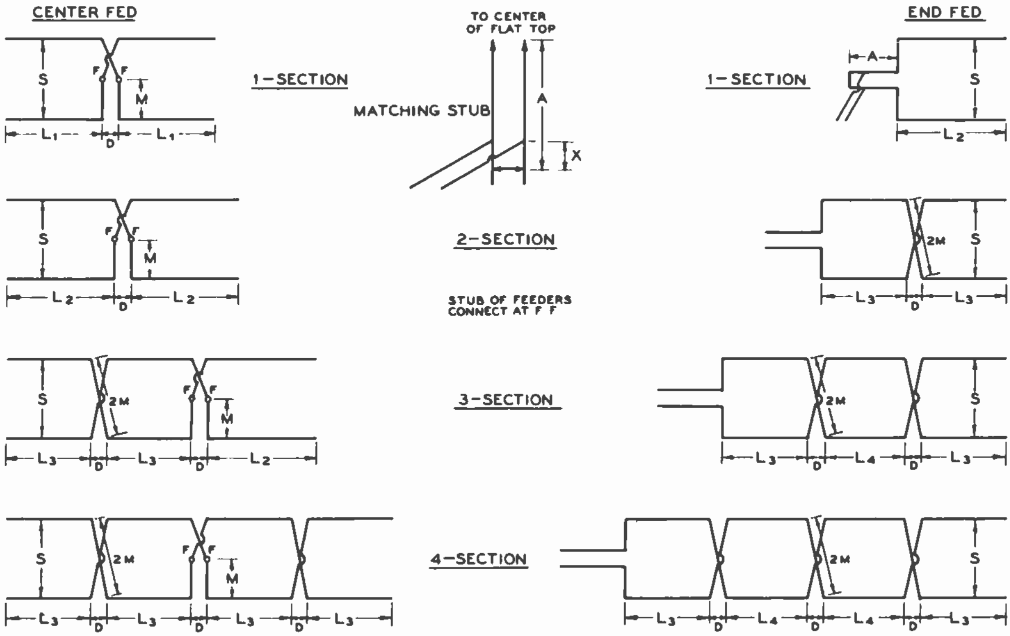


Figure 16.

FLAT-TOP BEAM (8JK ARRAY) DESIGN DATA.

FREQUENCY	Spacing	S	L_1	L_2	L_3	L_4	M	D	A (¼) approx.	A (½) approx.	A (¾) approx.	X approx.
7.0-7.2 Mc.	λ/8	17′4″	34′	60′	52′8″	44′	8′10″	4′	26′	60′	96′	4′
7.2-7.3	λ/8	17′0″	33′6″	59′	51′8″	43′1″	8′8″	4′	26′	59′	94′	4′
14.0-14.4	λ/8	8′8″	17′	30′	26′4″	22′	4′5″	2′	13′	30′	48′	2′
14.0-14.4	.15λ	10′5″	17′	30′	25′3″	20′	5′4″	2′	12′	29′	47′	2′
14.0-14.4	.20λ	13′11″	17′	30′	22′10″	.....	7′2″	2′	10′	27′	45′	3′
14.0-14.4	λ/4	17′4″	17′	30′	20′8″	.....	8′10″	2′	8′	25′	43′	4′
28.0-29.0	.15λ	5′2″	8′6″	15′	12′7″	10′	2′8″	1′6″	7′	15′	24′	1′
28.0-29.0	λ/4	8′8″	8′6″	15′	10′4″	.....	4′5″	1′6″	5′	13′	22′	2′
29.0-30.0	.15λ	5′0″	8′3″	14′6″	12′2″	9′8″	2′7″	1′6″	7′	15′	23′	1′
29.0-30.0	λ/4	8′4″	8′3″	14′6″	10′0″	.....	4′4″	1′6″	5′	13′	21′	2′

Dimension chart for flat-top beam antennas. The meanings of the symbols are as follows:

L_1, L_2 L_3 and L_4, the lengths of the sides of the flat-top sections as shown in Figure 37. L_1 is length of the sides of single-section center-fed, L_2 single-section end-fed and 2-section center-fed, L_3 4-section center-fed and end-sections of 4-section end-fed, and L_4 middle sections of 4-section end-fed.

S, the spacing between the flat-top wires.

M, the wire length from the outside to the center of each cross-over.

D, the spacing lengthwise between sections.

A (¼), the approximate length for a quarter-wave stub.

A (½), the approximate length for a half-wave stub.

A (¾), the approximate length for a three-quarter wave stub.

X, the approximate distance above the shorting wire of the stub for the connection of a 600-ohm line. This distance, as given in the table, is approximately correct only for 2-section flat-tops.

For single-section types it will be smaller and for 3- and 4-section types it will be larger.

The lengths given for a half-wave stub are applicable only to single-section center-fed flat-tops. To be certain of sufficient stub length, it is advisable to make the stub a foot or so longer than shown in the table, especially with the end-fed types. The lengths, A, are measured from the point where the stub connects to the flat-top.

Both the center and end-fed types may be used horizontally. However, where a vertical antenna is desired, the flat-tops can be turned on end. In this case, the end-fed types may be more convenient, feeding from the lower end.

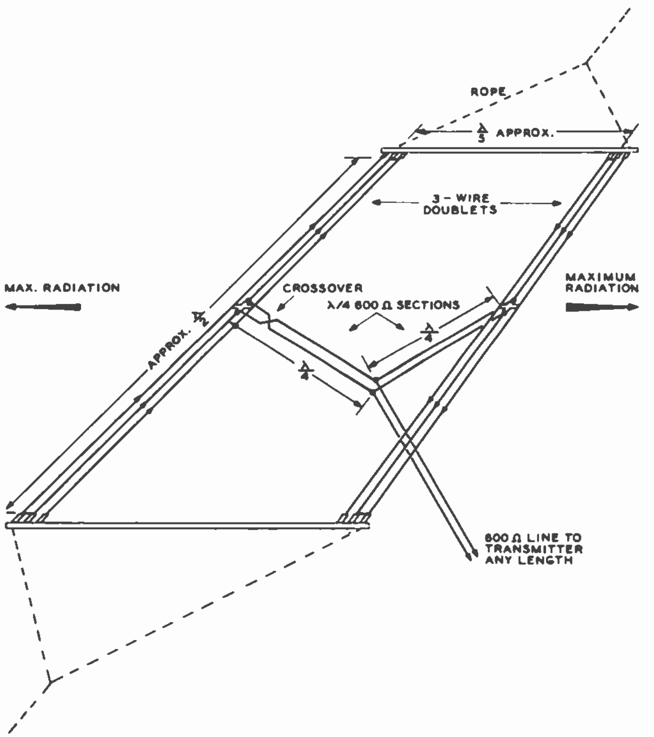


Figure 17.
THE "TWIN-THREE FLAT-TOP BEAM."

Through the use of three-wire radiating elements the feed-point impedance of the driven elements is increased to the point where the arrays may be fed through 600-ohm matching transformers directly from an open-wire line. The gain of such an antenna is slightly over 4 db. If desired the three wires which make up each element may be paralleled at each end and a single insulator used as support at the ends. Such an array is less subject to detuning in wet weather than the conventional flat-top beam.

especially good insulators should be used for best results in wet weather.

The exact lengths for the radiating elements are not especially critical, because slight deviations from the correct lengths can be compensated in the stub or tuned feeders. Proper stub adjustment is covered in Chapter Thirteen. Suitable radiator lengths and approximate stub dimensions are given in the accompanying design table.

Figure 16 shows *top views* of eight types of flat-top beam antennas. The dimensions for using these antennas on different bands are given in the design table. The 7- and 28-Mc. bands are divided into two parts, but the dimensions for either the low- or high-frequency ends of these bands will be satisfactory for use over the entire band.

In any case, the antennas are tuned to the frequency used, by adjusting the shorting wire on the stub, or tuning the feeders, if no stub is used. The data in the table may be extended to other bands or frequencies by applying the proper factor. Thus, for 50 to 52 Mc. operation, the values for 28 to 29 Mc. are divided by 1.8.

All of the antennas have a bidirectional horizontal pattern on their fundamental frequency. The maximum signal is broadside to the flat top. The single-section type has this pattern on both its fundamental frequency and second harmonic. The other types have four main lobes of radiation on the second and higher harmonics. The nominal gains of the different types over a half-wave comparison antenna are as follows: single-section, 4 db; two-section, 6 db; three-section, 7 db; four-section, 8 db.

The maximum spacings given make the beams less critical in their adjustments. Up to one-quarter wave spacing may be used on the fundamental for the one-section types and also the two-section center-fed, but it is not desirable to use more than 0.15 wavelength spacing for the other types.

Although the center-fed type of flat-top generally is to be preferred because of its symmetry, the end-fed type often is convenient or desirable. For example, when a flat-top beam is used vertically, feeding from the lower end is in most cases more convenient.

If a multisection flat-top array is end-fed instead of center-fed, and tuned feeders are used, stations off the ends of the array can be worked by tying the feeders together and working the whole affair, feeders and all, as a long-wire harmonic antenna. A single-pole double-throw switch can be used for changing the feeders and directivity.

CHAPTER FIFTEEN

V-H-F and U-H-F Antennas

The *very-high-frequency* or *v-h-f* frequency range is defined as that range falling between 30 and 300 Mc. The *ultra-high-frequency* or *u-h-f* range is defined as falling between 300 and 3000 Mc. Hence this chapter will be devoted to the design and construction of antenna systems for operation on the amateur 50-Mc., 144-Mc., 235-Mc., and 420-Mc. bands. Although the basic principles of antenna operation are the same for all frequencies, the shorter physical length of a wave in this frequency range and the differing modes of signal propagation make it possible and expedient to use antenna systems different in design from those used on the range from 3 to 30 Mc.

15-1 Antenna Requirements

Any type of antenna system usable on the lower frequencies *may* be used in the v-h-f and u-h-f bands. In fact, simple non-directive half-wave or quarter-wave vertical antennas are very popular for general transmission and reception from all directions, especially for short-range work. But for serious v-h-f or u-h-f work the use of some sort of directional antenna array is a necessity. In the first place, when the transmitter power is concentrated into a narrow beam the *apparent* transmitter power at the receiving station is increased many times. A "billboard" array or a Sterba curtain having a gain of 16 db will make a 25-watt transmitter sound like a kilowatt at the other station. Even a much simpler and smaller three- or four-element parasitic array having a gain of 7 to 10 db will produce a marked improvement in the received signal at the other station.

However, as all v-h-f and u-h-f workers know, the most important contribution of a high-gain antenna array is in reception. If a remote station cannot be heard it obviously is impossible to make contact. The limiting factor in v-h-f and u-h-f reception is in almost every case the noise generated within the receiver itself. Atmospheric noise is almost non-existent and ignition interference can almost invariably be reduced to a satisfactory level through the use of one of the effective noise limiters described in Chapter Six. Even with a grounded-grid or neutralized triode first stage in the receiver the noise contribution of the first tuned circuit in the receiver will be relatively large. Hence it is desirable to use an antenna system which will deliver the greatest signal voltage to the first tuned circuit for a given field strength at the receiving location.

Since the field intensity being produced at the receiving location by a remote transmitting station may be assumed to be constant, the receiving antenna which intercepts the greatest amount of wave front, assuming that the polarization and directivity of the receiving antenna is proper, will be the antenna which gives the best received signal-to-noise ratio. Hence an antenna which has two square wavelengths effective area will pick up twice as much signal power as one which has one square wavelength area, assuming the same general type of antenna and that both are directed at the station being received. Many instances have been reported where a frequency band sounded completely dead with a simple dipole receiving antenna but when the receiver was switched to a three-element or larger array a considerable amount of activity from 80 to 160 miles distant was heard.

Angle of Radiation The useful portion of the signal in the v-h-f and u-h-f range for short or medium distance communication is that which is radiated at a very low angle with respect to the surface of the earth; essentially it is that signal which is radiated parallel to the surface of the earth. A vertical antenna transmits a *portion* of its radiation at a very low angle and is effective for this reason; its radiation is not necessarily effective simply because it is vertically polarized. A simple horizontal dipole radiates very little low-angle energy and hence is not a satisfactory v-h-f or u-h-f radiator. Directive arrays which concentrate a major portion of the radiated signal at a low radiation angle (such as those described in the previous chapter) will prove to be effective radiators whether their signal is horizontally or vertically polarized.

In any event the radiating system for v-h-f and u-h-f work should be as high and in the clear as possible. Increasing the height of the antenna system will produce a very marked improvement in the number and strength of the signals heard, regardless of the actual type of antenna used.

Transmission Lines Transmission lines to v-h-f and u-h-f antenna systems may be either of the parallel-conductor or coaxial conductor type. Coaxial line is recommended for short runs and closely spaced open-wire line for longer runs. Wave guides may be used under certain conditions for frequencies greater than perhaps 1500 Mc. but their dimensions become excessively great for frequencies much below this value. Non-resonant transmission lines will be found to be considerably more efficient on these frequencies than those of the resonant type. In any event it is wise to use the very minimum length of transmission line possible since transmission line losses at frequencies above about 100 Mc. mount very rapidly.

Open lines should preferably be spaced closer than is common for longer wavelengths, as 6 inches is an appreciable fraction of a wavelength at 2 meters. Radiation from the line will be greatly reduced if 1-inch or 1½-inch spacing is used, rather than the more common 6-inch spacing.

Changeover Antenna It is strongly recommended that the same antenna be used for transmitting and receiving in the v-h-f and u-h-f range. An ever-present problem in this connection, however, is the antenna changeover relay. Reflections at the antenna changeover relay become of increasing importance as the frequency of transmission is increased. When coaxial cable is used as the antenna transmission line, satisfactory coaxial antenna changeover relays with low reflection can be used. One type manufactured by Advance Electric & Relay Co., Los Angeles 26, Calif., will give a satisfactorily low value of reflection.

An alternative system which will give very low reflection from the changeover system is shown in figure 1. This arrangement is an adaptation of the "TR" system used in radar work. Figure 1A shows the system used with an open-wire line. When the relays are not energized the short on the line to the transmitter one-quarter wave from the junction point appears as an open circuit at the point of junction so that all the received energy passes to the receiver. The reverse condition takes place when both relays are energized for transmission and all the transmitter energy passes to the antenna. The neon tube across the receiver input terminals is merely a protective measure in case the receiver relay fails to operate or has dirty contacts. A similar arrangement for use with coaxial transmission line is illustrated in figure 1B. In this case, since the velocity factor for polyethylene-filled coaxial cable is approximately 0.67 or 2/3, the actual physical length of the quarter-wave sections of line should be 2/3 of a quarter wave so that the electrical length will be one-quarter wave.

Effect of Feed System on Radiation Angle A vertical radiator for general coverage u-h-f use should be made either ¼ or ½ wave-

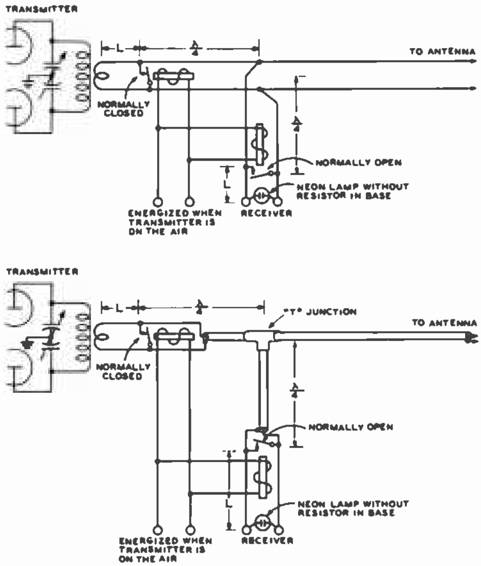


Figure 1.

ANTENNA CHANGEOVER SYSTEM FOR V.H.F.

Two quarter-wave sections of transmission line are used in conjunction with a pair of relays to accomplish antenna changeover between the transmitter and the receiver. Discussion of the circuit is given in the accompanying text.

length long. Longer vertical antennas do not have their maximum radiation at right angles to the line of the radiator (unless co-phased), and, therefore, are not practicable for use where greatest possible radiation parallel to the earth is desired.

Unfortunately, a feed system which is not perfectly balanced and does some radiating, not only robs the antenna itself of that much power, but *distorts the radiation pattern of the antenna.* As a result, the pattern of a vertical radiator may be so altered that the radiation is bent upwards slightly, and the amount of power leaving the antenna parallel to the earth is greatly reduced. A vertical half-wave radiator fed at the bottom by a quarter-wave stub is a good example of this; the slight radiation from the matching section decreases the power radiated parallel to the earth by nearly 10 db.

The only cure is a feed system which does not disturb the radiation pattern of the antenna itself. This means that if a 2-wire line is used, the current and voltages must be exactly the same (though 180° out of phase) at any point on the feed line. It means that if a concentric feed line is used, there should be no current flowing on the outside of the outer conductor.

Means for keeping the feed line out of strong fields where it connects to the radiator are discussed later in the chapter on descriptions of specific antenna systems. The unwanted currents induced in the feed line will be negligible when these precautions are taken.

Radiator Cross Section The statement was made in Chapter Thirteen that there is no point in using copper tubing for an antenna (on the medium frequencies). The reason is that considerable tubing would be required, and the cross section still would not be a sufficiently large fraction of a wavelength to improve the antenna bandwidth characteristics. At very high and ultra high frequencies, however, the radiator length is so short that the expense of large diameter conductor is relatively small, even though copper pipe of 1 inch cross section is used. With such conductors, the antenna will tune much more broadly, and often a broad resonance characteristic is desirable. This is particularly true when an antenna or array is to be used over an entire amateur band.

It should be kept in mind that with such large cross section radiators, the resonant length of the radiator will be somewhat shorter, being only slightly greater than 0.90 of a half wavelength for a dipole when heavy copper pipe is used above 100 Mc.

Insulation The matter of insulation is of prime importance at very high frequencies. Many insulators that have very low losses as high as 30 Mc. show up rather poorly at frequencies above 100 Mc. Even the low loss ceramics are none too good where the r-f voltage is high. One of the best and most practical insulators for use at this frequency is polystyrene. It has one disadvantage, however, in that it is subject to fracture and to deformation in the presence of heat.

It is common practice to design v-h-f and u-h-f antenna systems so that the various radiators are supported only at points of relatively low voltage; the best insulation, obviously, is air. The voltages on properly operated *untuned* feed lines are not high, and the question of insulation is not quite so important, though insulation still should be of good grade.

Antenna Polarization Commercial broadcasting in the U.S.A. for both FM and television in the v-h-f range has been standardized on horizontal polarization. One of the main reasons for this standardization is the fact that ignition interference is reduced through the use of a vertically polarized receiving antenna. Amateur practice, however, is divided between horizontal and vertical polarization in the v-h-f and u-h-f range. Mobile stations are invariably vertically polarized due to the physical limitations imposed by the automobile antenna installation. Most of the stations doing intermittent or occasional work on these frequencies use a simple ground-plane vertical antenna for both transmission and reception. However, those stations doing serious work and striving for maximum-range contacts on the 50-Mc. and 144-Mc. bands almost invariably use horizontal polarization.

Experience has shown that there is a great attenuation in signal strength when using crossed polarization (transmitting antenna with one polarization and receiving antenna with the other) for all normal ground-wave contacts on these bands. When contacts are being made through sporadic-*E* reflection, however, the use of crossed polarization seems to make no discernible difference in signal strength. So the operator of a station doing v-h-f work (particularly on the 50-Mc. band) is faced with a problem: If contacts are to be made with all stations doing work on the same band, provision must be made for operation on both horizontal and vertical polarization. This problem has been solved in many cases through the construction of an antenna array that may be revolved in the plane of polarization in addition to being capable of rotation in the azimuth plane. Several antennas of this type are described in Chapter Sixteen.

An alternate solution to the problem which involves less mechanical construction is simply to install a good ground-plane vertical antenna for all vertically-polarized work, and then to use a multi-element horizontally-polarized array for dx work.

TABLE OF WAVELENGTHS

Frequency in Mc.	¼ Wave Free Space	¼ Wave Antenna	½ Wave Free Space	½ Wave Antenna
50.0	59.1	55.5	118.1	111.0
50.5	58.5	55.0	116.9	109.9
51.0	57.9	54.4	115.9	108.8
51.5	57.4	53.9	114.7	107.8
52.0	56.8	53.4	113.5	106.7
52.5	56.3	52.8	112.5	105.7
53.0	55.7	52.4	111.5	104.7
54.0	54.7	51.4	109.5	102.8
144	20.5	19.2	41.0	38.5
145	20.4	19.1	40.8	38.3
146	20.2	18.9	40.4	38.0
147	20.0	18.8	40.0	37.6
148	19.9	18.6	39.9	37.2
235	12.6	11.8	25.2	23.6
236	12.5	11.8	25.1	23.5
237	12.5	11.7	25.0	23.5
238	12.4	11.7	24.9	23.4
239	12.4	11.6	24.8	23.3
240	12.3	11.6	24.6	23.2
420	7.05	6.63	14.1	13.25
425	6.95	6.55	13.9	13.1
430	6.88	6.48	13.8	12.95

All dimensions are in inches. Lengths have in most cases been rounded off to three significant figures. "½-Wave Free-Space" column shown above should be used with Lecher wires for frequency measurement.

15-2 Horizontally-Polarized Arrays

As has been mentioned before, antenna systems which do not concentrate radiation at the very low elevation angles are not recommended for v-h-f and u-h-f work. It is for this reason that the horizontal dipole and horizontally-disposed colinear arrays are generally unsuitable for work on these frequencies. Arrays using broadside or end-fire elements do concentrate radiation at low elevation angles and are recommended for v-h-f work. Arrays such as the lazy-H, Sterba curtain, flat-top beam, and arrays with parasitically excited elements are recommended for this work. Dimensions for the first three types of arrays may be determined from the data given in the previous chapter, and reference may be made to the *Table of Wavelengths* given in this chapter.

Arrays using vertically-stacked horizontal dipoles, such as are used by commercial television and FM stations, are capable of giving very high gain *without* a sharp horizontal radiation pattern. If sets of crossed dipoles, as shown in figure 2A, are fed 90° out of phase the resulting system is called a "turnstile" antenna. The 90° phase difference between sets of dipoles may be obtained by feeding one set of dipoles with a feed line which is one-quarter wave longer than the feed line to the other set

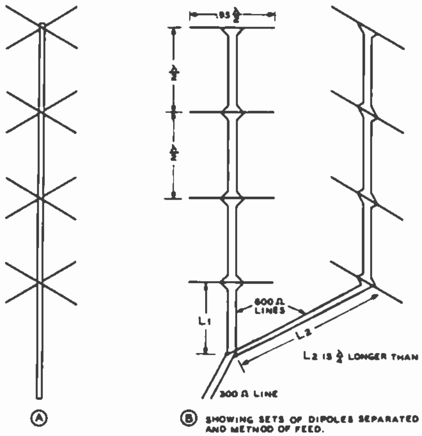


Figure 2.

THE STACKED TURNSTILE ARRAY.

An antenna system such as illustrated above will concentrate the radiated signal at the low radiation angles which are useful in the v-h-f range, but provides no horizontal directivity. Folded elements may be used, if desired, for the radiating sections.

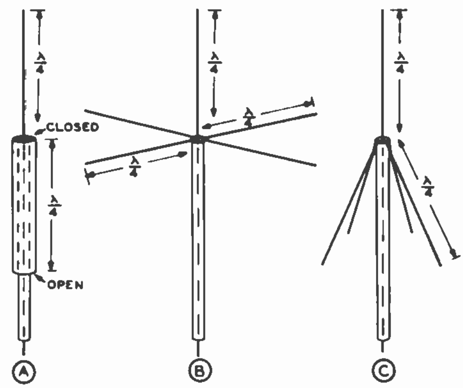


Figure 3.

THREE VERTICALLY-POLARIZED LOW-ANGLE RADIATORS.

Shown at (A) is the "sleeve" or "hypodermic" type of radiator. At (B) is shown the ground-plane vertical, and (C) shows a modification of this antenna system which increases the feed-point impedance to a value such that the system may be fed directly from a coaxial line with no standing waves on the feed line.

of dipoles. The free-space theoretical gain of an antenna such as shown is about 5 db over a half-wave dipole, but in actual practice on the v-h-f bands a considerably greater effective gain will be obtained in *all* directions simultaneously over a dipole. If the second set of four dipoles is placed one-quarter wave behind the first set and parasitically excited from the dipoles as a reflector (or as a director) approximately 10 db gain will be obtained and the horizontal pattern of the array will still be moderately broad. The majority of the gain in arrays of this type comes from concentration of substantially all radiation from the array at the useful low angles of radiation.

The array with several parasitically-excited elements is meeting with increasing favor for operation on the v-h-f's. Detailed discussion of the element lengths, method of feed, installation and tuning of such arrays is given in Chapter Sixteen.

15-3 Vertically-Polarized Antennas and Arrays

For general coverage with a single antenna, a single vertical radiator is commonly employed. A two-wire open transmission line is not suitable for use with this type antenna, and coaxial polyethylene feed line such as RG-8/U is to be recommended. Three practical methods of feeding the radiator with concentric line, with a minimum of current induced in the outside of the line, are shown in figure 3. Antenna (A) is known as the "sleeve" or "hypodermic" antenna, the lower half of the radiator being a large piece of pipe up through which the concentric feed line is run. At (B) is shown the ground-plane vertical, and at (C) a modification of this latter antenna.

The radiation resistance of the ground-plane vertical is approximately 30 ohms, which is not a standard impedance for coaxial line. To obtain a good match, the first quarter wavelength of feeder may be of 52 ohms surge impedance, and the remainder of the line of approximately 75 ohms impedance. Thus, the first quarter-wave section of line is used as a matching transformer, and a good match is obtained.

In actual practice the antenna would consist of a quarter-wave rod, mounted by means of insulators atop a pole or pipe mast. Elaborate insulation is not required, as the voltage at the lower end of the quarter-wave radiator is very low. Self-supporting rods from 0.25

to 0.28 wavelength would be extended out, as in the illustration, and connected together. As the point of connection is effectively at ground potential, no insulation is required; the horizontal rods may be bolted directly to the supporting pole or mast, even if of metal. The coaxial line should be of the low loss type especially designed for v-h-f use. The outside connects to the junction of the radials, and the inside to the bottom end of the vertical radiator.

The modification at (C) permits matching to a standard 50- or 70-ohm flexible coaxial cable without a linear transformer. If the lower rods hug the line and supporting mast rather closely, the feed-point impedance is about 70 ohms. If they are bent out to form an angle of about 30° with the support pipe the impedance is about 50 ohms.

Construction of the Ground-Plane Vertical Figure 4 is a photograph of a relatively simple ground-plane vertical antenna suitable for the v-h-f range. The mechanical construction details of the antenna are given in figure 5. An antenna of this type is moderately simple to construct and will give a good account of itself when fed at the lower end of the radiator directly by the 52-ohm RG-8/U coaxial cable. Theoretically the standing-wave ratio will be approximately 1.5-to-1 but in practice this moderate s-w-r produces no deleterious effects, even on coaxial cable.

An antenna design for a ground-plane vertical which will give a more accurate match to a 70-ohm coaxial cable is illustrated in figure 6 and sketched in figure 7. This type of ground-plane antenna is often called the *folded-unipole* antenna. The improvement in the match between the feed point on the dipole and the antenna transmission line is obtained by folding the radiator, in the same general manner as used with the folded dipole, grounding one end, and connecting the antenna transmission-line inner conductor to the ungrounded end of the radiator.

The use of a folded dipole (or unipole) where both conductors have the same diameter will result in a multiplication of the feed-point impedance by a factor of 4. Since the feed-point impedance at the lower end of a ground-plane vertical is approximately 30 ohms, the use of a folded dipole with the same conductor diameter would give a feed-point im-

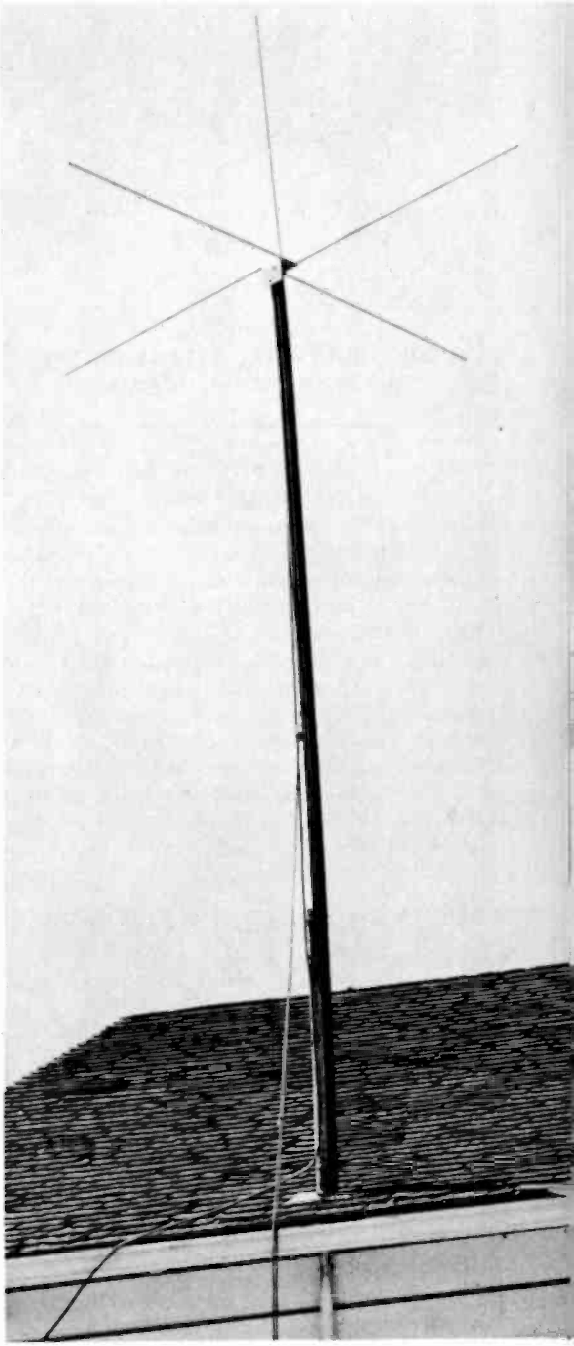


Figure 4.
50-MC. GROUND-PLANE VERTICAL ANTENNA.
Construction of this antenna system is illustrated in figure 5.

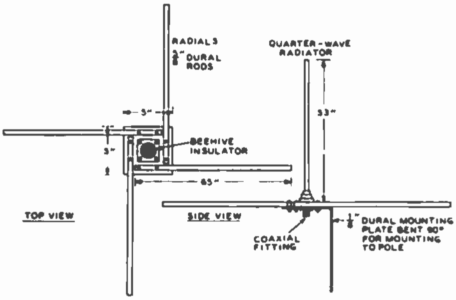


Figure 5.
CONSTRUCTIONAL DETAILS OF THE GROUND-PLANE VERTICAL.

pedance of about 120 ohms. Since standard polyethylene coaxial-cable impedances are 52 ohms and 70 ohms, we must use an impedance step up of less than four. If the diameter of the half of the radiator connected to the feed line is made larger than the diameter of the half whose lower end is grounded, the impedance multiplication will be less than four. A detailed discussion of the calculation of conductor sizes for obtaining varying impedance step up ratios is given in Chapter Sixteen. However, suffice to say here that it is impracticable to obtain the small impedance step up from 30 to 52 ohms by this method. It is better merely to tolerate the small standing wave that will be formed on the cable, or to build a quarter-wave coaxial transformer inside the support pipe of the antenna having a characteristic impedance of 38.5 ohms.

It is practicable, however, to match the 30-ohm basic impedance of the antenna to a 70-ohm coaxial cable through the use of the folded-unipole system. The diameter of the grounded half of the folded unipole should be one-quarter inch, and the diameter of the half of the unipole which goes to the inner conductor of the coaxial cable should be ⅝ inch. The center-to-center spacing of the two rods should be one to one and one-half inches. These are the dimensions used in the antenna whose photograph is shown in figure 6.

The number of radial legs used in a ground-plane antenna of either type has an important effect on the feed-point impedance and upon the radiation characteristics of the antenna system. Experiment has shown that three radials is the minimum number that should be used, and that increasing the number of radials above four adds substantially nothing to the effectiveness of the antenna and has no effect on the feed-point impedance. Experiment has shown, however, that the radials should be slightly longer than one-quarter wave for best results. A length of 0.28 wavelength has been shown to be the optimum value. This means that the radials for a 50-Mc. ground-plane vertical antenna should be 65" in length.

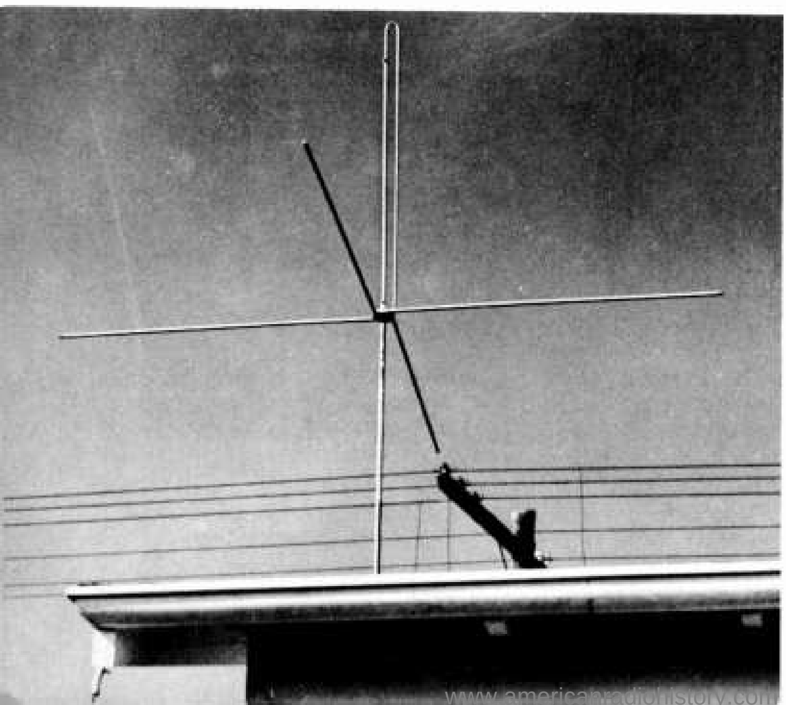


Figure 6.
THE "FOLDED UNIPOLE" GROUND-PLANE VERTICAL ANTENNA.
Constructional details of this antenna are given in figure 7.

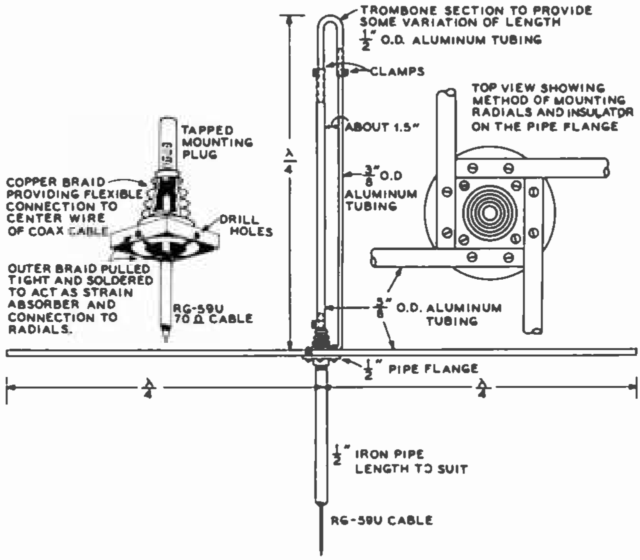


Figure 7.
CONSTRUCTIONAL DETAILS OF THE "FOLDED UNIPOLE" GROUND-PLANE VERTICAL ANTENNA.

15-4 The Discone Antenna

The Discone antenna is a vertically polarized omnidirectional radiator which has very broad band characteristics and permits a simple, rugged structure. This antenna presents a substantially uniform feed-point impedance, suitable for direct connection of a coaxial line, over a range of several octaves. Also, the vertical pattern is suitable for ground-wave work over several octaves, the gain varying only slightly over a very wide frequency range.

Commercial versions of the Discone antenna for various applications are manufactured by the Federal Telephone and Radio Corporation. A Discone type antenna for amateur work can be fabricated from inexpensive materials with ordinary hand tools.

A Discone antenna suitable for multi-band amateur work in the v-h/u-h-f range is shown schematically in figure 8. The distance D should be made approximately equal to a free-space quarter wavelength at the lowest operating frequency. The antenna then will perform well over a frequency range at least of 8 to 1. At certain frequencies within this range the vertical pattern will tend to "lift" slightly, causing a slight reduction in gain at zero angular elevation, but the reduction is very slight.

Below the frequency at which the slant height of the conical skirt is equal to a free-space quarter wavelength the standing-wave ratio starts to climb, and below a frequency

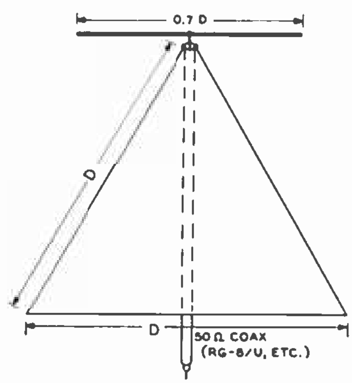


Figure 8.
THE "DISCONE" BROAD-BAND RADIATOR.

This antenna system radiates a vertically polarized wave over a very wide frequency range. The "disc" may be made of solid metal sheet, a group of radials, or wire screen; the "cone" may best be constructed by forming a sheet of thin aluminum. A single antenna may be used for operation on the 50, 144, and 220 Mc. amateur bands. The dimension D is determined by the lowest frequency to be employed, and is given in the chart of figure 9.

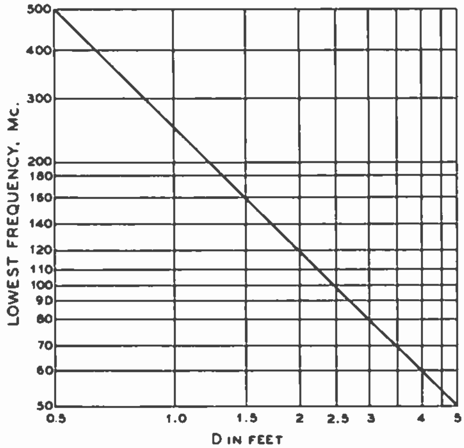
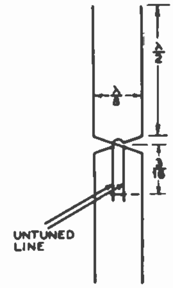


Figure 9.
DESIGN CHART FOR THE "DISCONE" ANTENNA.

Figure 10.
FLAT-TOP BEAM ORIENTED FOR VERTICAL POLARIZATION.

For design data on this antenna refer to figure 16 of the preceding chapter. The stub and feed line should be equidistant from the two lower radiating elements.



approximately 20 per cent lower than this the standing-wave ratio climbs very rapidly. This is termed the "cut off" frequency of the antenna. By making the slant height approximately equal to a free-space quarter wavelength at the lowest frequency employed (refer to chart), a VSWR of less than 1.5 will be obtained throughout the operating range of the antenna.

The Discone antenna may be considered as a cross between an electromagnetic horn and an inverted ground plane "unipole" antenna. It looks to the feed line like a properly terminated high-pass filter.

Construction Details The top disk and the conical skirt may be fabricated either from sheet metal, screen (such as "hardware cloth"), or 12 or more "spine" radials. If screen is used a supporting framework of rod or tubing will be necessary for mechanical strength except at the higher frequencies. If spines are used, they should be terminated on a stiff ring for mechanical strength except at the higher frequencies.

The top disk is supported by means of three insulating pillars fastened to the skirt. Either polystyrene or low-loss ceramic is suitable for the purpose. The apex of the conical skirt is grounded to the supporting mast and to the outer conductor of the coaxial line. The line is run down through the supporting mast. An alternative arrangement, one suitable for certain mobile applications, is to fasten the base of the skirt directly to an effective ground plane such as the top of an automobile.

15-5 Vertically-Polarized Arrays

Antenna arrays such as the flat-top beam and the lazy-H (when the latter is fed in the center instead of at one end) may be used with the elements vertically oriented to produce vertically-polarized radiation. Typical examples are shown in figures 10 and 11. Two other types of arrays, which are especially designed for vertical polarization, are shown in figure 12. It is important in the case of all these arrays that the stub and feed line be brought directly away from the antenna in a plane at right angles to the array for a distance of at least two wavelengths. If the stub or line is closer to one radiator than the other undesired currents will be induced in the feed line.

15-6 Helical Beam Antennas

Most v-h-f and u-h-f antennas are either vertically polarized or horizontally polarized (plane polarization). However, *circularly* polarized antennas have interesting characteristics which may be useful for certain applications. The installation of such an antenna can effectively solve the problem of horizontal vs. vertical polarization.

A circularly polarized wave has its energy divided equally between a vertically polarized component and a horizontally polarized com-

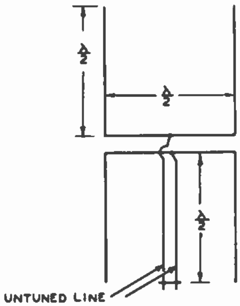


Figure 11.

H-TYPE ARRAY ARRANGED FOR VERTICAL POLARIZATION.

The half-wave matching stub feeds the center of the phasing section. The stub should be equidistant from the two lower radiators.

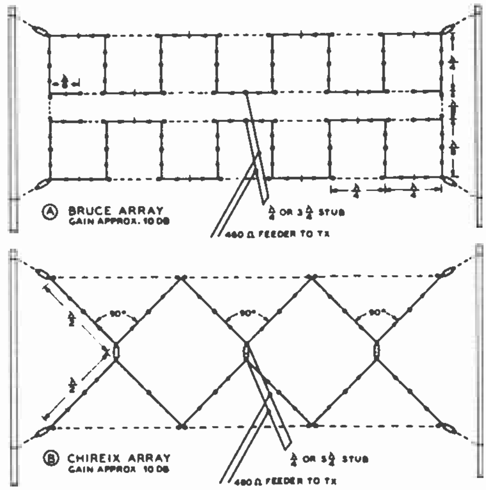


Figure 12.

TWO VERTICALLY POLARIZED HIGH-GAIN ARRAYS.

A pair of stacked Bruce arrays is illustrated at (A), and a pair of stacked Chireix arrays is shown at (B).

ponent, the two being 90 degrees out of phase. The circularly polarized wave may be either "left handed" or "right handed," depending upon whether the vertically polarized component leads or lags the horizontal component.

A circularly polarized antenna will respond to any plane polarized wave whether horizontally polarized, vertically polarized, or diagonally polarized. Also, a circular polarized wave can be received on a plane polarized antenna, regardless of the polarization of the latter. When using circularly polarized antennas at *both* ends of the circuit, however, both must be left handed or both must be right handed. This offers some interesting possibilities with regard to reduction of QRM. At the time of writing, there has been no standardization of the "twist" for general amateur work.

Perhaps the simplest antenna configuration for a directional beam antenna having circular polarization is the helical beam popularized by Dr. John Kraus, W8JK. The antenna consists simply of a helix working against a ground plane and fed with coaxial line. In the u-h-f and the upper v-h-f range the physical dimensions are sufficiently small to permit construction of a rotatable structure without much difficulty.

When the dimensions are optimized, the characteristics of the helical beam antenna are such as to qualify it as a "broad band" antenna. An optimized helical beam shows little variation in the pattern of the main lobe and a fairly uniform feed point impedance averaging approximately 125 ohms over a frequency range of as much as 1.7 to 1. The direction of "electrical twist" (right or left handed) depends upon the direction in which the helix is wound.

A six-turn helical beam is shown schematically in figure 13. The dimensions shown will give good performance over a frequency range of plus or minus 20 per cent of the design frequency. This means that the dimensions are not especially critical when the array is to be used at a single frequency or over a narrow band of frequencies, such as an amateur band. At the design frequency the beam width is about 50 degrees and the power gain about 12 db referred to a non-directional circularly polarized antenna.

The Ground Screen For the frequency range 100 to 500 Mc. a suitable ground screen can be made from "chicken wire" poultry netting of 1-inch mesh, fastened to a round or square frame of either metal or wood. The netting should be of the type that is galvanized *after* weaving. A small, sheet metal ground plate of diameter equal to approximately D/2 should be centered on the

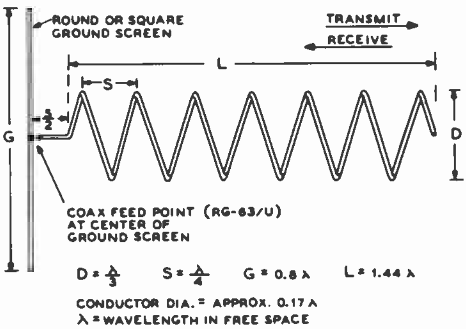


Figure 13.

THE "HELICAL BEAM" ANTENNA.

This type of directional antenna system gives excellent performance over a frequency range of 1.7 to 1.8 to 1. Its dimensions are such that it ordinarily is not practicable, however, for use as a rotatable array on frequencies below about 100 Mc. The center conductor of the feed line should pass through the ground screen for connection to the feed point. The outer conductor of the coaxial line should be grounded to the ground screen.

screen and soldered to it. Tin, galvanized iron, or sheet copper is suitable. The outer conductor of the RG-63/U (125 ohm) coax is connected to this plate, and the inner conductor contacts the helix through a hole in the center of the plate. The end of the coax should be taped with "Scotch" electrical tape to keep water out.

The Helix It should be noted that the beam proper consists of six full turns. The start of the helix is spaced a distance of S/2 from the ground screen, and the conductor goes directly from the center of the ground screen to the start of the helix.

Aluminum tubing in the "SO" (soft) grade is suitable for the helix. Alternatively, lengths of the relatively soft aluminum electrical conduit may be used. In the v-h-f range it will be necessary to support the helix on either two or four wooden longerons in order to achieve sufficient strength. The longerons should be of as small cross section as will provide sufficient rigidity, and should be given several coats of varnish. The ground plane butts against the longerons and the whole assembly is supported from the balance point if it is to be rotated.

Aluminum tubing in the larger diameters ordinarily is not readily available in lengths greater than 12 feet. In this case several lengths can be spliced by means of short telescoping sections and sheet metal screws.

The tubing is close wound on a drum and then spaced to give the specified pitch. Note that the length of one complete turn when spaced is somewhat greater than the circumference of a circle having the diameter D. However, it is useless anyhow to attempt to calculate how much to allow for the decrease in diameter when the turns are spaced because the tubing will spring to an unpredictably larger diameter when it is removed from the drum used as a winding form. The increase in diameter will depend upon the hardness, diameter, and wall thickness of the particular tubing used. It probably will be necessary to experiment with a single turn on drums of various diameters to find the diameter of winding form which will give the desired helix diameter when the turns are spaced to the proper pitch.

Broad-Band 144 to 225 Mc. Helical Beam A highly useful v-h-f helical beam which will receive signals with good gain over the complete frequency range from 144 through 225 Mc. may be constructed by using the following dimensions (180 Mc. design center):

D	22 in.
S	16½ in.
G	53 in.
Tubing o.d.	1 in.

The D and S dimensions are to the center of the tubing. These dimensions must be held rather closely, since the range from 144 through 225 Mc. represents just about the practical limit of coverage of this type of antenna system.

High-Band TV Coverage Note that an array constructed with the above dimensions will give unusually good high-band TV reception in addition to covering the 144-Mc. and 220-Mc. amateur bands and the taxi and police services.

On the 144-Mc. band the beam width is approximately 60 degrees to the half-power points, while the power gain is approximately 11 db over a non-directional circularly polarized antenna. For high-band TV coverage the gain will be 12 to 14 db, with a beam width of about 50 degrees. And on the 220-Mc. amateur band the beam width will be about 40 degrees with a power gain of approximately 15 db.

The antenna system will receive vertically polarized or horizontally polarized signals with equal gain over its entire frequency range. Conversely, it will transmit signals over the same range, which then can be received with equal strength on either horizontally polarized or vertically polarized receiving antennas. The standing-wave ratio will be very low over the complete frequency range if RG-63/U coaxial feed line is used. If the beam is arranged for rotation, care should be taken not to flex this type of transmission line on too small a diameter.

15-7 The Corner-Reflector Antenna

The corner-reflector antenna is a good directional radiator for the v-h-f and u-h-f region. The antenna may be used with the radiating element vertical, in which case the directivity is in the horizontal or azimuth plane, or the system may be used with the driven element horizontal in which case the radiation is horizontally polarized and most of the directivity is in the vertical plane. With the antenna used as a horizontally polarized radiating system the array is a very good low-angle beam array although the nose of the horizontal pattern is still quite sharp. When the radiator is oriented vertically the corner reflector operates very satisfactorily as a direction-finding antenna.

Design data for the corner-reflector antenna is given in figure 14 and in the chart *Corner-Reflector Design Data.* The planes which make up the reflecting corner may be made of solid sheets of copper or aluminum for the u-h-f bands, although spaced wires with the ends soldered together at top and bottom may be used as the reflector on the lower frequencies. Copper screen may also be used for the reflecting planes.

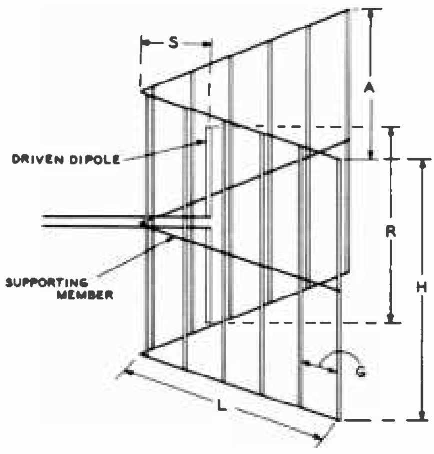


Figure 14.
CONSTRUCTION OF THE "CORNER REFLECTOR" ANTENNA.
Such an antenna is capable of giving high gain with a minimum of complexity in the radiating system. It may be used either with horizontal or vertical polarization. Design data for the antenna is given in the Corner-Reflector Design Table.

The values of spacing given in the corner-reflector chart have been chosen such that the center impedance of the driven element would be approximately 70 ohms. This means that the element may be fed directly with 70-ohm coaxial line, or a quarter-wave matching transformer such as a "Q" section may be used to provide an impedance match between the center-impedance of the element and a 460-ohm line constructed of no. 12 wire spaced 2 inches.

CORNER-REFLECTOR DESIGN DATA

Corner Angle	Freq. Band, Mc.	R	S	H	A	L	G	Feed Imped.	Approx. Gain, db
90	50	110"	82"	140"	200"	230"	18"	72	10
60	50	110"	115"	140"	230"	230"	18"	70	12
60	144	38"	40"	48"	100"	100"	5"	70	12
60	220	24.5"	25"	30"	72"	72"	3"	70	12
60	420	13"	14"	18"	36"	36"	screen	70	12

NOTE: Refer to figure 14 for construction of corner-reflector antenna.

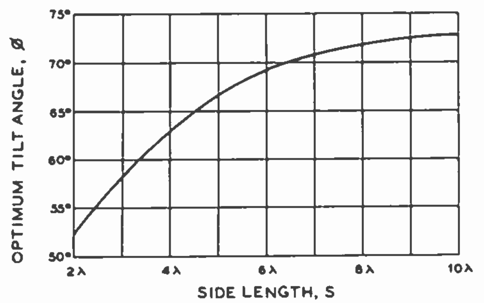


Figure 15.
V-H-F RHOMBIC ANTENNA DESIGN CHART.
The optimum tilt angle (see figure 16) for "zero-angle" radiation depends upon the length of the sides.

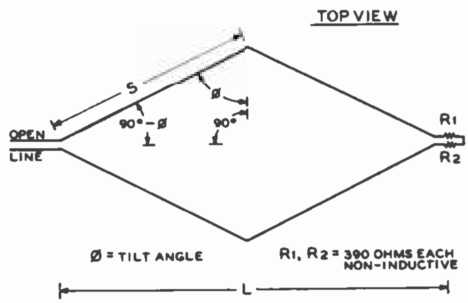


Figure 16.
V-H-F RHOMBIC ANTENNA CONSTRUCTION.

15-8 VHF Horizontal Rhombic Antenna

For v-h-f transmission and reception in a fixed direction, a horizontal rhombic permits 10 to 16 db gain with a simpler construction than does a phased dipole array, and has the further advantage of being useful over a wide frequency range.

Except at the upper end of the v-h-f range a rhombic array having a worthwhile gain is too large to be rotated. However, in locations 75 to 150 miles from a large metropolitan area a rhombic array is ideally suited for working into the city on extended (horizontally polarized) ground-wave while at the same time making an ideal antenna for TV reception.

The useful frequency range of a v-h-f rhombic array is about 2 to 1, or about plus 40% and minus 30% from the *design frequency*. This coverage is somewhat less than that of a high-frequency rhombic used for sky-wave communication. For ground-wave transmission or reception the only effective vertical angle is that of the horizon, and a frequency range greater than 2 to 1 cannot be covered with a rhombic array without an excessive change in the vertical angle of maximum radiation or response.

The dimensions of a v-h-f rhombic array are determined from the design frequency and figure 15, which shows the proper *tilt angle* (see figure 16) for a given leg length. The gain of a rhombic array increases with leg length. There is not much point in constructing a v-h-f rhombic array with legs shorter than about 4 wavelengths, and the beam width begins to become excessively sharp for leg lengths greater than about 8 wavelengths. A leg length of 6 wavelengths is a good compromise between beam width and gain.

The tilt angle given in figure 15 is based upon a "wave angle" of zero degrees. For leg lengths of 4 wavelengths or longer, it will be necessary to elongate the array a few per cent (pulling in the sides slightly) if the horizon elevation exceeds about 3 degrees.

Table I gives dimensions for two "dual purpose" rhombic arrays. One covers the 6-meter amateur band and the "low" television band. The other covers the 2-meter amateur band, the "high" television band, and the 1¼-meter amateur band. The gain is approximately 12 db over a matched half wave dipole and the beam width is about 6 degrees.

The Feed Line The recommended feed line is an open-wire line having a surge impedance between 450 and 600 ohms. With such a line the VSWR will be less than 2 to 1. A line with two-inch spacing is suitable for frequencies below 100 Mc., but one-inch spacing (such as used in the "Gonset Line" for TV installations) is recommended for higher frequencies.

Such a line can be fed directly into a television receiver with 300-ohm input, because the match at the antenna end will be sufficiently good to suppress ghosts due to "line echoes" even though there is a moderate mismatch at the receiver. If the array is to be

	6 METERS AND LOW BAND TV	2 METERS, HIGH BAND TV, AND 1¼ METERS
S (side)	90'	32'
L (length)	166' 10"	59' 4"
W (Width)	67' 4"	23' 11"

S=6 wavelenths at design frequency
Tilt angle =68°

TABLE I.

used only for receiving, 300-ohm ribbon line may be used if the feed line is not unusually long. For long line lengths an open-wire line is recommended for reception because of lower losses.

The small rhombic of Table I can be strung inside the larger rhombic without bad effects. Separate feed lines are recommended with this arrangement.

The Termination If the array is to be used only for reception, a suitable termination consists of two 390-ohm-carbon resistors in series. If 2-watt resistors are employed, this termination also is suitable for transmitter outputs of 10 watts or less. For higher powers, however, resistors having greater dissipation with negligible reactance in the upper v-h-f range are not readily available.

For powers up to several hundred watts a suitable termination consists of a "lossy" line consisting of stainless steel wire (corresponding to no. 24 or 26 B&S gauge) spaced 2 inches, which in turn is terminated by two 390-ohm 2-watt carbon resistors. The dissipative line should be at least 6 wavelengths long.

Space Tapered Legs A slight improvement in the characteristics of a rhombic array can be realized by using two wires for each leg, spaced vertically at the side apices by means of spreaders. The spacing at the side apices may be made about 0.1 wavelength.

CHAPTER SIXTEEN

Rotatable Antenna Arrays

The rotatable antenna array has become almost standard equipment for operation on the 28-Mc. and 50-Mc. bands and is very commonly used on the 14-Mc. band and on those frequencies above 144 Mc. The rotatable array offers many advantages for both military and amateur use. The directivity of the antenna types commonly employed, particularly the unidirectional arrays, offers a worthwhile reduction in interference from undesired directions. Also, the increase in the ratio of low-angle radiation plus the theoretical gain of such arrays results in a relatively large increase in both the transmitted signal and the signal intensity from a station being received.

A significant advantage of a rotatable antenna array in the case of the normal station is that a relatively small amount of space is required for erection of the antenna system. In fact, one of the best types of installations uses a single telephone pole with the rotating structure holding the antenna mounted atop the pole. To obtain results in all azimuth directions from fixed arrays comparable to the gain and directivity of a single rotatable three-element parasitic beam would require several acres of surface.

There are two normal configurations of radiating elements which, when horizontally polarized, will contribute to obtaining a low angle of radiation. These configurations are the end-fire array and the broadside array. The conventional three- or four-element rotary beam may properly be called a *unidirectional parasitic end-fire array,* and is actually a type of *yagi* array. The flat-top beam is a type of *bidirectional end-fire array.* The *broadside type* of array is also quite effective in obtaining low-angle radiation, and although widely used in FM and TV broadcasting has seen little use by amateur stations in rotatable arrays. All three of these types of arrays, and their use on rotating structures, will be described in this chapter.

16-1 Unidirectional Parasitic End-Fire Arrays (Yagi Type)

If a single parasitic element is placed on one side or the other of a driven dipole at a distance of from 0.1 to 0.25 wavelength the parasitic element can be tuned to make the array substantially unidirectional.

Two-Element Array The optimum spacing for a reflector in a two-element array is approximately 0.15 wavelength and with optimum adjustment of the length of the reflector a gain of approximately 5 db will be obtained, with a feed-point resistance of about 30 ohms.

If the parasitic element is to be used as a director the optimum spacing between it and the driven element is 0.1 wavelength. The gain will theoretically be slightly greater than with the optimum adjustment for a reflector (about 5.5 db) but the radiation resistance will be in the vicinity of 15 ohms.

In both the case of the director and the reflector in a two-element array the point of adjustment for maximum forward gain will be found to be somewhat different from that for maximum front-to-back ratio. The two adjustments are quite close together and either one may be chosen, depending upon the operating conditions desired. A sacrifice of approximately 1.0 db in forward gain is involved when the two-element array is adjusted for maximum front-to-back ratio.

The two-element array is most frequently used on the 14-Mc. band where the size of the supporting structure would become prohibitive with a larger number of elements. A representative two-element 14-Mc. array with the parasitic element operating as a director at 0.125 wavelength spacing is shown in figure 1.

Length of the Elements Due to the mutual impedance between the driven element and the parasitic element, the resonant frequency of the driven element will be changed from that value which would exist if the parasitic element were not present. With a single parasitic *director,* the driven element must be *longer* than the usual length to obtain resonance. With a *reflector,* the driven element must be *shorter* than the usual length. Appropriate dimensions for the elements are given in figure 2.

Three-Element Array The three-element array using a director, driven element, and reflector will exhibit as much as 30 db front-to-back ratio and 20 db front-to-side ratio for *low-angle radiation.* The theoretical gain is about 9 db over a dipole in free space. In actual practice, the array will often show 7 to 10 db apparent gain over a horizontal dipole placed the same height above ground (at 28 and 14 Mc.).

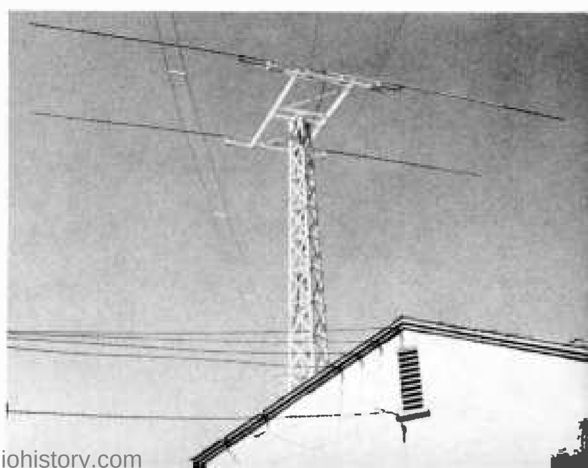
The use of more than three elements is desirable when the length of the supporting structure is such that spacings of approximately 0.2 wavelength between elements becomes possible. Four-element arrays are quite common on the 28-Mc. and 50-Mc. bands, and five elements are sometimes used for increased gain and discrimination. As the number of elements is increased the gain and front-to-back ratio increases but the radiation resistance decreases and the bandwidth or frequency range over which the antenna will operate without reduction in effectiveness is decreased.

Material for Elements While the elements may consist of wire supported on a wood framework, self-supporting elements of tubing are much to be preferred. The latter type array is easier to construct, looks better, is no more expensive, and avoids the problem of getting sufficiently good insulation at the ends of the elements. The voltages reach such high values towards the ends of the elements that losses will be excessive, unless the insulation is excellent.

The elements may be fabricated of thin-walled steel conduit, or hard drawn thin-walled copper tubing, but dural tubing is much better. Or, if you prefer, you may purchase tapered copper-plated steel tubing elements designed especially for the purpose. Kits are available complete with rotating mechanism and direction indicator, for those who desire to purchase the whole system ready to put up.

Element Spacing The optimum spacing for a two-element array is, as has been mentioned before, approximately 0.1 wavelength for a director and 0.15 wavelength for a reflector. However, when both a director and a reflector are combined with the driven element to make up a three-element array the optimum spacing is established by the bandwidth which the antenna will be required to cover. Wide spacing (of the order of 0.25 wavelength between elements) will result in greater bandwidth for a specified maximum standing-wave ratio on the antenna transmission line. Smaller spacings may be used when boom length is an important consideration,

Figure 1.
TWO-ELEMENT 14-MC. ROTATABLE ARRAY WITH T-MATCH.



ANTENNA TYPE	DRIVEN ELEMENT LENGTH	REFLECTOR LENGTH	1ST DIRECTOR LENGTH	2ND DIRECTOR LENGTH	3RD DIRECTOR LENGTH	SPACING BETWEEN ELEMENTS, λ	APPROX. GAIN, DB	APPROX. RADIATION RESISTANCE, OHMS
2-ELEMENT USING REFLECTOR	$\frac{462}{F_{MC}}$	$\frac{490}{F_{MC}}$	—	—	—	0.15	5.0	30
2-ELEMENT USING DIRECTOR	$\frac{482}{F_{MC}}$	—	$\frac{455}{F_{MC}}$	—	—	0.1	5.5	15
3-ELEMENT	$\frac{468}{F_{MC}}$	$\frac{500}{F_{MC}}$	$\frac{445}{F_{MC}}$	—	—	0.1 D, 0.2 R	7.0	20
3-ELEMENT	$\frac{468}{F_{MC}}$	$\frac{495}{F_{MC}}$	$\frac{450}{F_{MC}}$	—	—	0.25 D, 0.25 R	8.0	50
4-ELEMENT	$\frac{468}{F_{MC}}$	$\frac{492}{F_{MC}}$	$\frac{442}{F_{MC}}$	$\frac{438}{F_{MC}}$	—	0.2	9.0	13
5-ELEMENT	$\frac{468}{F_{MC}}$	$\frac{492}{F_{MC}}$	$\frac{442}{F_{MC}}$	$\frac{438}{F_{MC}}$	$\frac{434}{F_{MC}}$	0.2	10.0	10

Figure 2.

DESIGN CHART FOR PARASITIC ARRAYS.

The values of gain and effective radiation resistance given for the multi-element arrays are subject to considerable variation as a result of the tuning of the elements, but the quantities given can be assumed to be average values. Dimensions are in feet.

but for a specified standing-wave ratio and forward gain the frequency coverage will be smaller. Thus the Q of the antenna system will be *increased* as the spacing between the elements is *decreased,* resulting in smaller frequency coverage, and at the same time the feed-point impedance of the driven element will be decreased.

For broad-band coverage, such as the range from 26.96 to 29.7 Mc. or from 50 to 54 Mc., 0.25 wavelength spacing from the driven element to each of the parasitic elements is recommended. For narrower bandwidth, such as would be adequate for the 14.0 to 14.4 Mc. band or the 144 to 148 Mc. band, the recommended lengths based on the experience of many operators is 0.1 wavelength spacing to the director and 0.2 wavelength spacing to the reflector. Design data on antennas using these spacings is given in the chart of figure 2.

Length of the Driven Element Experience has shown that the driven element in a conventional three-element parasitic array, assuming a diameter-to-length ratio of 200 to 400, will have just about the same length as a conventional half-wave antenna. Hence, the standard formula — $L = 468/F_{mc.}$ — will prove satisfactory for determining the length of the driven element. If the driven element consists of a cage, a folded dipole, or a yoke match for increasing the feed-point impedance of the element, the length of the element will be reduced up to several per cent as a result of the increased effective diameter of the driven element.

The presence of the parasitic elements *normally* will not result in any detuning of the driven element. This assumes that the parasitic elements are spaced approximately the same distance from the driven element if they are wide spaced (greater than 0.2 wavelength spacing), or that the director is somewhat closer than the reflector if the elements are close spaced (such as the 0.1D-0.2R or the 0.1D-0.15R combinations). This also assumes that the parasitic elements are both detuned about the same amount from the resonant length of the driven element. Under such conditions the shortening effect of the director upon the driven element will be just about compensated by the lengthening effect of the reflector upon the driven element.

Length of the Parasitic Elements Experience has shown that it is quite practical to cut the parasitic elements of a three-element beam to length, using the formulas given in figure 2, before the installation of such an antenna. Such an array will give good signal gain, adequate front-to-back ratio, and will give frequency coverage approximately as discussed earlier. True, the performance at some specified frequency may be improved a few db by carefully pruning the elements for best front-to-back ratio or best forward gain, but such a procedure will almost invariably result in reduced bandwidth of the antenna system.

The closer the lengths of the parasitic elements are to the resonant length of the driven element, the lower will be the feed-point resistance of the driven element, and the lower will be the bandwidth of the array. Hence, for wide frequency coverage the director should be considerably shorter, and the reflector considerably longer than the driven element. For example, the director should still be less than a resonant half wave at the upper frequency limit of the range wherein the antenna is to be operated, and the reflector should still be long enough to act as a reflector at the lower frequency limit. Another way of stating the same thing is to say, in the case of an array to cover a wide frequency range such as the amateur range from 26.96 to 29.7 Mc. or the width of a low-band TV channel, that the director should be cut for the upper end of the band and the reflector for the lower end of the band. In the case of the 26.96 to 29.7 Mc. range this means that the director should be about 8 per cent shorter than the driven element and the reflector should be about 8 per cent longer. Such an antenna will show a relatively constant gain of about 6 db over its range of coverage, and the pattern will not reverse at any point in the range.

Where the frequency range to be covered is somewhat less, such as a high-band TV channel, the 14.0 to 14.4 Mc. amateur band, or the lower half of the amateur 28-Mc. phone band, the reflector should be about 5 per cent longer than the driven element, and the director about 4 per cent shorter. Such an antenna will perform well over its rated frequency band, will not reverse its pattern over this band, and will show a signal gain of 7 to 8 db.

More Than Three Elements A small amount of additional gain may be obtained through use of more than two parasitic elements, at the expense of reduced feed-point impedance and lessened bandwidth. One additional director will add about 1 db, and a second additional director (making a total of five elements including the driven element) will add slightly less than one db more. In the v-h-f range, where the additional elements may be added without much difficulty, and where required bandwidths are small, the use of more than two parasitic elements is quite practicable.

Stacking of Yagi Arrays Parasitic arrays (yagi's) may be stacked to provide additional gain in the same manner that dipoles may be stacked, as discussed in Chapter Fourteen. Thus if an array of six dipoles would give a gain of 10 db, the substitution of yagi arrays for each of the dipoles would add the gain of *one* yagi array to the gain obtained with the dipoles. However, the yagi arrays *must be more widely spaced* than the dipoles to obtain this theoretical improvement. As an example, if six 5-element yagi arrays having a gain of about 10 db were substituted for the dipoles, with appropriate increase in the spacing between the arrays, the gain of the whole system would approach the sum of the two gains, or 20 db. A group of arrays of yagi antennas, with recommended spacings and approximate gains, are illustrated in figure 3.

16-2 Feed Systems for Parasitic (Yagi) Arrays

The table of figure 2 gives, in addition to other information, the approximate radiation resistance referred to the center of the driven element of multi-element parasitic arrays. It is obvious, from these low values of radiation resistance, that especial care must be taken in materials used and in the construction of the elements of the array to insure that ohmic losses in the conductors will not be an appreciable percentage of the radiation resistance. It is also obvious that some method of impedance transformation must be used in many cases to match the low radiation resistance of these antenna arrays to the normal range of characteristic impedance used for antenna transmission lines.

A group of possible methods of impedance matching is shown in figures 4, 5, 6, and 7. All these methods have been used but certain of them offer advantages over some of the other methods. Generally speaking it is not mechanically desirable to break the center of the driven element of an array for feeding the system. Breaking the driven element rules out the practicability of building an all-metal or "plumber's delight" type of array, and imposes mechanical limitations with any type of construction. However, when continuous rotation is desired, an arrangement such as shown in figure 6D, utilizing a broken driven element with a rotatable transformer for coupling from the antenna transmission line to the driven element has proven to be quite satisfactory. In fact the method shown in figure 6D is

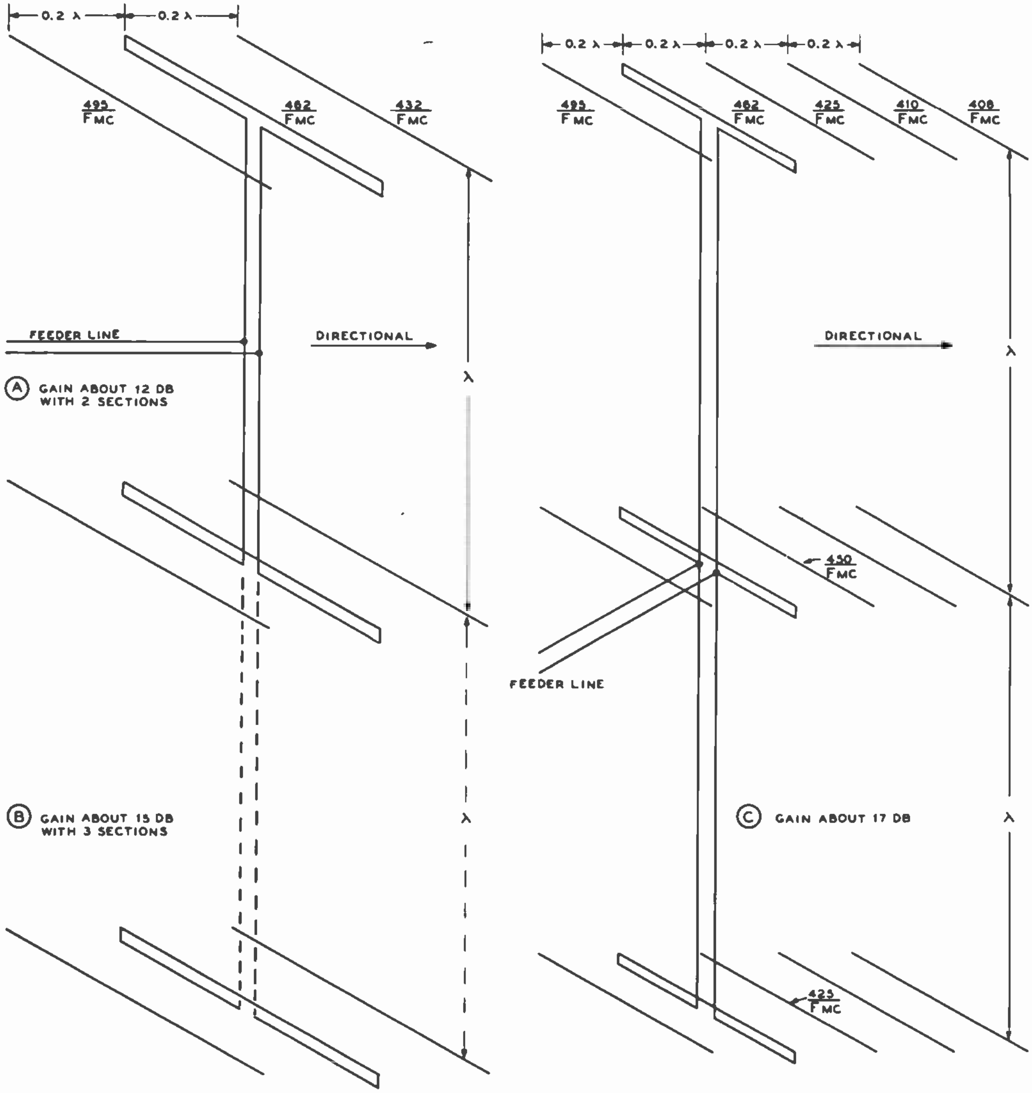


Figure 3.

STACKED YAGI ARRAYS.

It is possible to attain a relatively large amount of gain over a limited bandwidth with stacked yagi arrays. The two-section array at (A) will give a gain of about 12 db, while adding a third section will bring the gain up to about 15 db. Adding two additional parasitic directors to each section, as at (C), will bring the gain up to about 17 db.

probably the most practicable method of feeding the driven element when continuous rotation of the antenna array is required.

The feed systems shown in figure 4 will, under normal conditions, show the lowest losses of any type of feed system since the currents flowing in the matching network are the lowest of all the systems commonly used. The "Folded Element" match shown in figure 4A and the "Yoke" match shown in figure 4B are the most satisfactory electrically of all standard feed methods. However, both methods require the extension of an additional conductor out to the end of the driven element

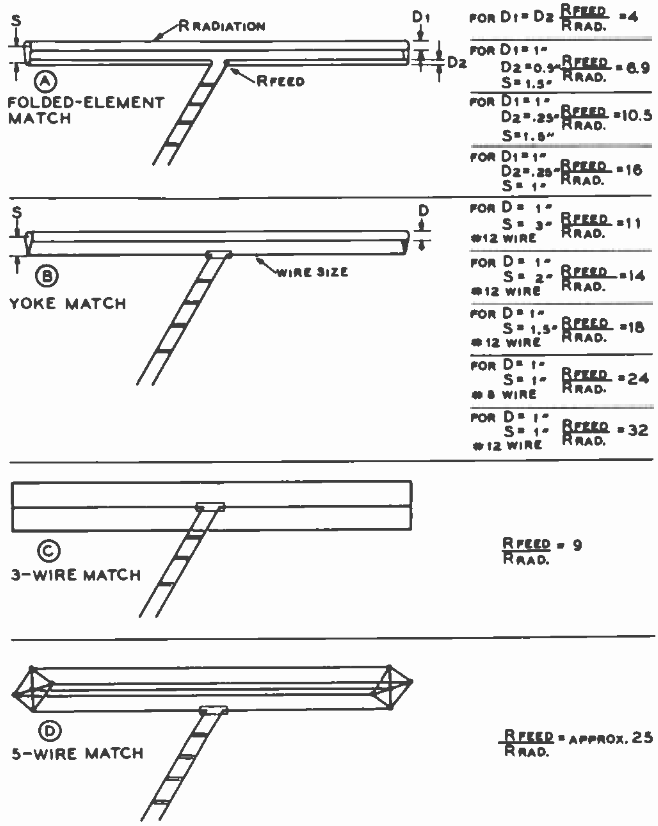


Figure 4.
DATA FOR FOLDED-ELEMENT MATCHING SYSTEMS.
In all normal applications of the data given the main element as shown is the driven element of a multi-element parasitic array. Directors and reflectors have not been shown for the sake of clarity.

as a portion of the matching system. The folded-element match is best on the 50-Mc. band and higher where the additional section of tubing may be supported below the main radiator element without undue difficulty. The yoke-match is more satisfactory mechanically on the 28-Mc. and 14-Mc. bands since it is only necessary to suspend a wire below the driven element proper. The wire may be spaced below the self-supporting element by means of several small strips of polystyrene which have been drilled for both the main element and the small wire and threaded on the main element.

The Folded-Element Match Calculations The calculation of the operating conditions of the folded-element matching system and the yoke match, as shown in figures 4A and 4B is relatively simple. A selected group of operating conditions has been shown on the drawing of figure 4. In applying the system it is only necessary to multiply the ratio of feed to radiation resistance (given in the figures to the right of the suggested operating dimensions in figure 4) by the radiation resistance of the antenna system to obtain the impedance of the cable to be used in feeding the array. Approximate values of radiation resistance for a number of commonly used parasitic-element arrays are given in figure 2.

As an example, suppose a 3-element array with 0.1D-0.2R spacing between elements is to be fed by means of a 465-ohm line constructed of no. 12 wire spaced 2 inches. The approximate radiation resistance of such an antenna array will be 20 ohms. Hence we need a ratio of impedance step up of 23 to obtain a match between the characteristic impedance of the transmission line and the radiation resistance of the driven element of the antenna array. Inspection of the ratios given in figure 4 shows that the fourth set of dimensions

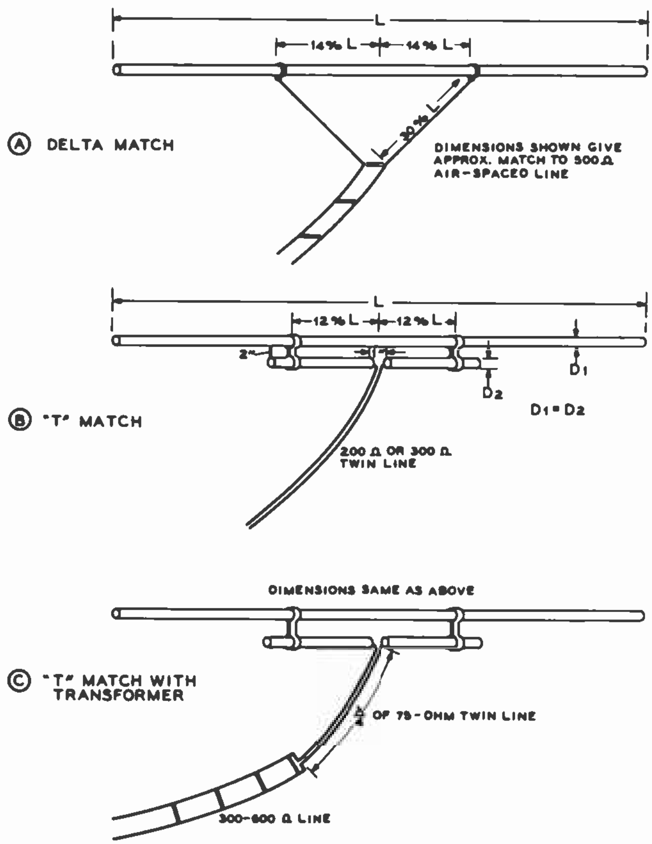


Figure 5.
AVERAGE DIMENSIONS FOR THE DELTA AND "T" MATCH.

given under figure 4B will give a 24-to-1 step up, which is sufficiently close. So it is merely necessary to use a 1-inch diameter driven element with a no. 8 wire spaced 1 inch centers (½ inch below the outside wall of the 1-inch tubing) below the 1-inch element. The no. 8 wire is broken and a 2-inch insulator placed in the center. The feed line then carries from this insulator down to the transmitter. The center insulator should be supported rigidly from the 1-inch tube so that the spacing between the piece of tubing and the no. 8 wire will be accurately maintained.

In many cases it will be desired to use the folded-element or yoke matching system with different sizes of conductors or different spacings than those shown on figure 4. Note, then, that the impedance transformation ratio of these types of matching systems is dependent *both upon the ratio of conductor diameters* and *upon their spacing.* The following equation has been given by Roberts (*RCA Review,* June, 1947) for the determination of the impedance transformation when using different diameters in the two sections of a folded element:

$$\text{Transformation ratio} = \left(1 + \frac{Z_1}{Z_2}\right)^2$$

In this equation Z_1 is the characteristic impedance of a line made up of the smaller of the two conductor diameters spaced the center-to-center distance of the two conductors in the antenna, and Z_2 is the characteristic impedance of a line made up of two conductors the size of the larger of the two. This assumes that the feed line will be connected in series with the *smaller* of the two conductors so that an impedance step up of greater than four will be obtained. If an impedance step up of less

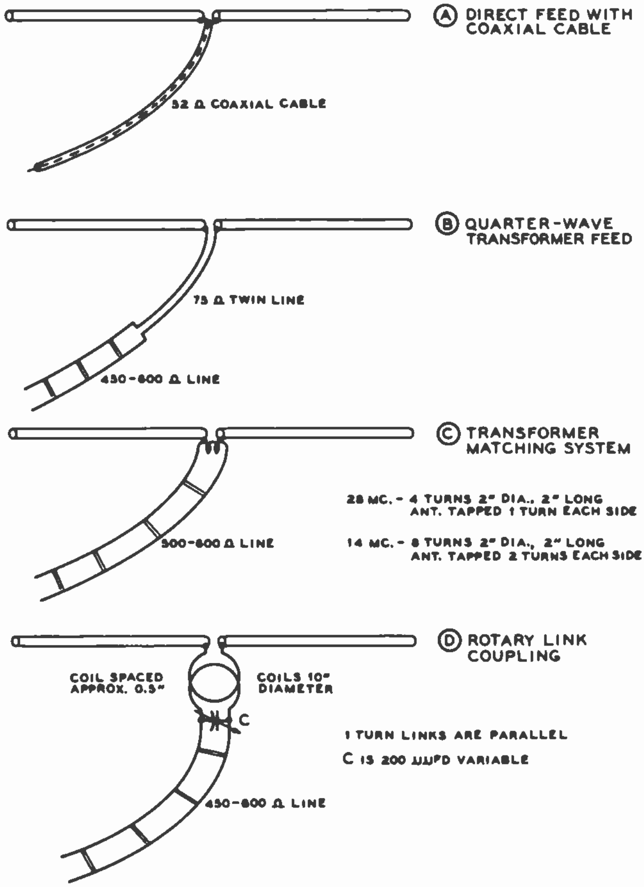


Figure 6.
ALTERNATE FEED METHODS WHERE THE DRIVEN ELEMENT MAY BE BROKEN IN THE CENTER.

than four is desired, the feed line is connected in series with the *larger* of the two conductors and Z_1 in the above equation becomes the impedance of a hypothetical line made up of the larger of the two conductors and Z_2 is made up of the smaller. The folded unipole described in the previous chapter is a case where the feed line is connected in series with the larger of the two conductors.

The conventional 3-wire match to give an impedance multiplication of 9 and the 5-wire match to give a ratio of approximately 25 are shown in figures 4C and 4D. The 4-wire match, not shown, will give an impedance transformation ratio of approximately 16.

Delta Match and "T" Match The delta match and the "T" match are shown in figure 5. Both these systems are widely used and can be adjusted to give a reasonable standing-wave ratio on 300 to 600 ohm feed line. In the case of all three of the systems shown it will be necessary to make adjustments in the tapping distance along the driven radiator until minimum standing waves on the antenna transmission line are obtained. Since it is sometimes impracticable to eliminate completely the standing waves from the antenna transmission line when using these matching systems, it is common practice to cut the feed line, after standing waves have been reduced to a minimum, to a length which will give satisfactory loading of the transmitter over the desired frequency range of operation.

In cases where it does not prove practicable to obtain a satisfactorily low standing wave ratio when using the "T" match to the driven element the arrangement shown at figure 5C has proven very helpful. In those cases where the standing-wave ratio cannot be reduced to a sufficiently low value it has been found that the impedance at the feed point in the "T" section is *lower* than that of the antenna transmission line. Hence the inclusion of a quarter-wave transformer between this feed point and the feed line will present a higher impedance to the antenna transmission line. In all cases when using polyethylene-filled line for a matching transformer, the length of the transformer should be *shorter* than ¼ wave. The physical length will be ¼ wave times the velocity factor of the cable or line being used.

Feed Systems Using a Driven Element with Center Feed Four methods of exciting the driven element of a parasitic array are shown in figure 6. The system shown at (A) has proven to be quite satisfactory in the case of an antenna-reflector two-element array or in the case of a three-element array with 0.2 to 0.25 wavelength spacing between the elements of the antenna system. The feed-point impedance of the center of the driven element is close enough to the characteristic impedance of the 52-ohm coaxial cable so that the standing-wave ratio on the 52-ohm coaxial cable is less than 2-to-4. (B) shows an arrangement for feeding an array with a broken driven element from an open-wire line with the aid of a quarter-wave matching transformer. With 465-ohm line from the transmitter to the antenna this system will give a close match to a 12-ohm impedance at the center of the driven element. (C) shows an arrangement which uses an untuned transformer with lumped inductance for matching the transmission line to the center impedance of the driven element.

Rotary Link Coupling In many cases it is desirable to be able to allow the antenna array to rotate continuously without regard to snarling of the feed line. If this is to be done some sort of slip rings or rotary joint must be made in the feed line. One relatively simple method of allowing unrestrained rotation of the antenna is to use the method of rotary link coupling shown in figure 6D. The two coupling rings are 10 inches in diameter and are usually constructed of ¼-inch copper tubing supported one from the rotating structure and one from the fixed structure by means of standoff insulators. The capacitor C in figure 6D is adjusted, after the

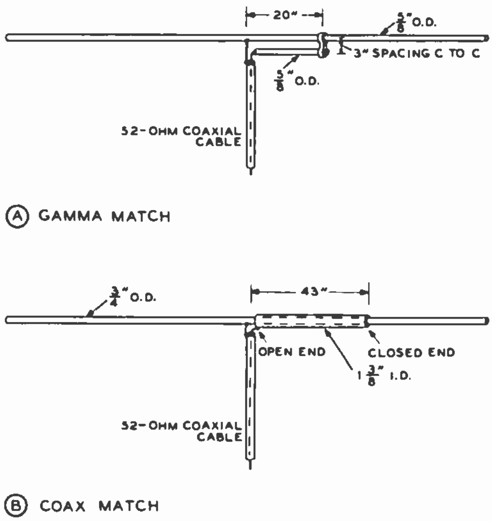


Figure 7.
THE GAMMA AND COAX MATCHING SYSTEMS.
Dimensions given are for the 28-Mc. band. Parasitic elements have not been shown.

antenna has been tuned, for minimum standing-wave ratio on the antenna transmission line. The dimensions shown will allow operation with either 14-Mc. or 28-Mc. elements, with appropriate adjustment of the capacitor C. The rings must of course be parallel and must lie in a plane normal to the axis of rotation of the rotating structure.

The Gamma Match and the Coax Match The use of coaxial cable to feed the driven element of a yagi array is becoming increasingly popular. One reason for this increased popularity lies in the fact that the TVI-reduction problem is simplified when coaxial feed line is used from the transmitter to the antenna system. Radiation from the feed line is minimized when coaxial cable is used, since the outer conductor of the line may be grounded at several points throughout its length and since the intense field is entirely confined within the outer conductor of the coaxial cable. Other advantages of coaxial cable as the antenna feed line lie in the fact that coaxial cable may be run within the structure of a building without danger, or the cable may be run underground without disturbing its operation. Also, trans-

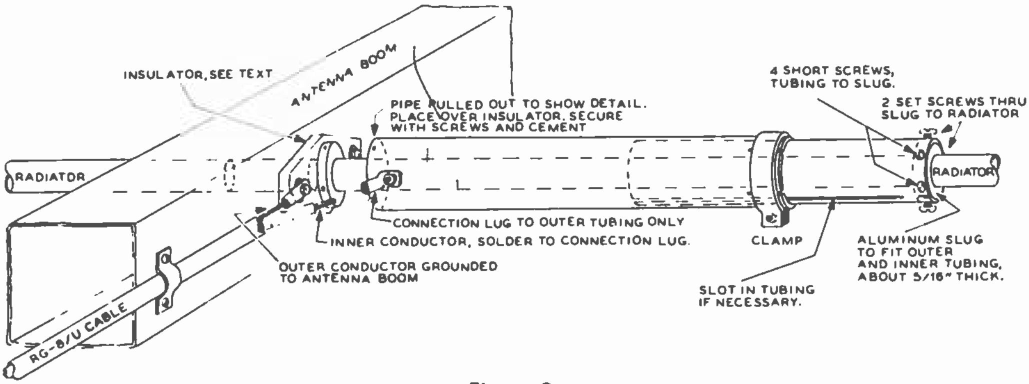


Figure 8.
CONSTRUCTIONAL DRAWING OF THE COAX MATCH.

mitting-type low-pass filters for a 52-ohm impedance are more widely available and are less expensive than equivalent filters for two-wire line.

The simple feed method shown in figure 6A may be used with coaxial cable when the center of the driven element may be broken, and when the feed-point resistance is fairly close to the characteristic impedance of the coaxial cable to be used. Two alternative feed methods for use with coaxial cable when the driven element may not be broken, and when the feed-point resistance may be considerably different from the line impedance are illustrated in figure 7. The method using a single parallel conductor illustrated in figure 7A is commonly called the "gamma" match, while the method using coaxial sections illustrated in figure 7B is called the "coax" match.

A drawing of the coax matching section, which slides over the driven element of the parasitic array, is illustrated in figure 8. Shown in figure 9 is a photograph of the end of the coax-match section which attaches to the support boom of the array by means of an insulator fabricated from polystyrene or other plastic material. A four-element parasitic array for the 28-Mc. band which uses the coax match system is described and illustrated in Section 16-5 of this chapter.

16-3 Unidirectional Driven Arrays

Three types of unidirectional driven arrays are illustrated in figure 10. The array shown in figure 10A is an end-fire system which may be used in place of a parasitic array of similar dimensions when greater frequency coverage than is available with the yagi type is desired. Figure 10B is a combination end-fire and colinear system which will give approximately the same gain as the system of figure 10A, but which requires less boom length and greater total element length. Figure 10C illustrates the familiar lazy-H with driven reflectors (or directors, depending upon the point of view) in a combination which will show wide bandwidth with a considerable amount of forward gain and good front-to-back ratio over the entire frequency coverage.

Figure 9.
SHOWING THE BOOM END OF THE COAX MATCHING SECTION.

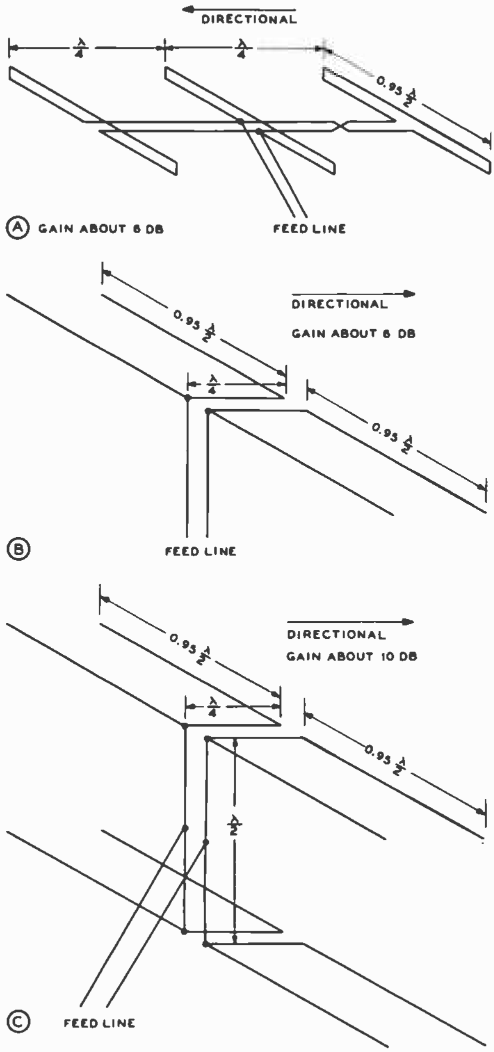


Figure 10.
UNIDIRECTIONAL ALL-DRIVEN ARRAYS.

A unidirectional all-driven end-fire array is shown at (A). (B) shows an array with two half waves in phase with driven reflectors. A Lazy-H array with driven reflectors is shown at (C). Note that the directivity is through the elements with the greatest total feed-line length in arrays such as shown at (B) and (C).

Unidirectional Stacked Broadside Arrays Three practicable types of unidirectional stacked broadside arrays are shown in figure 11. The first type, shown at figure 11A, is the simple "lazy H" type of antenna with parasitic reflectors for each element. (B) shows a simpler antenna array with a pair of folded dipoles spaced one-half wave vertically, operating with reflectors. In figure 11C is shown a more complex array with six half waves and six reflectors which will give a very worthwhile amount of gain.

In all three of the antenna arrays shown the spacing between the driven elements and the reflectors has been shown as one-quarter wavelength. This has been done to eliminate the requirement for tuning of the reflector, as a result of the fact that a half-wave element spaced exactly one-quarter wave from a driven element will make a unidirectional array when both elements are the same length. Using this procedure will give a gain of 3 db with the reflectors over the gain without the reflectors, with only a moderate decrease in the radiation resistance of the driven element. Actually, the radiation resistance of a half-wave dipole goes down from 73 ohms to 60 ohms when an identical half-wave element is placed one-quarter wave behind it.

A very slight increase in gain for the entire array (about 1 db) may be obtained at the expense of lowered radiation resistance, the necessity for tuning the reflectors, and decreased bandwidth by placing the reflectors 0.15 wavelength behind the driven elements and making them somewhat longer than the driven elements. The radiation resistance of each element will drop approximately to one-half the value obtained with untuned half-wave reflectors spaced one-quarter wave behind the driven elements.

Antenna arrays of the type shown in figure 11 require the use of some sort of lattice work for the supporting structure since the arrays occupy appreciable distance in space in all three planes.

Feed Methods The requirements for the feed systems for antenna arrays of the type shown in figure 11 are less critical than those for the close-spaced parasitic arrays shown in the previous section. This is a natural result of the fact that a larger number of the radiating elements are directly fed with energy, and of the fact that the effective radiation resistance of each of the driven elements of the array is much higher than the feed-point resistance of a parasitic array. As a consequence of this fact, arrays of the type shown in figure

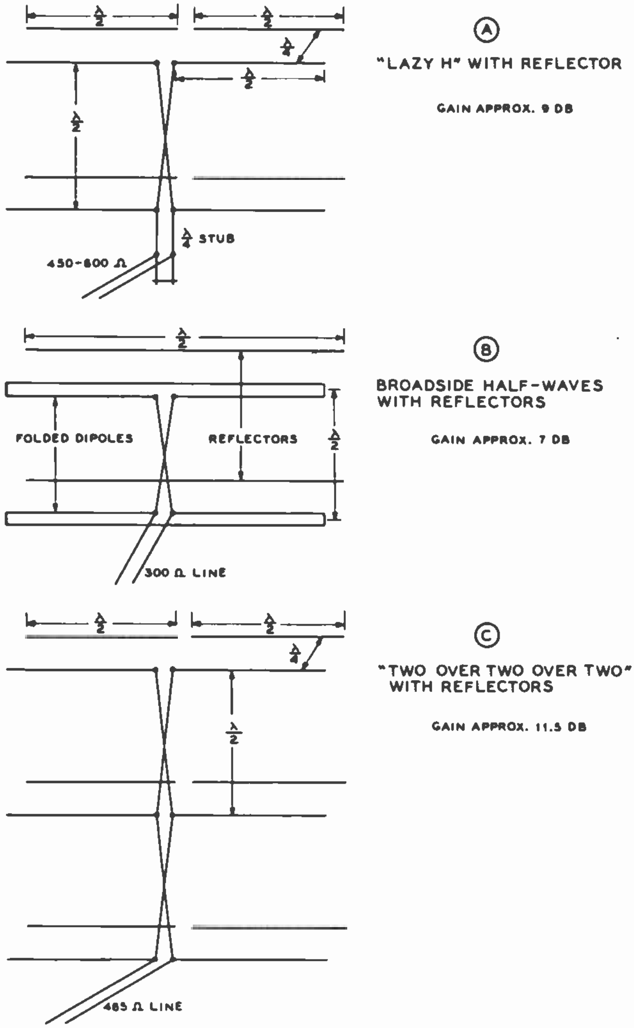


Figure 11.
BROADSIDE ARRAYS WITH PARASITIC REFLECTORS.
The apparent gain of the arrays illustrated will be greater than the values given due to concentration of the radiated signal at the lower elevation angles.

11 can be expected to cover a somewhat greater frequency band for a specified value of standing-wave ratio than the parasitic type of array.

In most cases a simple open-wire line may be coupled to the feed point of the array without any matching system. The standing-wave ratio with such a system of feed will often be less than 2-to-1. However, if a more accurate match between the antenna transmission line and the array is desired a conventional quarter-wave stub, or a quarter-wave matching transformer of appropriate impedance, may be used to obtain a low standing-wave ratio.

16-4 Bi-Directional Rotatable Arrays

The bi-directional type of array is sometimes used on the 28-Mc. and 50-Mc. bands where signals are likely to be coming from only one general direction at a time. Hence the sacrifice of discrimination against signals arriving from the opposite direction is likely to be of little disadvantage. Figure 12 shows two general types of bi-directional arrays. The flat-top beam, which has been described in detail in Chapter Fourteen, is well adapted to installa-

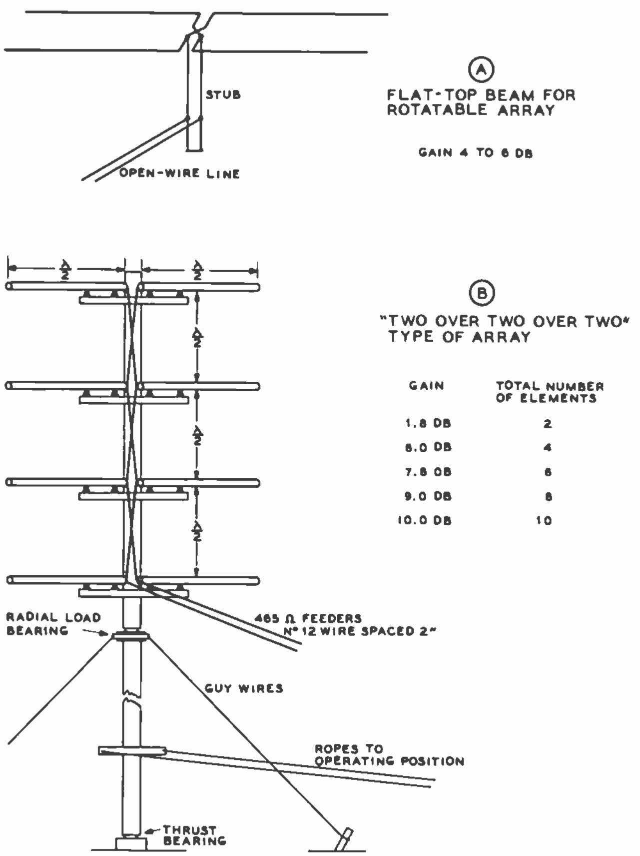


Figure 12.
TWO GENERAL TYPES OF BI-DIRECTIONAL ARRAYS.
Average gain figures are given for both the flat-top beam type of array and for the broadside-colinear array with different numbers of elements.

tion atop a rotating structure. When self-supporting elements are used in the flat-top beam the problem of losses due to insulators at the ends of the elements is somewhat reduced. With a single-section flat-top beam a gain of approximately 4 db can be expected, and with two sections a gain of approximately 6 db can be obtained.

Another type of bi-directional array which has seen less use than it deserves is shown in figure 12B. This type of antenna system has a relatively broad azimuth or horizontal beam, being capable of receiving signals with little diminution in strength over approximately 40°, but it has a quite sharp elevation pattern since substantially all radiation is concentrated at the lower angles of radiation if more than a total of four elements is used in the antenna system. Figure 12B gives the approximate gain over a half-wave dipole at the height of the center of the array which can be expected. Also shown in this figure is a type of "rotating mast" structure which is well suited to rotation of this type of array.

If six or more elements are used in the type of array shown in figure 12B no matching section will be required between the antenna transmission line and the feed point of the antenna. When only four elements are used the antenna is the familiar "lazy H" and a quarter-wave stub should be used for feeding from the antenna transmission line to the feed point of the antenna system.

If desired, and if mechanical considerations permit, the gain of the arrays shown in figure 12B may be increased by 3 db by placing a half-wave reflector behind each of the elements at a spacing of one-quarter wave. The array then becomes essentially the same as that shown in figure 11C and the same considerations in regard to reflector spacing and tuning will apply. However, the factor that a bidirectional array need be rotated through an angle of less than 180° should be considered in this connection.

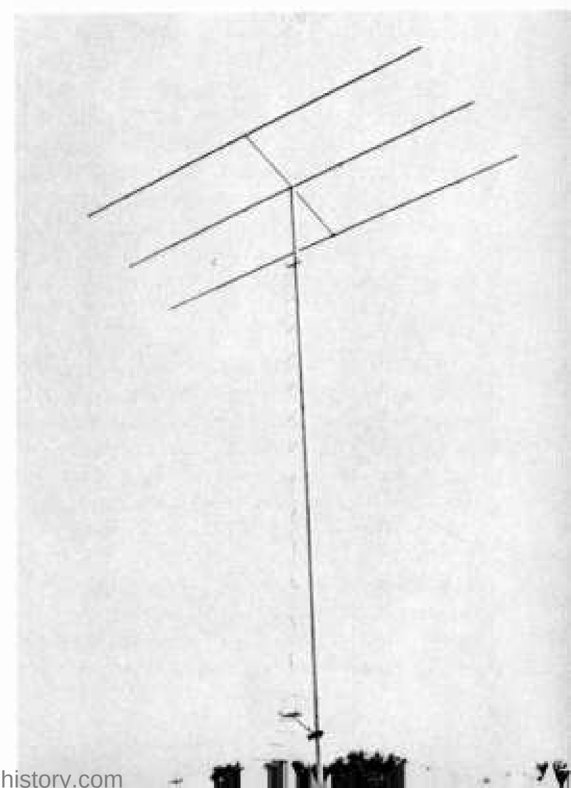
16-5 Construction of Rotatable Arrays

A considerable amount of ingenuity may be exercised in the construction of the supporting structure for a rotatable array. Every person has his own ideas as to the best method of construction. Often the most practicable method of construction will be dictated by the availability of certain types of constructional materials. But in any event be sure that sound mechanical engineering principles are used in the design of the supporting structure. There are few things quite as discouraging as the picking up of pieces, repairing of the roof, etc., when a newly constructed rotary comes down in the first strong wind. If the principles of mechanical engineering are understood it is wise to calculate the loads and torques which will exist in the various members of the structure with the highest wind velocity which may be expected in the locality of the installation. If this is not possible it will usually be worth the time and effort to look up a friend who understands these principles.

Radiating Elements One thing more or less standard about the construction of rotatable antenna arrays is the use of dural tubing for the self-supporting elements. Other materials may be used but an alloy known as 24ST has proven over a period of time to be quite satisfactory. Copper tubing is too heavy for a given strength, and steel tubing, unless copper plated, is likely to add an undesirably large loss resistance to the array. Also, steel tubing, even when plated, is not likely to withstand salt atmosphere such as encountered along the seashore for a satisfactory period of time. Do not use a soft aluminum alloy for the elements unless they will be quite short; 24ST is a hard alloy and is best although there are several other alloys ending in "ST" which will be found to be satisfactory. Do not use an alloy ending in "SO" or "S" in a position in the array where structural strength is important, since these letters designate a metal which has not been heat treated for strength and rigidity. However, these softer alloys, and aluminum electrical conduit, may be used for short radiating elements such as would be used for the 50-Mc. band or as interconnecting conductors in a stacked array.

"Plumber's Delight" Construction It is characteristic of the conventional type of multi-element parasitic array such as discussed previously and outlined in figure 2 that the centers of all the elements are at zero r-f potential with respect to ground. Hence it is possible to use a metallic structure without insulators for supporting the various elements of the array. A 28-Mc. three-element array of this type is shown in figure 13. Mechanical information on the construction of this particular array is given in figure 14. In this particular array pipe-fitting "T's" have been used at either end to support the 1-inch dural tubing reflector and director, with pieces of standard water pipe as spacers on either side between the parasitic elements and the driven element. The fitting at the center of the structure was made

Figure 13.
A "PLUMBER'S DELIGHT" ARRAY FOR 28 MC.
Dimensions for the array are given in figure 14.



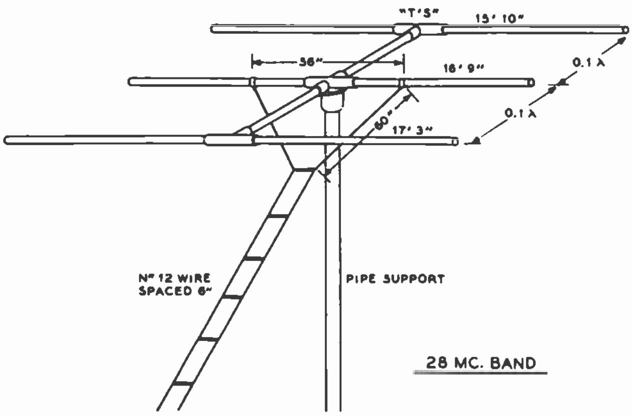


Figure 14.
AN APPLICATION OF THE "DELTA MATCH" SYSTEM FOR FEEDING A THREE-ELEMENT "PLUMBER'S DELIGHT" TYPE OF ROTATABLE ARRAY.

especially for the purpose, with threads at the bottom for the steel pipe which supports the entire structure, threads on two sides for receiving the two pieces of 1-inch water pipe which support the reflector and director, and another hole at right angles through which the driven element is passed. The parasitic elements are held fast in the T's at either end by splitting the top of the T the long way and running a bolt through the T and the dural element at each end of the T. The center element is held in position in the center piece by means of two bolts tapped into the center piece which are tightened down against the driven element.

The center piece in this particular array was turned from a large piece of rod and later drilled and tapped. A somewhat less expensive center piece of adequate strength could be constructed by welding an appropriate flange to the bottom of a steel plate, welding a piece of pipe to the top to hold the driven element, and welding either two pieces of pipe, or fittings to receive them, for the two members which support the parasitic elements.

A piece of hardened steel drill rod such as used in oil-well construction was used as the rotating-mast support for the array shown in figure 13. The rotating mast is supported a little more than half way up by a radial-load bearing attached to a telephone pole, and a heavy thrust bearing is used at the bottom.

The Simple Girder Supporting Structure In many cases when metal-working tools are not available a relatively simple supporting structure made of wood is used to hold the radiating elements. Figures 15, 16, 17, and 18 show photographs and constructional details for two 50-Mc. arrays in which a single piece of wood is used to hold the radiating elements. The main problem in constructing this type of a structure is to make sure that the holes through the supporting member are drilled true. If the holes are out of alignment the completed array is likely to have a somewhat haphazard appearance, although its operation probably will not be impaired. The array shown in figures 15 and 16 is a very simple affair which may be constructed in a relatively short period of time.

The antenna array shown in figures 17 and 18 is an answer to the problem of providing both horizontally-polarized and vertically-polarized operation with the same directive array. The supporting structure is constructed so as to be capable of rotation in a plane perpendicular to the axis of the member so that the elements may be either horizontal or vertical. A pair of ropes attached to a bridle bolted to the supporting member make it possible to change the polarization of the array without changing the direction of radiation.

Figure 19 is a drawing of a relatively simple three-element vertically-polarized array for the 50-Mc. band. The basis of the array is a "hypodermic" or sleeve-type dipole radiator. This array, as well as those shown in figures 15 and 17, is directly fed at the center of the driven element by means of 52-ohm coaxial cable.

Although a very simple wood structure can be used to support the elements of a 50-Mc. array, a somewhat more rigid structure must be

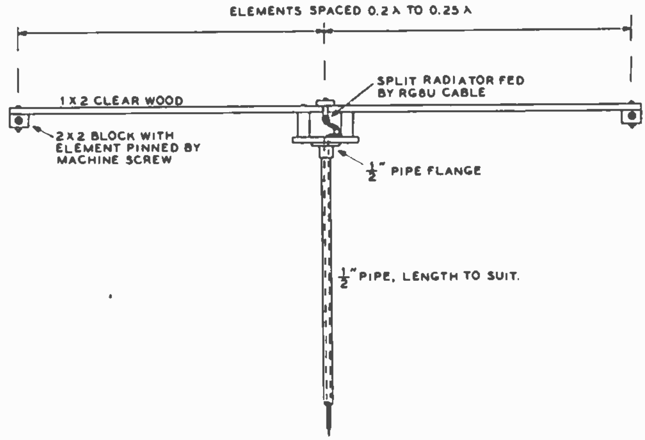


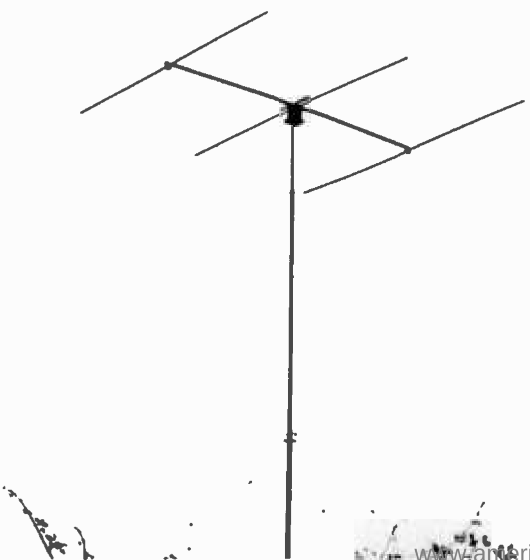
Figure 15.
DRAWING OF A SIMPLE THREE-ELEMENT ARRAY SUITABLE FOR 50-MC. OPERATION.

used as the main boom in a 28-Mc. or 14-Mc. antenna. Figure 20 shows two commonly used types of center main boom for a larger array such as is required on the lower-frequency bands. Figure 20A shows a metal-boom type of construction which is quite satisfactory for construction of a plumber's delight type of structure. If the rectangular type of tubing is available it will be found somewhat easier to manage than the round dural tubing, but both types are relatively simple to use in making such a structure. For anchoring the radiating elements to the dural boom either a set of collars on either side of the boom may be used, or bolts may be run through both the boom and the elements. Any of the shunt feeding systems shown in figures 4, 5 or 7 may be used to feed an array of this type.

A conventional ladder makes a satisfactory supporting boom for an array in the general manner illustrated in figure 20B. Ladders are relatively inexpensive, and produce a strong and stable type of mounting platform. The ladder, and for that matter any type of wood supporting structure, should be given several coats of a good grade of outside paint to protect it from the elements.

Supporting structures for more complex arrays involving elements in several planes may effectively be constructed of lattice work. Main members should be constructed of well-seasoned pieces of straight lumber. The entire structure should be both glued (using a waterproof glue) and bolted into place.

Figure 16.
PHOTOGRAPH OF THE ARRAY OF FIGURE 15.



Four-Element Array Using the Coax Match Figures 21 and 22 illustrate a four-element close-spaced array for the 28-Mc. band which uses the coax match for feeding the array from 52-ohm coaxial cable. The matching system is discussed in Section 16-2 and illustrated in figure 7B and figure 8. The complete array, which had 0.1-wavelength spacing between elements, showed a standing-wave ratio of 1:1 at 28.6-Mc., and the s.w.r. was below 2:1 over the range from 28 to 29.2-Mc. The final element lengths were: reflector, 17'3"; driven

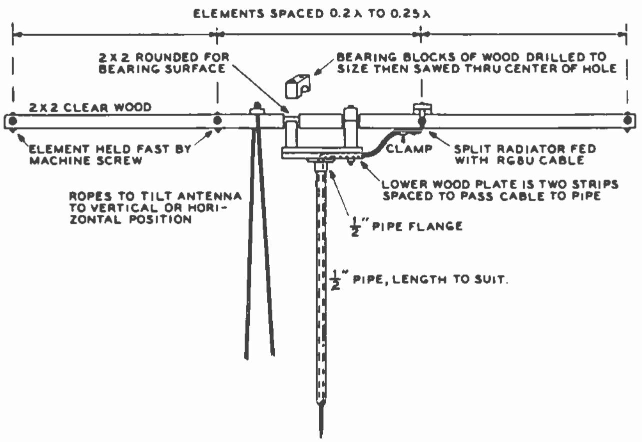
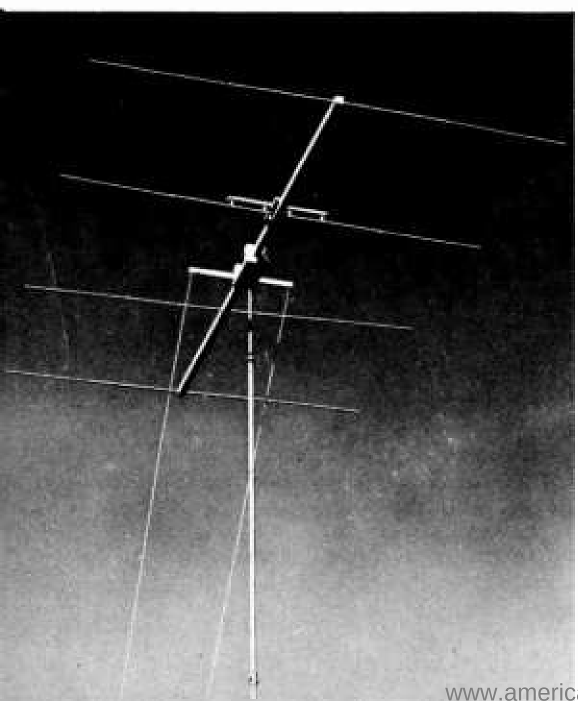


Figure 17.
DRAWING OF A "TIPPABLE" FOUR-ELEMENT ARRAY FOR THE 50-MC. BAND.

element, 16'7"; outer tube of matching section, 43⅜"; first director, 15'9"; second director, 15'9".

A closeup of the matching section is shown in figure 22, and is sketched in figure 8. If two sizes of tubing in the range between 1¼ and 1½ inches which will nest can be found, construction will be much simplified. But in constructing the model illustrated the only size available was about 1⅜" i.d. Hence it was necessary to cut a slot about 5/16" wide from the wall of one section of tubing so that it would slide inside the other piece. After the correct length of the matching section has been found, a single section of tubing the correct length may be substituted for the nested lengths.

Figure 18.
PHOTOGRAPH OF THE "TIPPABLE" 50-MC. ARRAY IN THE HORIZONTAL POSITION.



The ¼" thick aluminum disc at the end of the matching section, and the compression clamp at the end of the rigidly mounted section of tubing posed no problem. But the insulator at the boom end of the matching section, as shown in figures 8 and 9, was fabricated from two sections of ¼" thick polystyrene sheet, cemented together with polystyrene cement. However, since the r-f voltage at this point is quite low, material such as micarta or laminated bakelite would be quite adequate from the standpoint of losses. A clearance hole for the antenna element is drilled through both sections of the insulator, and small holes for sheet-metal screws to hold the end of the matching section are drilled in the outer piece of the assembled insulator.

Before the tubing which makes up the matching section is slid over the antenna element, the large "solder pot" type of lug which will take the center conductor of the coaxial cable should be mounted. The lug should be fastened with a screw whose head is inside the tubing, and with a nut and lockwasher so that the screw will not loosen. The matching section may then be slid on the antenna element

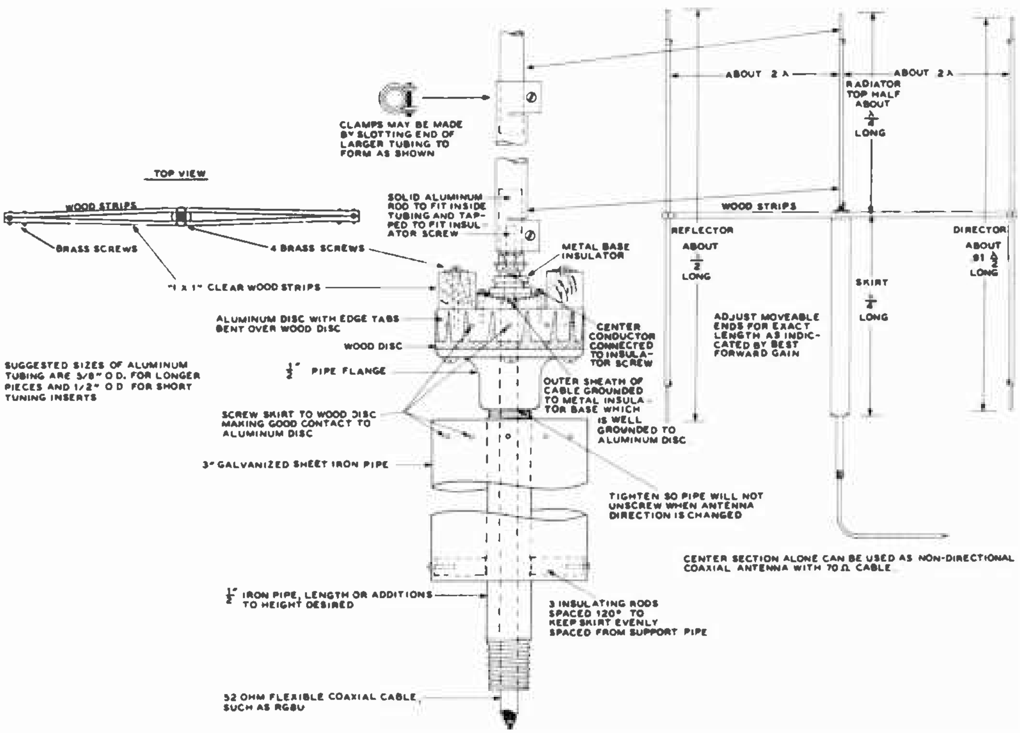


Figure 19.
DRAWING OF A THREE-ELEMENT VERTICALLY-POLARIZED ARRAY.

and the coaxial cable connected as shown in figure 8. After assembly, the end of the coaxial cable should be wrapped tightly with "Scotch" electrical tape to prevent the entry of moisture.

Dual-Band Array for 28 and 50 Mc.

Figure 23 shows a dual-band 11-10 meter and 6-meter antenna array. The antenna system is an excellent example of the neat and lightweight structure which can result from careful overall planning of the entire structure. The system represents a very moderate cash outlay, yet gives excellent results over both frequency ranges.

The antenna structure is of the rotating-mast type, with coaxial feed lines for both antennas being fed from the base of the structure to the driven elements of the arrays through the rotating mast. Constructional details on the rotating-mast structure are given in figure 28.

Construction details of the antenna system are given in figure 24. It may be seen that the 28-Mc. array is mounted to a short length of 1-inch galvanized iron pipe. Two angle-iron brackets are welded to this short length of pipe, and the boom for the antenna then is welded to the lengths of angle iron. This length of pipe then is joined to the main 1-inch support pipe, with lock nuts thoroughly tightened to keep the assembly from unscrewing with use or in a heavy wind. It was found most convenient to drill the angle irons to pass quarter-inch carriage bolts before the irons are welded to the short length of pipe.

All materials used in the antenna system are widely available, since standard plumbing supplies are used throughout. No machine work is required, and the welding may be done anywhere. If it is desired to install a 50-Mc. array atop the one for 28-Mc., the smaller

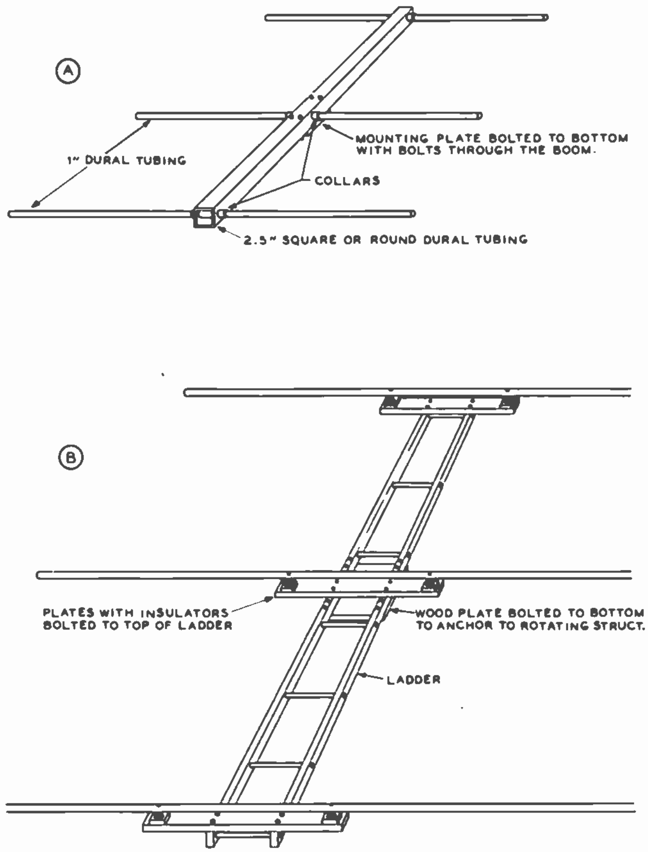


Figure 20.
ALTERNATIVE SUPPORTING BOOM ARRANGEMENTS.

(A) shows the use of a section of dural tubing (either rectangular or round cross section) for supporting a moderate size array. At (B) is shown the use of a ladder for supporting a relatively large array.

array may be supported about 4 feet above the larger antenna by means of a length of ½-inch pipe. The ½-inch pipe is screwed to the stub of the 1-inch pipe projecting above the larger array by means of a standard 1-inch to ½-inch pipe reducer. If the smaller array is not to be used, a 1-inch pipe cap may be placed over the end of the 1-inch stub of pipe. All pipe joints should be treated with pipe compound to prevent rusting of the threads, and all lock nuts must be thoroughly tightened.

Construction of the Antenna The boom longerons were made from a single length of 1" by 4" selected stock in each case. After the length is selected for each boom at the lumber yard, have each length ripped to make two matched 1" by 2" pieces. When the two pieces thus obtained are placed on edge it will probably be found that they have a slight curve, but the curve should be matched for both pieces. Place the two pieces so that the curve is in an upward direction at the ends. Then the weight of the parasitic elements will just about compensate for the initial droop in the boom. Strips of ½-inch oak flooring material will make good mountings at the ends of the boom for the parasitic elements.

The boom length in each case is equal to one-half wavelength. Stiffening blocks halfway out on each side are used on the 28-Mc. boom, while no stiffening blocks are required for the 50-Mc. array. With quarter-wave spacing of the parasitic elements as used in these antennas, the frequency response is quite broad. The lengths are cut to standard dimensions as given in figure 2. (Driven element, 468/F; reflector, 495/F; and director, 450/F, with frequencies in megacycles and lengths in feet.) Satisfactory operation over the active portion of the 50-Mc. band, and

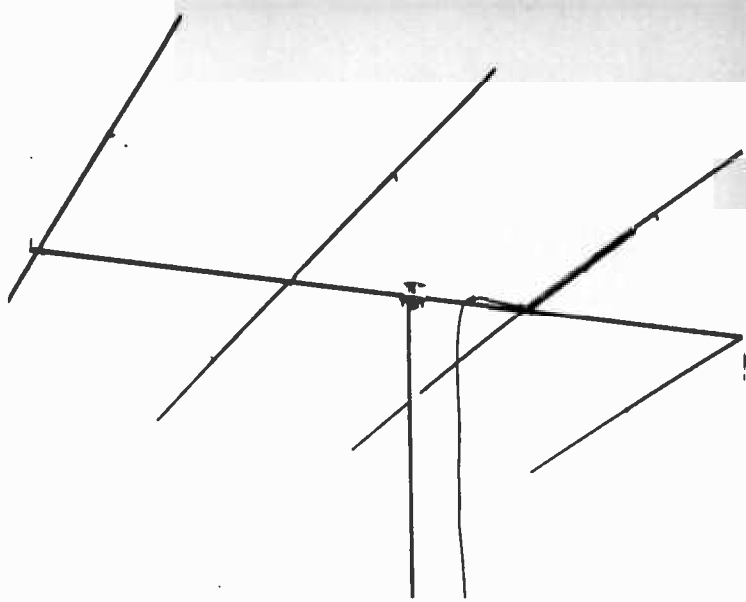


Figure 21.
FOUR-ELEMENT 28-MC. ARRAY WITH COAX MATCHING SECTION.
Dimensions for this array are given in the accompanying text.

over the entire 26.96 to 29.7 Mc. range is thus obtained, thanks to the relatively wide spacing of the parasitic elements.

Experience has shown that the elements may be constructed of the relatively soft but easily available electrical conduit. The ten-foot lengths may be used as the center portions of the 28-Mc. elements, with the adjustable ends made of aluminum tubing of the next smaller size. Both antennas are quite light in weight. It is not difficult to install the 28-Mc. array atop the rotating pipe, after the pipe has been lowered as far as it will travel. Then the 50-Mc. array may be installed, and all the appropriate lock nuts tightened. The installation is essentially a one-man job, but some help may be needed to pull the coaxial feed lines down the pipe as the antennas are being lifted into place.

The driven element of both arrays is split in the center, with each half being supported by a pair of standoff insulators. The driven element then is fed directly by the 52-ohm RG-8/U coaxial line; the inner conductor is connected directly to one side of the radiator while the outer conductor of the coaxial line is connected to the other side. The feed-point impedance of such a three-element array is close enough to 52 ohms to give a very low s.w.r. on the coaxial line.

8-Element "Tippable" Array for 144 Mc. Figures 25, 26, and 27 illustrate an 8-element rotatable array for use

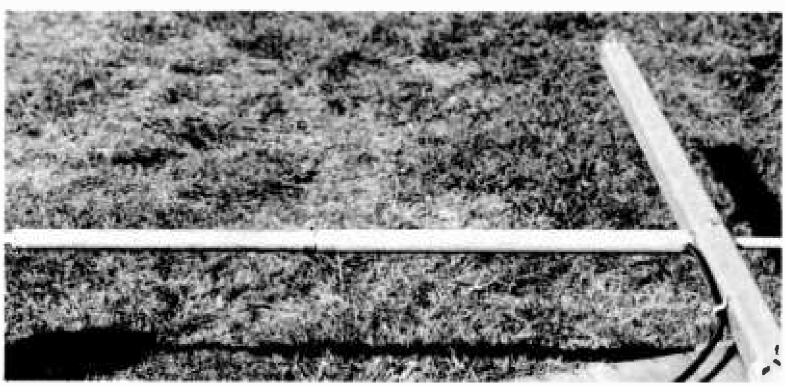


Figure 22.
CLOSE-UP PHOTO GRAPH OF THE COAX MATCHING SECTION, THE BOOM, AND THE DRIVEN ELEMENT OF THE ARRAY ILLUSTRATED IN FIGURE 21.

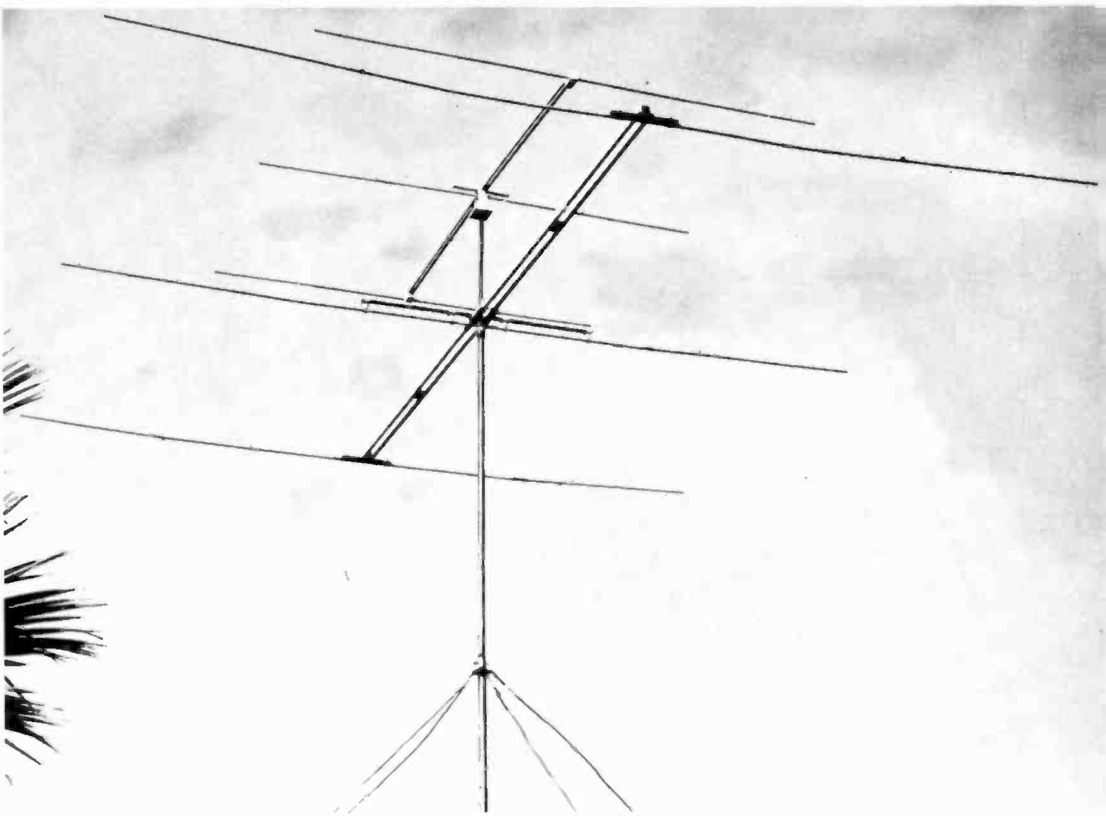


Figure 23.
LIGHTWEIGHT DUAL 28-MC. AND 50-MC. ANTENNA ARRAY.
Both antennas are supported by the lightweight rotating mast structure illustrated in figure 28. The two coaxial feed lines pass down the rotating support pipe to the operating position.

on the 144-Mc. amateur band which is "tippable" to obtain either horizontal or vertical polarization. It is necessary that the transmitting and receiving station use the same polarization for the ground-wave signal propagation which is characteristic of this frequency range. Although polarization has been loosely standardized in various areas of the country, exceptions are frequent enough so that it is desirable that the polarization of antenna radiation be easily changeable from horizontal to vertical.

The antenna illustrated has shown a signal gain of about 13 db, representing a power gain of about 20. Although the signal gain of the antenna is the same whether it is oriented for vertical or horizontal polarization, the horizontal beam width is smaller when the antenna is oriented for vertical polarization. Conversely, then, the vertical pattern is the sharper when the antenna system is oriented for horizontal polarization.

The changeover from one polarization to the other is accomplished simply by pulling on the appropriate cord. Hence, the operation is based on the offset head sketched in figure 25. Although a wood mast has been used, the same system may be used with a pipe mast.

The 40-inch lengths of RG-59/U cable (electrically ¾ wavelength) running from the center of each folded dipole driven element to the coaxial T-junction allow enough slack to permit free movement of the main boom when changing polarity. Type RG-8/U cable is run

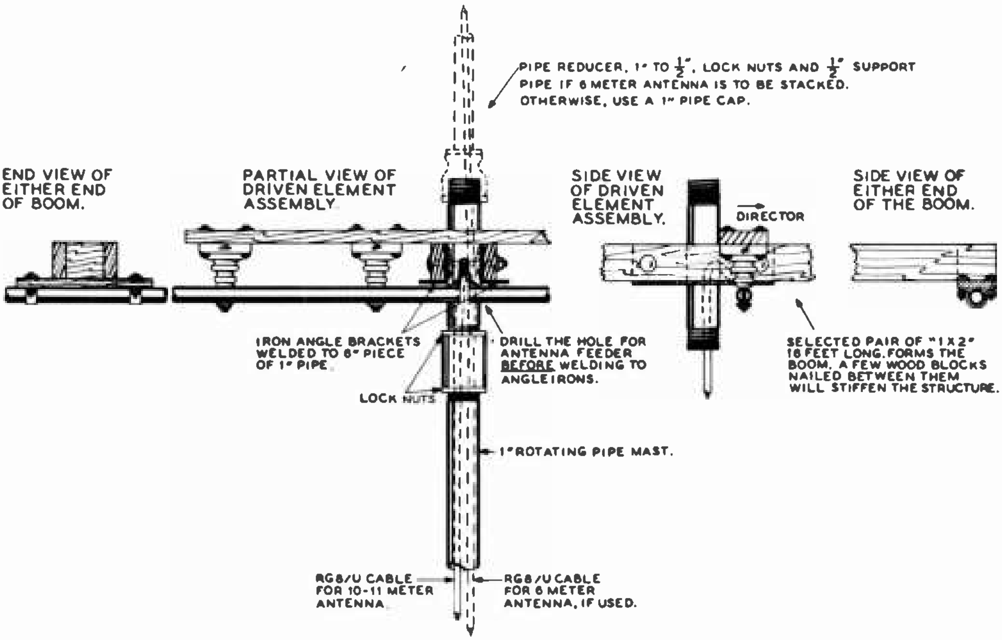


Figure 24.
CONSTRUCTIONAL DETAILS OF THE LIGHTWEIGHT ARRAY.

from the T-junction to the operating position. Measured standing-wave ratio was less than 2:1 over the 144 to 148 Mc. band, with the lengths and spacings given in figure 25.

Construction of the Array Most of the constructional aspects of the antenna array are self-evident from figure 25. However, the pointers given in the following paragraphs will be of assistance to those wishing to reproduce the array.

The drilling of holes for the small elements should be done carefully on accurately marked centers. A small angular error in the drilling of these holes will result in a considerable misalignment of the elements after the array is assembled. The same consideration is true of the filing out of the rounded notches in the ends of the main boom for the fitting of the two antenna booms.

Short lengths of wood dowel are used freely in the construction of the array. The ends of the small elements are plugged with an inch or so of dowel, and the ends of the antenna booms are similarly treated with larger discs pressed into place. A nice touch is added to these points and to the junctions of the elements and the antenna booms by the use of "Plastic Metal" for filling. The material is available from auto supply stores, and is used for filling fender dents before repainting.

The wood plugs in the ends of the elements and booms are pressed in about an eighth inch past the ends of the tubing. Then the filling compound is brushed in, layer by layer, until a smoothly rounded end is obtained. Similarly, the element junctions, which have been aligned and the element held in place with a single sheet-metal screw, are brushed with the compound, allowing each layer to dry, until the desired thickness and contour has been obtained. When thoroughly dry the built-up compound may be dressed smooth with fine sandpaper to give an appearance similar to a wiped solder joint.

The ends of the folded dipoles are made in the following manner: Drive a length of dowel into the short connecting lengths of aluminum tubing. Then drill down the center of the dowel with a clearance hole for the connecting screw. Then shape the ends of the connecting pieces to fit the sides of the element ends.

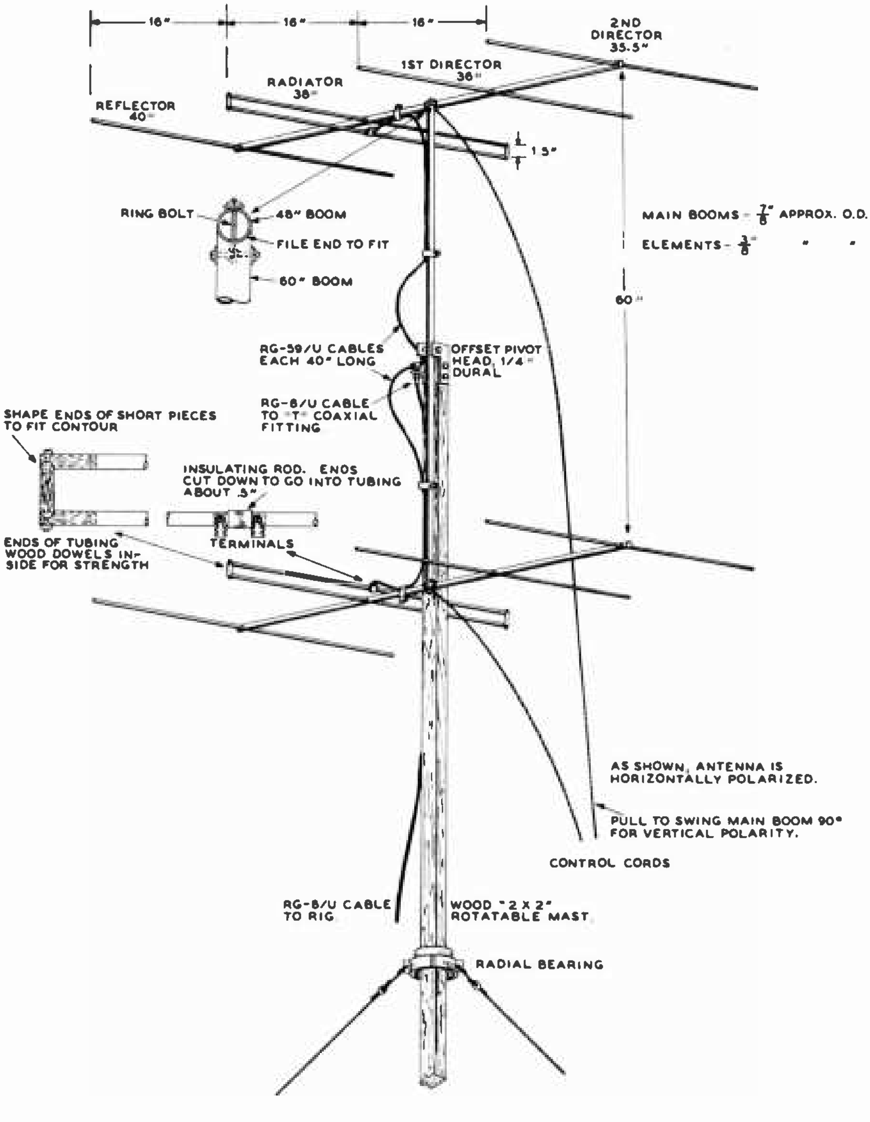


Figure 25.
CONSTRUCTIONAL DRAWING OF AN EIGHT-ELEMENT TIPPABLE 144-MC. ARRAY.

After assembly the junctions may be dressed with a file and sandpaper until a smooth fit is obtained.

The mast used for supporting the array is a 30-foot spliced 2 by 2. A large discarded ball bearing is used as the radial load bearing and guy-wire termination. Enough of the upper-mast corners were removed with a drawknife to permit sliding the ball bearing down about 9 feet from the top of the mast. The bearing then was encircled by an assembly of three pieces of dural ribbon to form a clamp, with ears for tightening screws and attachment of the guy wires. The bearing then was greased and covered with a piece of auto inner tube to serve as protection from the weather. Another junk-box bearing was used at the bottom of the mast as a thrust bearing.

The main booms were made from ¾-inch aluminum electrical conduit. Any size of small tubing will serve for making the elements. Note that the main boom is mounted at the balance center and not necessarily at the physical center. The pivot bolt in the offset head should be tightened sufficiently that there will be adequate friction to hold the array in position. Then an additional nut should be placed on the pivot bolt as a lock.

In connecting the phasing sections between the T-junction and the centers of the folded dipoles, it is important that the center conductors of the phasing sections be connected to the same side of the driven elements of the antennas. In other words, when the antenna is oriented for horizontal polarization and the center of the coaxial phasing section goes to the left side of the top antenna, the center conductor of the other coaxial phasing section should go to the left side of the bottom antenna.

16-6 Tuning the Array

Although many arrays may be constructed, installed, and operated with substantially no tuning process, there is always some doubt in the mind of the operator as to whether or not the array is delivering optimum results. So many operators make a check on the operation of the array before calling the job complete.

The process of tuning an array may fairly satisfactorily be divided into two more or less distinct steps: the actual tuning of the array for best front-to-back ratio or for maximum forward gain, and the project of obtaining the best possible impedance match between the antenna transmission line and the feed point of the array.

Tuning the Array Proper The actual tuning of the array for best front-to-back ratio or maximum forward gain may best be accomplished with the aid of a low-

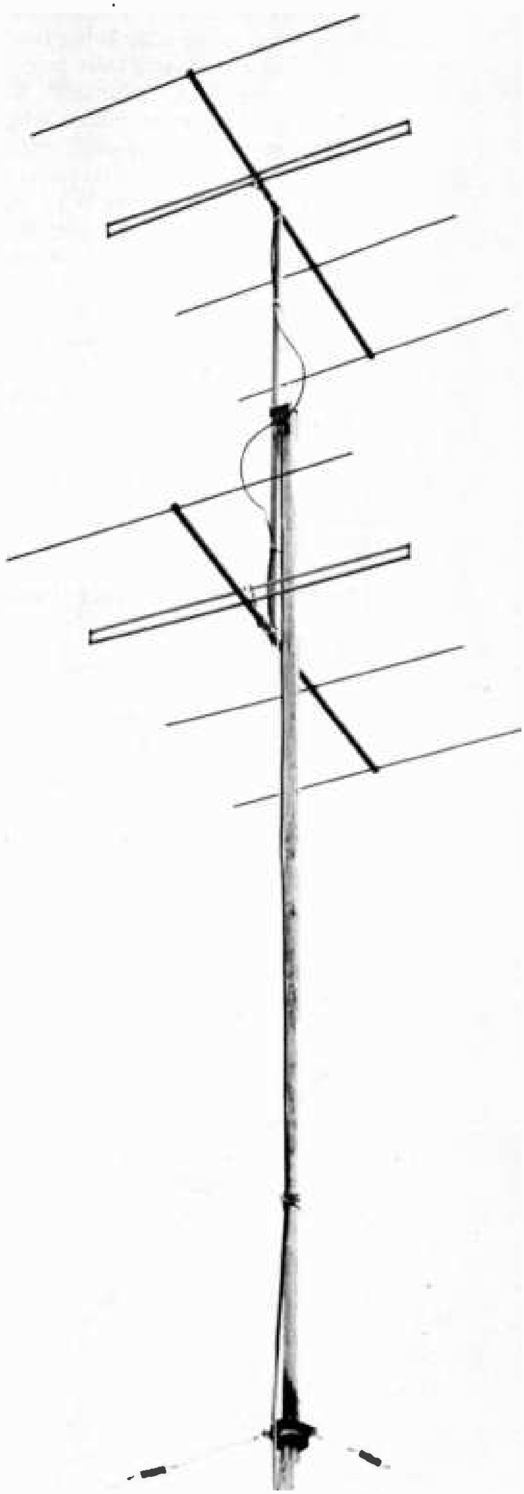


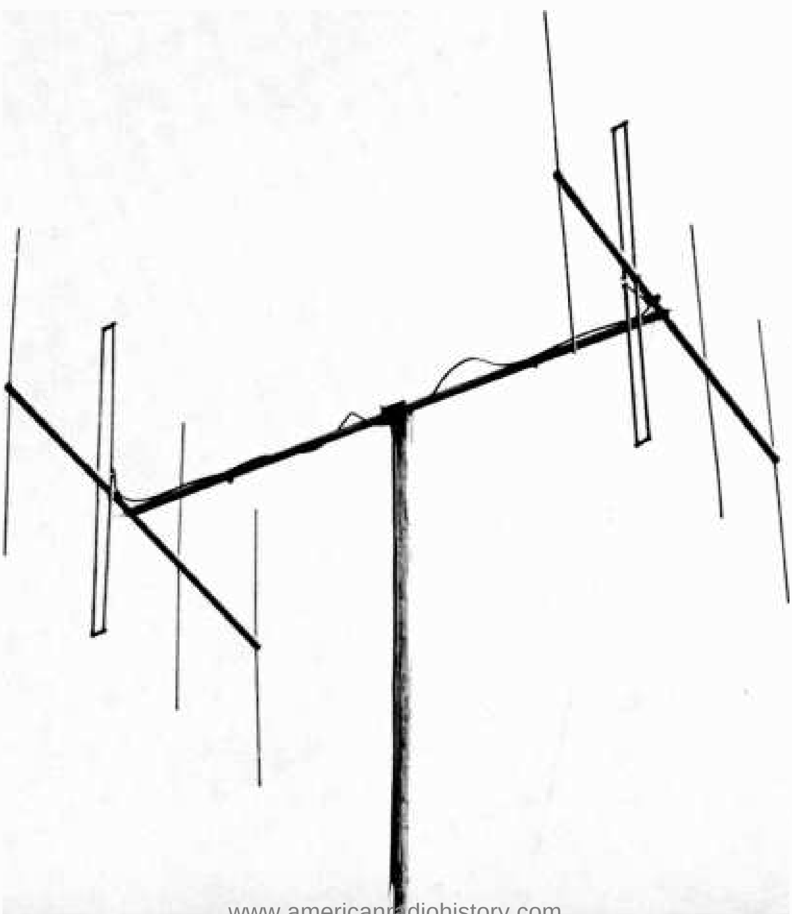
Figure 26.
SHOWING THE TIPPABLE 144-MC. ARRAY IN THE HORIZONTAL POSITION.
The radial-load bearing may be seen in this photograph.

power transmitter feeding a dipole antenna (polarized the same as the array being tuned) at least four or five wavelengths away from the antenna being tuned. A calibrated field-strength meter of the remote-indicating type is then coupled to the feed point of the antenna array being tuned. The transmissions from the portable transmitter should be made as short as possible and the call sign of the station making the test should be transmitted at least every ten minutes.

It is, of course, possible to tune an array with the receiver connected to it and with a station a mile or two away making transmissions on your request. But this method is more cumbersome and is not likely to give complete satisfaction. It is also possible to carry out the tuning process with the transmitter connected to the array and with the field-strength meter connected to the remote dipole antenna. In this event the indicating instrument of the remote-indicating field-strength meter should be visible from the position where the elements are being tuned. However, when the array is being tuned with the transmitter connected to it there is always the problem of making continual adjustments to the transmitter so that a constant amount of power will be fed to the array under test. Also, if you use this system, use very low power (5 or 10 watts of power is usually sufficient) and make sure that the antenna transmission line is effectively grounded as far as d-c plate voltage is concerned. The use of the method described in the previous paragraph of course eliminates these problems.

One satisfactory method for tuning the

Figure 27.
SHOWING THE TIPPABLE 144-MC. ARRAY IN THE VERTICAL POSITION.



array proper, assuming that it is a system with several parasitic elements, is to set the directors to the dimensions given in figure 2 and then to adjust the reflector for maximum forward signal. Then the first director should be varied in length until maximum forward signal is obtained, and so on if additional directors are used. Then the array may be reversed in direction and the reflector adjusted for best front-to-back ratio. Subsequent small adjustments may then be made in both the directors and the reflector for best forward signal with a reasonable ratio of front-to-back signal. The adjustments in the directors and the reflector will be found to be interdependent to a certain degree, but if small adjustments are made after the preliminary tuning process a satisfactory set of adjustments for maximum performance will be obtained. It is usually best to make the end sections of the elements smaller in diameter so that they will slip inside the larger tubing sections. The smaller sliding sections may be clamped inside the larger main sections.

In making the adjustments described, it is best to have the rectifying element of the remote-indicating field-strength meter directly at the feed point of the array, with a resistor at the feed point of the estimated value of feed-point impedance for the array.

Matching to the Antenna Transmission Line The problem of matching the impedance of the antenna transmission line to the array is much simplified if the process of tuning the array is made a substantially separate process as just described. After the tuning operation is complete, the resonant frequency of the driven element of the antenna should be checked, directly at the center of the driven element if practicable, with a grid-dip meter. It is important that the resonant frequency of the antenna be at the *center* of the frequency band to be covered. If the resonant frequency is found to be much different from the desired frequency, the length of the driven element of the array should be altered until this condition exists. A relatively small change in the length of the driven element will have only a second order effect on the tuning of the parasitic elements of the array. Hence, a moderate change in the length of the driven element may be made without repeating the tuning process for the parasitic elements.

When the resonant frequency of the antenna system is correct, the antenna transmission line, with the calculated value of impedance-matching transformer or network between the line and antenna feed point, is then attached to the array and coupled to a low-power exciter unit or transmitter. Then, preferably, a standing-wave meter is connected in series with the antenna transmission line at a point relatively much more close to the transmitter than to the antenna. However, for best indication there should be 10 to 15 feet of line between the transmitter and the standing-wave meter. If a standing-wave meter is not available the standing-wave ratio may be checked approximately by means of a neon lamp or a short fluorescent tube if twin transmission line is being used, or it may be checked with a thermomilliammeter and a loop, a neon lamp, or an r-f ammeter and a pair of clips spaced a fixed distance for clipping onto one wire of a two-wire open line.

If the standing-wave ratio is below 2 to 1 it is satisfactory to leave the installation as it is. If the ratio is greater than this range it will be best when twin line or coaxial line is being used, and advisable with open-wire line, to attempt to decrease the s.w.r.

The condition of the match may be checked in the following manner with open line: measure the current in one leg of the feeder starting directly at the point where the antenna transmission line connects to the antenna (or antenna matching system such as the delta, "T," or yoke) and check the current values as you proceed toward the transmitter. If the current *increases* as you proceed away from the antenna the feed-point impedance is *higher* than that of the transmission line. If the current *decreases* as you proceed toward the transmitter the feed-point impedance is *lower* than the characteristic impedance of the transmission line. The ratio of the current maximum to the current minimum should be noted as the standing-wave ratio *r*.

Since the points of minimum current can be found with more accuracy than the points of maximum current, measure the distance from the antenna end of the feed line to the first current minimum. If this distance is one-half wavelength the feed line is operating into a pure resistance of *r* times the characteristic impedance of the feed line; in other words, the load impedance which the transmission line "sees" is resistive, meaning that the antenna is resonant, and is greater than the characteristic impedance of the transmission line by the value of the standing-wave ratio. If the distance to the first minimum is one-quarter wavelength the antenna is also resonant and is presenting a resistive load to the line but the value of this resistance is equal to the characteristic impedance of the transmission line *divided* by r (the s.w.r. on the line). If the distance from the antenna end of the transmission line to the first current minimum is any value *other* than ½ wave or ¼ wave the antenna system is presenting a reactive load to the antenna transmission line; hence the antenna should be retuned to resonance before proceeding further.

Assuming that the antenna system is presenting a resistive load of the wrong value to the antenna feed line, and assuming also that all adjustments have been made on the delta match, "T" match, yoke match, or folded-element match in an effort to obtain the lowest possible value of standing-wave ratio, the easiest procedure for obtaining an accurate impedance match is to insert a quarter-wave transformer into the feed line. The feed line should be cut at a point of current *maximum;* the points of current maximum are of course ¼ wave in either direction from the more easily measured points of current minimum. If the current minimum is one-quarter wave from the antenna the current maximum will be directly at the point where the feed line connects to the antenna system. If the current *minimum* is one-half wave from the end of the transmission line the feeder should be cut *one-quarter* wave from the antenna since this will be a point of current maximum. A quarter-wave section of transmission line should then be inserted at this point. The impedance of this quarter-wave section should be equal to the geometric mean between the characteristic impedance of the main feed line and the impedance that the feed line is "seeing" at the point where it was cut. However, the required impedance for the feed line can also be determined very easily from the standing-wave ratio r and the characteristic impedance of the main antenna feed line Z_0. The expression for determining the proper value of impedance for the quarter-wave section, Z_Q, is as follows:

$$Z_Q = \sqrt{\frac{Z_0^2}{r}}$$

To take an example, suppose a 465-ohm line constructed of no. 12 wire on 2" feeder spreaders is being used to feed a rotary affair with "T" match. The first current minimum on the feed line came just one-quarter wave from the "T" and the standing-wave ratio of current measured on the line was found to be 4-to-1. The proper impedance of the quarter-wave section Z_Q would then be:

$$Z_Q = \sqrt{\frac{(465)^2}{4}} = \sqrt{54{,}056} = 232 \text{ ohms}$$

Since the first current minimum came one-quarter wave from the "T," the first current maximum is right at the "T" so the 465-ohm feed line should be removed from the "T" and a quarter-wave section of 232-ohm transmission line inserted. This line may be made up in the same manner as a set of "Q" bars with ½-inch aluminum tubing spaced 1¾", or a four-wire line, as described in Chapter Twelve, may be made up with four no. 14 wires equally spaced around a 3.44" circle; the spacing between wires on the corners would be 2-7/16 inches.

Matching When the Feed Point is Reactive If the antenna is presenting a reactive load to the transmission line at its feed point, an attempt should first be made to make the antenna perfectly resonant. This will probably involve a slight adjustment in the length of the driven element in a parasitic array, or a change in the dimensions of all the driven elements in a stacked array. If the reactance cannot be eliminated the antenna system may still be made to present a resistive load to the transmission line in the following manner: Measure the standing-wave ratio r as before. Locate the current *maximum* on the feed line which is closest to the antenna. As before the current maximums will be located one-quarter wave from the current minimums, and when the approximate location of the current maximum has been found, determine accurately the location of the closest current minimum and measure one-quarter wave from the minimum to obtain the accurate location of the current maximum. Cut the main antenna transmission line at the point of this current *maximum,* and insert a quarter-wave section of transmission line having a characteristic impedance determined in the same manner from the Z_0 of the main transmission line and the s.w.r. r as discussed in the previous paragraph. In other words, the antenna system is presenting a pure resistive load at a point of current maximum on the main feed line so we can use the same procedure, with this point as reference, as was discussed in the previous paragraph.

Matching with Feeder other than Open-Wire Line The problem of matching the feed line to the antenna is simplest when using open-wire line. When twin line is being used it is sometimes possible to obtain a satisfactory indication of the relative current values in the line through the use of a flat loop of wire about 6" long attached to a thermomilliammeter, with the straight part of the loop placed on the side of the twin line directly alongside one of the conductors. A relative indication may be obtained through the use of a fluorescent tube placed against the flat of the line. In this case the length of the glow will be approximately equal to the voltage between the two conductors making up the line.

When coaxial cable is being used as feed line on the v-h-f and u-h-f bands a slotted line may be used to measure standing-wave ratio and the position of the current maxima and minima. But the use of a slotted line is impracticable due to the size required on frequencies below perhaps 144 Mc.

Standing-wave meters may be used with coaxial line, twin line, or open-wire line. These instruments are very satisfactory for determining the magnitude of the s.w.r., but they do not indicate the position of the current maximums and minimums on the line. Sometimes a s-w-r meter may be used in conjunction with a fluorescent tube in the tuning process with ribbon-type line. The fluorescent tube will indicate the position of the voltage maxima and minima and the s-w-r meter will give the magnitude of the s.w.r. on the line, thus giving all the information necessary to obtain an accurate match between the antenna system and the transmission line.

Raising and Lowering the Array A practical problem always present when tuning up and matching an array is the physical location of the structure. If the array is atop the mast it is inaccessible for adjustment, and if it is located on step-ladders where it can be adjusted easily it cannot be rotated. One encouraging factor in this situation is the fact that experience has shown that if the array is placed 6 or 8 feet above ground on some step-ladders for the preliminary tuning process, the raising of the system to its full height will not produce a serious change in the adjustments. So it is usually possible to make preliminary adjustments with the system located slightly greater than head height above ground, and then to raise the antenna to a position where it may be rotated for final adjustments. If the position of the sliding sections as determined near the ground is marked so that the adjustments will not be lost, the array may be raised to rotatable height and the fastening clamps left loose enough so that the elements may be slid in by means of a long bamboo pole. After a series of trials a satisfactory set of lengths can be obtained. But the end results usually come so close to the figures given in figure 2 that a subsequent array is usually cut to the dimensions given and installed as is.

The matching process does not require rotation, but it does require that the antenna proper be located at as nearly its normal operating position as possible. However, on a particular installation the positions of the current minimums on the transmission line near the transmitter may be checked with the array in the air, and then the array may be lowered to ascertain whether or not the positions of these points have moved. If they have not, and in most cases if the feeder line is strung out back and forth well above ground as the antenna is lowered they will not change, the positions of the last few toward the antenna itself may be determined. Then the calculation of the matching quarter-wave section may be made, the section installed, the standing-wave ratio again checked, and the antenna re-installed in its final location.

16-7 Antenna Rotation Systems

Structures for the rotation of antenna arrays may be sub-divided into two general classes: the rotating mast and the rotating platform. The rotating-mast type of structure is probably the most satisfactory arrangement for a home constructed system.

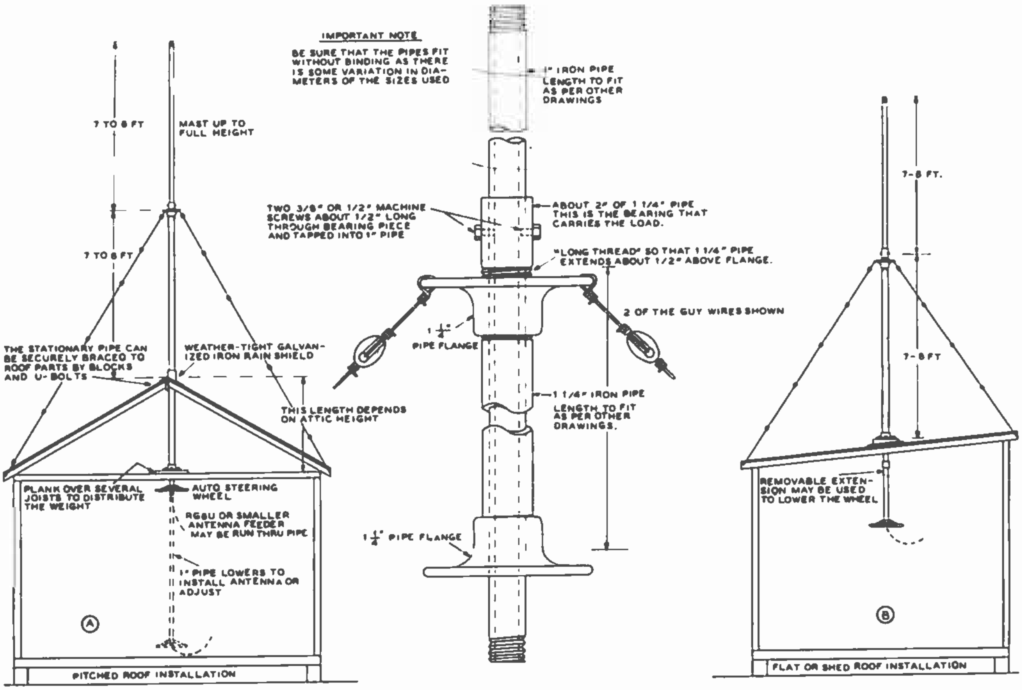


Figure 28.

ALL-PIPE ROTATING-MAST STRUCTURE FOR ROOF INSTALLATION.

An installation suitable for a building with a pitched roof is shown at (A). At (B) is shown a similar installation for a flat or shed roof. The arrangement as shown is strong enough to support a lightweight 3-element 28-Mc. array and a light 3-element 50-Mc. array above the 28-Mc. array on the end of a 4-foot length of ½-inch pipe.

The lengths of pipe shown were chosen so that when the system is in the lowered position one can stand on a household ladder and put the beam in position atop the rotating pipe. The lengths may safely be revised upward somewhat if the array is of a particularly lightweight design with low wind resistance. The several types of antennas shown in this chapter with pipe flange mounting provision all have been used with this type of mast and rotating structure.

Just before the mast is installed it is a good idea to give the rotating pipe a good smearing of cup grease or waterproof pump grease. To get the lip of the top of the stationary section of 1¼-inch pipe to project above the flange plate, it will be necessary to have a plumbing shop cut a slightly deeper thread inside the flange plate, as well as cutting an unusually long thread on the end of the 1¼-inch pipe. It is relatively easy to waterproof this assembly through the roof since the 1¼-inch pipe is stationary at all times. Be sure to use pipe compound on all the joints and then really tighten these joints with a pair of pipe wrenches.

The rotating-platform system is probably best if the rotating device is to be purchased. A number of excellent rotating platform devices are available on the market for varying prices. The larger and more expensive rotating devices are suitable for the rotation of a rather sizeable array for the 14-Mc. band while the smaller structures, such as those designed for rotating a TV antenna are designed for less load and should be used only with a 28-Mc. or 50-Mc. array. Most common practice is to install the rotating device atop a platform built at the top of a telephone pole or on the top of a lattice mast of sizeable cross section so that the mast will be self-supporting and capable of withstanding the torque imposed upon it by the rotating platform.

Turning the Rotating Structure If the rotating-mast type of structure is used it is a relatively simple process to work out a drive method. If the rotating mast comes down through the top of the radio shack as does the mast of the antenna

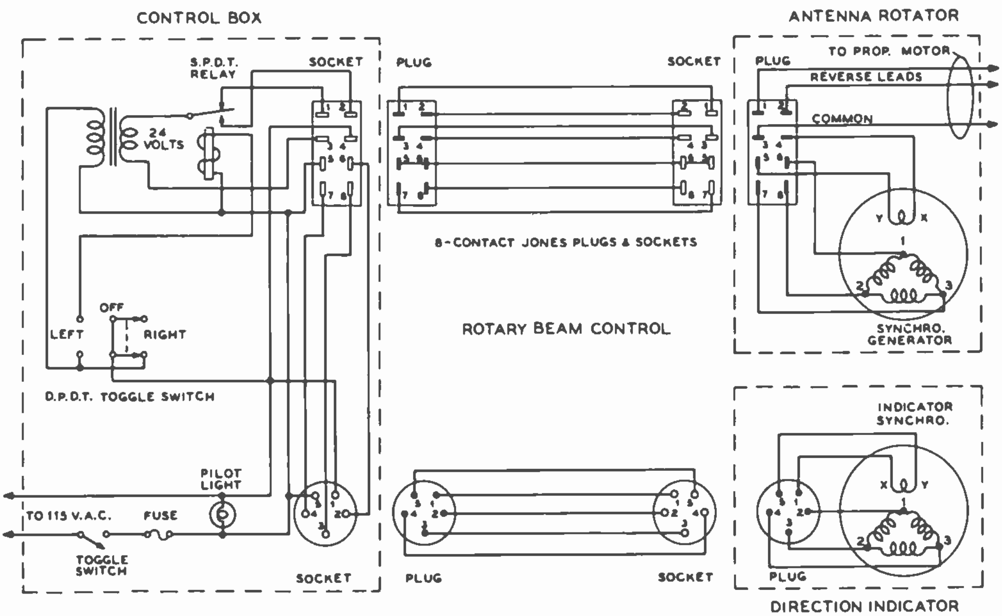


Figure 29.
SCHEMATIC OF THE COMPLETE ANTENNA CONTROL SYSTEM.

system shown in figure 23, a very satisfactory arrangement is merely to install a large steering wheel on the bottom of the rotating mast, with the thrust bearing for the structure located above the roof. This system is sketched in figure 28. In the example shown a piece of 1¼-inch pipe is fixed permanently to the roof of the shack and the rotary array is mounted on a piece of 1-inch pipe which comes down into the shack through the larger diameter pipe.

If the rotating mast is located a reasonable distance from the operating position a system of pulleys and ropes may be used to drive the antenna. Of course the most satisfactory drive system is that which uses an electric motor.

A system which is widely used for driving rotatable antennas for radar work uses a *servomechanism* for accomplishing the actual rotation of the antenna. In this type of system a small *synchro* (or Selsyn) motor (5G for example) is coupled to a handwheel and an indicator dial. The output of the control synchro is coupled to another synchro of special design (5CT for example) in the base of the pedestal or rotating structure of the antenna in such a manner that any error between the relative position of the two synchros is coupled into an electronic device called a servoamplifier. The output of the servoamplifier system is then fed to a motor in the base of the pedestal of the antenna. The polarity and strength of the energy fed from the servoamplifier to the driving motor are such that the motor will turn the rotating structure until the synchro (which is geared to the rotating platform) is in the same relative position as the rotor of the synchro which is geared to the handwheel. This type of drive for a rotatable antenna is ideal since only the slightest force on the handwheel is adequate to position even the largest rotatable antenna to the exact direction indicated by the pointer on the driving synchro. However, such systems are complex and expensive. Nevertheless a number of the components for such a driving system are available on the surplus market and may be adapted to the job of driving an amateur rotatable array.

Indication of Direction The most satisfactory method for indicating the direction of transmission of a rotatable array is that which uses Selsyns or synchros for the transmission of the data from the rotating structure to the indicating pointer at the operating position. A number of synchros and Selsyns of various types are available on the surplus market. Some of them are designed for operation on 115 volts at 60 cycles, some are designed for operation on 60 cycles but at a lowered voltage, and some are designed for operation from 400-cycle or 800-cycle energy. This latter type of high-frequency synchro is the most generally available type, and the high-frequency units are smaller and lighter than the 60-cycle units. Since the indicating synchro must deliver an almost negligible amount of power to the pointer which it drives, the high-frequency types will operate quite satisfactorily from 60-cycle power if the voltage on them is reduced to somewhere between 6.3 and 20 volts. In the case of many of many of the units available, a connection sheet is provided along with a recommendation in regard to the operating voltage when they are run on 60 cycles. In any event the operating voltage should be held as low as it may be and still give satisfactory transmission of data from the antenna to the operating position. Certainly it should not be necessary to run such a voltage on the units that they become overheated.

Systems using a potentiometer capable of continuous rotation and a milliammeter, along with a battery or other source of direct current, may also be used for the indication of direction. A commercially-available potentiometer (Ohmite RB-2) may be used in conjunction with a 0-1 d-c milliammeter having a hand-calibrated scale for direction indication.

Rotary-Beam Control System

The following description gives the circuit details and some of the mechanical considerations of a more or less standard rotary-beam antenna control system. The installation uses the common propeller pitch-change motor as the driving motor and reduction gear. A pair of 115-volt synchros (Selsyns) repeat back antenna position information to the operating position.

The installation consists of three units: the motor and synchro generator at the base of the antenna, the control box, and the rotating-globe direction indicator. An 8-wire cable runs a distance of about 50 feet from the control box to the base of the antenna, and a 5-wire cable runs from the control box to the direction indicator. Power for the installation is obtained through a standard line cord from the control box which plugs into an outlet.

The components which go to make up the installation are illustrated in figures 29, 30, and 31. The three-element 28-Mc. antenna is supported by a rotating mast of 1½-inch diameter steel tubing with a radial load bearing about 10 feet above the ground. The thrust

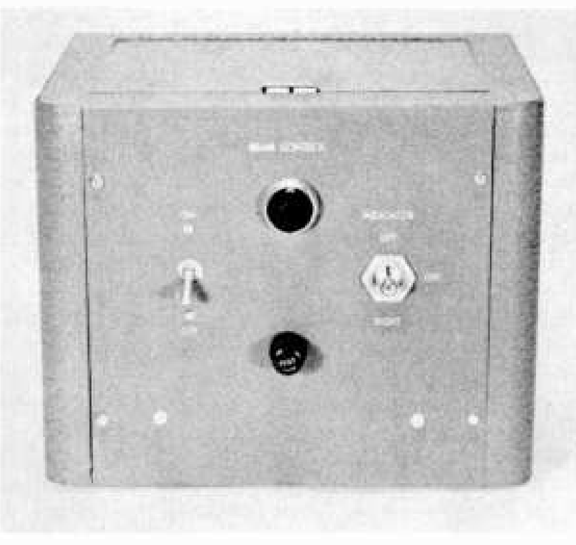


Figure 30.
CONTROL BOX FOR THE ANTENNA ROTATION SYSTEM.

of the antenna system is taken by the pitch-change mechanism at the base of the mast.

The rotating support mast for the array is constructed of sections from a surplus steel-tubing mast. The mast comes in 5-foot sections with the end 6 inches of each section reduced in diameter so that successive sections may be telescoped together. Quarter-inch holes are drilled through each junction between sections so that a bolt may be run through the junction. The bolt serves to tighten the junction and to pin the sections together so that torque from the rotator at the base will be transmitted to the antenna at the top.

The 1½-inch diameter of the base of the mast fits neatly in the center of the rotating portion of the pitch-change motor. Also, holes are conveniently provided in this portion of the pitch-change motor for pinning to the base of the mast, by drilling matching holes in the bottom section of tubing.

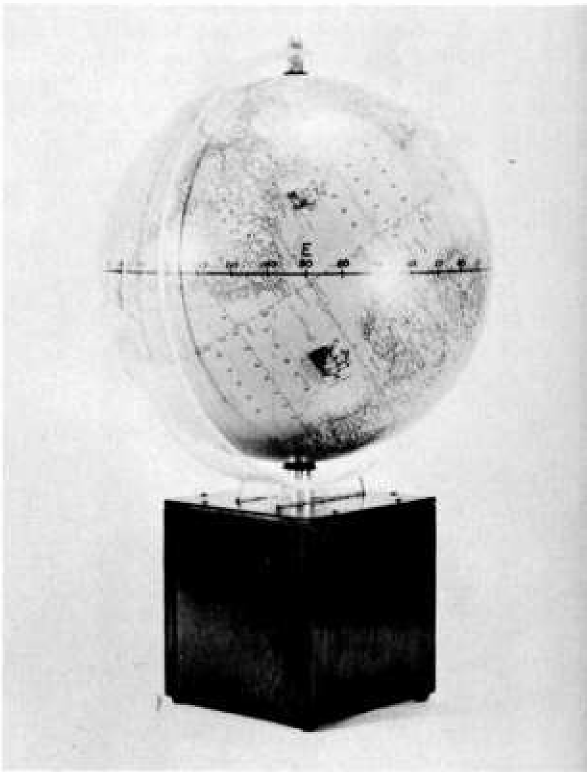
The Control Box A photograph of the control box is given in figure 30, while the schematic diagram of the control box is given at the left in figure 29. The control box includes a main switch, fuse, pilot light, and the direction control for antenna rotation. When the main switch is turned on the pilot lamp lights and primary voltage is applied to the two synchro units in the indicator system. The two synchros tend to pull into step as soon as voltage is applied to their primary windings. The rotor of the synchro at the base of the antenna is fixed, since it is geared to the rotating mast of the antenna, so the synchro which is connected to the rotating globe indicator will rotate until its rotor is in the same relative position as the rotor of the synchro at the base of the antenna.

The direction control is a three-way toggle switch. It is spring loaded so that as soon as the force is released the switch returns to the off position. Note that the switch contacts are not required to carry the rather heavy current in the 24-volt a-c line. One section of the switch applies voltage to the primary of the 24-volt transformer in either of the LEFT (counter-clockwise) or the RIGHT (clockwise) positions. The second section of the switch actuates a s.p.d.t. relay with the 115-volt line in the LEFT position, but leaves the relay unactuated in the RIGHT position. In the unactuated position the relay feeds current to the RIGHT (clockwise) lead of the motor; the common lead of the motor being connected at all times. When the relay is actuated by placing the control switch in the LEFT position, current is fed to the LEFT lead of the motor.

Cables Cables are run between the major units of the control and indicator system. The use of cables and plugs allows rapid and convenient interconnection or separation of the units. An 8-wire cable with 8-contact plugs at each end, runs from the control box to the antenna rotator. The plugs and receptacles are Jones P-408 series. Similarly, a 5-wire cable is run from the control box to the rotating-globe direction indicator. Amphenol type 86 series plugs and connectors are used on the indicator cable.

Note that the "hot" receptacles or ends of cables always are terminated in female plugs

Figure 31.
THE ROTATING-GLOBE DIRECTION INDICATOR.



or receptacles, while the "cold" ends always are terminated in the appropriate male fittings. Through this procedure in the use of cable connectors it is always safe to handle a cable or a connector without the danger of electrical shock.

The Rotating-Globe Indicator The use of a rotating-globe type of direction indicator is convenient in that it immediately gives an idea of the optimum direction and of the coverage of the directional antenna. Also, the indication is given in terms of the familiar globe rather than in terms of a great-circle projection with its attendant distortion in the relative areas of the land and ocean masses.

The unit illustrated in figure 31 was constructed from a 12-inch globe obtained from a book store. The globe was dismounted from its carrier and a small hole drilled at the home city of the station. Then another small hole was drilled at exactly the opposite point on the globe, and both holes enlarged to clear a ¼-inch screw. An adapter 1-inch in diameter and about 1½ inches long was machined of dural rod by a machine shop. One end of the adapter was machined to take the shaft of the synchro motor and the other end was bored to take a ¼-inch rod. Provision for set screws at each end of the adapter was made by the machine shop.

The shaft for the rotating globe is a 14-inch length of ¼-inch welding rod, with one end of the rod threaded about 1½ inches. The shaft then was assembled to the globe with rubber washers at each end, using shellac to cement the rubber washers to the globe.

The mounting box for the globe, which has the synchro motor mounted inside and the cable fitting on the rear, was constructed of ¾-inch plywood. A false bottom inside the mounting box holds the flange for mounting the synchro motor. The marker and top bearing is made of ¼-inch Plexiglas. The Plexiglas first was cut and filed to shape, and the hairline marker drawn. Then the material was heated in the oven and bent to shape around a drum which worked out to give the correct clearance of about one-half inch between the marker and the globe. Holes were drilled in the completed marker and the base was cemented to the marker with Plexiglas cement.

Alignment The alignment of the indicator system was accomplished by pointing the antenna at the north star, with the voltage applied to both synchros. Then the set screw in the adapter which holds the motor shaft was loosened and the globe rotated by hand until the marker aligned with north. Then the set screw was tightened. The two synchros used in the indicator system are of the 115-volt 60-cycle type.

Television and Broadcast Interference

The problem of interference to television reception is one of the most difficult and most serious ever to confront the radio amateur. But the problem can be solved if tackled in an orderly manner, as is attested by the fact that thousands of amateurs have eliminated the TV interference originally caused by their transmitters.

In an area of high TV-signal field intensity the TVI problem is capable of complete solution with the taking of routine measures both at the amateur transmitter and at the affected receivers. But in fringe areas of low TV-signal field strength the complete elimination of TVI is a difficult and challenging problem. But, as is attested by the work of W1DBM and many others, it still is a problem capable of solution. An extensive bibliography of material pertaining to TVI is included at the end of this chapter.

17-1 Types of Television Interference

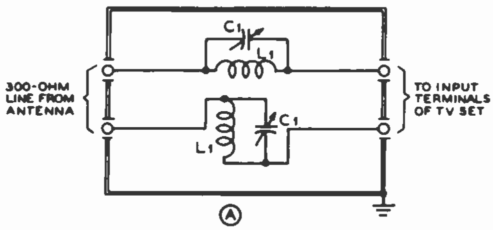
There are three main types of TVI which may be caused singly or in combination by the emissions from an amateur transmitter. These types of interference are:

(1) Overloading of the TV set by the transmitter fundamental
(2) Impairment of the picture by spurious emissions
(3) Impairment of the picture by the radiation of harmonics

TV Set Overloading Even if the amateur transmitter were perfect and had no harmonic radiation or spurious emissions whatever, it still would be likely to cause overloading of TV sets whose antennas were within a few hundred feet of the transmitting antenna. This type of overloading is essentially the same as the common type of BCI encountered when operating a medium-power or high-power amateur transmitter within a few hundred feet of the normal type of BCL receiver. The field intensity in the immediate vicinity of the transmitting antenna is sufficiently high that the amateur signal will get into the BC or TV set either through overloading of the front end, or through the i-f, video, or audio system. A characteristic of this type of interference is that it always will be eliminated when the transmitter temporarily is operated into a dummy antenna. Another characteristic of this type of overloading is that its effects will be substantially continuous over the entire frequency coverage of the BC or TV receiver. Channels 2 through 13 will be affected in approximately the same manner.

With the overloading type of interference the problem is simply to keep the *fundamental* of the transmitter out of the affected receiver. Other types of interference may or may not show up when the fundamental is taken out of the TV set (they probably will appear), but at least the fundamental *must* be eliminated first.

The elimination of the transmitter fundamental from the TV set is normally the only



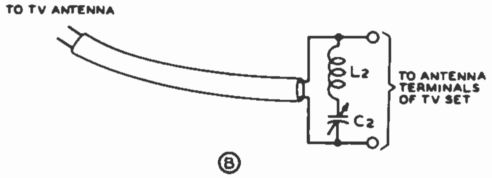


Figure 1.

TUNED TRAPS FOR THE TRANSMITTER FUNDAMENTAL.

The arrangement at (A) has proven to be effective in eliminating the condition of general blocking as caused by a 28-Mc. transmitter in the vicinity of a TV receiver. The tuned circuits L_1-C_1 are resonated separately to the frequency of transmission. The adjustment may be done at the station, or it may be accomplished at the TV receiver by tuning for minimum interference on the TV screen.

Shown at (B) is an alternative arrangement with a series-tuned circuit across the antenna terminals of the TV set. The tuned circuit should be resonated to the operating frequency of the transmitter. This arrangement gives less attenuation of the interfering signal than that at (A); the circuit has proven effective with interference from transmitters on the 50-Mc. band, and with low-power 28-Mc. transmitters.

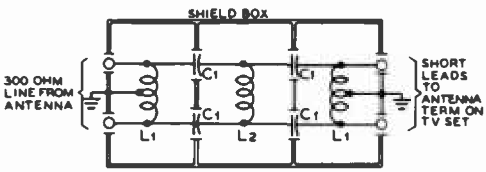
operation performed on or in the vicinity of the TV receiver. After the fundamental has been eliminated as a source of interference to reception, work may then be begun on or in the vicinity of the transmitter toward eliminating the other two types of interference.

Taking Out the Fundamental More or less standard BCI-type practice is most commonly used in taking out fundamental interference. Wavetraps and filters are installed, and the antenna system may or may not be modified so as to offer less response to the signal from the amateur transmitter. In regard to a comparison between wavetraps and filters, the same considerations apply as have been effective in regard to BCI for many years; wavetraps are quite effective when properly installed and adjusted, but they must be readjusted whenever the band of operation is changed, or even when moving from one extreme end of a band to the other. Hence, wavetraps are not recommended except when operation will be confined to a relatively narrow portion of one amateur band. However, figure 1 shows two of the most common signal trapping arrangements.

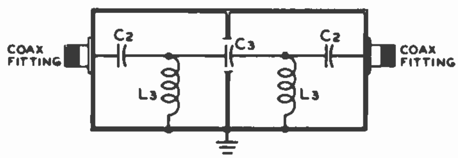
High-Pass Filters High-pass filters in the antenna lead of the TV set have proven to be quite satisfactory as a means of eliminating TVI of the overloading type. In many cases when the interfering transmitter is operated only on the bands below 7.3 Mc., the use of a high-pass filter in the antenna lead has completely eliminated all TVI. In some cases the installation of a high-pass filter in the antenna transmission line and an a-c line filter of a standard variety has proven to be completely effective in eliminating the interference from a transmitter operating in one of the lower frequency amateur bands.

In general, it is suggested that commercially manufactured high-pass filters be purchased. Such units are available from a number of manufacturers at a relatively moderate cost. However, such units may be home constructed; suggested designs are given in figures 2 and 3. Types for use both with coaxial and with balanced transmission lines have been shown. In most cases the filters may be constructed in one of the small shield boxes which are now on the market. Input and output terminals may be standard connectors, or the inexpensive type of terminal strips usually used on BC and TV sets may be employed. Coaxial terminals should of course be employed when a coaxial feed line is used to the antenna. In any event the leads from the filter box to the TV set should be very short, including both the antenna lead and the ground lead to the box itself. If the leads from the box to the set have much length, they may pick up enough signal to nullify the effects of the high-pass filter.

Blocking from 50-Mc. Signals Operation on the 50-Mc. amateur band in an area where channel 2 is in use for TV imposes a special problem in the



(A) FOR 300-OHM LINE, SHIELDED OR UNSHIELDED



(B) FOR 50-75 OHM COAXIAL LINE

Figure 2.
HIGH-PASS TRANSMISSION LINE FILTERS.

The arrangement at (A) will stop the passing of all signals below about 45 Mc. from the antenna transmission line into the TV set. Coils L_1 are each 1.2 microhenrys (17 turns no. 24 enam. closewound on 1/4-inch dia. polystyrene rod) with the center tap grounded. It will be found best to scrape, twist, and solder the center tap before winding the coil. The number of turns each side of the top may then be varied until the tap is in the exact center of the winding. Coil L_2 is 0.6 microhenry (12 turns no. 24 enam. closewound on 1/4-inch dia. polystyrene rod). The capacitors should be about 16.5 μμfd., but either 15 or 20 μμfd. ceramic capacitors will give satisfactory results. A similar filter for coaxial antenna transmission line is shown at (B). Both coils should be 0.12 microhenry (7 turns no. 18 enam. spaced to 1/2 inch on 1/4-inch dia. polystyrene rod). Capacitors C_2 should be 75 μμfd. midget ceramics, while C_3 should be a 40-μμfd. ceramic.

matter of blocking. The input circuits of most TV sets are sufficiently broad so that an amateur signal on the 50-Mc. band will ride through with little attenuation. Also, the normal TV antenna will have a quite large response to a signal in the 50-Mc. band since the lower limit of channel 2 is 54 Mc.

High-pass filters of the normal type simply are not capable of giving sufficient attenuation to a signal whose frequency is so close to the necessary pass band of the filter. Hence, a resonant circuit element, as illustrated in figure 1, must be used to trap out the amateur field at the input of the TV set. The trap must be tuned or the section of transmission line cut, if a section of line is to

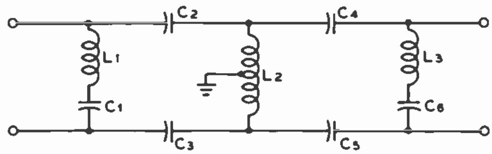


Figure 3.
SERIES-DERIVED HIGH-PASS FILTER.

This filter is designed for use in the 300-ohm transmission line from the TV antenna to the TV receiver. Nominal cutoff frequency is 36 Mc. and maximum rejection is at about 29 Mc.
C_1,C_6—15-μμfd. zero-coefficient ceramic
C_2,C_3,C_4,C_5—20-μμfd. zero-coefficient ceramic
L_1,L_3—2.0 μh. About 24 turns no. 28 d.c.c. wound to 5/8" on 1/4" diameter polystyrene rod. Turns should be adjusted until the coil resonates to 29 Mc. with the associated 15-μμfd. capacitor.
L_2—0.66 μh., 14 turns no. 28 d.c.c. wound to 5/8" on 1/4" dia. polystyrene rod. Adjust turns to resonate externally to 20 Mc. with an auxiliary 100-μμfd. capacitor whose value is accurately known.

be used for a particular frequency in the 50-Mc. band—and this frequency will have to be near the lower frequency limit of the 50-Mc. band to obtain adequate rejection of the amateur signal while still not materially affecting the response of the receiver to channel 2.

Elimination of Spurious Emissions All spurious emissions from amateur transmitters (ignoring harmonic signals for the time being) must be eliminated to comply with FCC regulations. But in the past many amateur transmitters have emitted spurious signals as a result of key clicks, parasitics, and overmodulation transients. In most cases the operators of the transmitters were not aware of these emissions since they were radiated only for a short distance and hence were not brought to his attention. But with one or more TV sets in the neighborhood it is probable that such spurious signals will be brought quickly to his attention.

The elimination of parasitic oscillation and overmodulation transients has been discussed in detail in Chapter Eleven with further information given in Chapters Five and Ten.

Two new wrinkles recently suggested for the elimination of self-oscillation and parasitic oscillation from beam tetrode amplifier stages are illustrated in figure 4. The use of a short section of coaxial transmission line as an additional means of holding the screen at ground potential is shown in figures 4A and 4B. This measure has proven effective, with a 6 to 15 inch length of coaxial cable at Z, in the taming of a single-ended 807 stage operating in the v-h-f range. A means of raising the resonant frequency of the grid circuit for eliminating parasitics is shown at figure 4C.

Harmonic Radiation After any condition of blocking at the TV receiver has been eliminated, and when the transmitter is completely free of transients and parasitic oscillations, it is probable that TVI will be eliminated in certain cases. Certainly general interference should be eliminated, particularly if the transmitter is a well designed affair operated on one of the lower frequency bands, and the station is in a high-signal TV area. But when the transmitter is to be operated on one of the higher frequency bands, and particularly in a marginal TV area, the job of TVI-proofing will just have begun. The elimination of harmonic radiation from the transmitter is a difficult and tedious job which must be done in an orderly manner if completely satisfactory results are to be obtained.

First it is well to become familiar with the TV channels presently assigned, with the TV intermediate frequencies commonly used, and with the channels which will receive interference from harmonics of the various amateur bands. Figures 5 and 6 give this information.

Even a short inspection of figures 5 and 6 will make obvious the seriousness of the interference which can be caused by harmonics of amateur signals in the higher frequency bands. With any sort of reasonable precautions in the design and shielding of the transmitter it is not likely that harmonics higher than the 6th will be encountered. Hence the main offenders in the way of harmonic interference will be those bands above 14-Mc.

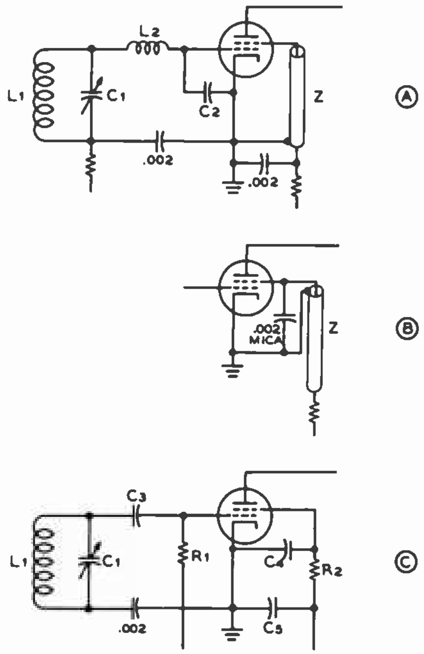


Figure 4.
ANTI-PARASITIC CIRCUIT SUGGESTIONS.

Shown at (A) is the combination of a low-pass filter in the grid circuit plus a de-tuning trap in the screen return. C_2 should be a small ceramic capacitor of about 15 µµfd. connected directly from grid to cathode on the tube socket. A starting value for L_2 is 6 turns of no. 18 enam. wound on a 1/4-inch diameter insulating rod. The largest inductance which will still permit full excitation on the highest frequency band should be employed. Suggestions for the short section of coaxial line at Z as a means of de-tuning the screen circuit are included in the text. At (C) is shown the use of a 15-µµfd. series ceramic capacitor as C_3 for raising the resonant frequency of the grid circuit as a means of stopping parasitic oscillations. Also shown at (C) is the use of a double screen by-pass system. Capacitor C_4 may be a small ceramic trimmer capacitor of about 50 µµfd. maximum, R may be a 10-ohm 2-watt carbon, while R_1 is the conventional grid leak for the stage. Capacitor C_5 is the conventional 0.002-µfd. to 0.01-µfd. mica.

Nature of Harmonic Interference Investigations into the nature of the interference caused by amateur signals on the TV screen, assuming that blocking has been eliminated as described earlier in this chapter, have revealed the following facts:

1. An unmodulated carrier, such as a c-w signal with the key down or an AM signal without modulation, will give a cross-hatch or herringbone pattern on the TV screen. This same general type of picture also will occur in the case of a narrow-band FM signal either with or without modulation.
2. A relatively strong AM signal will give in addition to the herringbone a very serious succession of light and dark bands across the TV picture.

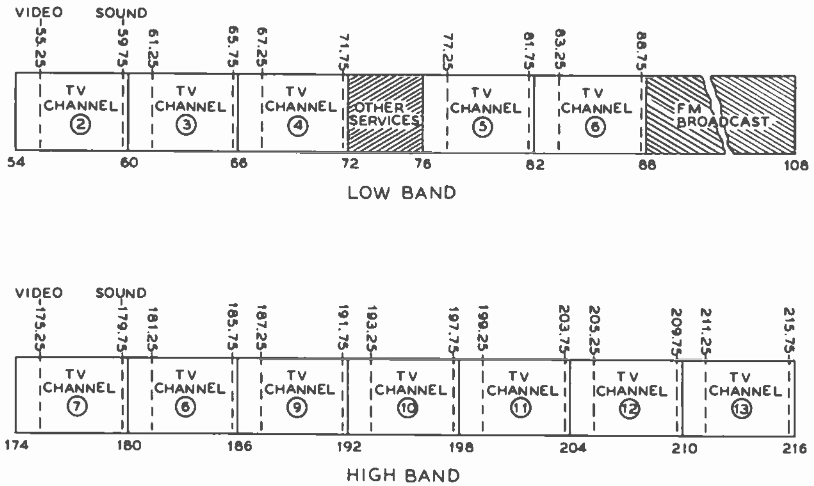


Figure 5.
FREQUENCIES OF THE V-H-F TV CHANNELS.
Showing the frequency ranges of TV channels 2 through 13, with the picture carrier and sound carrier frequencies also shown.

3. A moderate strength c-w signal without transients, in the absence of overloading of the TV set, will result merely in the turning on and off of the herringbone on the picture.

To discuss condition (1) above, the herringbone is a result of the beat note between the TV video carrier and the amateur harmonic. Hence the higher the beat note the less obvious will be the resulting cross-hatch. Further, it has been shown that a much stronger signal is required to produce a discernible herringbone when the interfering harmonic is as far away as possible from the video carrier, without running into the sound carrier. Thus, as a last resort, or to eliminate the last vestige of interference after all corrective measures have been taken, operate the transmitter on a frequency such that the interfering harmonic will fall as far as possible from the picture carrier. The worst possible interference to the picture from a continuous carrier will be obtained when the interfering signal is very close in frequency to the video carrier.

Isolating the Source of the Interference Throughout the testing procedure it will be necessary to have some sort of indicating device as a means of determining harmonic field intensities. The best indicator for field intensities some distance from the transmitting antenna will probably be the TV receiver of some neighbor with whom friendly relations are still maintained. This person then will be able to give a check, occasionally, on the relative nature of the interference. But it will probably be necessary to go and check yourself periodically the results obtained, since the neighbor probably will not be able to give any sort of a quantitative analysis of the progress which has been made.

An additional device for checking relatively high field intensities in the vicinity of the transmitter will be almost a necessity. A simple indicating wavemeter (made from a Silver 903 wavemeter, a 1N34, and a microammeter) will accomplish this function. Also, it will be very helpful to have a receiver, with an S meter, capable of covering at least the 50 to 100 Mc. range and preferably the range to 216 Mc. This device may consist merely of the station receiver and a simple converter using the two halves of a 6J6 as oscillator and mixer.

The first check can best be made with the neighbor who is receiving the most serious or the most general interference. Turn on

TRANSMITTER FUNDAMENTAL	2ND	3RD	4TH	5TH	6TH	7TH	8TH	9TH	10TH
7.0 – 7.3		21-21.9 TV I.F.			42-44 NEW TV I.F.		56-58.4 CHANNEL 2	63-65.7 CHANNEL 3	70-73 CHANNEL 4
14.0 – 14.4		42-43 NEW TV I.F.	56-57.6 CHANNEL 2	70-72 CHANNEL 4	84-86.4 CHANNEL 6	98-100.8 FM BROAD-CAST			
21.0 – 21.45 (TV I.F.)		63-64.35 CHANNEL 3	84-85.8 CHANNEL 6	105-107.25 FM BROAD-CAST				189-193 CHANNELS 9 10	210-214.5 CHANNEL 13
26.96 – 27.23	53.92-54.46 CHANNEL 2 ABOVE 27 MC ONLY	80.88-81.69 CHANNEL 5	107.84-108.92 FM BROAD-CAST			189 CHANNEL 9	216 CHANNEL 13		
28.0 – 29.7	56-59.4 CHANNEL 2	84-89.1 CHANNEL 6			168-178.2 CHANNEL 7	196-207.9 CHANNELS 10 11 12			
50.0 – 54.0	100-108 FM BROAD-CAST		200-216 CHANNELS 11 12 13				POSSIBLE INTERFERENCE TO U-H-F CHANNELS	450-486	500-540

Figure 6.

HARMONICS OF THE AMATEUR BANDS.

Shown are the harmonic frequency ranges of the amateur bands between 7 and 54 Mc., with the TV channels (and TV i-f systems) which are most likely to receive interference from these harmonics. Under certain conditions amateur signals in the 1.8 and 3.5 Mc. bands can cause interference as a result of direct pickup in the video systems of TV receivers which are not adequately shielded.

the transmitter and check all channels to determine the extent of the interference and the number of channels affected. Then disconnect the antenna and substitute a group of 100-watt lamps as a dummy load for the transmitter. Experience has shown that 8 100-watt lamps connected in two seriesed groups of four in parallel will take the output of a kilowatt transmitter on 28 Mc. if connections are made symmetrically to the group of lamps. Then note the interference. Now remove plate voltage from the final amplifier and determine the extent of interference caused by the exciter stages.

In the average case, when the final amplifier is a beam tetrode stage and the exciter is relatively low powered and adequately shielded, it will be found that the interference drops materially when the antenna is removed and a dummy load substituted. It will also be found in such an average case that the interference will stop when the exciter only is operating.

General Considerations for Reduced Harmonic Radiation

It might be well at this point to discuss the general conditions which will result in reduced harmonic generation in the amateur transmitter. In the first place, it is best to have only one high-level r-f range in the transmitter. This consideration has been discussed in some detail in Chapter Ten, but it should be obvious that if there is only one high-level stage in the transmitter and all low-level stages are shielded, there will be only one stage from which harmonic signals of any sizeable amplitude will be obtained.

Second, it is best whenever practicable that the high-level stage in the transmitter be operated in the class B region rather than class C. Contrary to popular opinion, the output stage in any amateur transmitter may be operated class B for all modes of transmission other than amplitude modulation of the output stage. A reduction in the conversion efficiency of the output stage of about 10 per cent will result from dropping the grid bias and excitation to obtain class B operation. But the small reduction in output power certainly will not be noticeable on the air, while the driving power requirements for the stage and the harmonic signal output of the stage will have been materially reduced.

If the output stage must be operated class C, for high-level amplitude modulation, and if a fairly sizeable driver stage must be used to obtain adequate excitation to the final amplifier, it is quite important that the driver be operated as a straight-through amplifier. Under no conditions should a high-level driver be operated as a frequency multiplier; the operation of a high-level driver as a frequency multiplier is the surest way of having a high level of harmonic signal on hand in the transmitter. Consequently, all frequency multiplication should be done in the low-level stages of the transmitter — preferably with relatively small receiving tubes operating well within their transmitting ratings.

17-2 Suppression of Harmonic Radiation

Much has been written in recent years about the problem of suppressing harmonic radiation from medium power transmitters operating in the frequency range covered by the h-f amateur bands. Many means for attacking the problem have been suggested. But with the passing of time and the gaining of experience it becomes increasingly apparent that harmonic interference may be eliminated in nearly every case by doing only two things:

(1) SHIELD
(2) FILTER.

The above statement may seem to be an oversimplification. But if the facts implied by the statement are kept in mind while attacking a harmonic-reduction problem, that problem will at least be clarified and much lost motion may be eliminated. By inference, the statement suggests that harmonic plate traps, special low-inductance by-pass capacitors from plate to ground, and other such means of reducing the generation of harmonics *within the transmitter,* are not necessary. This is not strictly true. But it is true in the great majority of cases that such means of reducing harmonic *generation* will *not* eliminate TVI of the type caused by harmonic radiation.

Means of reducing harmonic *generation* may accomplish a reduction of 10, 20, or even 40 db in the level of harmonic energy within the transmitter enclosure. But such a level of harmonic reduction will seldom eliminate harmonic-type TVI; much greater degree of harmonic attenuation normally is required. Hence, it is much more appropriate, and more straightforward, to attack the harmonic reduction problem from the standpoint of *shielding* and *filtering*. In the great majority of cases the shielding of the equipment and the filtering of all leads into the transmitter enclosure will afford sufficient attenuation of harmonic radiation. In those special cases where *additional* harmonic reduction is required, means which reduce the generation of harmonics within the transmitter will be of assistance. But it is better to shield and filter in the first place, since these measures will be required in any event.

Transmitter Power Level It should be made clear at this point that the level of power used at the transmitter is not of great significance in the basic harmonic reduction problem. The difference in power level between a 20-watt transmitter and one rated at a kilowatt is only a matter of about 17 db. Yet the degree of harmonic attenuation required to eliminate interference caused by harmonic radiation is from 80 to 120 db, depending upon the TV signal strength in the vicinity. This is not to say that it is not a simpler job to eliminate harmonic interference from a low-power transmitter than from a kilowatt equipment. It is simpler to suppress harmonic radiation from a low-power transmitter simply because it is a much easier problem to shield a low-power unit, and the filters for the leads which enter the transmitter enclosure may be constructed less expensively and smaller for a low-power unit.

Shielding The matter of shielding is one about which a fair amount of information has been published recently. Certainly it is impossible to enclose the transmitter and exciter completely in a soldered copper box. The question then becomes, how much deviation from the ideal case may be tolerated before the shielding loses its effectiveness. The answer must, of course, be given in relative terms. But it is safe to say that

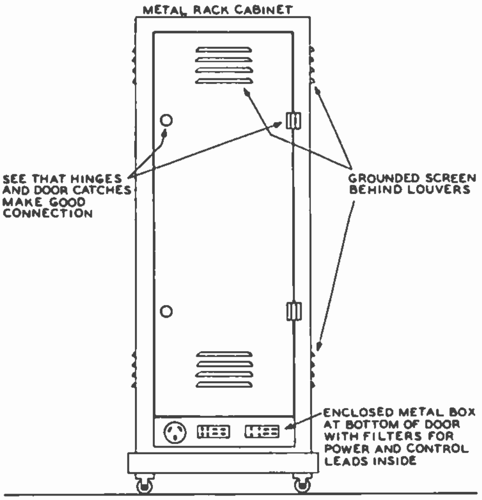


Figure 7.

SHIELDING MEASURES FOR THE RACK-ENCLOSED TRANSMITTER.

It may be found that the hinges and door catches do not offer good enough bonding of the rear door to the transmitter frame. In such an event a number of wiping contacts may be installed around the inside of the rear opening in such a position that the contacts will bear against scraped regions on the rear door. The use of shielding braid on the hinged edge, and thumbscrews on the other three sides, also may be found necessary.

slots, holes, and louvers in the shielding are most likely to be found the offenders as far as harmonic leakage is concerned.

It is important to remember that a slot in a piece of metal is exactly as good a radiator as a piece of wire having the dimensions of the slot. This may be a new concept to some, but the slot between the rear door and the housing of a 6-foot enclosed relay rack can radiate *horizontally* polarized waves in the vicinity of 80 Mc. (channel 5) as effectively as a horizontal half-wave dipole. Several hinges or door catches will serve to break up the slot, thus raising its resonant frequency out of the low TV band at least. Louvers in the side of the cabinet can be effective radiators in the vicinity of the high-band TV range. The soldering or bolting of copper screen in place behind the louvers will be found effective in completing the shielding.

Meter faces and glass-front meter panels can be effective v-h-f radiators. Meters should not be installed in such a posiiton that they can destroy the shielding of a compartment. The leads which connect to the meters should be carefully filtered, a by-pass capacitor should be placed across the meter terminals, and the meters should be shielded either in front of the face or behind the case. The use of bonded copper screen in place of, or behind, the glass of a glass-front meter panel will complete the shielding.

Viewing windows in the front panel of a power amplifier stage should have copper screen installed in place of or behind the glass. Where the esthetic effect of copper window screen would not be desirable, ⅛-inch or ¼-inch mesh brass or bronze screen may be substituted with a considerable improvement in the appearance of the opening.

In general it may be stated that small holes, or groups of small holes as would be the case when screen is used, will not seriously impair the shielding of a compartment. But large holes or slots will have an appreciable radiation resistance in the v-h-f range. When screen is used to cover a slot or a hole it is important that the metal of the screen be firmly bonded to the metal of the housing at intervals of not more than a few inches. The paint should be scraped from the region in the cabinet where the screen is to be bolted to insure good electrical contact.

Any conductor which leaves through a hole in the shielding, and which is not solidly grounded at the point where it passes through the hole, can act as a very effective radiator in the upper portion of the v-h-f range. The tuning shafts on variable capacitors have been found to be particularly bad in this regard. All tuning capacitors used on a transmitter or exciter which is to be shielded should have a shaft bearing which is firmly bolted to the panel—or else the shaft of the capacitor should be extended so that the extension which passes through the panel has a panel bearing. Many of the newer v-h-f type (including APC-type) capacitors are particularly bad in this regard since the capacitor unit is mounted behind the panel with studs while the tuning shaft is designed to pass through the panel without a panel bearing. The shafts of such capacitors can completely destroy the shielding effect of an otherwise excellent enclosure. The only cure seems to be to replace the offending capacitor with a type having a sturdy panel bearing.

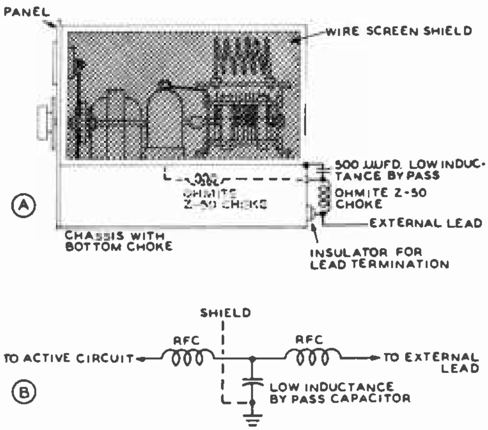


Figure 8.
LEAD-FILTER CIRCUIT.

Showing the physical and electrical details of an effective method for stopping the passage of r-f signals down power and control leads. The T-type low-pass filter has been found most effective for such applications. An alternative arrangement is to use a conventional low-frequency r-f. choke for the inductor inside the shielded compartment, and an inductor composed of about 40 turns of no. 26 d.c.c. wire wound on a quarter-inch polystyrene rod for the inductor mounted outside the shielded compartment.

Filtering of Leads The shielding of a compartment can be completely destroyed if wires and leads leaving the compartment can carry energy out of the shielded enclosure. A lead 2½ or 3 feet long can act as a very effective radiator of harmonic signals in the low TV band. Hence it is quite important that r-f signals be stopped from travelling along the leads which leave the transmitter or exciter enclosure.

The filtering of leads within a transmitter may be accomplished either on a deck-by-deck basis, or in terms of the whole transmitter. Where the transmitter is completely enclosed in a metal housing which includes the shielding provisions mentioned in the previous section, it will be more satisfactory to shield only those leads which leave the transmitter enclosure. All the leads from the transmitter may be brought out in the conventional manner at the bottom rear, with an isolating and filtering compartment at this spot. The filtering compartment may then act as the termination of all the external leads to the transmitter.

The wavemeter-microammeter "gimmick" will be found to be quite helpful in locating the offending leads and in determining the frequencies being radiated. In many cases, especially when the transmitter is running relatively high power, it will be found that the indicator will show a reading regardless of the setting of the wavemeter dial. This condition usually is the result of the rather intense fundamental-frequency field inside the transmitter. This trouble may be eliminated by placing the measuring device outside the transmitter, with a probe loop link coupled to the indicator. A line of small-size Twin-Lead about 4 feet long, or a similar length of twisted hookup wire, with a loop around the wavemeter coil at one end and a loop about 1 inch in diameter at the other end usually will give adequate isolation of the fundamental from the resonant circuit of the indicator.

In any event be *exceedingly careful* when probing to determine the location and amplitude of harmonic fields. Mount the pickup loop for the "gimmick" on the end of about a 2-foot length of dowel rod. Then the loop may be moved about without danger of contact with dangerous potentials while watching the indicating meter.

There are two types of leads entering the high-field intensity portion of the transmitter which may require filters. These are control leads, bias and plate voltage leads, and all other types of leads which carry a relatively low current, all taken as one group, and a-c line power leads which carry a relatively heavy current. Filters for the two types of leads are somewhat different, due to the difference in current carrying capability.

Figure 8 shows in semi-schematic form one method which has proven effective in stopping r-f energy from travelling down supply wires and thus being radiated. Note that a "T" section filter has been used rather than the more common "pi" section type. The logic behind this selection is as follows: The harmonic voltage appearing on the active leads is relatively low, but the current through the by-pass capacitor may be moderately high in value. If the external lead were attached directly to the by-pass capacitor, the lead still could radiate effectively as an odd-quarter-wave wire being fed from a low-impedance source. But by connecting an additional r-f choke in series from the lead to the by-pass

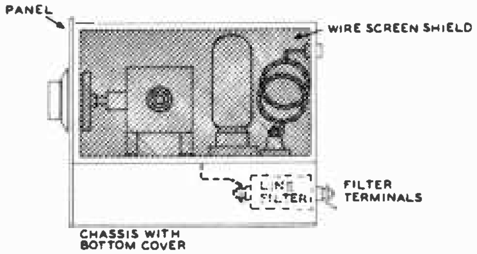


Figure 9.
INSTALLATION OF A LINE FILTER.
Power leads, such as those for 115-volt line current, may be filtered by installing a conventional metal-case line filter in such a manner that its output leads project through the chassis. Sometimes the addition of a low-inductance capacitor (such as the disc type) and a small r-f choke as shown in figure 8 will afford improved filtering of the power leads.

capacitor, the current passing through the by-pass capacitor is effectively stopped from passing on to the lead wire.

The r-f chokes used in the filter may be manufactured units such as suggested in figure 8, or they may be wound on such small coil forms as available. The inductance of the chokes should be from 10 to 25 microhenries, they should be wound as single-layer solenoids with as large a ratio of length to diameter as is practicable, and the turns should be spaced to reduce capacitive coupling from one end of the choke to the other.

The by-pass capacitors used in the filter should have as low lead inductance as possible. The Sprague "Hy-Pass" type of capacitor has been found to be quite effective for by-passing applications. Disc ceramic capacitors, and the tubular high-k ceramic type also are satisfactory, provided that the minimum possible lead length is used. For use in high-voltage circuits, the 500-$\mu\mu$fd. 10,000-volt ceramic filter capacitors designed for application in high-voltage TV power supplies have proven to be satisfactory as by-pass capacitors.

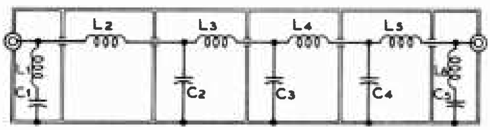
High-Current Lead Filters In some cases it will be necessary to install line filters in series with the a-c line leads which feed the transmitter. Commercially manufactured line filters may do an adequate job if installed as shown in figure 9. Alternatively, the filter may be assembled in the same general manner as shown in figure 8, using high-current solenoid-type chokes in series with the line. Suitable chokes are the Ohmite Z-20, Z-21, and Z-22 with current ratings of 5, 10, and 20 amperes respectively. Two of these dual-winding chokes will be required; one can be installed in series with both sides of the line inside the shielded compartment, and one installed outside for connection of the external leads.

Antenna Couplers Antenna couplers, which have been described in considerable detail in Chapter Eleven, will afford some harmonic reduction, but their main function should be to match the impedance of the antenna transmission line to the characteristic impedance of the low-pass filter used in the lead between the transmitter and the load.

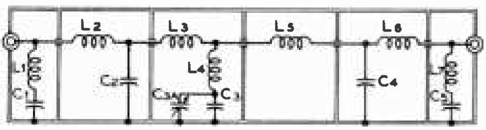
Low-Pass Filters After the transmitter has been shielded, and all power leads have been filtered in such a manner that the transmitter shielding has not been rendered ineffective, the only remaining available exit for harmonic energy lies in the antenna transmission line. Hence the main burden of harmonic attenuation will fall on the low-pass filter installed between the output of the transmitter and the antenna system.

Experience has shown that the low-pass filter can best be installed externally to the main transmitter enclosure, and that the transmission line from the transmitter to the low-pass filter should be of the coaxial type. Hence the majority of low-pass filters are designed for a characteristic impedance of 52 ohms, so that RG-8/U cable (or RG-58/U for a small transmitter) may be used between the output of the transmitter and the antenna transmission line or the antenna tuner.

Transmitting-type low-pass filters for amateur use usually are designed in such a manner as to pass frequencies up to about 30 Mc. without attenuation. The nominal cutoff frequency of the filters is usually between 38 and 45 Mc., and m-derived sections with maximum attenuation in channel 2 usually are included. Well-designed filters capable of carrying any power level up to one kilowatt are available commercially from several manufacturers. Alternatively, filters in kit form are available from several manufacturers at a somewhat lower price. Effective filters may be home



(A)



(B)

Figure 10.
LOW-PASS FILTER SCHEMATIC DIAGRAMS.

The filter illustrated at (A) uses m-derived terminating half sections at each end, with three constant-k mid-sections. The filter at (B) is essentially the same except that the center section has been changed to act as an m-derived section which can be designed to offer maximum attenuation to channels 2, 4, 5, or 6 in accordance with the constants given below. Cutoff frequency is 45 Mc. in all cases. All coils, except L_4 in (B) above, are wound ½" i.d. with 8 turns per inch.

The (A) Filter
C_1,C_5—41.5μμfd. (40 μμfd. will be found suitable.)
C_2,C_3,C_4—136 μμfd. (130 to 140 μμfd. may be used.)
L_1,L_6—0.2 μh.; 3½ t. no. 14
L_2,L_5—0.3 μh.; 5 t. no. 12
L_3,L_4—0.37 μh.; 6½ t. no. 12

The (B) Filter with Mid-Section tuned to Channel 2 (58 Mc..)
C_1,C_5—41.5 μμfd.
C_2,C_4—136 μμfd.
C_3—87 μμfd. (50 μμfd. fixed and 75 μμfd. variable in parallel.)
L_1,L_7—0.2 μh.; 3½ t. no. 14
L_2,L_3,L_5,L_6—0.3 μh.; 5 t. no. 12
L_4—0.09 μh.; 2 t. no. 14 ½" dia. by ¼" long

The (B) Filter with Mid-Section tuned to Channel 4 (71 Mc.). All components same except that:
C_3—106 μμfd.
L_3,L_5—0.33 μh.; 6 t. no. 12
L_4—0.05 μh.; 1½ t. no. 14, ⅜" dia. by ⅜" long.

The (B) Filter with Mid-Section tuned to Channel 5 (81 Mc.). Change the following:
C_3—113 μμfd.
L_3,L_5—0.34 μh.; 6 t. no. 12
L_4—0.033 μh.; 1 t. no. 14 ⅜" dia.

The (B) Filter with Mid-Section tuned to Channel 6 (86 Mc.). All components are essentially the same except that the theoretical value of L_4 is changed to 0.03 μh., and the capacitance of C_3 is changed to 117 μμfd.

constructed, if the test equipment is available and if sufficient care is taken in the construction of the assembly.

Construction of Low-Pass Filters Figures 10, 11 and 12 illustrate high-performance low-pass filters which are suitable for home construction. All are constructed in slip-cover aluminum boxes (ICA no. 29110 or L.M. Bender no. 19) with dimensions of 17 by 3 by 2⅝ inches. Five aluminum baffle plates have been installed in the chassis to make six shielded sections within the enclosure. Feed-through bushings between the shielded sections are Johnson no. 135-55.

Both the (A) and (B) filter types are designed for a nominal cut-off frequency of 45 Mc., with a frequency of maximum rejection at about 57 Mc. as established by the terminating half-sections at each end. Characteristic impedance is 52 ohms in all cases. The alternative filter designs diagrammed in figure 10B have provision for an additional rejection trap in the center of the filter unit which may be designed to offer maximum rejection in channel 2, 4, 5, or 6, depending upon which channel is likely to be received in the area in question. The only components which must be changed when changing the frequency of the maximum rejection notch in the center of the filter unit are inductors L_3, L_4, and L_5, and capacitor C_3. A trimmer capacitor has been included as a portion of C_3 so that the frequency of maximum rejection can be tuned accurately to the desired value. Reference to figures 5 and 6 will show the amateur bands which are most likely to cause interference to specific TV channels.

Either high-power or low-power components may be used in the filters diagrammed in figure 10. With the small Centralab TCZ zero-coefficient ceramic capacitors used in the filter units of figure 10A or figure 10B, power levels up to 200 watts output may be used without danger of damage to the capacitors,

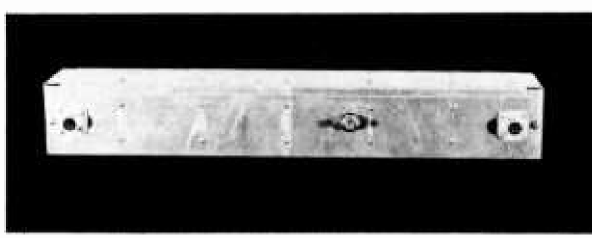


Figure 11.
PHOTOGRAPH OF THE (B) FILTER WITH THE COVER IN PLACE.

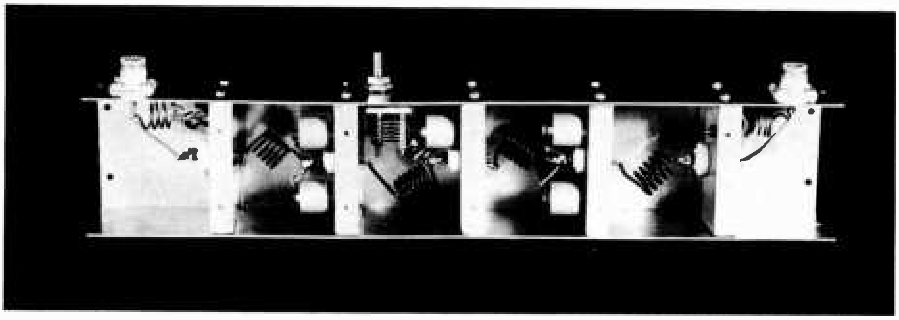


Figure 12.
PHOTOGRAPH OF THE (B) FILTER WITH COVER REMOVED.
The mid-section in this filter is adjusted for maximum rejection of channel 4. Note that the main coils of the filter are mounted at an angle of about 45 degrees so that there will be minimum inductive coupling from one section to the next through the holes in the aluminum partitions. Mounting the coils in this manner was found to give a measurable improvement in the attenuation characteristics of the filter.

provided the filter is feeding a 52-ohm resistive load. It may be practicable to use higher levels of power with this type of ceramic capacitor in the filter, but at a power level of 200 watts on the 28-Mc. band the capacitors run just perceptibly warm to the touch. As a point of interest, it is the current rating which is of significance to the capacitors used in filters such as illustrated. Since current ratings for small capacitors such as these are not readily available, it is not possible to establish an accurate power rating for such a unit. The high-power unit illustrated in figure 12, which uses Centralab type 850S and 854S capacitors, has proven quite suitable for power levels up to one kilowatt.

Capacitors C_1, C_2, C_4, and C_5 can be standard manufactured units with normal 5 per cent tolerance. The coils for the end sections can be wound to the dimensions given (L_1, L_6, and L_7). Then the resonant frequency of the series resonant end sections should be checked with a grid-dip meter, after the adjacent input or output terminal has been shorted with a very short lead. The coils should be squeezed or spread until resonance occurs at 57 Mc.

The intermediate *m*-derived section in the filter of figure 10B may also be checked with a grid-dip meter for resonance at the correct rejection frequency, after the hot end of L_4 has been temporarily grounded with a low-inductance lead. The variable capacitor portion of C_3 can be tuned until resonance at the correct frequency has been obtained. Note that there is so little difference between the constants of this intermediate section for channels 5 and 6 that variation in the setting of C_3 will tune to either channel without materially changing the operation of the filter.

The coils in the intermediate sections of the filter (L_2, L_3, L_4, and L_5 in figure 10A, and L_2, L_3, L_5, and L_6 in figure 10B) may be checked most conveniently outside the filter unit with the aid of a small ceramic capacitor of known value and a grid-dip meter. The ceramic capacitor is paralleled across the small coil with the shortest possible leads. Then the assembly is placed atop a cardboard box and the resonant frequency checked with a grid-dip meter. A Shure reactance slide rule may be used to ascertain the correct resonant frequency for the desired L-C combination (or one of the equations in Chapter Three may be used), and the coil altered until the desired resonant frequency is attained. The coil may then be installed in the filter unit, making sure that it is not squeezed or compressed as it is being installed. However, if the coils are wound exactly as given under figure 10, the filter may be assembled with reasonable assurance that it will operate as designed.

Using Low-Pass Filters The low-pass filter connected in the output transmission line of the transmitter is capable of affording an enormous degree of harmonic attenuation. However, the filter must be operated in the correct manner

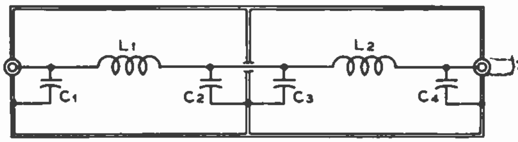


Figure 13.
SCHEMATIC OF THE SINGLE-SECTION HALF-WAVE FILTER.

The constants given below are for a characteristic impedance of 52 ohms, for use with RG-8/U and RG-58/U cable. Coil L_1 should be checked for resonance at the operating frequency with C_1, and the same with L_2 and C_4. This check can be made by soldering a low-inductance grounding strap to the lead between L_1 and L_2 where it passes through the shield. When the coils have been trimmed to resonance with a grid-dip meter, the grounding strap should of course be removed. This filter type will give an attenuation of about 30 db to the second harmonic, about 48 db to the third, about 60 db to the fourth, 67 db to the fifth, and so on increasing at a rate of about 30 db per octave.

C_1,C_2,C_3,C_4—Silver mica or small ceramic for low power, transmitting type ceramic for high power. Capacitance for different bands is given below:
160 meters—1700 μμfd.
80 meters—850 μμfd.
40 meters—440 μμfd.
20 meters—220 μμfd.
10 meters—110 μμfd.
6 meters—60 μμfd.

L_1,L_2—May be made up of sections of B&W Miniductor for power levels below 250 watts, or of no. 12 enam. for power up to one kilowatt. Approximate dimensions for the coils are given below, but the coils should be trimmed to resonate at the proper frequency with a grid-dip meter as discussed above. All coils except the ones for 160 meters are wound 8 turns per inch.
160 meters—4.2 μh.; 22 turns no. 16 enam., 1" dia. 2" long
80 meters—2.1 μh.; 13 t. 1" dia. (No. 3014 Miniductor or no. 12)
40 meters—1.1 μh.; 8 t. 1" dia. (No. 3014 or no. 12 at 8 t.p.i.)
20 meters—0.55 μh.; 7 t. ¾" dia. (No. 3010 or no. 12 at 8 t.p.i.)
10 meters—0.3 μh.; 6 t. ½" dia. (No. 3002 or no. 12 at 8 t.p.i.)
6 meters—0.17 μh.; 4 t. ½" dia. (No. 3002 or no. 12 at 8 t.p.i.)

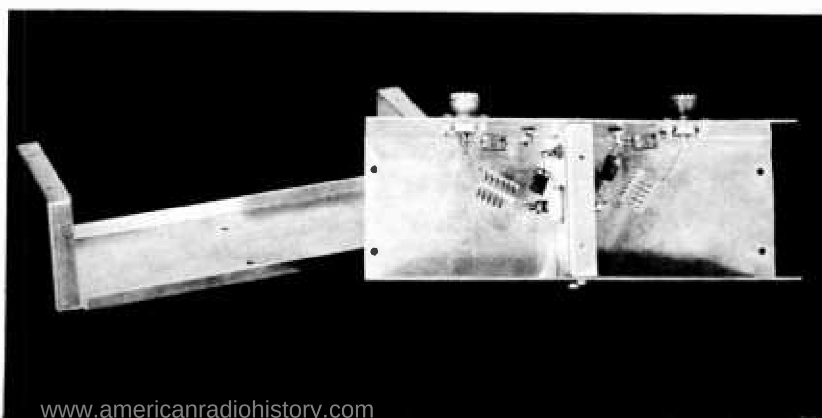
or the results obtained will not be up to expectations.

In the first place, all direct radiation from the transmitter and its control and power leads must be suppressed. This subject has been discussed in the previous section. Secondly, the filter must be operated into a load impedance approximately equal to its design characteristic impedance. The filter itself will have very low losses (usually less than 0.5 db) when operated into its nominal value of resistive load. But if the filter is mis-terminated its losses will become excessive, and it will not present the correct value of load impedance to the transmitter.

If a filter, being fed from a high-power transmitter, is operated into an incorrect termination it may be damaged; the coils may be overheated and the capacitors destroyed as a result of excessive r-f currents. Hence it is wise, when first installing a low-pass filter, to check the standing-wave ratio of the load being presented to the output of the filter with a standing-wave meter of any of the conventional types. Then the antenna termination or the antenna coupled should be adjusted, with low power on the transmitter, until the s.w.r. of the load being presented to the filter is less than 2.0, and preferably below 1.5.

Half-Wave Filters Half-wave filters ("Harmonikers") have been discussed in various publications including the Nov.-Dec. 1949 GE *Ham News*. Such filters are relatively simple and offer the advantage that they present the same value of impedance at their input terminals as appears as load across their output terminals. Such filters normally are used as one-band affairs, and they offer high attenuation only to the third and

Figure 14.
HALF-WAVE FILTER FOR THE 28-MC. BAND.
Showing one possible type of construction of a 52-ohm half-wave filter for relatively low power operation on the 28-Mc. band.



higher harmonics. Design data on the half-wave filter is given in figure 13. Construction of half-wave filters is illustrated in figure 14 and in the 50-Mc. transmitter shown in Chapter Twenty-four.

17-3 Broadcast Interference

Interference to the reception of signals in the broadcast band (540 to 1600 kc.) or in the FM broadcast band (88 to 108 Mc.) by amateur transmissions is a serious matter to those amateurs living in densely populated areas. Although broadcast interference has recently been overshadowed by the seriousness of television interference, the condition of BCI is still present.

In general, signals from a transmitter operating properly are not picked up by receivers tuned to other frequencies unless the receiver is of inferior design, or is in poor condition. Therefore, if the receiver is of good design and is in good repair, the burden of rectifying the trouble rests with the owner of the interfering station.

Phone and c-w stations both are capable of causing broadcast interference, key-click annoyance from the code transmitters being particularly objectionable. The elimination of key clicks has been discussed in detail in Chapter Ten.

A knowledge of each of the several types of broadcast interference, their cause, and methods of eliminating them is necessary to the successful disposition of this trouble. An effective method of combatting one variety of interference is often of no value whatever in the correction of another type. Broadcast interference seldom can be cured by "rule of thumb" procedure.

Broadcast interference, as covered in this section refers primarily to standard (amplitude modulated, 550-1600 kc.) broadcast. Interference with FM broadcast reception is much less common, due to the wide separation in frequency between the FM broadcast band and the more popular amateur bands, and due also to the limiting action which exists in all types of FM receivers. Occasional interference with FM broadcast by a harmonic of an amateur transmitter has been reported; if this condition is encountered, it may be eliminated by the procedures discussed in the first portion of this chapter under Television Interference.

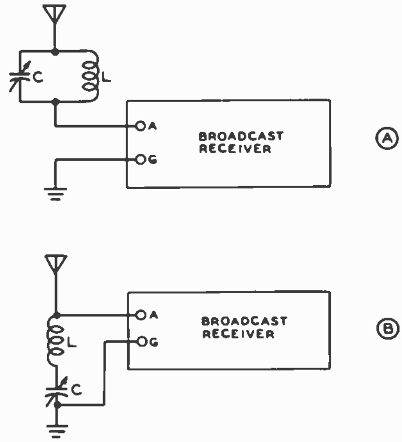


Figure 15.
WAVE-TRAP CIRCUITS.
The circuit at (A) is the most common arrangement, but the circuit at (B) may give improved results under certain conditions. Manufactured wave traps for the desired band of operation may be purchased or the traps may be assembled from the data given in figure 17.

The use of frequency-modulation transmission by an amateur station is likely to result in much less interference to broadcast reception than either amplitude-modulated telephony or straight keyed c.w. This is true because, insofar as the broadcast receiver is concerned, the amateur FM transmission will consist of a plain unmodulated carrier. There will be no key clicks or voice reception picked up by the b-c-l set (unless it happens to be an FM receiver which might pick up a harmonic of the signal), although there might be a slight click when the transmitter is put on or taken off the air. This is one reason why narrow-band FM has become so popular with phone enthusiasts who reside in densely populated areas.

Interference Classifications Depending upon whether it is traceable directly to causes within the *station* or within the *receiver*, broadcast interference may be divided into two main classes. For example, that type of interference due to transmitter over-modulation is at once listed as being

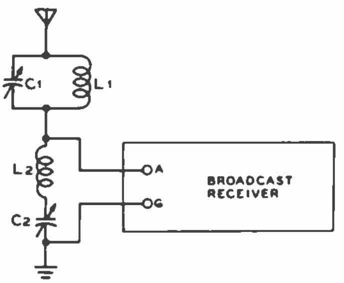


Figure 16.
HIGH-ATTENUATION WAVE-TRAP CIRCUIT.
The two circuits may be tuned to the same frequency for highest attenuation of a strong signal, or the two traps may be tuned separately for different bands of operation.

BAND	COIL, L	CAPACITOR, C
1.8 Mc.	1 inch no. 30 enam. closewound on 1" form	75-$\mu\mu$fd. var.
3.5 Mc.	42 turns no. 30 enam. closewound on 1" form	50-$\mu\mu$fd. var.
7.0 Mc.	23 turns no. 24 enam. closewound on 1" form	50-$\mu\mu$fd. var.
14 Mc.	10 turns no. 24 enam. closewound on 1" form	50-$\mu\mu$fd. var.
21 Mc.	7 turns no. 24 enam. closewound on 1" form	50-$\mu\mu$fd. var.
28 Mc.	4 turns no. 24 enam. closewound on 1" form	25-$\mu\mu$fd. var.
50 Mc.	3 turns no. 24 enam. spaced ½" on 1" form	25-$\mu\mu$fd. var.

Figure 17.
COIL AND CAPACITOR TABLE FOR AMATEUR-BAND WAVETRAPS.

caused by improper operation, while an interfering signal that tunes in and out with a broadcast station is probably an indication of cross modulation or image response in the receiver, and the poorly-designed input stage of the receiver is held liable. The various types of interference and recommended cures will be discussed in the following paragraphs.

Blanketing This is not a tunable effect, but a total blocking of the receiver. A more or less complete "washout" covers the entire receiver range when the carrier is switched on. This produces either a complete blotting out of all broadcast stations, or else knocks down their volume several decibels—depending upon the severity of the interference. Voice modulation of the carrier causing the blanketing will be highly distorted or even unintelligible. Keying of the carrier which produces the blanketing will cause an annoying fluctuation in the volume of the broadcast signals.

Blanketing generally occurs in the immediate neighborhood (inductive field) of a powerful transmitter, the affected area being directly proportional to the power of the transmitter. Also it is more prevalent with transmitters which operate in the 160-meter and 80-meter bands, as compared to those on the higher frequencies.

The remedies are to (1) shorten the receiving antenna and thereby shift its resonant frequency, or (2) remove it to the interior of the building, (3) change the direction of either the receiving or transmitting antenna to minimize their mutual coupling, or (4) keep the interfering signal from entering the receiver input circuit by installing a wave-trap tuned to the signal frequency, (See figure 15.) or a low-pass filter as shown in figure 24.

A suitable wave-trap is quite simple in construction, consisting only of a coil and midget variable capacitor. When the trap circuit is tuned to the frequency of the interfering signal, little of the interfering voltage reaches the grid of the first tube. Commercially manufactured wave-traps are available from several concerns, including the J. W. Miller Co. in Los Angeles. However, the majority of amateurs prefer to construct the traps from spare components selected from the "junk box."

The circuit shown in figure 16 is particularly effective because it consists of two traps. The shunt trap blocks or rejects the frequency to which it is tuned, while the series trap across the antenna and ground terminals of the receiver provides a very low impedance path to ground at the frequency to which it is tuned and by-passes the signal to ground. In moderate interference cases, either the shunt or series trap may be used alone, while similarly, one trap may be tuned to one of the frequencies of the interfering transmitter and the other trap to a different interfering frequency. In either case, each trap is effective over but a small frequency range and must be readjusted for other frequencies.

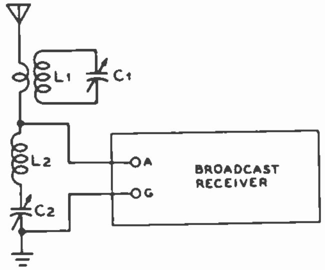


Figure 18.
MODIFICATION OF THE FIGURE 16 CIRCUIT.

In this circuit arrangement the parallel-tuned tank is inductively coupled to the antenna lead with a 3 to 6 turn link instead of being placed directly in series with the antenna lead.

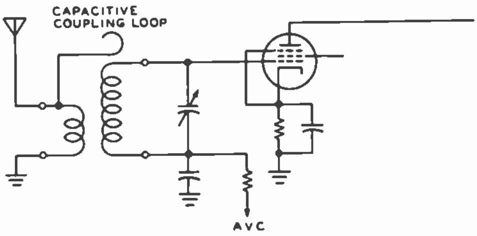


Figure 19.
CAPACITIVE BOOST COUPLING CIRCUIT.

Such circuits, included within the broadcast receiver to bring up the stage gain at the high-frequency end of the tuning range, have a tendency to increase the susceptibility of the receiver to interference from amateur-band transmissions.

The wave-trap must be installed as close to the receiver antenna terminal as practicable, hence it should be as small in size as possible. The variable capacitor may be a midget air-tuned trimmer type, and the coil may be wound on a 1-inch dia. form. The table of figure 17 gives winding data for wave-traps built around standard variable capacitors. For best results, both a shunt and a series trap should be employed as shown.

Figure 18 shows a two-circuit coupled wave-trap that is somewhat sharper in tuning and more efficacious. The specifications for the secondary coil L_1 may be obtained from the table of figure 17. The primary coil of the shunt trap consists of 3 to 5 closewound turns of the same size wire wound in the same direction on the same form as L_1 and separated from the latter by ⅛ of an inch.

Overmodulation A carrier modulated in excess of 100 per cent acquires sharp cutoff periods which give rise to transients. These transients create a broad signal and generate spurious responses. Transients caused by overmodulation of a radiotelephone signal may at the same time bring about impact or shock excitation of nearby receiving antennas and power lines, generating interfering signals in that manner.

Broadcast interference due to overmodulation is frequently encountered. The remedy is to reduce the modulation percentage or to use a clipper-filter system or a high-level splatter suppressor in the speech circuit of the transmitter.

Cross Modulation Cross modulation or "cross talk" is characterized by the amateur signal "riding in" on top of a strong broadcast signal. There is usually no heterodyne note, the amateur signal being tuned in and out with the program carriers.

This effect is due frequently to a faulty input stage in the affected receiver. Modulation of the interfering carrier will swing the operating point of the input tube. This type of trouble is seldom experienced when a variable-μ tube is used in the input stage.

Where the receiver is too ancient to incorporate such a tube, and is probably poorly shielded at the same time, it will be better to attach a wave trap of the type shown in figure 15 rather than to attempt rebuilding of the receiver. The addition of a good ground and a shield can over the input tube often adds to the effectiveness of the wave trap.

Transmission via Capacitive Coupling A small amount of capacitive coupling is now widely used in receiver r.f. and antenna transformers as a gain booster at the high-frequency end of the tuning range. The coupling capacitance is obtained by means of a small loop of wire cemented close to the grid end of the secondary winding, with one end directly connected to the plate or antenna end of the primary winding. (See figure 19.)

It is easily seen that a small capacitor at this position will favor the coupling of the higher frequencies. This type of capacitive coupling in the receiver coils will tend to pass amateur high-frequency signals into a receiver tuned to broadcast frequencies.

The amount of capacitive coupling may be reduced to eliminate interference by moving the coupling turn further away from the secondary coil. However, a simple wave trap of the type shown in figure 15, inserted at the antenna input terminal, will generally accomplish the same result and is more to be recommended than reducing the amount of capacitive coupling (which lowers the receiver gain at the high-frequency end of the broadcast band). Should the wave trap alone not suffice, it will be necessary to resort to a reduction in the coupling capacitance.

In some simple broadcast receivers, capacitive coupling is obtained by closely coupled primary and secondary coils, or as a result of running a long primary or antenna lead close to the secondary coil of an unshielded antenna coupler.

Phantoms With two strong local carriers applied to a non-linear impedance, the beat note resulting from cross-modulation between them may fall on some frequency within the broadcast band and will be audible at that point. If such a "phantom" signal falls on a local broadcast frequency, there will be heterodyne interference as well. This is a common occurrence with broadcast receivers in the neighborhood of two amateur stations, or an amateur and a police station. It also sometimes occurs when only one of the stations is located in the immediate vicinity.

As an example: an amateur signal on 3514 kc. might beat with a local 2414-kc. police carrier to produce a 1100-kc. phantom. If the two carriers are strong enough in the vicinity of a circuit which can cause rectification, the 1100-kc phantom will be heard in the broadcast band. A poor contact between two oxidized wires can produce rectification.

Two stations must be transmitting simultaneously to produce a phantom signal; when either station goes off the air the phantom disappears. Hence, this type of interference is apt to be reported as highly intermittent and might be difficult to duplicate unless a test oscillator is used "on location" to simulate the missing station. Such interference cannot be remedied at the transmitter, and often the rectification takes place some distance from the receivers. In such occurrences it is most difficult to locate the source of the trouble.

It will also be apparent that a phantom might fall on the intermediate frequency of a simple superhet receiver and cause interference of the untunable variety if the manufacturer has not provided an i-f wave-trap in the antenna circuit.

This particular type of phantom may, in addition to causing i-f interference, generate harmonics which may be tuned in and out with heterodyne whistles from one end of the receiver dial to the other. It is in this manner that "birdies" often result from the operation of nearby amateur stations.

When one component of a phantom is a steady, unmodulated carrier, only the intelligence present on the other carrier is conveyed to the broadcast receiver.

Phantom signals almost always may be identified by the suddenness with which they are interrupted, signalizing withdrawal of one party to the union. This is especially baffling to the inexperienced interference-locater, who observes that the interference suddenly disappears, even though his own transmitter remains in operation.

If the mixing or rectification is taking place in the receiver itself, a phantom signal may be eliminated by removing either one of the contributing signals from the receiver input circuit. A wave-trap of the type shown in figure 15, tuned to either signal, will do the trick. If the rectification is taking place outside the receiver, the wave-trap should be tuned to the frequency of the phantom, instead of to one of its components. I-f wave-traps may be built around a 2.5-millihenry r-f choke as the inductor, and a compression-type mica padding capacitor. The capacitor should have a capacitance range of 250—525 $\mu\mu$fd. for the 175- and 206-kc. intermediate frequencies; 65—175 $\mu\mu$fd. for 260 kc. and other intermediates lying between 250 and 400 kc; and 17—80 $\mu\mu$fd. for 456, 465, 495, and 500 kc. Slightly more capacitance will be required for resonance with a 2.1 millihenry choke.

Spurious Emissions This sort of interference arises from the transmitter itself. The radiation of any signal (other than the intended carrier frequency) by an amateur station is prohibited by FCC regulations. Spurious radiation may be traced to imperfect neutralization, parasitic oscillations in the r-f or modulator stages, or to "broadcast-band" variable-frequency oscillators or e.c.o.'s.

Low-frequency parasitics may actually occur on broadcast frequencies or their near sub-harmonics, causing direct interference to programs. An all-wave monitor operated in the vicinity of the transmitter will detect these spurious signals.

The remedy will be obvious in individual cases. Elsewhere in this book are discussed methods of complete neutralization and the suppression of parasitic oscillations in r-f and audio stages.

A-c/d-c Receivers Inexpensive table-model a-c/d-c receivers are particularly susceptible to interference from amateur transmissions. In fact, it may be said with a fair degree of assurance that the majority of BCI encountered by amateurs operating in the 1.8-Mc. to 29-Mc. range is a result of these inexpensive receivers. In most cases the receivers are at fault; but this does not absolve the amateur of his responsibility in attempting to eliminate the interference.

Stray Receiver Rectification In most cases of interference to inexpensive receivers, particularly those of the a-c/d-c type, it will be found that stray receiver rectification is causing the trouble. The offending stage usually will be found to be a high-mu triode as the first audio stage following the second detector. Tubes of this type are quite non-linear in their grid characteristic, and hence will readily rectify any r-f signal appearing between grid and cathode. The r-f signal may get to the tube as a result of direct signal pickup due to the lack of shielding, but more commonly will be fed to the tube from the power line as a result of the series heater string.

The remedy for this condition is simply to insure that the cathode and grid of the high-mu audio tube (usually a 12SQ7 or equivalent) are at the same r-f potential. This is accomplished by placing an r-f by-pass capacitor with the shortest possible leads directly from grid to cathode, and then adding an impedance in the lead from the volume control to the grid of the audio tube. The impedance may be an amateur band r-f choke (such as a National R-100U) for best results,

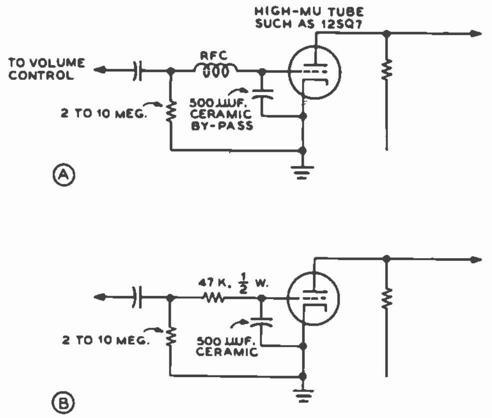


Figure 20.
CIRCUITS FOR ELIMINATING AUDIO-STAGE RECTIFICATION.

but for a majority of cases it will be found that a 47,000-ohm ½-watt resistor in series with this lead will give satisfactory operation. Suitable circuits for such an operation on the receiver are given in figure 20.

In many a.c.-d.c. receivers there is no r-f by-pass included across the plate supply rectifier for the set. If there is an appreciable level of r-f signal on the power line feeding the receiver, r-f rectification in the power rectifier of the receiver can cause a particularly bad type of interference which may be received on other broadcast receivers in the vicinity in addition to the one causing the rectification. The soldering of a 0.01-μfd. disc ceramic capacitor directly from anode to cathode of the power rectifier (whether it is of the vacuum-tube or selenium-rectifier type) usually will by-pass the r-f signal across the rectifier and thus eliminate the difficulty.

"Floating" Volume Control Shafts Several sets have been encountered where there was only a slightly interfering signal; but, upon placing one's hand up to the volume control, the signal would greatly increase. Investigation revealed that the volume control was installed with its shaft insulated from ground. The control itself was connected to a critical part of a circuit, in many instances to the grid of a high-gain audio stage. The cure is to install a volume control with *all* the terminals insulated from the shaft, and then to ground the shaft.

Spray-Shield Tubes Although they are no longer made, there are yet quite a few sets in use which employ spray-shield tubes. These are used in both r-f and in audio circuits. In some audio applications of this type of tube, the cathode and the spray-shield (to which the cathode is connected) are not at ground potential, but are bypassed to ground with an electrolytic capacitor of large capacitance. This type of capacitor is a very poor r-f filter, and in a strong r-f field, some rectification will take place, producing interference. The best cure is to install a standard glass tube with a glove shield, which is then actually grounded, and also to shield the grid leads to these tubes. As an alternative, bypassing the electrolytic cathode capacitor with a .05-μfd. tubular paper capacitor may be tried.

Power-Line Pickup When radio-frequency energy from a radio transmitter enters a broadcast receiver through the a-c power lines, it has either been fed back into the lighting system by the offending transmitter, or picked up from the air by overhead power lines. Underground lines are seldom responsible for spreading this interference.

To check the path whereby the interfering signals reach the line, it is only necessary to replace the transmitting antenna with a dummy antenna and adjust the transmitter for maximum output. If the interference then ceases, overhead lines have been picking up the energy. The trouble can be cleared up by installing a wave trap or a commercial line filter in the power lines at the receiver. If the receiver is reasonably close to the transmitter, it is very doubtful that changing the direction of the transmitting antenna to right angles with the overhead lines will eliminate the trouble.

If, on the contrary, the interference continues when the transmitter is connected to the dummy antenna, radio-frequency energy is being fed directly into the power line by the transmitter, and the station must be inspected to determine the cause.

One of the following reasons for the trouble will usually be found: (1) the r-f stages are not sufficiently bypassed and/or choked, (2) the antenna coupling system is not performing efficiently, (3) the power transformers have no electrostatic shields; or, if shields are present, they are ungrounded, (4) power lines are running too close to an antenna or r-f circuits carrying high currents. If none of these causes apply, wave-traps must be installed in the power lines at the transmitter to remove r-f energy passing back into the lighting system.

BAND	COIL, L	CAPACITOR, C
3.5 Mc.	17 turns no. 14 enameled 3-inch diameter 2¼-inch length	100-$\mu\mu$fd. variable
7.0 Mc.	11 turns no. 14 enameled 2½-inch diameter 1½-inch length	100-$\mu\mu$fd. variable
14 and 21 Mc.	4 turns no. 10 enameled 3-inch diameter 1⅛-inch length	100-$\mu\mu$fd. variable
27 and 28 Mc.	3 turns ¼-inch o.d. copper tubing 2-inch diameter 1-inch length	100-$\mu\mu$fd. variable

Figure 21.
COIL AND CAPACITOR TABLE FOR A-C LINE TRAPS.

The wave-traps used in the power lines at transmitter or receiver must be capable of passing relatively high current. The coils are accordingly wound with heavy wire. Figure 21 lists the specifications for power line wave-trap coils, while figure 22 illustrates the method of connecting these wave-traps. Observe that these traps are enclosed in a shield box of heavy iron or steel, well grounded.

All-Wave Receivers Each complete-coverage home receiver is a potential source of annoyance to the transmitting amateur. The novice short-wave broadcast-listener who tunes in an amateur station often considers it an interfering signal, and complains accordingly.

Neither selectivity nor image rejection in most of these sets is comparable to those properties in a communication receiver. The result is that an amateur signal will occupy too much dial space and appear at more than

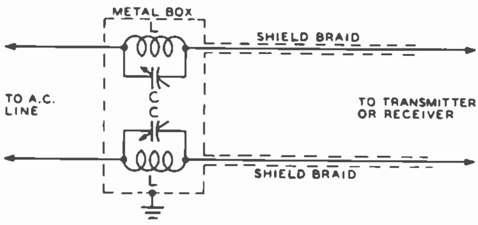


Figure 22.
RESONANT POWER-LINE WAVE-TRAP CIRCUIT.
The resonant type of power-line filter is more effective than the more conventional "brute force" type of line filter, but requires tuning to the operating frequency of the transmitter.

one point, giving rise to interference on adjacent channels and distant channels as well.

If carrier-frequency harmonics are present in the amateur transmission, serious interference will result at the all-wave receiver. The harmonics may, if the carrier frequency has been so unfortunately chosen, fall directly upon a favorite short-wave broadcast station and arouse warranted objection.

The amateur is apt to be blamed, too, for transmissions for which he is not responsible, so great is the public ignorance of short-wave allocations and signals. Owners of all-wave receivers have been quick to ascribe to amateur stations all signals they hear from tape machines and V-wheels, as well as stray tones and heterodyne flutters.

The amateur cannot be held responsible when his carrier is deliberately tuned in on an all-wave receiver. Neither is he accountable for the width of his signal on the receiver dial, or for the strength of image repeat points, if it can be proven that the receiver design does not afford good selectivity and image rejection.

If he so desires, the amateur (or the owner of the receiver) might sharpen up the received signal somewhat by shortening the receiving antenna. Set retailers often supply quite a sizeable antenna with all-wave receivers, but most of the time these sets perform almost as well with a few feet of inside antenna.

The amateur *is* accountable for harmonics of his carrier frequency. Such emissions are unlawful in the first place, and he must take all steps necessary to their suppression. Practical suggestions for the elimination of harmonics have been given earlier in this chapter under Television Interference.

Image Interference In addition to those types of interference already discussed, there are two more which are common to superhet receivers. The prevalence of these types is of great concern to the amateur, although the responsibility for their existence more properly rests with the broadcast receiver.

The mechanism whereby image production takes place may be explained in the following manner: when the first detector is set to the frequency of an incoming signal, the high-frequency oscillator is operating on another frequency which differs from the signal by the number of kilocycles of the intermediate frequency. Now, with the setting of these two stages undisturbed, there is another signal which will beat with the high-frequency oscillator to produce an i-f signal. This other signal is the so-called image, which is separated from the desired signal by twice the intermediate frequency.

Thus, in a receiver with 175-kc. i.f., tuned to 1000 kc.: the h-f oscillator is operating on 1175 kc., and a signal on 1350 kc. (1000 kc. plus 2 x 175 kc.) will beat with this 1175 kc. oscillator frequency to produce the 175-kc. i-f signal. Similarly, when the same receiver is tuned to 1400 kc., an amateur signal on 1750 kc. can come through. The dial point where any 160-meter signal will produce an image can be determined from the equation:

$$F_b = (F_{am} - 2 \text{ i.f.})$$

Where F_b = receiver dial frequency
F_{am} = amateur transmitter frequency, and
i.f. = receiver intermediate frequency.

If the image appears only a few cycles or kilocycles from a broadcast carrier, heterodyne interference will be present as well. Otherwise, it will be tuned in and out in the manner of a station operating in the broadcast band. Sharpness of tuning will be comparable to that of broadcast stations producing the same a-v-c voltage at the receiver.

The second variety of superhet interference is the result of harmonics of the receiver h-f oscillator beating with amateur carriers to produce the intermediate frequency of the receiver. The amateur transmitter will always be

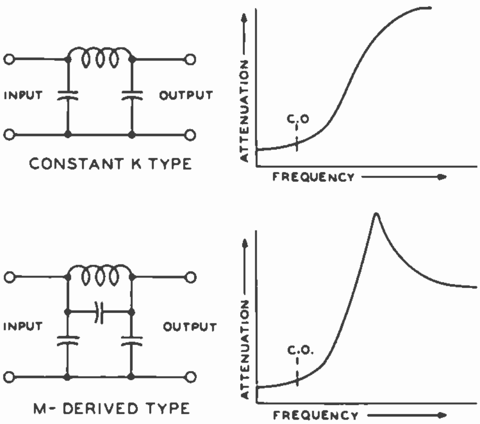


Figure 23.
TYPES OF LOW-PASS FILTERS.
Filters such as these may be used in the circuit between the antenna and the input of the receiver.

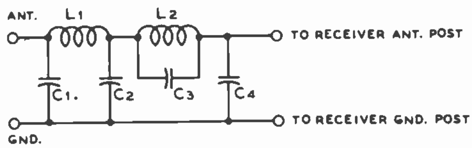


Figure 24.
COMPOSITE LOW-PASS FILTER CIRCUIT.
This filter is highly effective in reducing broadcast interference from all high frequency stations, and requires no tuning. Constants for 400 ohm terminal impedance and 1600 kc. cutoff are as follows: L_1, 65 turns no. 22 d.c.c. closewound on 1½ in. dia. form. L_2, 41 turns ditto, not coupled to L_1. C_1, 250 μμfd. fixed mica capacitor. C_2, 400 μμfd. fixed mica capacitor. C_3 and C_4, 150 μμfd. fixed mica capacitors, former of 5% tolerance. With some receivers, better results will be obtained with a 200 ohm carbon resistor inserted between the filter and antenna post on the receiver. With other receivers the effectiveness will be improved with a 600 ohm carbon resistor placed from the antenna post to the ground post on the receiver. The filter should be placed as close to the receiver terminals as possible.

found to be on a frequency equal to some harmonic of the receiver h-f oscillator, *plus or minus the intermediate frequency.*

As an example: when a broadcast superhet with 465-kc. i.f. is tuned to 1000 kc., its high-frequency oscillator operates on 1465 kc. The third harmonic of this oscillator frequency is 4395 kc., which will beat with an amateur signal on 3930 kc. to send a signal through the i-f amplifier. The 3930 kc. signal would be tuned in at the 1000-kc. point on the dial.

Some oscillator harmonics are so related to amateur frequencies that more than one point of interference will occur on the receiver dial. Thus, a 3500-kc. signal may be tuned in at six points on the dial of a nearby broadcast superhet having 175 kc. i.f. and no r-f stage.

Insofar as remedies for image and harmonic superhet interference are concerned, it is well to remember that *if* the amateur signal did not in the first place reach the input stage of the receiver, the annoyance would not have been created. It is therefore good policy to try to eliminate it by means of a wave-trap or low-pass filter. Broadcast superhets are not always the acme of good shielding, however, and the amateur signal is apt to enter the circuit through channels other than the input circuit. If a wave-trap or filter will not cure the trouble, the only alternative will be to attempt to select a transmitter frequency such that neither image nor harmonic interference will be set up on favorite stations in the susceptible receivers. The equation given earlier may be used to determine the proper frequencies.

Low Pass Filters The greatest drawback of the wave-trap is the fact that it is a single-frequency device; i.e.—it may be set to reject at one time only one frequency (or, at best, an extremely narrow band of frequencies). Each time the frequency of the interfering transmitter is changed, every wave-trap tuned to it must be retuned. A much more satisfactory device is the *wave filter* which requires no tending. One type, the low-pass filter, passes all frequencies below one critical frequency, and eliminates all higher frequencies. It is this property that makes the device ideal for the task of removing amateur frequencies from broadcast receivers.

A good low pass filter designed for maximum attenuation around 1700 kc. will pass all broadcast carriers, but will reject signals originating in any amateur band. Naturally such a device should be installed only in standard broadcast receivers, never in all-wave sets.

Two types of low-pass filter sections are shown in figure 23. A composite arrangement comprising a section of each type is more effective than either type operating alone. A composite filter composed of one K-section and one shunt-derived M-section is shown in figure 24, and is highly recommended. The M-section is designed to have maximum attenuation at 1700 kc., and for that reason C_3 should be of the "close tolerance" variety. Likewise, C_3 should not be stuffed down inside L_2 in the interest of compactness, as this will alter the inductance of the coil appreciably, and likewise the resonant frequency.

If a fixed 150 μμfd. mica capacitor of 5 per cent tolerance is not available for C_3, a compression trimmer covering the range of 125—175 μμfd. may be substituted and adjusted to give maximum attenuation at about 1700 kc.

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CHAPTER EIGHTEEN

Workshop Practice

With a few possible exceptions, such as fixed air capacitors, neutralizing capacitors and transmitting coils, it hardly pays one to attempt to build the components required for the construction of an amateur transmitter. This is especially true when the parts are of the type used in construction and replacement work on broadcast receivers, as mass production has made these parts very inexpensive.

Transmitters Those who have and wish to spend the necessary time can effect considerable monetary saving in their transmitters by building them from the component parts. The necessary data are given in the construction chapters of this handbook.

To many builders, the construction is as fascinating as the operation of the finished transmitter; in fact, many amateurs get so much satisfaction out of building a well-performing piece of equipment that they spend more time constructing and rebuilding equipment than they do operating the equipment on the air.

Those who are not mechanically minded and are more interested in the pleasures of working dx and rag chewing than in experimentation and construction will find on the market many excellent transmitters which require only line voltage and an antenna. If you are one of those amateurs, you will find little to interest you in this chapter.

Receivers There is room for argument as to whether one can save money by constructing his own communications receiver. The combined demand for these receivers by the government, amateurs, airways, short-wave listeners, and others has become so great that it may be argued that there is no more point in building such a receiver than in building a regular broadcast set. Yet, many amateurs still prefer to construct their own receivers—in spite of the fact that it costs almost as much to build a receiver as to purchase an equivalent factory-made job—either because they enjoy construction work and take pride in the fruits of their efforts, or because the receiver must meet certain specifications and yet cost as little as possible.

The only factory-produced receiver that is sure to meet the requirements of every amateur or short-wave listener is the rather expensive de luxe type having every possible refinement. A person primarily interested in specialized performance on a few bands usually can obtain better results, with less expense, by constructing special converters or adapters to operate with an inexpensive manufactured receiver. Thus the v-h-f man may construct crystal-controlled converters to precede a surplus communications receiver, the person concerned with voice communication or single-sideband work may construct a single-sideband adapter to follow his communications receiver, and so forth in each specialized case.

18-1 Types of Construction

Layout and mounting of the component parts usually require the most time and work.

Various methods of mounting can be used, ranging from simple wooden boards to elaborate metal racks and panels.

Breadboard The simplest method of constructing equipment is to lay it out in breadboard fashion, which consists of fastening the various components to a board of suitable size with wood screws or machine bolts, arranging the parts so that important leads will be as short as possible.

Breadboard construction is suitable for testing an experimental layout, or sometimes for assembling an experimental unit of test equipment. But no permanent item of station equipment should be left in the breadboard form. Breadboard construction is dangerous, since components carrying dangerous voltages are left exposed. Also, breadboard construction is never suitable for any r-f portion of a transmitter, since it would be substantially impossible to shield such an item of equipment for the elimination of TVI resulting from harmonic radiation.

Metal Chassis Though quite a few more tools and considerably more time will be required for metal chassis construction, much neater and more satisfactory equipment can be built by mounting the parts on sheet metal chassis instead of breadboards. This type of construction is necessary when shielding of the apparatus is required. A front panel and a back shield minimizes the danger of shock and completes the shielding of the enclosure.

Dish type construction is practically the same as metal chassis construction, the main difference lying in the manner in which the chassis is fastened to the panel.

Special Frameworks For high-powered r-f stages, many amateur constructors prefer to discard the more conventional types of construction and employ instead special metal frameworks and brackets which they design specially for the parts which they intend to use. These are usually arranged to give the shortest possible r-f leads and to fasten directly behind a relay rack panel by means of a few bolts, with the control shafts projecting through corresponding holes in the panel.

18-2 Tools

Beautiful work can be done with metal chassis and panels with the help of only a few inexpensive tools. However, the time required for construction will be greatly reduced if a fairly complete assortment of metal-working tools is available. Thus, it can be seen that while an array of tools will speed up the work, excellent results may be accomplished with but few tools, if one has the time and patience.

The investment one is justified in making in tools is dependent upon several factors. If you like to tinker, there are many tools useful in radio construction that you would probably buy anyway, or perhaps already have, such as screwdrivers, hammer, saws, square, vise, files, etc. This means that the money taken for tools from your radio budget can be used to buy the more specialized tools, such as socket punches or hole saws, taps and dies, etc.

The amount of construction work one does determines whether buying a large assortment of tools is an economical move. It also determines if one should buy the less expensive type offered at surprisingly low prices by the familiar mail order houses, "five and ten" stores and chain auto-supply stores, or whether one should spend more money and get first-grade tools. The latter cost considerably more and work but little better when new, but will outlast several sets of the cheaper tools. Therefore they are a wise investment for the experimenter who does lots of construction work (if he can afford the initial cash outlay). The amateur who constructs only an occasional piece of apparatus need not be so concerned with tool life, as even the cheaper grade tools will last him several years, if they are given proper care.

The hand tools and materials in the accompanying lists will be found very useful around the home workshop. Materials not listed but ordinarily used, such as paint, can best be purchased as required for each individual job.

ESSENTIAL HAND TOOLS AND MATERIALS

1 *Good* electric soldering iron, about 100 watts

1 Spool rosin-core wire solder

1 Each large, medium, small, and midget screwdrivers

1 *Good* hand drill (eggbeater type), preferably two speed
1 Pair regular pliers, 6 inch
1 Pair long nose pliers, 6 inch
1 Pair cutting pliers (diagonals), 5 inch or 6 inch
1 1⅛-inch tube-socket punch
1 "Boy Scout" knife
1 Combination square and steel rule, 1 foot
1 Yardstick or steel pushrule
1 Scratch awl or ice pick scribe
1 Center punch
1 Dozen or more assorted round shank drills (as many as you can afford between no. 50 and ¼ or ⅜ inch, depending upon size of hand drill chuck)
1 Combination oil stone
Light machine oil (in squirt can)
Friction tape
1 Hacksaw and blades
1 Medium file and handle
1 Cold chisel (½ inch tip)
1 Wrench for socket punch
1 Hammer

HIGHLY DESIRABLE HAND TOOLS AND MATERIALS

1 Bench vise (jaws at least 3 inch)
1 Spool plain wire solder
1 Carpenter's brace, ratchet type
1 Square-shank countersink bit
1 Square-shank taper reamer, small
1 Square-shank taper reamer, large, (The two reamers should overlap; ½ inch and ⅞ inch size will usually be suitable.)
1 ⅞ inch tube-socket punch (for electrolytic capacitors)
1 1-3/16 inch tube-socket punch
1 ⅝ inch tube-socket punch
1 Adjustable circle cutter for holes to 3 inch
1 Set small, inexpensive, open-end wrenches
1 Pair tin shears, 10 or 12 inch
1 Wood chisel (½ inch tip)
1 Pair wing dividers
1 Coarse mill file, flat, 12 inch
1 Coarse bastard file, round, ½ or ¾ inch
1 Set allen and spline-head wrenches.
6 or 8 Assorted small files: round, half-round, triangular, flat, square, rat-tail
4 Small "C" clamps
Steel wool, coarse and fine
Sandpaper and emery cloth, coarse, medium, and fine
Duco cement
File brush

USEFUL BUT NOT ESSENTIAL TOOLS AND MATERIALS

1 Jig or scroll saw (small) with metal-cutting blades
1 Small wood saw (crosscut teeth)
1 Each square-shank drills: ⅜, 7/16, and ½ inch
1 Tap and die outfit for 6-32, 8-32, 10-32 and 10-24 machine screw threads
4 Medium size "C" clamps
Lard oil (in squirt can)
Kerosene
Empire cloth
Clear lacquer ("industrial" grade)
Lacquer thinner
Dusting brush
Paint brushes
Sheet celluloid, Lucite, or polystyrene
1 Carpenter's plane
1 Each "Spintite" wrenches, ¼, 5/16, 11/32 to fit the standard 6-32 and 8-32 nuts used in radio work
1 Screwdriver for recessed head type screws

The foregoing assortment assumes that the constructor does not want to invest in the more expensive power tools, such as drill press, grinding head, etc. If power equipment is purchased, obviously some of the hand tools and accessories listed will be superfluous. A drill press greatly facilitates construction work, and it is unfortunate that a good one costs as much as a small transmitter. A booklet* available from the Delta Manufacturing Co. will be of considerable aid to those who have access to a drill press.

Not listed in the table are several special-purpose radio tools which are somewhat of a luxury, but are nevertheless quite handy, such as various around-the-corner screwdrivers and wrenches, special soldering iron tips, etc. These can be found in the larger radio parts stores and are usually listed in their mail order catalogs. It is not uncommon to find amateurs who have had sufficient experience as machinists to design and produce tools for special purposes.

If it is contemplated to use the newer and very popular miniature series of tubes (6AK5,

* "Getting the Most Out of Your Drill Press," James Tate, Delta Manufacturing Company, Milwaukee, Wis.

6C4, 6BA6, etc.) in the construction of equipment certain additional tools will be required to mount the smaller components. Miniature tube sockets mount in a ⅝-inch hole, while 9-pin sockets mount in a ¾-inch hole. Greenlee socket punches can be obtained in these sizes, or a smaller hole may be reamed to the proper size. Needless to say, the punch is much the more satisfactory solution. Mounting screws for miniature sockets are usually of the 4-40 size.

NUMBERED DRILL SIZES

DRILL NUMBER	Diameter (In.)	Clears Screw	Correct for Tapping Steel or Brass†
1	.228	—	—
2	.221	12-24	—
3	.213	—	14-24
4	.209	12-20	—
5	.205	—	—
6	.204	—	—
7	.201	—	—
8	.199	—	—
9	.196	—	—
10*	.193	10-32	—
11	.191	10-24	—
12*	.189	—	—
13	.185	—	—
14	.182	—	—
15	.180	—	—
16	.177	—	12-24
17	.173	—	—
18*	.169	8-32	—
19	.166	—	12-20
20	.161	—	—
21*	.159	—	10-32
22	.157	—	—
23	.154	—	—
24	.152	—	—
25*	.149	—	10-24
26	.147	—	—
27	.144	—	—
28*	.140	6-32	—
29*	.136	—	8-32
30	.128	—	—
31	.120	—	—
32	.116	—	—
33*	.113	4-36 4-40	—
34	.111	—	—
35*	.110	—	6-32
36	.106	—	—
37	.104	—	—
38	.102	—	—
39*	.100	3-48	—
40	.098	—	—
41	.096	—	—
42*	.093	—	4-36 4-40
43	.089	2-56	—
44	.086	—	—
45*	.082	—	3-48

†Use next size larger drill for tapping bakelite and similar composition materials (plastics, etc.).

*Sizes most commonly used in radio construction.

18-3 Construction Practice

Chassis Layout The chassis first should be covered with a layer of wrapping paper, which is drawn tightly down on all sides and fastened with scotch tape. This allows any number of measurement lines and hole centers to be spotted in the correct positions without making any marks on the chassis itself. Place on it the parts to be mounted and play a game of chess with them, trying different arrangements until all the grid and plate leads are made as short as possible, tubes are clear of coil fields, r-f chokes are in safe positions, etc. Remember, especially if you are going to use a panel, that a good mechanical layout often can accompany sound electrical design, but that the electrical design should be given first consideration.

All too often parts are grouped to give a symmetrical panel, irrespective of the arrangement behind. When a satisfactory arrangement has been reached, the mounting holes may be marked. The same procedure now must be followed for the underside, always being careful to see that there are no clashes between the two (that no top mounting screws come down into the middle of a paper capacitor on the underside, that the variable capacitor rotors do not hit anything when turned, etc.).

When all the holes have been spotted, they should be center-punched *through* the paper into the chassis. Don't forget to spot holes for leads which must also come through the chassis.

For transformers which have lugs on the bottoms, the clearance holes may be spotted by pressing the transformer on a piece of paper to obtain impressions, which may then be transferred to the chassis.

Punching In cutting socket holes, one can use either a fly-cutter or socket punches. These punches are easy to operate and only a few precautions are necessary. The guide pin should fit snugly in the guide hole. This increases the accuracy of location of the socket. If this is not of great importance, one may well use a drill of 1/32 inch larger diameter than the guide pin. Some of the punches will operate without guide holes, but the latter always make the punching operations simpler and easier. The only other precaution is to be sure the work is properly lined up before applying the hammer. If this is not done, the punch may slide sideways when you strike and thus not only shear the chassis but also take off part of the die. This is easily avoided by always making sure that the piece is parallel to the faces of the punch, the die, and the base. The latter should be an anvil or other solid base of heavy material.

A punch by *Greenlee* forces socket holes through the chassis by means of a screw turned with a wrench. It is noiseless, and works much more easily and accurately than most others.

The male part of the punch should be placed in the vise, cutting edge up and the female portion forced against the metal with a wrench as in figure 1. These punches can be obtained in sizes to accommodate all tube sockets and even large enough to be used for meter holes. In the octal socket sizes they require the use of a 3/8 inch center hole to accommodate the bolt.

Transformer Cutouts Cutouts for transformers and chokes are not so simply handled. After marking off the part to be cut, drill about a 1/4-inch hole on each of the inside corners and tangential to the edges. After burring the holes, clamp the piece and a block of cast iron or steel in the vise. Then, take your burring chisel and insert it in one of the corner holes. Cut out the metal by hitting the chisel with a hammer. The blows should be light and numerous. The chisel acts against the block in the same way that the two blades of a pair of scissors work against each other. This same process is repeated for the other sides. A file is used to trim up the completed cutout.

Another method is to drill the four corner holes large enough to take a hack saw blade, then saw instead of chisel. The four holes permit nice looking corners.

Still another method is shown in figure 2. When heavy panel steel is used and a drill press or electric drill is available, this is the most satisfactory method.

Removing Burrs In both drilling and punching, a burr is usually left on the work. There are three simple ways of removing these. Perhaps the best is to take a chisel (be sure it is one for use on metal) and set it so that its bottom face is parallel to the piece. Then gently tap it with a hammer. This usually will make a clean job with a little practice. If one has access to a counterbore, this will also do a nice job. A countersink will work, although it bevels the edges. A drill of several sizes larger is a much used arrangement. The third method is by filing off the burr, which does a good job but scratches the adjacent metal surfaces badly.

Mounting Components There are two methods in general use for the fastening of transformers, chokes, and similar

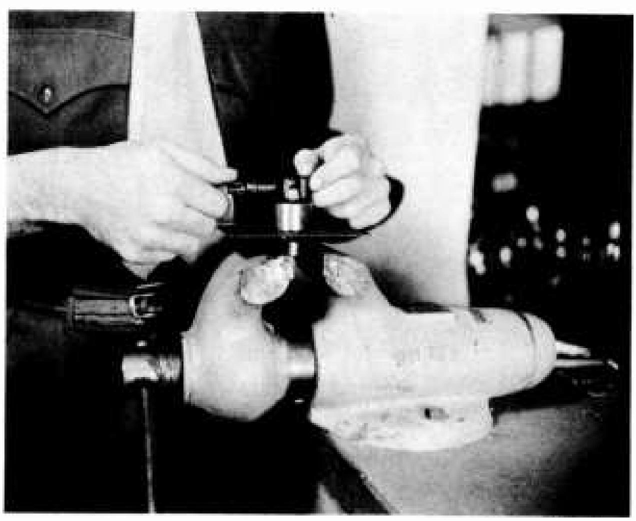
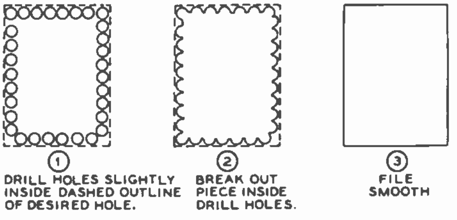


Figure 1.
PROPER METHOD OF USING A SOCKET PUNCH OF THE "GREENLEE" TYPE.



MAKING RECTANGULAR CUTOUT

Figure 2.

pieces of apparatus to chassis or breadboards. The first, using nuts and machine screws, is slow, and the commercial manufacturing practice of using self-tapping screws is gaining favor. For the mounting of small parts such as resistors and capacitors, "tie points" are very useful to gain rigidity. They also contribute materially to the appearance of finished apparatus.

Rubber grommets of the proper size, placed in all chassis holes through which wires are to be passed, will give a neater appearing job and also will reduce the possibility of short circuits.

Soldering Making a strong, low-resistance solder joint does not mean just dropping a blob of solder on the two parts to be joined and then hoping that they'll stick. There are several definite rules that *must* be observed.

All parts to be soldered must be absolutely clean. To clean a wire, lug, or whatever it may be, take your pocket knife and scrape it thoroughly, until fresh metal is laid bare. It is not enough to make a few streaks; scrape until the part to be soldered is bright.

Make a good mechanical joint before applying any solder. Solder is intended primarily to make a good *electrical* connection; mechanical rigidity should be obtained by bending the wire into a small hook at the end and nipping it firmly around the other part, so that it will hold well even before the solder is applied.

Keep your iron properly tinned. It is impossible to get the work hot enough to take the solder properly if the iron is dirty. To tin your iron, file it, while hot, on one side until a full surface of clean metal is exposed. Immediately apply rosin core solder until a thin layer flows completely over the exposed surface. Repeat for the other faces. Then take a clean rag and wipe off all excess solder and rosin. The iron should also be wiped frequently while the actual construction is going on; it helps prevent pitting the tip.

Apply the solder to the work, not to the iron. The iron should be held against the parts to be joined until they are thoroughly heated. The solder should then be applied against the parts, and the iron should be held in place until the solder flows smoothly and envelops the work. If it acts like water on a greasy plate, and forms a ball, the work is not sufficiently clean.

The completed joint must be held perfectly still until the solder has had time to solidify. If the work is moved before the solder has become *completely* solid, a "cold" joint will result. This can be identified immediately, because the solder will have a dull "white" appearance rather than one of shiny "silver." Such joints tend to be of high resistance, and will very likely have a bad effect upon a circuit. The cure is simple: merely reheat the joint and do the job correctly.

Wipe away all surplus flux when the joint has cooled if you are using a paste type flux. Be sure it is non-corrosive, and use it with plain (not rosin core) solder.

Finishes If the apparatus is constructed on a painted chassis (commonly available in black wrinkle and gray wrinkle), there is no need for application of a protective coating when the equipment is finished, assuming that you are careful not to scratch or mar the finish while drilling holes and mounting parts. However, many amateurs prefer to use unpainted (zinc or cadmium plated) chassis, because it is much simpler to make a chassis ground connection with this type of chassis. A thin coat of clear "linoleum" lacquer may be applied to the whole chassis after the wiring is completed to retard rusting. In localities near the sea coast it is a good idea to lacquer the various chassis cutouts even on a painted chassis, as rust will get a good start at these points unless the metal is protected where the drill or saw has exposed it. If too thick a coat is applied, the lacquer will tend to peel. It may be thinned with lacquer thinner to permit application of a light coat. A thin coat will adhere to any *clean* metal surface that is not too shiny.

An attractive dull gloss finish, almost velvety,

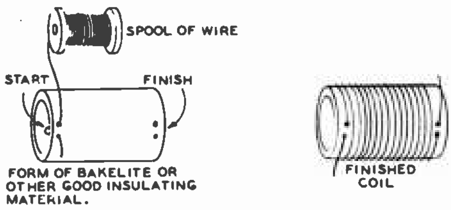


Figure 3.

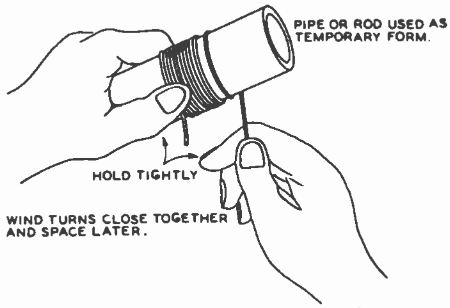


Figure 4.

can be put on aluminum by sand-blasting it with a very weak blast and fine particles and then lacquering it. Soaking the aluminum in a solution of lye produces somewhat the same effect as a fine grain sand blast.

There are also several brands of dull gloss black enamels on the market which adhere well to metals and make a nice appearance. Air-drying wrinkle finishes are sometimes successful, but a bake job is usually far better. Wrinkle finishes, properly applied, are very durable and are pleasing to the eye. If you live in a large community, there is probably an enamelling concern which can wrinkle your work for you at a reasonable cost. A very attractive finish, for panels especially, is to spray a wrinkle finish with aluminum paint. In any painting operation (or plating, either, for that matter), the work should be very thoroughly cleaned of all greases and oils.

To protect brass from tarnish, thoroughly cleanse and remove the last trace of grease by the use of potash and water. The brass must be carefully rinsed with water and dried; but in doing it, care must be taken not to handle any portion with the bare hands or anything else that is greasy. Then lacquer.

Drilling Glass This is done with a common drill by using a mixture of turpentine and camphor. When the point of the drill has come through, it should be taken out and the hole worked through with the point of a three-cornered file, having the edges ground sharp. Use the corners of the file, scraping the glass rather than using the file as a reamer. Great care must be taken not to crack the glass or flake off parts of it in finishing the hole after the point of the drill has come through. Use the mixture freely during the drilling and scraping. The above mixture will also be found useful in drilling hard cast iron. Drilling glass must be done very slowly. It is a good idea to practice by drilling several holes in scrap glass before tackling the actual piece to be drilled, to acquire the knack.

Etching Solution Add three parts nitric acid to one part muriatic acid. Cover the piece to be etched with beeswax. This can be done by heating the piece in a gas or alcohol flame and rubbing the wax over the surface. Use a sharp steel point or hard lead pencil point as a stylus. A pointed glass dropper can be used to put the solution at the place needed. After the solution foams for two or three minutes, remove with blotting paper and put oil on the piece, and then heat and remove the wax.

Chromium Polish So much chromium is now used in radio sets and on panels that it is well to know that this finish may be polished. The only materials required are absorbent cotton or soft cloth, alcohol, and ordinary lampblack.

A wad of cotton or the cloth is moistened in the alcohol and pressed into the lampblack. The chromium is then polished by rubbing the lampblack adhering to the cotton briskly over its surface. The mixture dries almost instantly and may be wiped off with another wad of cotton.

The alcohol serves merely to moisten the lampblack to a paste and make it stick to the cotton. The mixture cleans and polishes very quickly and cannot scratch the chromium surface. It polishes nickel-work just as effectively as it does chromium. Care should be taken to

Figure 5.
TOOLS FOR RADIO CONSTRUCTION.

see that the lampblack does not contain any hard, gritty particles which might produce scratches during the polishing.

Winding Coils Coils are of two general types, those using a form and "air-wound" types. Neither type offers any particular constructional difficulties. Figure 3 illustrates the procedure used in form winding a coil. If the winding is to be spaced, the spacing can be done either by eye or a string or another piece of wire may be wound simultaneously with the coil wire and removed after the winding is in place. The usual procedure is to clamp one end of the wire in a vise, attaching the other end to the coil form and with the coil form in hand, walk slowly towards the vise winding the wire but at the same time keeping a strong tension on the wire as the form is rotated. After the coil is wound, if there is any possibility of the turns slipping, the completed coil is either entirely coated with a coil or Duco cement or cemented in those spots where slippage might occur.

V-h-f and u-h-f coils are commonly wound of heavy enameled wire on a form and then removed from the form as in figure 4. If the coil is long or has a tendency to buckle, strips of polystyrene or a similar material may be cemented longitudinally inside the coil. Due allowance must be made for the coil springing out when removed from the form, when selecting the diameter of the form.

On air wound coils of this type, spacing between turns is accomplished after removal from the form, by running a pencil, the shank of a screwdriver or other round object spirally between the turns from one end of the coil to the other, again making due allowance for spring.

Air-wound coils approaching the appearance of commercially manufactured ones, can be constructed by using a round wooden form which has been sawed diagonally from end to end. Strips of insulating material are temporarily attached to this mandrel, the wire then being wound over these strips with the desired separation between turns and cemented to the strips. When dry, the split mandrel may be removed by unwedging it.

CHAPTER NINETEEN

Mobile Equipment and Installation

While mobile operation is permitted on certain other bands, operation at the present time is mostly confined to the 80, 20, 10-11, and 2 meter bands, and the following discussion will be confined to operation on those bands.

The problems involved in achieving a satisfactory two-way installation vary somewhat with the band, but many of the problems are common to all bands. For instance, ignition noise is more troublesome on 10 meters than on 75 meters, but on the other hand an efficient antenna system is much more easily accomplished on 10 meters than on 75 meters. And getting a worthwhile amount of transmitter output without excessive battery drain is a problem on all bands.

19-1 Mobile Reception

When a broadcast receiver is in the car, the most practical receiving arrangement on 75, 20, and 10 meters involves a converter feeding into the auto set. The advantages of good selectivity with good image rejection obtainable from a double conversion superheterodyne are achieved in most cases without excessive "birdie" troubles, a common difficulty with a double conversion superheterodyne constructed as an integral receiver in one cabinet. However, it is important that the b-c receiver employ an r-f stage in order to provide adequate isolation between the converter and the high frequency oscillator in the b-c receiver. The r-f stage also is desirable from the standpoint of image rejection if the converter does not employ a tuned output circuit (tuned to the frequency of the auto set, usually about 1500 kc.). A few of the late model auto receivers, even in the better makes, do not employ an r-f stage.

For 10 and/or 20 and 75 meter operation a large percentage of amateurs purchase a commercially built converter for reception and build their own transmitter. But for those who prefer to build their own, construction details for a simple one-tube 10-75 meter converter are in this section.

The usual procedure is to obtain converter plate voltage from the auto receiver. Experience has shown that if the converter does not draw more than about 15 or at most 20 ma. total plate current no damage to the auto set or loss in performance will occur other than a slight reduction in vibrator life. The converter drain can be minimized by avoiding a voltage regulator tube on the converter h-f oscillator. On 10 and lower frequencies it is possible to design an oscillator with sufficient stability that no voltage regulator is required in the converter.

With some cars satisfactory 75-meter operation can be obtained without a noise clipper if resistor type spark plugs (such as those made by Autolite) are employed. However, a noise clipper is helpful if not absolutely necessary, and it is recommended that a noise clipper be installed without confirming the necessity therefor. It has been found that quiet reception sometimes may be obtained on 75 meters simply by the use of resistor type plugs, but after a few thousand miles these plugs often become less effective and no longer do a fully adequate job. Also, a noise clipper insures against ignition noise from passing trucks and "un-suppressed" cars. On 10 meters a noise clipper is a "must" in any case.

Modifying the Auto Receiver There are certain things that should be done to the auto set when it is to be used with a converter, and they might just as well be done all at the same time, because "dropping" an auto receiver and getting into the chassis to work on it takes quite a little time.

First, however, check the circuit of the auto receiver to see whether it is one of the few receivers which employ circuits which complicate connection of a noise clipper or a converter. If the receiver is yet to be purchased, it is well to investigate these points ahead of time.

If the receiver uses a negative B resistor strip for bias (as evidenced by the cathode of the audio output stage being grounded), then the additional plate current drain of the converter will upset the bias voltages on the various stages and probably cause trouble. Because the converter is not on all the time, it is not practical simply to alter the resistance of the bias strip, and major modification of the receiver probably will be required.

If the receiver uses a separate "control head" with audio wires running between the receiver and control head volume control, tone control, etc., it is possible that stray coupling will permit ignition noise to "leak around" the noise clipper and greatly reduce its effectiveness.

The best type of receiver for attachment of a converter and noise clipper uses an r-f stage; permeability tuning; single unit construction (except possibly for the speaker); push button tuning rather than a tuning motor; a high vacuum rectifier such as a 6X4 (rather than an 0Z4 or a synchronous rectifier); a 6SQ7 (or miniature or Loctal equivalent) with *grounded cathode* as second detector, first audio, and a.v.c.; power supply negative grounded directly (no common bias strip); a PM speaker (to minimize battery drain); and an internal r-f gain control (indicating plenty of built-in reserve gain which may be called upon if necessary). Many current model auto radios have all of the foregoing features, and numerous models have most of them, something to keep in mind if the set is yet to be purchased.

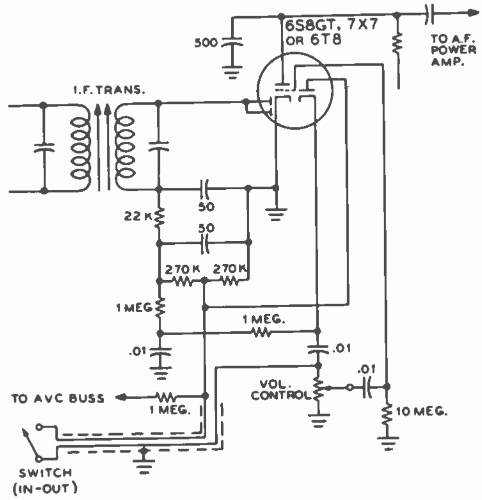


Figure 1.
SERIES-GATE NOISE LIMITER FOR AUTO RECEIVER.
Auto receivers using a 6SQ7, 7B6, 7X7, or 6AT6 as second detector and a.v.c. can be converted to the above circuit with but few wiring changes. The circuit has the advantage of not requiring an additional tube socket for the limiter diode.

Noise Limiters A noise limiter either may be built into the set or purchased as a commercially manufactured unit for "outboard" connection via shielded wires. If the receiver employs a 6SQ7 (or Loctal or miniature equivalent) in a conventional circuit, it is a simple matter to build in a noise clipper by substituting a 6S8 octal, 7X7 Loctal, or a 6T8 9-pin miniature as shown in figure 1. When substituting a 6T8 for a 6AT6 or similar 7-pin miniature, the socket must be changed to a 9-pin miniature type. This requires reaming the socket hole slightly.

If the receiver employs cathode bias on the 6SQ7 (or equivalent), and perhaps delayed a.v.c., the circuit usually can be changed to the grounded-cathode circuit of figure 1 without encountering trouble.

Some receivers take the r-f excitation for the a-v-c diode from the plate of the i-f stage. In this case, leave the a.v.c. alone and ignore the a-v-c buss connection shown in figure 1 (eliminating the 1-megohm decoupling resistor). If the set uses a separate a-v-c diode which receives r-f excitation via a small capacitor connected to the detector diode, then simply change the circuit to correspond to figure 1.

In case anyone might be considering the use of a crystal diode as a noise limiter in conjunction with the tube already in the set, it might be well to point out that crystal diodes perform quite poorly in series-gate noise clippers of the type shown.

It will be observed that no tone control is shown. Multi-position tone controls tied in with the second detector circuit often permit excessive "leak through." Hence it is recommended that the tone control components be completely removed unless they are confined to the grid of the a-f output stage. If removed, the highs can be attenuated any desired amount by connecting a mica capacitor from plate to screen on the output stage. Ordinarily from .005 to .01 μfd. will provide a good compromise between fidelity and reduction of background hiss on weak signals.

Usually the switch SW will have to be mounted some distance from the noise limiter components. If the leads to the switch are over approximately 1½ inches long, a piece of shield braid should be slipped over them and grounded. The same applies to the "hot" leads to the volume control if not already shielded. Closing the switch disables the limiter. This may be desirable for reducing distortion on broadcast reception or when checking the intensity of ignition noise to determine the effectiveness of suppression measures taken on the car. The switch also permits one to check the effectiveness of the noise clipper.

The 22,000-ohm decoupling resistor at the bottom end of the i-f transformer secondary is not critical, and if some other value already is incorporated inside the shield can it may be left alone so long as it is not over 47,000 ohms, a common value. Higher values must be replaced with a lower value even if it requires a can opener, because anything over 47,000 ohms will result in excessive loss in gain. There is some loss in a-f gain inherent in this type of limiter anyhow (slightly over 6 db), and it is important to minimize any additional loss.

It is important that the total amount of capacitance in the RC decoupling (r-f) filter not exceed about 100 $\mu\mu$fd. With a value much greater than this "pulse stretching" will occur and the effectiveness of the noise clipper will be reduced. Excessive capacitance will reduce the amplitude and increase the duration of the ignition pulses before they reach the clipper. The reduction in pulse amplitude accomplishes no good since the pulses are fed to the clipper anyhow, but the greater duration of the lengthened pulses increases the audibility and the "blanking interval" associated with each pulse. If a shielded wire to an external clipper is employed, the r-f by-pass on the "low" side of the RC filter may be eliminated since the capacitance of a few feet of shielded wire will accomplish the same result as the by-pass capacitor.

The switch SW is connected in such a manner that there is practically no change in gain with the limiter in or out. If the auto set does not have any reserve gain and more gain is needed on weak broadcast signals, the switch can be connected from the hot side of the volume control to the junction of the 22,000, 270,000 and 1 megohm resistors instead of as shown. This will provide approximately 6 db more gain when the clipper is switched out.

Many late model receivers are provided with an internal r-f gain control in the cathode of the r-f and/or i-f stage. This control should be advanced full on to provide better noise limiter action and make up for the loss in audio gain introduced by the noise clipper.

Installation of the noise clipper often detunes the secondary of the last i-f transformer. This should be repeaked before the set is permanently replaced in the car unless the trimmer is accessible with the set mounted in place.

Selectivity While not of serious concern on 10 meters, the lack of selectivity exhibited by a typical auto receiver will result in QRM difficulty on 20 and 75 meters. A typical auto set has only two i-f transformers of relatively low-Q design, and the second one is loaded by the diode detector. The skirt selectivity often is so poor that a strong local will depress the a.v.c. when listening to a weak station as much as 15 kc. different in frequency.

One solution is to add an outboard i-f stage employing two good quality double-tuned transformers (not the midget variety) connected "back-to-back" through a small coupling capacitance. The amplifier tube (such as a 6BA6) should be biased to the point where the gain of the outboard unit is relatively small (1 or 2), assuming that the receiver already has adequate gain. If additional gain is needed, it may be provided by the outboard unit. Low-capacitance shielded cable should be used to couple into and out of the outboard unit, and the unit itself should be thoroughly shielded.

Such an outboard unit will sharpen the nose selectivity slightly and the skirt selectivity greatly. Operation then will be comparable to a home-station communications receiver, though selectivity will not be as good as a receiver employing a 50-kc. or 85-kc. "Q-5'er."

Another solution to the selectivity problem is to obtain a surplus Motorola P69-18 police receiver. This receiver has very good selectivity, is crystal controlled, and has a built-in noise clipper (which may be modified if desired). The receiver also has a squelch circuit, but this should be disabled for amateur use. These receivers often may be obtained at a small fraction of their original price from police departments that have switched over to v.h.f.

If one of these receivers is to be used in conjunction with a standard amateur-band converter, no cutting into the auto receiver is necessary since the receiving installation is completely independent. This arrangement is particularly desirable for one who turns in his car on a new one every year. The earlier series of the Motorola police receiver, such as the P69-13, have one less i-f transformer and hence are less desirable from a selectivity standpoint. To determine the required crystal frequency for one of these receivers, take the converter i-f output frequency (which may be anything between 1430 and 2500 kc.) and add 262 kc.

Obtaining Power for the Converter While the set is on the bench for installation of the noise clipper, provision should be made for obtaining filament and plate voltage for the converter, and for the exciter and speech amplifier of the transmitter, if such an arrangement is to be used. To permit removal of either the converter or the auto set from the car without removing the other, a connector should be provided. The best method is to mount a small receptacle on the receiver cabinet or chassis, making connection via a matching plug. An Amphenol type 77-26 receptacle is compact enough to fit in a very small space and allows four connections (including ground for the shield braid). The matching plug is a type 70-26.

To avoid the possibility of vibrator hash being fed into the converter via the heater and plate voltage supply leads, it is important that the heater and plate voltages be taken from points well removed from the power supply portion of the auto receiver. If a single-ended audio output stage is employed, a safe place to obtain these voltages is at this tube socket, the high voltage for the converter being taken from the screen. In the case of a push-pull output stage, however, the screens sometimes are fed from the input side of the power supply filter. The ripple at this point, while sufficiently low for a push-pull audio output stage, is not adequate for a converter without additional filtering. If the schematic shows that the screens of a push-pull stage are connected to the input side instead of the output side of the power supply filter (usually two electrolytics straddling a resistor in an R-C filter), then follow the output of the filter over into the r-f portion of the set and pick it up there at a convenient point, before it goes through any additional series dropping or isolating resistors.

The voltage at the output of the filter usually runs from 200 to 250 volts with typical converter drain and the motor not running. This will increase perhaps 10 per cent when the generator is charging. The converter drain will drop the B voltage slightly at the output of the filter, perhaps 15 to 25 volts, but this reduction is not enough to have a noticeable effect upon the operation of the receiver. If the B voltage is higher than desirable or necessary for proper operation of the converter, a 2-watt carbon resistor of suitable resistance should be inserted in series with the plate voltage lead to the power receptacle. Usually something between 2200 and 4700 ohms will be found about right.

Receiver Disabling on Transmit When the battery drain is high on transmit, as is the case when a PE-103A is run at maximum rating and other drains such as the transmitter heaters and auto headlights must be considered, it is desirable to disable the vibrator power supply in the receiver during transmissions. The vibrator power supply usually draws several amperes, and as the receiver must be disabled in some manner anyhow during transmissions, opening the 6-volt supply to the vibrator serves both purposes. It has the further advantage of introducing a slight delay in the receiver recovery, due to the inertia of the power supply filter, thus avoiding the possibility of a feedback "yoop" when switching from transmit to receive.

To avoid troubles from vibrator hash, it is best to open the ground lead from the vibrator by means of a midget s.p.d.t. 6-volt relay and thus isolate the vibrator circuit from the external control and switching circuit wires. The relay is hooked up as shown in figure 2. Standard 8-ampere contacts will be adequate for this application.

The relay should be mounted as close to the vibrator as practicable. Ground one of the coil terminals and run a shielded wire from the other coil terminal to one of the power receptacle connections, grounding the shield at both ends. By-pass each end of this wire to ground with .01 μfd., using the shortest possible leads. A lead is run from the corresponding terminal on the mating plug to the control circuits, to be discussed later.

If the normally open contact on the relay is connected to the hot side of the voice coil winding as shown in figure 2 (assuming one side of the voice coil is grounded in accordance with usual practice), the receiver will be killed instantly when switching from receive to transmit, in spite of the fact that the power supply filter in the receiver takes a moment to discharge. However, if a "slow start" power supply (such as a dynamotor or a vibrator pack with a large filter) is used with the transmitter, shorting the voice coil probably will not be required.

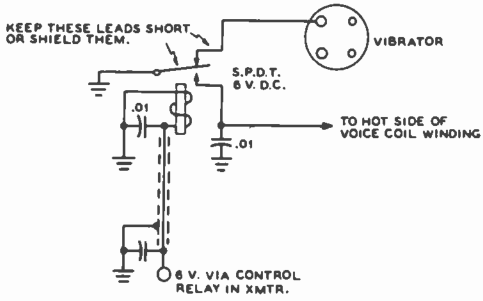


Figure 2.

METHOD OF ELIMINATING THE BATTERY DRAIN OF THE RECEIVER VIBRATOR PACK DURING TRANSMISSION.

If the receiver chassis has room for a midget s.p.d.t. relay, the above arrangement not only silences the receiver on transmit but saves several amperes battery drain.

Using the Receiver Plate Supply On Transmit An alternative and highly recommended procedure is to make use of the receiver B supply on transmit, instead of disabling it. One disadvantage of the popular PE-103A dynamotor is the fact that its 450-500 volt output is too high for the low power r-f and speech stages of the transmitter. Dropping this voltage to a more suitable value of approximately 250 volts by means of dropping resistors is wasteful of power, besides causing the plate voltage on the oscillator and any buffer stages to vary widely with tuning. By means of a midget 6-volt s.p.d.t. relay mounted in the receiver, connected as shown in figure 3, the B supply of the auto set is used to power the oscillator and other low power stages (and possibly screen voltage on the modulator). On transmit the B voltage is removed from the receiver and converter, automatically silencing the receiver. When switching to "receive" the transmitter oscillator is killed instantly, thus avoiding trouble from dynamotor "carry over."

The efficiency of this arrangement is good because the current drain on the main high voltage supply for the modulated amplifier and modulator plate(s) is reduced by the amount of current borrowed from the receiver. At least 80 ma. can be drawn from practically all auto sets, at least for a short period, without damage.

It will be noted that with the arrangement of figure 3, plate voltage is supplied to the audio output stage at all times. However, when the screen voltage is removed, the plate current drops practically to zero.

The 200-ohm resistor in series with the normally open contact is to prevent excessive sparking when the contacts close. If the relay feeds directly into a filter choke or large capacitor there will be excessive sparking at the contacts. Even with the arrangement shown, there will be considerable sparking at the contacts; but relay contacts can stand such sparking quite a while, even on d.c., before becoming worn or pitted enough to require attention. The 200-ohm resistor also serves to increase the effectiveness of the .01-μfd. r-f by-pass capacitor.

Auxiliary Antenna Trimmer One other modification of the auto receiver which may or may not be desirable depending upon the circumstances is the addition of an auxiliary antenna trimmer capacitor. If the converter uses an untuned output circuit and the antenna trimmer on the auto set is peaked with the converter cut in, then it is quite likely that the trimmer adjustment will not be optimum for broadcast-band reception when the converter is cut out. For reception of strong broadcast band signals this usually will not be serious, but where reception of weak broadcast signals is desired the loss in gain often cannot be tolerated, especially in view of the fact that the additional length of antenna cable required for the converter installation tends to reduce the strength of broadcast band signals.

If the converter has considerable reserve gain, it may be practicable to peak the antenna trimmer on the auto set for broadcast-band reception rather than resonating it to the converter output circuit. But oftentimes this results in insufficient converter gain, excessive image troubles from loud local amateur stations, or both.

The difficulty can be circumvented by incorporation of an auxiliary antenna trimmer connected from the "hot" antenna lead on the auto receiver to ground, with a switch in series for cutting it in or out. This capacitor and switch can be connected across either the converter end or the set end of the cable between the converter and receiver. This auxiliary trimmer should have a range of about 3 to 50 $\mu\mu$fd., and may be of the inexpensive compression mica type.

With the trimmer cut out and the converter turned off (by-passed by the "in-out" switch), peak the regular antenna trimmer

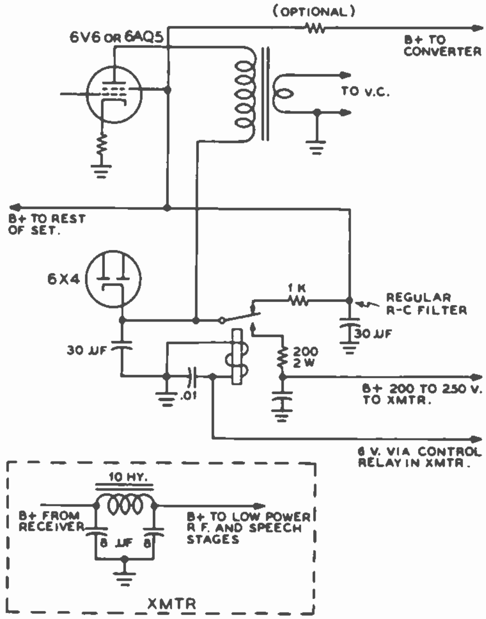


Figure 3.
USING THE RECEIVER PLATE SUPPLY FOR THE TRANSMITTER.

This circuit silences the receiver on transmit, and in addition makes it possible to use the receiver plate supply for feeding the exciter and speech amplifier stages in the transmitter.

on the auto set at about 1400 kc. Then turn on the converter, with the receiver tuned to 1500 kc., switch in the auxiliary trimmer, and peak this trimmer for maximum background noise. The auxiliary trimmer then can be left switched in at all times except when receiving very weak broadcast band signals. One of the newer mobile converters (Gonset Tri-Band) includes such a provision.

Some auto sets, particularly certain General Motors custom receivers, employ a high-Q high-impedance input circuit which is very critical as to antenna capacitance. Unless the shunt capacitance of the antenna (including cable) approximates that of the antenna installation for which the set was designed, the antenna trimmer on the auto set cannot be made to hit resonance with the converter cut out. This is particularly true when a long antenna cable is used to reach a whip mounted at the rear of the car. Usually the condition can be corrected by unsoldering the internal connection to the antenna terminal connector on the auto set and inserting in series a 100-μμfd. mica capacitor. Alternatively an adjustable trimmer covering at least 50 to 150 μμfd. may be substituted for the 100-μμfd. fixed capacitor. Then the adjustment of this trimmer and that of the regular antenna trimmer can be juggled back and forth until a condition is achieved where the input circuit of the auto set is resonant with the converter either in or out of the circuit. This will provide maximum gain and image rejection under all conditions of use.

Reducing Battery Drain of the Receiver When the receiving installation is used frequently, and particularly when the receiver is used with the car parked, it is desirable to keep the battery drain of the receiver-converter installation at an absolute minimum. A substantial reduction in drain can be made in many receivers, without appreciably affecting their performance. The saving, of course depends upon the design of the particular receiver and upon how much trouble and expense one is willing to go to. Some receivers normally draw (without the converter connected) as much as 10 amperes. In many cases this can be cut to about 5 amperes by incorporating all practicable modifications. Each of the following modifications is applicable to many auto receivers.

If the receiver uses a speaker with a field coil, replace the speaker with an equivalent PM type.

Practically all 0.3-ampere r-f and a-f voltage amplifier tubes have 0.15-ampere equivalents. In many cases it is not even necessary to change the socket wiring. However, when substituting i-f tubes it is recommended that the i-f trimmer adjustments be checked. Generally speaking it is not wise to attempt to substitute for the converter tube or a-f power output tube.

If the a-f output tube employs conventional cathode bias, substitute a cathode resistor of twice the value originally employed, or add an identical resistor in series with the one already in the set. This will reduce the B drain of the receiver appreciably without seriously reducing the maximum undistorted output. Because the vibrator power supply is much less than 100 per cent efficient, a saving of one watt of B drain results in a saving of nearly 2 watts of battery drain. This also minimizes the overload on the B supply when the converter is switched in, assuming that the converter uses B voltage from the auto set.

If the receiver uses push-pull output and if one is willing to accept a slight reduction in the maximum volume obtainable without distortion, changing over to a single ended stage is simple if the receiver employs conventional cathode bias. Just pull out one tube, double the value of cathode bias resistance, and add a 25-μfd. by-pass capacitor across the cathode resistor if not already by-passed. In some cases it may be possible to remove a phase inverter tube along with one of the a-f output tubes.

If the receiver uses a motor driven station selector with a control tube (d-c amplifier), usually the tube can be removed without upsetting the operation of the receiver. One then must of course use manual tuning.

While the changeover is somewhat expensive, the 0.6 ampere drawn by a 6X4 or 6X5 rectifier can be eliminated by substituting six 115-volt r-m-s 50-ma. selenium rectifiers (such as Federal type 402D3200). Three in series are substituted for each half of the full-wave rectifier tube. Be sure to observe the correct polarity. The selenium rectifiers also make a good substitution for an 0Z4 or 0Z4-GT which is causing hash difficulties when using the converter.

Offsetting the total cost of nearly $4.00 is the fact that these rectifiers probably will last for the entire life of the auto set. Before purchasing the rectifiers, make sure that there is room available for mounting them. While these units are small, most of the newer auto sets employ very compact construction.

Two-Meter Reception For reception on the 144-Mc. amateur band, and those higher in frequency, the simple converter-auto-set combination has not proven very satisfactroy. The primary reason for this is the fact that the relatively sharp i-f channel of the auto set imposes too severe a limitation on the stability of the high-frequency oscillator in the converter. And if a crystal-controlled beating oscillator is used in the converter, only a portion of the band may be covered by tuning the auto set.

The most satisfactory arrangement has been found to consist of a separately mounted i.f., audio, and power supply system, with the

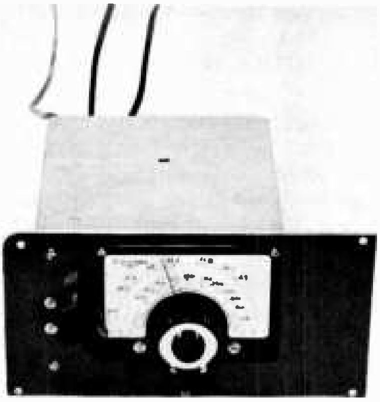


Figure 4.
FRONT OF THE TWO-BAND MOBILE CONVERTER.

control head of the converter mounted near the steering column. The i-f system should have a bandwidth of 30 to 100 kc. and may have a center frequency of 10.7 Mc. if standard i-f transformers are to be used. The control head may include the 144-Mc. r-f, mixer, and oscillator sections, and sometimes the first i-f stage. Alternatively, the control head may include only the h-f oscillator, with a broadband r-f unit included within the main receiver assembly along with the i.f. and audio system. Commercially manufactured kits and complete units using this general lineup are available.

An alternative arrangement is to build a converter, 10.7-Mc. i-f channel, and second detector unit, and then to operate this unit in conjunction with the auto-set power supply, audio system, and speaker. Such a system makes economical use of space and power drain, and can be switched to provide normal broadcast-band auto reception or reception through a converter for the h-f amateur bands.

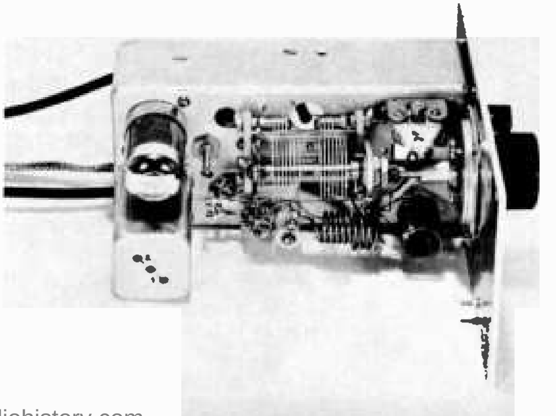
One-Tube Mobile Converter For 10 and 75

The simple converter shown in figures 4 through 7 provides reception over the frequency ranges from 26.5 through 30 Mc. and 3500 through 4000 kc. when operated into a standard broadcast-band auto radio receiver. The converter uses a single tube, a single-ended Loctal type 7S7, and operates into an intermediate frequency of 1500 to 1600 kc. Power supply requirements for the converter are low enough so that power may be furnished by the auto receiver.

Circuit of the Converter A type 7S7 tube was chosen for use in the converter primarily as a result of its relatively high conversion conductance of 525 micromhos. This value of conductance results in a slightly higher overall gain in the converter than could be obtained with the more standard types of tube such as the 6K8 or the 6SA7. Also, the single-ended Loctal construction of the tube contributes to circuit layout and to stability with respect to the mechanical vibration encountered.

The 7S7 tube is a triode-heptode converter with internal injection between the triode grid and the number 3 grid of the heptode. Seriesed tank circuits are employed both in the oscillator portion of the circuit and in the antenna circuit. Through the use of the series-tank arrangement a lower value of loss is obtainable since the high-potential end of the tank circuit for the high-frequency band is effectively isolated from the selector switch. Then, when the switch is placed in the position for the low-frequency band, the high-frequency tank is in series with the low-frequency tank circuit. The effect of having the high-frequency tank in series with that for the low-fre-

Figure 5.
SIDE VIEW OF THE CONVERTER.



quency band is relatively minor. Reference to the schematic diagram of figure 7 will indicate the circuit details of the bandswitching method.

The oscillator tank circuits, L_1 and L_2, must be solidly constructed and rigidly mounted. All components must be firmly held in place in such a manner that they cannot vibrate when the car is in motion. This requirement is particularly important in regard to the components for the 28-Mc. band. Any vibration in the oscillator tank circuit will show up as severe frequency modulation of an incoming signal when the car is in motion. Even a small amount of vibration of the components can result in a degree of frequency modulation which will make it completely impossible to read an incoming signal with the car moving. This effect will disappear, of course, as soon as the car is stopped.

The Antenna Circuit It will be noted that two tank circuits are connected in series in the antenna circuit the same as in the oscillator circuit. The tank circuit for the 3.5-Mc. band is at the low-potential end, and is shorted out when the selector switch is in the 28-Mc. position. The main tuning capacitor C_{1B}, is connected permanently to a tap on the 28-Mc. coil. The negligible impedance of this tuned circuit on the 3.5-Mc. band results in the tuning capacitor being effectively connected to the top end of the tank coil on the lower frequency 3.5-Mc. band.

Conventional inductive coupling from the antenna transmission line to the tank coil is used on the 28-Mc. band. Completely satisfactory results will be obtained with a conventional whip, 6 to 11 feet in length, as the receiving antenna on the 28-Mc. band. However, a quite different condition exists on the 3.5-Mc. band. Any conventional whip antenna will appear as a moderate value of capacitance with a negligible value of radiation resistance to the end of the antenna transmission line. The matter of the antenna for mobile operation on the 3.5-Mc. band has been discussed in some detail at the beginning of this chapter. But suffice to say that if a conventional unloaded whip is to be used for receiving, results will certainly not be very satisfactory as only the loudest signals on the band will be heard. Also, the feed line from the antenna terminal of the converter to the base of the antenna should be as short as possible and should be of the lowest capacitance cable obtainable. Even then, the entire feed line and antenna system must be treated as a capacitance with a small amount of signal pickup. Hence the tuning system indicated for the 3.5-Mc. band on the schematic diagram of figure 7.

If a good antenna is to be installed on the car for 3.5-Mc. mobile work (such as some type of loaded affair), then inductive coupling from the feed line to the antenna coil L_4 may be used, the same as has been done for the 28-Mc. band on L_3. The bottom end of capacitor C_4 should be grounded with this type of antenna coupling so that this capacitor may act as the trimmer for the antenna tank circuit. Much improved reception will be obtained with this type of antenna and the inductive method of antenna coupling. But any short loaded antenna for the 3.5-Mc. band will necessarily be rather sharp in its tuning characteristics. So only a selected portion of the amateur band will be received with maximum signal pickup for any given set of tuning conditions at the antenna loading coil.

Output Circuit of the 7S7 The output tank in the plate circuit of the 7S7 converter stage may be any i-f transformer with a primary winding capable of

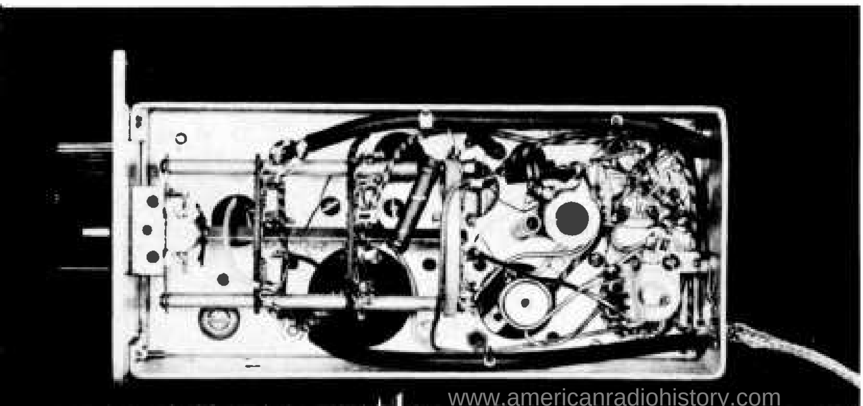


Figure 6. UNDERCHASSIS OF THE CONVERTER.

tuning to 1500 kc. Some experiment will be required in determining the number of turns on the secondary of L_3. The input circuit of most broadcast auto receivers is quite high in impedance, and is designed to operate from a signal source with a moderate amount of capacitance to ground. Hence the first test might be made with a conventional 1500-kc. double-tuned i-f transformer with the primary unchanged and with the tuning capacitance removed from the secondary. It will probably be found that there is too much circuit capacitance for resonance in the secondary winding under these conditions so some turns will have to be removed from the secondary. The actual tuning procedure of the installation will be described in the following paragraphs.

Alignment of the Converter The alignment procedure for the converter is relatively simple, but it can be carried out much more conveniently and successfully on a table in the operating room rather than with the converter installed in the car. Connect power supply voltages to the converter, and bring a short lead from the antenna post of the station communications receiver into the vicinity of the oscillator circuits of the converter. The intermediate frequency to be used with the converter should now be established. It is merely necessary to choose some frequency between 1450 and 1550 kc. which is not in use by some local broadcast station, and which can be tuned by the receiver in the car. If we assume that 1500 kc. is to be used, this value should be added to the frequency coverage of the oscillator in the converter. The oscillator is on the high side for both bands. This means that the oscillator must cover 5000 to 5500 kc. with some leeway at each end for the 3.5 to 4.0 Mc. band; and the oscillator must cover from about 28.0 to 31.5 Mc. to give reception from 26.5 Mc. through 30 Mc.

Adjustment of the oscillator coverage should be accomplished first on the 28-Mc. range. Only a variation in the setting of the zero-coefficient ceramic trimmer C_2 will be required if the oscillator coil has been made and tapped as specified. Then the bandswitch should be placed in the 3.5-Mc. position and the tuning slug in L_2 varied until coverage from about 4950 to 5550 kc. is obtained.

The output coaxial cable from the converter is now connected to the input of the communications receiver and the receiver tuned to the frequency which has been chosen as the i.f.—1500 kc. in the example mentioned. The tuned circuit in the plate of the 7S7 is peaked for maximum noise output. The bandswitch is now placed in the 28-Mc. position and capacitor C_3 peaked for maximum noise level on the 28-Mc. band. If a signal generator is available it may be used to check the tracking of the converter over the 28-Mc. band and to give an idea as to the relative sensitivity of the converter. If a signal generator is not available, the converter may be peaked with the aid of signals received in the 27 to 29.7 Mc. range.

Alignment of the 3.5-Mc. range of the converter must be accomplished with the equipment installed in the automobile, if the antenna input circuit of figure 7 has been used. However, if the feed line from the antenna to the converter actually is operating as a matched line instead of as a capacitance (as discussed in a previous paragraph) the inductive input circuit may be used. With this type of input circuit the converter may be tracked and aligned on the bench before installation in the car. With the type of input circuit shown in figure 7 the converter should be installed and capacitor C_1 peaked for maximum noise level in the center of the desired band, probably about 3925 kc., with the antenna extended the normal amount.

Matching the Converter to the Broadcast Receiver Automobile broadcast receivers are designed to operate from an antenna with a specified capacitance range to ground. A trimmer is provided on these receivers to compensate for variations in the capacitance to ground of the actual antenna in use. With the receiver tuned for a particular capacitance to ground, a large deterioration in performance will be encountered when the value of this capacitance is changed without a compensating change in the setting of the trimmer in the receiver. Consequently it is important that the input of the broadcast receiver see the same value of capacitance whether the converter is in operation or the receiver is being fed directly from the antenna.

After the converter is installed, with plate and heater voltage taken from within the broadcast receiver by means of a shielded cable and plug, the switch on the front of

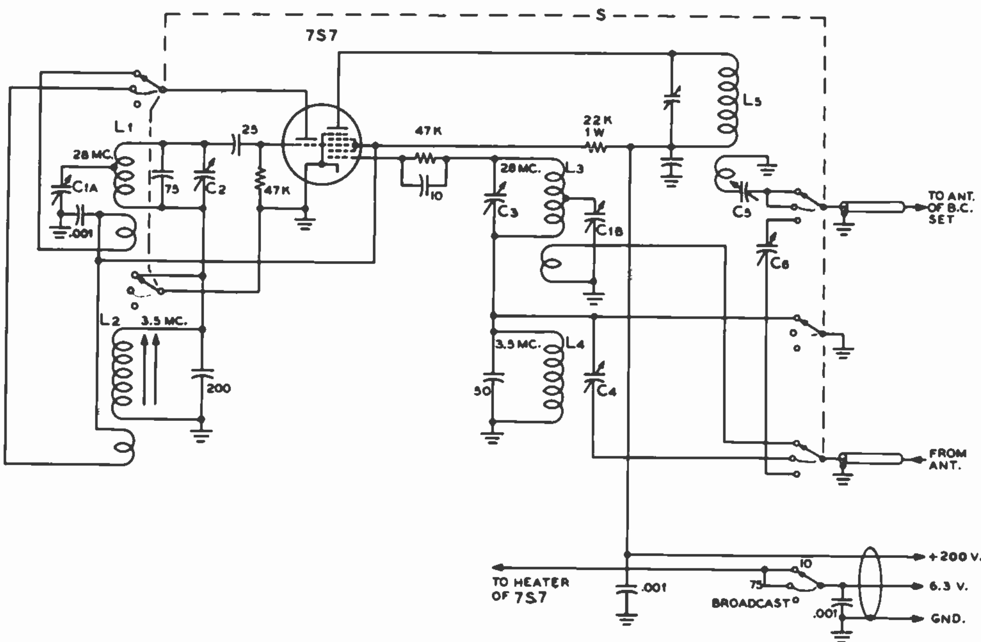


Figure 7.

SCHEMATIC OF THE TWO-BAND CONVERTER.

C_{1A}, C_{1B}—Dual 100-$\mu\mu$fd. variable (National STHD-100) with three plates removed from each stator section

C_2—4.5-25 $\mu\mu$fd. zero-coefficient ceramic trimmer (Centralab 822AZ)

C_3—7-35 $\mu\mu$fd. ceramic trimmer (Centralab 820C)

C_4—10-100 $\mu\mu$fd. ceramic trimmer (Centralab 823BN)

C_5—20-125 $\mu\mu$fd. ceramic trimmer (Centralab 823 AN)

C_6—Omit when whip antenna used—10-100 $\mu\mu$fd. ceramic trimmer

L_1—5 turns no. 14 tinned on ½-inch insulator spaced to ½", tapped at 3½ turns, with 4 turns no. 22 enam. on bottom for tickler

L_2—20 turns no. 22 enam. closewound on National XR-50 slug-tuned form

L_3—9 turns no. 14 tinned, air wound, ½" dia. by 1" long, tapped 4 turns from ground end, 4-turn link for antenna coupling

L_4—38 turns no. 22 enam. closewound on ¾-inch dia. form (Amphenol no. 24)

L_5—1500-kc. i-f transformer with trimmer removed from secondary. A 1415-kc. transformer of the type removed from a BC-454 command set was used in the model illustrated.

S—5-pole 3-position switch made up from Centralab "Switchkit." Ceramic deck used for coil switching; other decks are wafer type.

the converter is placed in the position which allows broadcast reception. If a whip antenna is being used for reception it will be possible to vary the trimmer in the front end of the broadcast receiver until maximum sensitivity on the broadcast band is obtained. If a low-impedance transmission line from an antenna on the rear of the car is to be used it probably will be necessary to install an additional capacitor, C_6, in series with the lead between the two switch sections. This capacitor is in series with the feed line on the broadcast band; variation in this capacitance about a value in the vicinity of 75 $\mu\mu$fd. will allow the car receiver to be peaked in spite of the rather large feed-line capacitance to the rear of the car. Further information on this problem has been given earlier under "Auxiliary Antenna Trimmer."

After the installation has been peaked for maximum sensitivity on the broadcast band with the existing antenna and feed line combination, the converter should be switched to the 28-Mc. position. The trimmer across L_5 should be peaked for maximum noise level on the 28-Mc. band or for maximum strength of a recieved signal. It is quite likely that insufficient overall gain will be obtained as a result of the detuning of the input circuit of the broadcast receiver. With step-by-step variation in the number of turns on the link winding of L_5 and in the capacitance of C_5 it will be possible to obtain optimum coupling of the signal from the converter into the broadcast receiver, and at the same time peak up the first tuned circuit in the broadcast set for maximum overall gain.

Construction The unit illustrated was constructed on a 3 by 6½ by 2 inch chassis of the type used in a surplus marker-beacon receiver. The case for the original receiver also serves as the shield cover for the converter. A moderate amount of variation in layout is permissible since only one tube is used and internal shielding is not critical. The form factor of the equipment can be varied considerably if it is necessary to fit the completed converter into the glove compartment, below the dash, or onto the steering column. The most important consideration is that the unit be completely shielded by a cover that completely encloses the assembly. Next most important is that the leads which supply plate and heater voltage to the equipment be properly shielded throughout their length from the car receiver to the converter, and that these leads be by-passed by low-inductance ceramic by-pass capacitors at each end. Also, make sure that the shield cover for the unit is securely fixed to the panel at the front and to the chassis at the rear. If the cover is permitted to vibrate with respect to the chassis it is probable that serious frequency modulation of incoming signals will occur whenever the automobile is in motion.

The factors involved in the installation of a noise limiter and in the deriving of plate and heater voltage for the converter from the broadcast receiver have been discussed earlier in this chapter.

19-2 Mobile Transmitters

As in the case of transmitters for fixed-station operation, there are many schools of thought as to the type of transmitter which is most suitable for mobile operation. One school states that the mobile transmitter should be of very low power drain, so that no modification of the electrical system of the automobile will be required, and so that the equipment may be operated without serious regard to discharging the battery when the car is stopped, or overloading the generator when the car is in motion. This point of view has much to commend it.

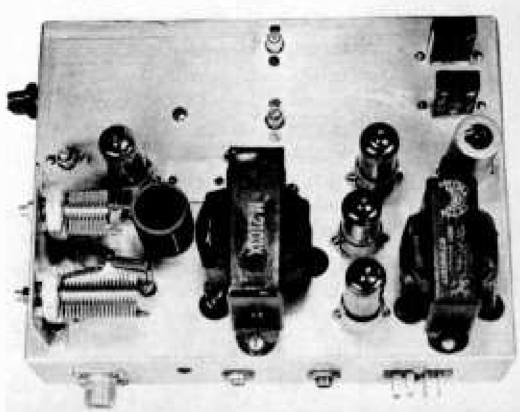
However, the majority of mobile transmitters are powered by a PE-103 dynamotor, which may be operated from a level somewhat below its nominal rating to a level somewhat above. A considerable amount of information on the conversion and operating characteristics of the PE-103 is given in later portions of this chapter.

12-Watt 3.9-Mc. and 28-Mc. Mobile Transmitter

With the opening of the 75-meter band to mobile radiophone operation, a large number of amateurs have had to revise their thinking about mobile operation. All previous operation has been on the 28-Mc. band and on the v.h.f.'s. The little transmitter shown in figures 8, 9, and 10 was designed to assist the average amateur in giving mobile operation a fling without too much capital investment. Both the component requirements and the power drain of this small transmitter are modest; yet it runs 12 watts input to the final stage and is capable of instant change from the 10-meter band to the 75-meter band at the flick of a switch.

The transmitter is designed for operation from a single 300-volt 100-ma. power sup-

Figure 8.
TOP OF THE 12-WATT TWO-BAND MOBILE TRANSMITTER.



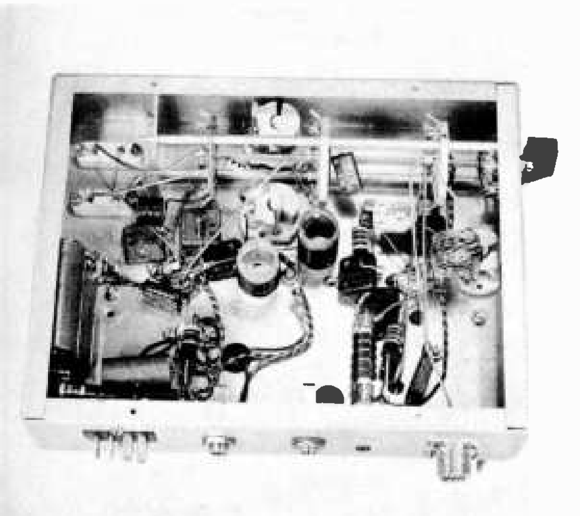


Figure 9.
UNDERCHASSIS OF THE 12-WATT MOBILE TRANSMITTER.

ply. Two power units which have proven satisfactory for use with the transmitter are the Mallory VP-552 Vibrapack and the *9-volt* dynamotor from a Model ABK or SCR-595 IFF transpondor. The 9-volt dynamotor was designed to deliver 450 volts, but when operated from a 6-volt supply it delivers about 280 volts at a 100-ma. load. A photograph of the separate power unit for this transmitter is shown in figure 11, while the schematic diagram is shown in figure 12. Note that an inexpensive automobile-horn relay is used as the contactor to control the dynamotor.

Circuit Design A unique feature of this transmitter which adds greatly to operating convenience is the fact that it can be completely pre-tuned for operation on both the 75-meter phone band and on either the 10-meter or the 11-meter bands. The selector switch, S_1, selects the crystal and the pre-tuned tank circuits for either band of operation. A 6AQ5 tube is used as a conventional crystal oscillator with a 3.9-Mc. crystal for the 75-meter band, while this same tube acts as a harmonic oscillator with a crystal in the 13.5 to 14.85 Mc. range to give excitation to the 6AQ5 final amplifier in the 27 to 29.7 range. Tank circuit C_1L_1 tunes to the 10-meter band while C_2L_2 tunes to the 3.9-Mc. band.

Shunt feed is used to the plate of the 6AQ5 final amplifier so that different types of output circuits may be used between the plate of the tube and the antenna circuit. Tank L_3C_3 for the 10-meter range is a conventional parallel-tuned circuit with a link wound over the low-potential end of the coil for antenna coupling. It may be found desirable to insert a 75-$\mu\mu$fd. APC capacitor (C_7) in series with the lead from the link coil to the antenna-selector portion of S_1. Through the use of this capacitor the power input to the final amplifier of the transmitter may be varied by varying the antenna coupling. This capacitor also can serve to tune out any reactance which may exist in the antenna circuit on the 28-Mc. band. This capacitor is shown dotted in figure 10. A conventional 8-foot to 12-foot whip antenna will be found best for operation on the 10-meter band. Coaxial cable such as RG-8/U, RG-58/U, or RG-11/U will prove satisfactory for the lead from the base of the antenna to the transmitter.

A conventional pi network is switched in as a coupler when the transmitter is changed over for operation on the 75-meter band. This type of coupler has been used since it allows a wide degree of flexibility in the value of antenna impedance which may be matched to the plate of the final-amplifier tube. Antennas for 75-meter mobile operation are discussed in Section 19-3 of this chapter.

The modulator for the transmitter is quite conventional, except in one respect; the carbon microphone is connected in the cathode circuit of the first speech tube. The 6C4 tube is operated as a grounded-grid amplifier, with its plate circuit driving the grids of the 6AQ5 modulator tubes through a push-pull input transformer. The 6C4 cathode current provides the microphone current for the carbon button. Surplus microphone type T-17B has proven quite adequate for use in this circuit. The circuit eliminates the microphone transformer and the microphone-voltage filter which is almost invariably required when the microphone is operated from the 6-volt line. The 6AQ5 tubes operate into a plate-to-plate load-impedance of about 10,000 ohms, while the final amplifier presents a load of about 7500 ohms to the secondary when it is running at 40 ma. with 300 volts on the plate. The 6AQ5's have more than adequate power output

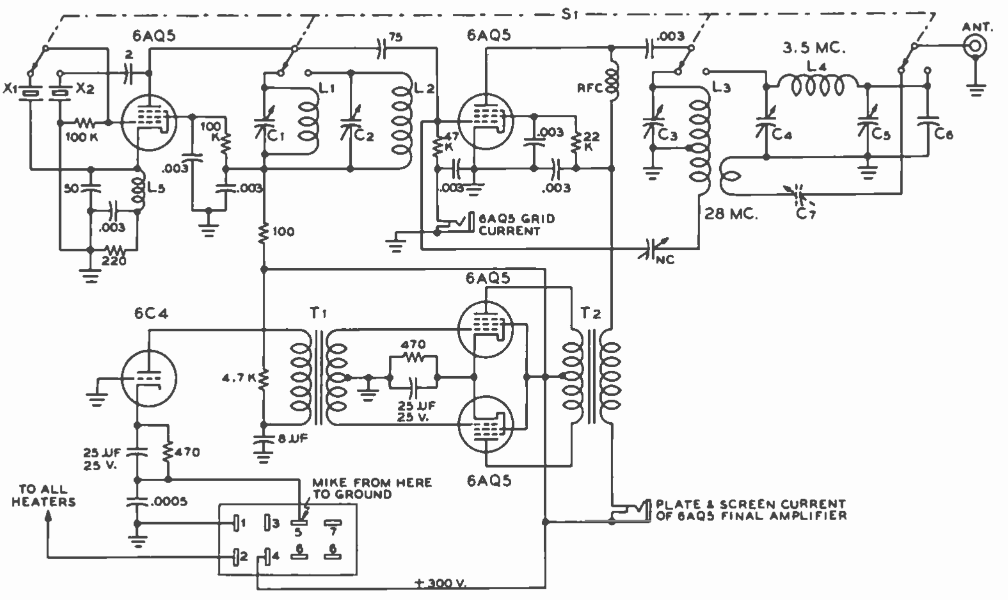


Figure 10.

SCHEMATIC DIAGRAM OF THE 10/75 METER MOBILE TRANSMITTER.

C_1—50-μμfd. APC capacitor
C_2—100-μμfd. APC capacitor
C_3—25-μμfd. APC capacitor
C_4—75-μμfd. APC capacitor
C_5—140-μμfd. APC capacitor
C_6—Ceramic padding capacitor (see text)
C_7—75-μμfd. APC capacitor
L_1—6 turns no. 18, ¾" dia., spaced to ¾"
L_2—42 turns no. 24, ¾" dia., closewound
L_3—6 turns no. 18, ¾" dia., ¾" long, with 3-turn link (B&W 3010 Miniductor), with 2 turns added from ground point for neutralizing
L_4—38 turns no. 22 enam., 1" dia., 1" long (National XR-2 form)
L_5—9 turns no. 18, ½" dia., closewound
NC—2.5-6-μμfd. ceramic trimmer
X_1—14-Mc. crystal (James Knights type H-73)
X_2—3.9-Mc. crystal (James Knights type H-73)
T_1—10,000-ohm plate to p-p grids driver (Stancor A-4734)
T_2—10-watt mod. trans. 10,000 ohm pri. to 8000 ohms (Peerless M-2103X)
RFC—2.5 mh. (National R-100U)
S_1—3-deck 4-pole double-throw ceramic (Made from Centralab Switchkit)
Power connector—Jones P-308AB

to modulate the 12 watts input to the final amplifier.

The transmitter is constructed on a 7 by 9 by 2 inch aluminum chassis. Steel can of course be used, but aluminum is capable of lower losses from the fields around the tank coils, and the aluminum is much easier to work. The construction of the unit is apparent from the photograph. The switch and tank circuits occupy more than one half of the chassis, while the modulator and tubes take up the balance.

Wiring is accomplished with conventional stranded hookup wire for all the d-c and audio circuits and with no. 14 tinned wire for all the r-f circuits. Coil L_5 is self-supporting, while L_1 and L_2 are wound on Amphenol no. 24 polystyrene coil forms. L_3 is a portion of a B&W no. 3010 Miniductor, while L_4 is wound on a National XR-2 coil form.

Tuning Procedure Two closed-circuit jacks are included in the chassis of the transmitter unit. One jack is used to measure grid current to the 6AQ5 amplifier, while the other jack is used to measure the sum of the plate and screen current to the same tube. A 0-50 d-c milliammeter may be plugged into either jack for measurement of the appropriate currents while tuning the transmitter. When tuning the antenna system it will probably be found most satisfactory to mount an insulated jack on the bumper apron at the rear of the car. Then a plug from this jack may be plugged into the plate-current jack on the transmitter. Through this arrange-

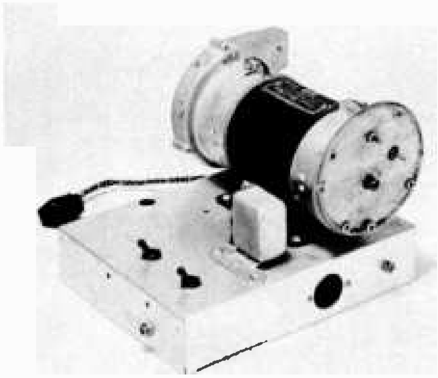


Figure 11.
SHOWING THE DYNAMOTOR UNIT FOR THE 12-WATT TRANSMITTER.
The dynamotor is a 9-volt to 450-volt unit removed from a surplus IFF transpondor. The blower assembly has been removed from the far end, although half the blower housing still remains. The reduction-gear assembly has been removed from the near end of the dynamotor. The plug and cable in the rear connect the dynamotor to the 12-watt mobile transmitter.

ment the trunk may be closed for tuning the antenna system, yet plate current to the final amplifier may be measured by a milliammeter plugged into the external jack.

Normal grid current to the 6AQ5 final amplifier will run about 3 ma., although 2 to 2.5 ma. is adequate for satisfactory operation of the stage. On the 3.9-Mc. band, where the crystal stage operates as a conventional tuned-plate oscillator, it will be found necessary to detune C_2 on the plate tank of the oscillator slightly on the low-capacitance side of maximum output. This procedure is standard for a straight crystal oscillator to insure that the crystal will start reliably. On the 28-Mc. band the tank circuit in the plate of the crystal oscillator is peaked for maximum grid current to the final amplifier. It will be found best to listen to the output frequency of the oscillator plate tank in a 28-Mc. receiver, to insure that the correct harmonic of the crystal oscillator is being tuned.

Neutralization of the Final Amplifier After excitation to the final amplifier has been obtained on both bands, the next procedure is to neutralize the 6AQ5 stage on the 28-Mc. band. Neutralization is not required on the 3.9-Mc. band.

The neutralizing procedure is as follows: Plug an *open-circuited* phone plug into the plate-current jack. Plug the milliammeter into the grid-current jack. Then apply plate voltage to the transmitter and tune for maximum grid current to the 6AQ5 on the 28-Mc. band. Now vary the adjustment of the 28-Mc. plate tank capacitor, C_3, while noting the indication on the grid milliammeter. It probably will be noticed that there is a considerable kick in the grid current as the plate tank circuit is tuned through resonance. Now change the setting of NC and retune C_1 again for maximum grid current. Again tune C_3 through resonance and see if the kick in grid current has been reduced. If the kick has been reduced, advance NC further in the same direction; if the kick was increased, move NC in the opposite direction. After a few tries, a setting for NC will be found where the tuning of C_3 through resonance will cause no change in the grid current of the 6AQ5. The final amplifier stage has been neutralized when this condition is reached.

Tuning the Output Circuit The loading procedure for the final amplifier on the 28-Mc. band is standard. The plate tank is dipped to resonance without the antenna connected. Then the antenna is connected, the tank tuned slightly to make sure that it is at resonance, and the plate current is noted. The number of link turns on L_3 and the setting of the series capacitor C_7 are varied until the indication on the plate-and-screen milliammeter is about 45 ma. at resonance.

A pi network of the type indicated in figure 10 for use on the 3.9-Mc. band is satisfactory for feeding an antenna system which has a relatively low value of feed-point impedance. Such an antenna might consist of a center-loaded radiator with the value of loading inductance varied until the impedance at the base of the antenna is an approximate match for a coaxial feed line.

If the pi-network loading arrangement shown in figure 10 is used in conjunction with an antenna system which presents a relatively low value of load impedance, it will be found necessary to experiment with the value

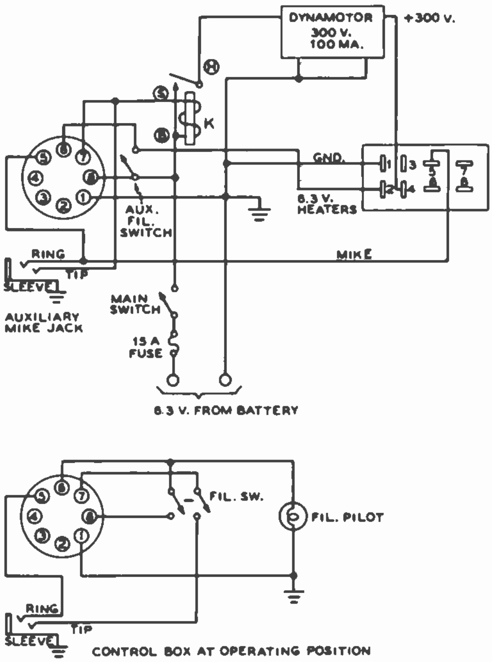


Figure 12.
SCHEMATIC OF THE DYNAMOTOR AND TRANSMITTER CONTROL CIRCUIT.

Although the main control for the transmitter is at the operating position, provision is made for alignment at the transmitter by closing the auxiliary filament switch and plugging the microphone into the auxiliary mike jack. The two-circuit filament switch at the operating position prevents turning on the dynamotor unless the filaments are lighted. An inexpensive automotive horn relay is used at K to control the dynamotor. The three connections to the usual type of horn relay are indicated by the three letters encircled on the drawing. The filament pilot lamp will light whether the heaters have been turned on at the transmitter or at the operating position.

of capacitor used at C_6. For a first test it is best to place a 500-$\mu\mu$fd. ceramic capacitor at this position. Then, with the load connected, dip C_1 to resonance and note the minimum plate current. The objective of the tuning procedure is to obtain the rated value of about 45 ma. of plate current plus screen current to the 6AQ5 with the largest possible value of total capacitance at C_3 plus C_6. If the minimum dip in plate current is too low, decrease the setting of C_3 and again dip C_1. If the plate current dip is still too low, assuming that the out-of-resonance plate current exceeds the desired value, decrease the value of the shunting capacitor C_6. If the antenna is presenting a very low value of load impedance to the transmitter it may be found necessary to reduce the number of turns in L_1 to obtain proper loading. When the transmitter is loaded properly it will be found that the plate current at resonance is only about 10 per cent less than the out-of-resonance plate current.

No difficulty should be encountered in getting the modulator to operate properly. However, it will be found that close talking is required to obtain full modulation with a T-17 microphone. Other single-button carbon microphones, including the "F-1 unit" and the military "lip mike" type will give greater modulation for the same voice level. A suggested control circuit for the transmitter is indicated in the schematic of figure 12.

De-Luxe 50-Watt Portable or Mobile

After an extended period of operation with the usual type of mobile transmitter the operator of such a unit usually will reach a set of conclusions as to desirable features to be included in his next mobile transmitter. Such a set of conclusions probably will include the following:

(1) The transmitter should *not* be installed under the rear deck. A transmitter installed in an inaccessible position such as this is difficult to tune at any time and quite awkward to adjust during a period of inclement weather. Also, there always is some doubt as to whether the output circuit is properly tuned when the deck is closed and the automobile is out in open country. Loaded antennas are quite critical as to tuning, and may be detuned materially by the presence of buildings and overhead wires at the location where the transmitter output circuit was tuned. The preferable location for the transmitter is in the front compartment, probably on the fire wall of the automobile.

(2) The transmitter should be capable of operation on the 10, 20, and 75 meter amateur phone bands, preferably by switching from the front panel of the unit and not by changing coils.

(3) The transmitter should include a v.f.o.,

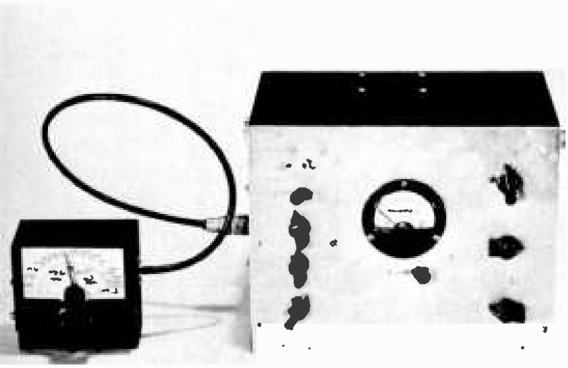


Figure 13.
FRONT OF THE DE-LUXE MOBILE TRANSMITTER.

The v-f-o tuning head is shown alongside. Mike gain, power cable connector, and mike connector are on the left side, with the antenna cable connector on the right. The crystal socket can be seen at top left on the front panel. Panel controls are, from top to bottom along the left: v.f.o.-crystal selector switch, 14-Mc. multiplier tuning, 28-Mc. multiplier tuning, exciter bandswitch. The center control is the meter switch. On the right at top is the pi-network output capacitor, below it is the pi-network bandswitch, and at bottom is the 807 plate tuning capacitor.

and should be capable of v.f.o. control or crystal control on all three bands. Crystal control is desirable for spot frequency operation and v-f-o control should be available for moving from beneath QRM and for moving to any desired frequency in the 10, 20, and 75 meter bands for calling purposes. It is quite disconcerting to be travelling through a town where a group of amateurs are in local contact, yet to be unable to break these stations due to inability to move to the operating frequency of the stations.

(4) The transmitter should be completely enclosed, and should be easily removable for bench testing. As an extension of this thought, the transmitter also should be usable as a portable equipment merely by plugging in an appropriate power unit.

(5) The transmitter should be TVI proof. Hit-and-run TVI must be eliminated as TV reception spreads to include all the major population areas of the country. This also is important for some types of portable operation and for intermittent use of the transmitter at the home station.

(6) The control circuits of the transmitter should be flexible. It should be possible to tune the exciter to the desired frequency without applying plate voltage to any of the higher level stages. Also, it should be possible to turn on the v.f.o. for frequency spotting without disabling the receiver or turning on the output stage of the transmitter.

(7) Trick circuits requiring special high-frequency crystals are not desirable for obtaining 10-meter excitation to the final stage. A standard exciter unit which will use standard 3.5-Mc. and 7-Mc. crystals would be desirable, but the exciter should require a minimum of heater and plate power.

The transmitter described in the following paragraphs and illustrated in figures 13 through 22 meets all the requirements listed above, and offers several additional operating

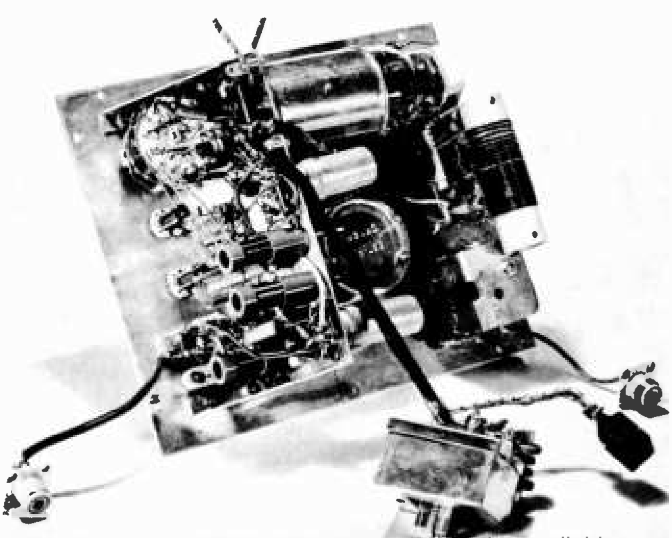


Figure 14.
THE R-F UNIT OF THE TRANSMITTER.

The various connectors and the power-lead filter have been removed from the case for this photograph. Note the shield on the rear of the milliammeter. The small 4-prong plug supplies power to the speech amplifier. The major portion of the exciter assembly is mounted on the bent aluminum vertical chassis. The 6AK6 28-Mc. multiplier tube is mounted between the 807 tube shield and the front panel.

conveniences, such as a pi-network output circuit on all bands, provision for operating the complete exciter unit from the 200 to 250 volt plate supply of the auto receiver, and provision for using a dynamic or ceramic-crystal microphone.

The Exciter Unit The r-f exciter portion of the transmitter is the heart of the whole equipment and is the result of an extended amount of development work. Although the exciter requires a total of four tubes, these are all low-current types so that the entire exciter requires a plate supply of only 40 ma. at 225 volts for v-f-o operation on the 28-Mc. band. Also, the total heater current requirement for the exciter, when all tubes are operating as on the 28-Mc. band, is 0.75 amperes. Each of the tubes in the exciter requires about 10 ma. at 225 volts. Hence, dropping the v.f.o. tube (6CB6) for 28-Mc. crystal operation decreases the drain to about 30 ma., and so forth down to a total current drain of the exciter of 10 ma. for crystal operation on the 75-meter band; or about 20 ma. for 75-meter v-f-o operation.

The heaters of the 6CB6 v-f-o tube and the 6AK6 crystal stage or first multiplier are on at all times. The 6AK6 14-Mc. and 28-Mc. multipliers have their heaters lighted only when the transmitter is being operated on either of the latter two bands. Plate voltage is removed from the 6CB6 when v-f-o operation is not being used. Type 6AK6 tubes have been used throughout the exciter (except for the v-f-o tube where a higher G_m was required) as a result of their low heater and plate current requirements (150 ma. heater current and about 10 ma. plate current as mentioned above) and because these tubes require only a relatively small amount of excitation voltage for efficient operation as frequency multipliers. Type 6C4 miniature triodes also were tried; while these tubes required the same heater and plate current for the same output, they required considerably more excitation for efficient operation as frequency multipliers, thus precluding the use of broadly resonant tank circuits in the 3.5-Mc. and 7-Mc. stages.

The 6AK6 Oscillator-Multiplier Stage With switch S_1 in the lower position the first 6AK6 operates as a Colpitts harmonic crystal oscillator. The circuit will oscillate with any crystal whose fundamental frequency is between 2 and 10 Mc., regardless of the conditions in the plate circuit of the tube. It is merely necessary to plug in the crystal and apply plate voltage; if the crystal is within that range and has normal activity, it will oscillate.

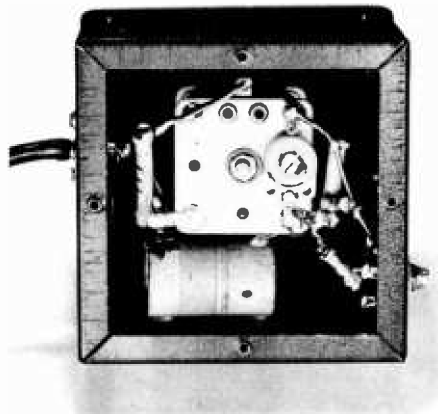


Figure 15.
INSIDE VIEW OF THE V-F-O TUNING HEAD.

Crystals in the 3.5-Mc. to 4-Mc. range normally would be used for operation on the 80-meter or 75-meter bands, and both 3.5-Mc. and 7-Mc. crystals may be used for operation on the 14-Mc. and 28-Mc. bands. Coils L_2 and L_3 in the plate circuit of this tube are broadly resonant and slug tuned. With switch S_2 in the 80-meter position, broadly-resonant coil L_3 is connected to feed plate current to the 6AK6, and excitation from the plate of this tube is fed directly to the grid of the 807 final amplifier. No tuning of the exciter portion of the transmitter is required for operation on any frequency within the 3.5 to 4-Mc. range. The grid current to the 807 will be between 3.5 and 5 ma. for normal operation of the transmitter, with either crystal or v-f-o excitation, at any frequency within this range.

The 14-Mc. and 28-Mc. Multipliers The heaters of the next two 6AK6 tubes are lighted only when the exciter bandswitch S_2 is in such a position that the operation of the tubes is required. Hence, when either of these tubes is picked up by the bandswitch, a period of 10 seconds or so should be allowed for the heaters to come to operating temperature.

With S_2 in either the 20 or 10 meter position, broadly resonant 7-Mc. coil L_2 is in the plate circuit of the 6AK6 oscillator-multiplier, and the output of this tube is fed to the grid of the 6AK6 14-Mc. multiplier. The 6AK6 oscillator-multiplier acts as a doubler for v-f-o operation on these bands, or when a 3.5-Mc. crystal is being used; it operates straight through with a 7-Mc. crystal.

For operation with the excitation switch S_2 in the 14-Mc. position, the heater circuit of the 28-Mc. doubler is open, and the plate circuit of the 14-Mc. doubler is capacitively coupled to the grid of the 807. The tuning capacitor across L_4 in the plate of the 14-Mc. 6AK6 doubler is resonated to the operating frequency by a panel control. But since the input capacitance of the 807 with the extra switching circuit and the longer grid lead is about 7 $\mu\mu$fd. greater than that of the following 6AK6, a trimmer capacitor is connected from the grid of the 6AK6 28-Mc. multiplier to ground. This trimmer is adjusted to such a setting that the tuning capacitor across L_4 need not be changed when the bandswitch is changed from the 14-Mc. to the 28-Mc. position. Coil L_5 in the plate circuit of the 28-Mc. multiplier also is tuned by a panel control. Grid current to the 807 under normal operating conditions with 225 volts on the exciter will be between 3 and 4.5 ma. for operation on any frequency within the 20, 11, and 10 meter bands.

The 6CB6 Remotely-Tuned V.F.O. The Clapp oscillator v.f.o. of the transmitter is unique in that the tuned circuit of the v.f.o. is remote from the transmitter itself, and is connected to it by means of coaxial cable. This is a great operating convenience, especially in mobile work, since the transmitter proper may be mounted some distance from the driver's seat, but the tuning head of the v.f.o. may be placed alongside the receiver. In the unit illustrated the cable between the tuning unit and the transmitter is only about 20 inches long, since the transmitter is mounted under the dash. But in a test installation the coaxial cable between the tuning unit and the transmitter was made 25 *feet* long without deleterious effects. It was necessary only to compensate for cable capacitance at the *transmitter* end by decreasing the capacitance of the grid-cathode and cathode-ground capacitors of the 6CB6 oscillator.

The principle of remote-tuning of the v.f.o., which is made possible through the use of the Clapp oscillator circuit, is of considerable convenience in the home-station installation also, since the small tuning head may be

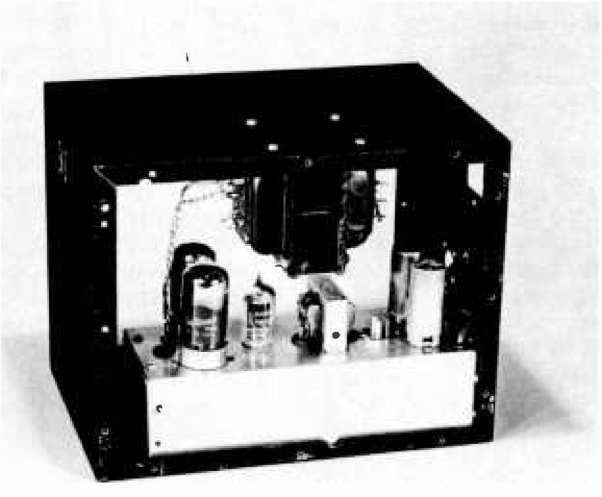


Figure 16.
REAR VIEW SHOWING THE MODULATOR INSTALLED
The r-f unit had been removed from the case when this photograph was taken.

placed alongside the receiver on the operating table, with a coaxial cable running from it to the shielded transmitter at any place in the room or even in another room. In this way the tuning head is effectively isolated from the heat and intense fields of the transmitter, thus simplifying the temperature-compensation and shielding problems. Also, due to the small size requirement of the tuning head, it may be placed in a convenient location on the operating table.

As discussed in Chapter Seven, the stability of the Clapp oscillator insofar as tube variations are concerned is a function of the grid-cathode and cathode-ground capacitors of the oscillator tube. The larger the capacitance at these points, the better will be the stability of the oscillator with respect to tube variations. But a large capacitance at this point requires a high G_m tube for the circuit to oscillate. Hence the 6AG7 tube with a G_m of about 11,000 is ideal for this application. But for mobile work the 6AG7 requires excessive heater and plate current, so the 6CB6 was chosen. This relatively new TV pentode has a G_m of about 6200, and requires about 10 ma. plate and screen current with a heater drain of 300 ma. In the unit illustrated a broadly-resonant coil in the plate of the 6CB6 delivers excitation to the first 6AK6 when S_1 is in the v-f-o position. This coil is peaked at 3750 kc., and delivers adequate excitation to the 6AK6 over the range between 3.5 and 4.0 Mc.

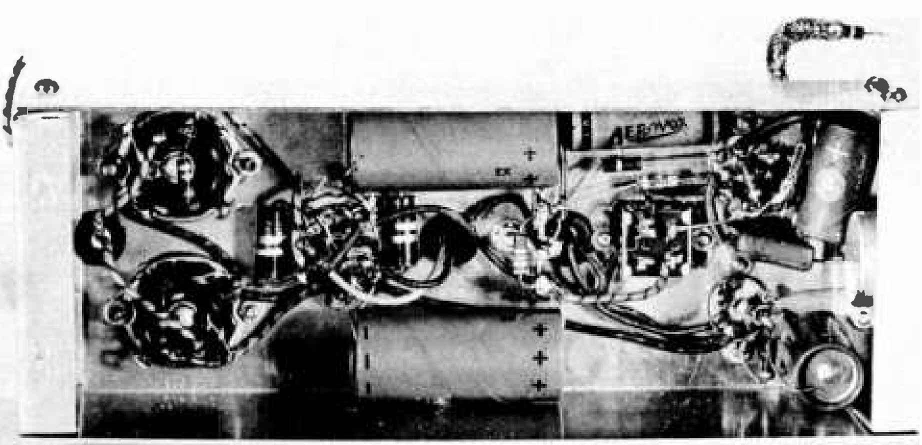
The V.F.O Tuning Head Figure 15 is an inside photograph of the v-f-o tuning head. The tuning head includes the components shown in figure 18, including a switch to change the frequency coverage of the v.f.o. In the 75-meter position the tuning head covers from 1910 to 2015 kc. to give good bandspread on the 75-meter phone band (full-scale on the dial covers 3810 to 4030 kc.). In the other switch position the tuning head covers from 1760 to 1870 kc., to give a dial calibration of 3520 to 3740 kc. This tuning range spreads the 10-meter phone band over nearly all the tuning scale. Hence, the 20-meter phone band occupies only a small portion of the range.

The tuning head includes a 40-μμfd. Centralab temperature compensating capacitor, a ceramic bandset capacitor with access hole for the adjusting screwdriver in the back cover of the tuning head, and the zero-coefficient ceramic padding capacitors. Also included within the tuning head is a zero-coefficient ceramic capacitor across the coaxial output terminal of the tuning head. This capacitor must be decreased for increased length of coaxial line to the transmitter—with 12 feet of RG-58/U line this capacitor is removed

Figure 17.

UNDERCHASSIS VIEW OF THE MODULATOR UNIT.

The modulation transformer is mounted externally to the modulator chassis, on the top of the main housing as shown in figure 16. The shielded lead with the resistor at the end extending from the top of the chassis is soldered to the microphone receptacle.



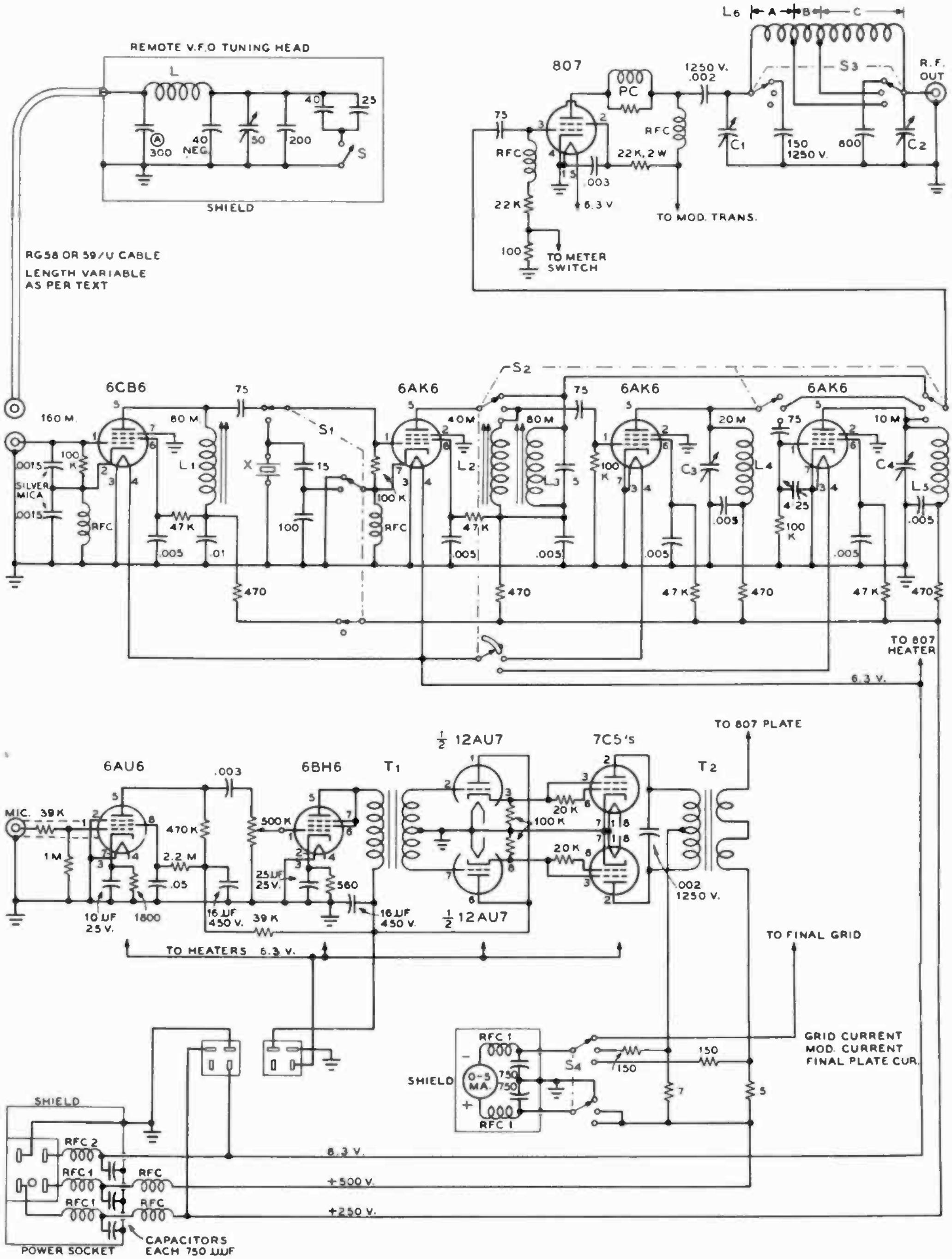

REMOTE V.F.O TUNING HEAD
L
300
40 NEG.
50
200
40
25
S
SHIELD
RG58 OR 59/U CABLE
LENGTH VARIABLE
AS PER TEXT
807
PC
1250 V.
.002
RFC
22 K, 2 W
75
RFC
.003
22 K
6.3 V
100
TO METER SWITCH
TO MOD. TRANS.
L6
A
B
C
S3
R.F. OUT
C1
150
1250 V.
800
C2
S2
6CB6
160 M.
80 M
L1
75
S1
X
15
100
47 K
.005
.01
470
.0015
SILVER MICA
.0015
100 K
RFC
6AK6
40 M
L2
80 M
L3
100 K
RFC
47 K
.005
470
6AK6
75
20 M
C3
L4
100 K
.005
47 K
470
6AK6
75
10 M
C4
L5
4-25
100 K
.005
47 K
470
TO 807 HEATER
6.3 V.
6AU6
MIC.
39 K
1 M
470 K
2.2 M
.05
10 UF 25 V.
1800
.003
16 UF 450 V.
39 K
6BH6
500 K
25 UF 25 V.
560
16 UF 450 V.
T1
1/2 12AU7
100 K
1/2 12AU7
20 K
20 K
7C5's
T2
.002 1250 V.
TO 807 PLATE
TO HEATERS 6.3 V.
TO FINAL GRID
RFC 1
0-5 MA.
750
750
RFC 1
SHIELD
S4
150
150
7
5
GRID CURRENT
MOD. CURRENT
FINAL PLATE CUR.
SHIELD
RFC 2
RFC 1
RFC
RFC 1
RFC
6.3 V.
+500 V.
+250 V.
POWER SOCKET
CAPACITORS EACH 750 UUF

Figure 18.
SCHEMATIC OF THE TRANSMITTER UNIT.

L—30 μh.; 35 turns no. 28 enam. closewound on 1" form, coil waxed after winding
L_1—National XR-50 form with one layer closewound no. 32 enam.
L_2—36 t. no. 26 enam. on National XR-50 form
L_3—Same as L_1
L_4—22 t. 5/8" dia. 16 t.p.i. (B&W no. 3007)
L_5—8 t. 5/8" 16 t.p.i. (B&W no. 3007)
L_6—33 turns no. 16 enam. on National XR-13A form. Section A is 6 turns, section B is 4 turns, and section C is 23 turns. Twisted taps are brought out at junction between A and B, and between B and C.
T_1—3:1 pri. to total sec. push-pull input (Stancor A-53-C)
T_2—15 to 25 watt multiple-match mod. trans. (Merit A-3104 used with taps as indicated)
RFC—2.5-mh. 125-ma. r-f chokes
RFC_1—V-h-f type chokes
RFC_2—High-current v-h-f type choke
S—S.p.s.t. midget wafer switch
S_1—4-pole double-throw lever switch (3 poles used) Centralab 1458
S_2—Assembled from Centralab Switchkit. Rear ceramic wafer 3-pole 3-position. Front phenolic wafer Centralab G to pick up successive heater circuits.
S_3—2-pole 4-position 90-degree ceramic type (Centralab 2543)
S_4—2-pole 3-position non-shorting lever type (Centralab 1454)
C_1—100-μμfd. 1500-volt variable (Johnson 100H15)
C_2—Dual-section midget broadcast variable with sections in parallel (must be good quality)
C_3, C_4—20-μμfd. miniature variables (Johnson 20M11)
PC—6 t. no. 18 tinned wound around 2-watt 47-ohm Ohmite

completely. With a line length shorter than 20 inches this capacitor must be increased proportionately (the capacitance of RG-58/U line is 28.5 μμfd. per foot.). By proportionate reduction in the grid-cathode and cathode-ground capacitors of the 6CB6 much greater cable length between the tuning head and the oscillator tube may be employed. However, the coaxial cable used in this position should be of the solid-molded polyethylene variety such as RG-58/U or RG-59/U. Do not use so-called low-capacitance microphone cable or auto-antenna lead-in cable; in such types of cable the inner conductor is not rigidly centered, so that oscillator instability with cable movement or vibration may be encountered. Also, it is important that all components within the tuning head be mounted in such a manner that they will not vibrate.

The 807 Final Amplifier The shield for the 807 output is a baby-food can with the top removed and the bottom cut out to pass the top of the tube socket. The most common size of baby-food can is just right for an 807 shield. If an 807W is to be used in place of the 807, there is a smaller size can which is adequate for this newer tube. However, in the unit illustrated it was found necessary only to include the standard 6-turn coil of no. 18 tinned wire around a 47-ohm 2-watt Ohmite carbon resistor in series with the lead to the plate cap of the 807 for complete suppression of all types of parasitic oscillation. With this parasitic choke the stage is completely stable on all bands, using either the standard 807 or the 807W. Without this suppressor the 807 showed strong parasitic oscillations, while the 807W showed oscillations only slightly less strong. Hence there is little to choose between the two tubes on this basis. The 807W does require slightly higher screen voltage and slightly less excitation, however, for the same output as the standard 807.

The 807 normally operates with a plate voltage of 500 at a plate current of 100 ma. from a PE-103A for mobile work or from a standard a-c operated power pack for non-mobile operation. Higher or lower voltage may be used on the 807, just so that the maximum plate current rating of 100 ma. is not exceeded.

The Pi-Network Plate Circuit After experimenting with several types of output circuits in the plate of the 807 the pi-network was finally chosen since it affords fair harmonic reduction within itself, it provides a convenient loading adjustment, and it is convenient to use with a single tapped coil for operation on all three bands. The main tuning capacitor for the plate of the 807 is a 100-μμfd. Johnson 100H15. This is paralleled with a 150-μμfd. 1200-volt mica capacitor for operation on the 4-Mc. band. The plate tank coil is wound of no. 14 enameled wire on a National XR-13A form, with twisted taps brought out at the appropriate points. The output capacitor of the pi-network is a dual-section BC-band variable capacitor with both sections in parallel to give about 600 μμfd. maximum capacitance. This capacitor is paralleled with a fixed 800-μμfd. ceramic