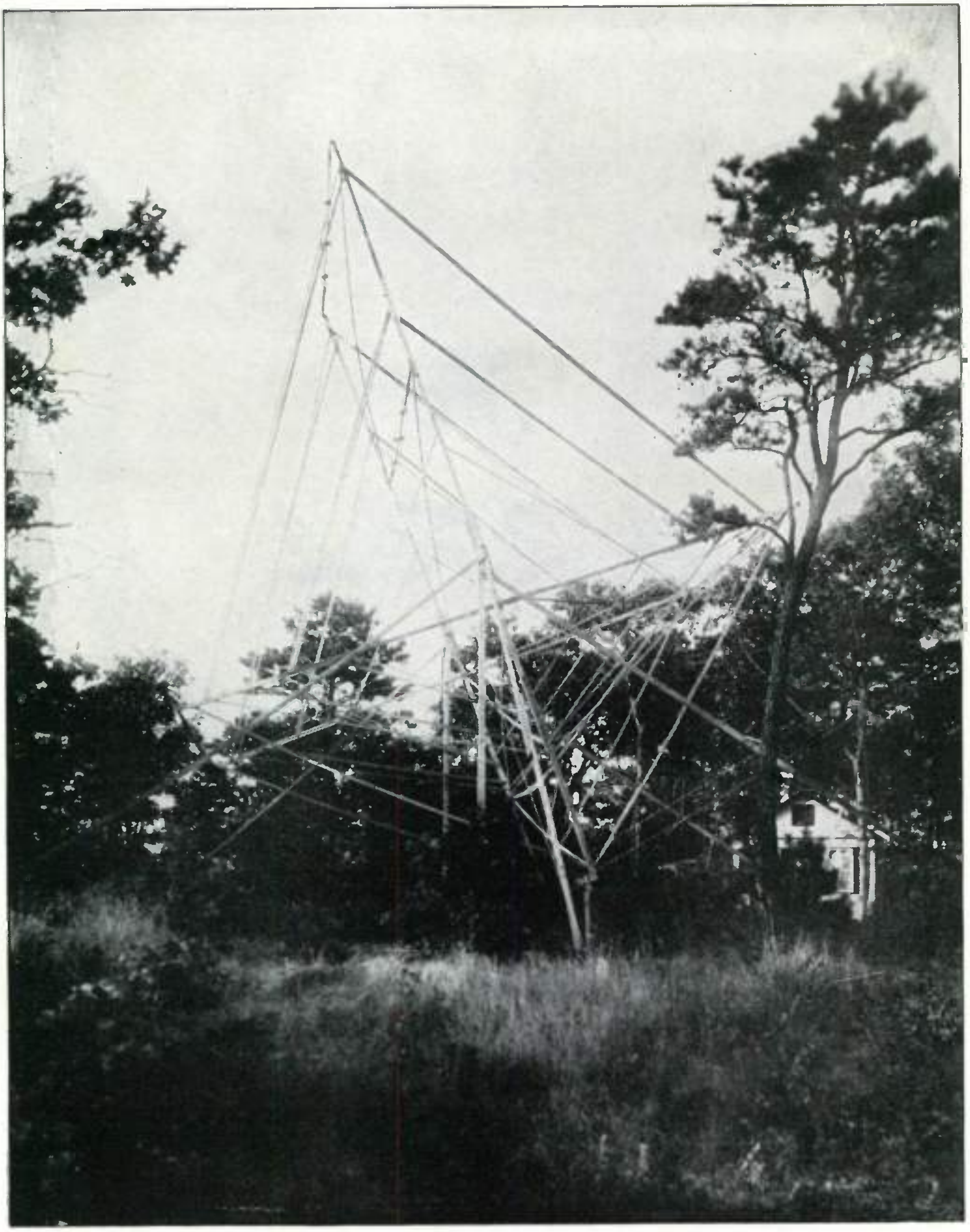
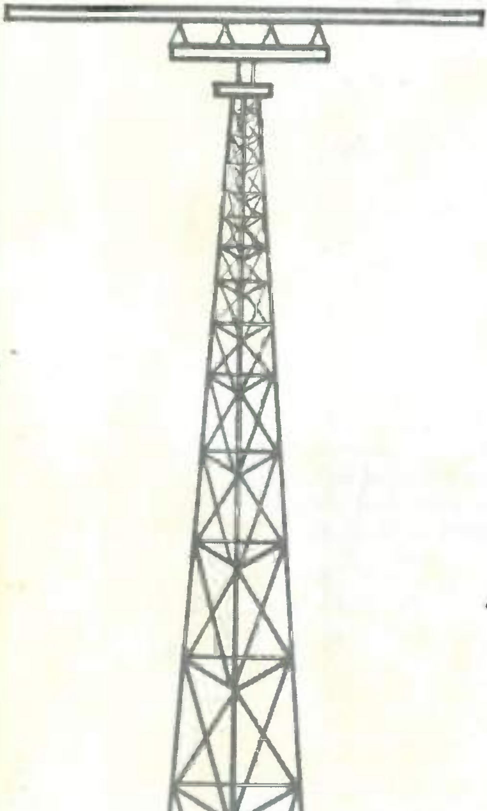
The A.R.R.L. Antenna Book



The very latest in amateur beam antennas — a few decades ago! This intriguing device was a 28-Mc. beam antenna used at W1CCZ in 1928 to test the effectiveness of high-angle radiation at this frequency. The system used three reflectors and two directors in the traditional Yagi configuration. It was pointed toward Australia.

Tests with W6UF and ZL2AC indicated that low-angle was more desirable than high-angle radiation, and all later tests have confirmed these findings.



The A.R.R.L. Antenna Book

1949

The AMERICAN RADIO RELAY LEAGUE, INC

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Foreword . . .

A decade has passed since the first edition of this volume was offered to the radio amateur. In view of its gratifying acceptance it would, in the normal course of events, have gone through successive revisions, keeping pace with antenna developments as they occurred. But the war made this impossible; in consequence, the present edition is essentially not a revision, but a complete rewriting.

The ten-year period just passed has brought a clearer understanding of the principles of antennas and transmission lines, a growing volume of useful design data, and the development of methods and devices for determining and optimizing the performance of an antenna system. It is with these aspects of antenna operation that the new material in this book is largely concerned. There may be some who will be surprised that no treatment of microwave antennas is included, but at this writing there has been so little of the amateur type of reduction to practice, at frequencies above 1000 Mc., that any such material would lack the authenticity that comes from first-hand knowledge. That authenticity, we believe, is a valuable ingredient of these pages. It is composed of practical experience with antenna systems—not the authors' alone, but that of the hundreds of amateurs who, through the pages of *QST*, have contributed to the body of practical know-how that this book expresses. The authors' task has been that of interpreting the basic theory and methods of design, and of selecting from the many antenna systems in use those which meet the most pressing needs of the communicating amateur.

The book has two principal divisions. Chapters One through Five deal with the principles of antennas and transmission lines, wave propagation and its relationship to antenna design, and the performance characteristics of directive antenna systems. These five chapters, prepared by George Grammer, technical editor of *QST*, might be called a textbook on antennas; they enable the reader to design a system of his own to fit his particular needs. Beginning with Chapter Six, there is a series of chapters in which complete data are given on specific designs for the various amateur bands. The amateur who has not studied the first section, or who wishes to avoid the necessity for making his own calculations, will find in these chapters the information necessary for putting up the system that appeals to him. The remaining chapters deal with the highly-important mechanical features of construction and related subjects such as determining geographical directions. With the exception of Chapter Nine, which was contributed by *QST's* v.h.f. editor, Edward P. Tilton, Chapters Six through Fourteen are the work of Byron Goodman, one of *QST's* assistant technical editors.

It is sincerely hoped that you will find the work helpful. If there are some things left untreated that you wish to know more about, some things you do not understand, we shall be grateful if you, the user of the book, will give us your suggestions on how a future edition can be made even more valuable to you.

West Hartford, Conn.
March 1, 1949

A. L. Budlong,
Acting Secretary, A.R.R.L.

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Wave Propagation

Because radio communication is carried on by means of electromagnetic waves traveling through the earth's atmosphere, it is desirable to know something about the characteristics of waves and the way in which their behavior is influenced by the conditions they meet in their trip from the radio transmitter to the receiver. While detailed knowledge of wave propagation is not at all essential to the amateur who wants to put up an effective antenna, a few facts must be understood before the principles of antenna design can be intelligently applied. Although an antenna may — and usually does — radiate the power applied to it with a high degree of efficiency, if that power does not travel to the desired receiving point but goes somewhere else instead, the antenna is a failure.

Our purpose in this chapter, therefore, is to discuss those features of wave propagation that have some bearing on the design of an antenna system. In doing this we do not mean to infer that there is nothing more that the wide-awake amateur will want or need to know about the subject. Effective radio communication results from a combination of equipment, antenna system, *and* operating skill; the latter must include that ability to anticipate "conditions" that in turn is based on an understanding of the vagaries of wave propagation. But the latter, fascinating though they are, are somewhat outside the scope of this book when they do not directly affect antenna design.

● WAVE CHARACTERISTICS

The reader who has some knowledge of electricity has already been introduced to the idea of electric and magnetic fields. A radio wave is a special combination of both types of field, with the energy divided equally between the two. If the waves could originate at a point source in "free" space — empty space such as occurs, for all practical purposes, in the interplanetary and interstellar stretches of the universe — they would spread out in ever-growing spheres with the source as the center. The speed at which the spheres expand would be the same as the speed of light, since light also is an electromagnetic wave. In empty space this speed is 300,000,000 meters per second, or approximately 186,000 miles in one second. The path of a ray from the source to any point on the spherical surface always is a straight line — a radius of the sphere.

It is obvious that in a remarkably short time a sphere growing outward from the center would be very large indeed. An observer on such a spherical surface would conclude, if he could "see" the wave in his vicinity, that it does not appear to be spherical at all but instead seems like a flat surface — just as the earth seems to human beings to be flat rather than spherical. A wave that is far enough from the source to appear flat is called a **plane wave.** The radio waves with which we deal in communication always meet this condition, at least after they have traveled a short distance from the transmitting antenna.

A typical representation of the lines of electric and magnetic force in a plane wave is given in Fig. 1-1. The nature of wave propagation is such

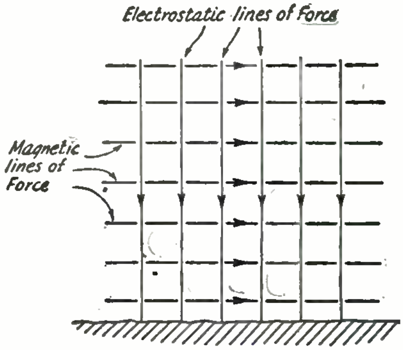


Fig. 1-1 — Representation of the magnetic and electric fields of a vertically-polarized plane wave traveling along the ground. The arrows indicate the instantaneous directions of the fields for a wave traveling perpendicularly out of the page toward the reader. Reversal of the direction of *one* set of lines reverses the direction of travel. There is no change in direction when *both* sets are reversed. Such a dual reversal occurs, in fact, once each half cycle.

that the electric and magnetic lines always are mutually perpendicular, as indicated in the drawing. The plane containing the set of crossed lines represents the **wave front.** The direction of wave travel always is perpendicular to the wave front, but whether the direction is "forward" or "backward" is determined by the relative directions of the electric and magnetic forces.

If the wave is traveling through anything other than empty space its speed is not 300,000,000 meters per second but is something less. Just how much less depends on the substance or **medium** through which the wave is traveling. If the medium is air instead of empty space, the reduction in speed is so small that it can be ignored in most calculations. In solid insulating materials the speed is in general much slower; for example, in distilled water (which is a good insulator) the waves travel only one-ninth as fast as they do in space. In good conductors such as metals the speed is so low that opposing fields (which are set up by currents induced in the conductor by the wave itself) occupy practically the same space as the original wave and thus almost cancel it out. This is the reason for the skin effect in conductors at high frequencies and also the reason why thin metal enclosures form good shields for electrical circuits at radio frequencies.

Phase and Wavelength

Because the speed at which radio waves travel is so great, we are likely to fall into the habit of ignoring the time that elapses between the instant at which a wave leaves the transmitting antenna and the instant at which it arrives at the receiving antenna. It is true that it takes only one-seventh of a second for a wave to travel around the world, and from a communication standpoint that is hardly worth worrying about. But there is another consideration that makes this factor of *time* extremely important.

The wave is brought into existence because an alternating current flowing in a conductor (which is usually an antenna) sets up the necessary electric and magnetic fields. The alternating currents used in radio work may have frequencies anywhere from a few tens of thousands to several billion cycles per second. Suppose that the frequency is 30 megacycles per second — that is, 30,000,000 cycles per second. One of those cycles will be completed in 1/30,000,000 second, and since the wave is traveling at a speed of 300,000,000 meters per second it will have moved only 10 meters during the time that the current is going through one complete cycle. To put it another way, the electromagnetic field 10 meters away from the antenna is caused by the current that *was* flowing in the antenna *one cycle earlier* in time; the field 20 meters away is caused by the current that *was* flowing *two* cycles earlier, and so on.

Now if each cycle of current is simply a repetition of the one that preceded it, the currents at corresponding instants in each cycle will be identical, and the fields caused by those identical currents also will be identical. As the fields move outward they become more thinly spread over larger and larger surfaces, so their amplitudes decrease with distance from the antenna. But they do not lose their identity with respect to the instant of the cycle at which they were generated. That is, the **phase** of the outwardly-moving surface remains constant. It follows, then, that at intervals of 10 meters (in the example above) measured outward from the antenna the phase of the waves at any given instant is identical.

In this fact we have the means for defining rather precisely both "wave front" and "wavelength." The wave front is simply a surface in every part of which the wave is in the same phase. The **wavelength** is simply the distance between two wave fronts having identical phase at any given instant. In the example, the wavelength is 10 meters because the distance between two wave fronts having the same phase is, as we found, 10 meters. This distance, incidentally, always must be measured perpendicular to the wave fronts; in other words, along the same line that represents the direction in which the wave is traveling. Measurements made along any other line between the two wave fronts would lead to the erroneous conclusion that the wave is longer than it really is. Expressed in a formula, the length of a wave is

$$\lambda = \frac{v}{f}$$

where λ = wavelength
v = velocity of wave
f = frequency of current causing the wave

The wavelength will be in the same length units as the velocity when the frequency is expressed in the same time units as the velocity. For waves traveling in free space (and nearly enough for waves traveling through air) the wavelength is

$$\lambda \text{ (meters)} = \frac{300}{f \text{ (Mc.)}}$$

We shall continually be encountering this idea of phase in succeeding chapters in this book, because it is fundamental to the operation of antenna systems. It is essential, therefore, to have a clear understanding of what it means if antenna behavior is to be appreciated. Basically, "phase" means "time," but when something goes through periodic variations with time in the way that an alternating current does, corresponding instants in succeeding cycles are said to have the same phase even though the actual time difference is one cycle. In using the word phase in this fashion we are inherently using the cycle as the unit of time measurement, just as we use one 24-hour day as a unit of time measurement. Four o'clock this afternoon corresponds to four o'clock yesterday afternoon in much the same way that an instant in an a.c. cycle corresponds to the identical instant in the preceding cycle.

In Fig. 1-2 points A, B and C are all in the same phase because they are corresponding instants in each cycle. This is a conventional drawing of a sine-wave alternating current with time progressing to the right. It also represents an instantaneous "snapshot" of the distribution of intensity of the traveling fields if distance is substituted for time in the horizontal axis. In that case the distance between A and B or between B and C represents one wavelength. This shows that the field-intensity distribution follows the sine curve, both as to amplitude and polarity, to correspond exactly to the time variations in current that produced the fields. It must be remembered that this is an *instantaneous* picture; the actual wave

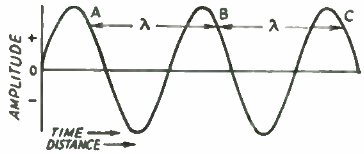


Fig. 1-2 — The instantaneous amplitude of both fields (electric and magnetic) varies sinusoidally with time as shown in this graph. Since the fields travel at constant velocity, the graph also represents the instantaneous distribution of field intensity along the wave path. The distance between two points of equal phase, such as *A–B* and *B–C*, is the length of the wave.

travels along just as a wave in water does. To an observer at any fixed point along the wave path, the field intensity goes through *time* variations corresponding to the time variations of the current that initiated the wave.

Field Intensity

The strength of a wave is measured in terms of the voltage between two points lying on an electric line of force in the plane of the wave front. The unit of length is the meter, and since the voltage in a wave usually is quite low, the measurement is made in microvolts per meter. The voltage so measured goes through time variations just like those of the original current that caused the wave, and so is measured like any other a.c. voltage — that is, in terms of the effective value or, sometimes, the peak value.

There are few, if any, occasions in amateur work where a measurement of actual field strength is necessary. This is fortunate, because the equipment required is elaborate. It is comparatively easy, however, to make measurements of *relative* field strength, and thus determine whether an adjustment to an antenna system has resulted in an improvement or not.

Polarization

A wave such as is shown in Fig. 1-1 is said to be **polarized** in the direction of the electric lines of force. In this drawing the polarization is vertical because the electric lines are perpendicular to the earth. A wave "with its feet on the ground" as shown in Fig. 1-1 is, as a matter of fact, usually vertically polarized. This is because the ground acts as a rather good conductor, particularly at frequencies below about 10 Mc., and it is one of the laws of electromagnetic action that electric lines touching the surface of a good conductor must do so perpendicularly. Over semiconducting ground there may be a forward "tilt" to the wave front; this tilt in the electric lines is greater as the energy loss in the ground becomes greater.

Waves traveling in contact with the surface of the earth are of little usefulness in amateur communication because as the frequency is raised the distance over which the surface wave will travel without excessive loss of energy or **attenuation** becomes smaller and smaller. The surface wave is of most utility at low frequencies and through the standard broadcast band. At high frequencies the wave reaching the receiving antenna ordinarily has not had much contact with the earth and its polarization is not necessarily vertical. If the electric lines of force are horizontal, the wave is said to be **horizontally polarized.** However, the polarization can be anything between horizontal and vertical. In many cases, the polarization is not fixed but continually rotates. When this occurs the wave is said to be **elliptically polarized.**

Attenuation

In free space the field intensity of the wave decreases directly with the distance from the source. That is, if the field strength one mile from the source has a value, let us say, of 100 microvolts per meter, the field strength at two miles will be 50 microvolts per meter, at 100 miles will be 1 microvolt per meter, and so on. This decrease in field strength is caused by the fact that the energy in the wave has to spread out over larger and larger spheres as the distance from the source is increased.

In actual communication by radio the attenuation of the wave may be much greater than this "inverse-distance" law would indicate. For one thing, the wave is not traveling in empty space. For another, the receiving antenna seldom is situated so that there is a clear "line of sight" between it and the transmitting antenna. Since the earth is spherical and the waves do not penetrate its surface to any considerable extent, communication has to be by some means that will bend the waves around the curvature of the earth. These means exist, but they usually involve additional energy losses that increase the attenuation of the wave with distance.

Reflection, Refraction and Diffraction

It has been mentioned that radio waves and light waves are the same type of wave; the only difference is in the scale of wavelength. We are all familiar with the reflection of light; radio waves are reflected in much the same way. Frequently, however, the reflecting surfaces are small (in terms of wavelength) compared with the surfaces from which we see light waves reflected. An object the size of an automobile, for instance, will not reflect much of the energy in an 80-meter wave. On the other hand, it may be a very good reflector of waves only a meter or two in length. The thickness of the object is of some importance because the waves penetrate it to an extent depending on its characteristics. In a material of given conductivity, for example, longer waves will penetrate farther than shorter ones and so require a greater thickness for good reflection. Thin metal is a good reflector even at quite long wavelengths, but in poorer conductors such as the earth — which certainly meets the requirement of having a large surface — the longer wavelengths may penetrate quite a few feet.

Reflection may also take place from any surface that represents a change in the dielectric constant of the medium in which the wave is moving. A familiar example in optics is the reflection of light from the surface of a pane of glass that is itself quite transparent to light waves. When viewed from certain angles, it is practically impossible to see through the pane of glass because of the reflected light.

Another phenomenon that has a rather familiar counterpart in optics is **refraction**, or the bending that takes place when the wave enters (at an angle) a medium having a different dielectric constant than the medium it has just left. This bending is caused by the fact that the wave travels at a different speed when the dielectric constant is changed. The part of the wave that enters the new medium first is either slowed down or speeded up (depending on the relative dielectric constants) and so tends to get ahead of or fall behind the sections of the wave that enter later. The effect is to change the direction in which the wave is moving. The classic example in optics is the apparent sharp bend in a stick held partly in and partly out of a body of water at an angle.

Most of the optical examples of refraction are based on two homogeneous substances having a very definite common boundary, as between air and glass. In that case the rays travel in straight lines *inside* either medium and the bending takes place at the common surface. In radio transmission it is frequently the case that the boundary between the two areas of differing dielectric constant is not at all sharp; the dielectric constant simply changes gradually over quite a distance of the wave path. This causes the wave bending also to be gradual, and the wave path becomes curved.

A somewhat less familiar optical phenomenon that has its radio counterpart is **diffraction.** To the eye, the shadows cast by a pin-point source of light appear to be quite sharp. However, close examination shows that light bends around the edge of an object to some extent, depending on the thickness of the edge. This effect becomes greater as the wavelength is increased, and can be of some importance at radio frequencies. For example, with waves traveling in a straight line one would expect that no signal could be heard behind a hill, but the bending caused by diffraction does produce a signal in the "shadow area." At high radio frequencies the diffracted signal is weak compared with the direct ray, and frequently is masked by stronger signals reaching the same spot by other means such as reflection or refraction in the atmosphere.

Both reflection and refraction can take place in various parts of the atmosphere, and the mechanisms by which they occur are likewise varied. The result is that radio waves frequently are "scattered," just as light is scattered in the atmosphere. Such scattering accounts for the reception of signals under conditions when they would not be expected from the simplified pictures of wave travel now to be discussed.

The Ground Wave

Waves travel close to the earth in several different ways, some of which involve relatively little contact with the ground itself. The selection of proper nomenclature therefore becomes somewhat confusing, but more or less by common consent the term **ground wave** is applied to waves that stay close to the earth and do not reach the receiving point by reflection or refraction from the much higher region of the atmosphere known as the "ionosphere." The ground wave therefore can be a wave traveling in actual contact with the ground, such as the wave pictured in Fig. 1-1, or it can be a wave that goes directly from the transmitting antenna to the receiving antenna when the two antennas are high enough so that they can "see" each other. It can also be a wave that is refracted or reflected in the atmosphere near the earth (the **troposphere**).

THE SURFACE WAVE

A wave that travels in contact with the earth's surface is called a **surface wave.** It is the type of wave that provides reception up to distances of 100 miles or more in the standard broadcast band during the daytime. The attenuation of this type of wave is rather high, so the intensity dies off rapidly with distance from the transmitter. The attenuation also increases rapidly with frequency, with the result that the surface wave is of little value in amateur communication with the possible exception of distances up to perhaps 50 miles on the 3.5-Mc. band. As explained earlier, the surface wave must be essentially vertically polarized. The transmitting and receiving antennas therefore must generate and receive vertically-polarized waves, if the surface wave is to be utilized to advantage. In general terms, this means that both antennas must be vertical.

THE SPACE WAVE

The conditions that exist when the transmitting antenna and receiving antenna are within line of sight of each other are shown in Fig. 1-3. One ray travels directly from the transmitter to the receiver and consequently is attenuated in about the same way as a wave in free space. However, the wave from the transmitting antenna also strikes the ground between the two antennas, and the ray that does so at the proper angle to reach the receiving antenna (the angle of incidence being equal to the angle of reflection, as in optics) combines with the **direct ray** to produce the actual signal at the receiving antenna.

Fig. 1-3 — The ray traveling directly from the transmitting antenna to the receiving antenna combines with a ray reflected from the ground to form the space wave.

In most practical cases where the communication is between two stations on the ground (as contrasted with communication between ground and an airplane, or between two airplanes) the angle at which the **ground-reflected ray** strikes the earth and is reflected will be very small. That is, the ray strikes the earth at almost grazing incidence. Now it happens that such a reflection reverses the phase of the wave, so if the distance traveled by the direct wave and the distance traveled by the ground-reflected ray were exactly the same, the two rays would arrive out of phase and would cancel each other. Actually, the ground-reflected ray has to travel a little farther, and so the phase difference between the two rays depends on the difference in path length as measured in terms of wavelength. If the difference in the length of the two paths is 3 meters, for example, the phase difference from this cause will be only 3 degrees if the wave is 360 meters long. This is only a negligible shift in phase from the 180 degrees caused by the reflection, and so the signal strength would be small. On the other hand, if the wavelength is 6 meters the phase shift caused by the same difference in path length would be 180 degrees — enough to overcome completely the 180-degree reversal caused by the reflection, so the two rays would add at the receiving antenna. In short, the **space wave** is a negligible factor in communication at *low* frequencies, because the difference in the distance traveled by the two rays always is very small, when measured in terms of wavelength. The space wave therefore is canceled out at such frequencies. But as the frequency is raised (wavelength shortened) the space wave becomes increasingly important. It is the dominating factor in communication at v.h.f. and u.h.f.

The space-wave picture presented is a simplified one and, as usual, there are practical complications that modify it. There is some loss of energy when the ray strikes the ground, so the reflected ray does not arrive at the receiving antenna with the same intensity as the direct ray. Because of the ground loss the phase of the reflected ray is not shifted exactly 180 degrees. For both these reasons the two waves never cancel completely at the receiving antenna. Also, at frequencies in the u.h.f. region it is possible to form the wave into a beam, much like the light beam from a flashlight. Such a beam puts most of the energy into the direct ray and reduces the amount that can strike the ground, particularly when the transmitting and receiving antennas are both at high elevations. Thus the effect of the ground-reflected ray is avoided.

Strictly speaking, the description above applies only to a horizontally-polarized wave and perfectly-conducting earth. Practically, the polarization does not make much difference because the earth is neither a perfect conductor nor a perfect dielectric. The over-all result is that at frequencies below, say, 20 Mc., the space wave is inconsequential. But at v.h.f. it is readily possible to transmit to the horizon by means of the space wave.

"Line-of-Sight" Propagation

From inspection of Fig. 1-3 it appears that use of the space wave for communication between two points depends on having a line of sight between the two locations. This is not quite literally true. The structure of the atmosphere near the earth is such that under "normal" conditions (a theoretical normal, rather than an actual one; in many parts of the world, at least, the "normal" is an average which is statistically useful but seldom represents the actual condition of the atmosphere) the waves are bent into a curved path that keeps them nearer to the earth than true straight-line travel would. This effect can be approximated by assuming that the waves travel in straight lines but that the earth's radius is increased in dimension by one-third. On this assumption, the distance from the transmitting antenna to the horizon is given by the following formula:

$$D\ \text{(miles)} = 1.41\sqrt{H\ \text{(feet)}}$$

where H is the height of the transmitting antenna, as shown in Fig. 1-4. The formula assumes that

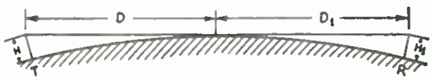


Fig. 1-4 — The distance, D, to the horizon from an antenna of height H is given by the formula in the text. The maximum line-of-sight distance between two elevated antennas is equal to the sum of their distances to the horizon, as indicated in the drawing.

the earth is perfectly smooth out to the horizon; of course, any obstructions that rise along any given path must be taken into consideration. The point at the horizon is assumed to be on the ground. If the receiving antenna also is elevated, the maximum line-of-sight distance between the two antennas is equal to $D + D_1$; that is, the sum of the distance to the horizon from the transmitting antenna and the distance to the horizon from the receiving antenna. The distances are given in chart form in Fig. 1-5. Two stations on a flat plain, one having an antenna on a tower 60 feet high and the other having an antenna supported 40 feet in the air, could be separated approximately 20 miles for line-of-sight communication.

In addition to the "normal" refraction or bending, the waves also are diffracted around the curvature of the earth, so that the actual distance that can be covered does exceed the line-of-sight distance. However, the amount of diffraction at v.h.f. and u.h.f., where the space wave is of chief importance, is rather small and the signal strength drops off very rapidly in a short distance beyond the earth's "shadow".

To make maximum use of the ordinary space wave discussed here it is necessary that the antenna be as high as possible above the surrounding country. A hill that juts above the adjacent terrain is usually an excellent location. However,

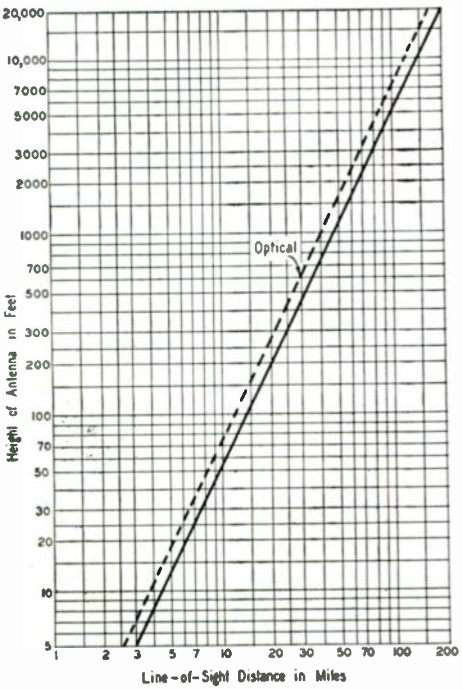


Fig. 1-5 — **Distance to the horizon from an antenna of given height. The solid curve includes the effect of atmospheric refraction. The optical line-of-sight distance is given by the broken curve.**

the peak of a hill is not necessarily the best spot, particularly if it is of the nature of a plateau. Arriving waves may have to be diffracted over the brow of the hill to reach the antenna unless the latter is placed on a high pole or tower; in other words, the brow of the hill may shield the antenna from waves arriving from a desired direction. Also, it is advantageous to have the ground drop off fairly sharply in front of the antenna, as this frequently prevents the ground-reflected ray from approaching at such a flat angle as it would over level ground. Generally speaking, a location just below the peak of a hill is the optimum one for transmitting and receiving in a desired direction, as indicated in Fig. 1-6.

Since the space wave goes essentially in a straight line from the transmitter to the receiver, the antenna used for radiating it should concentrate the energy toward the horizon. That is, the antenna should be a "low-angle" radiator, because energy radiated at angles *above* the horizon obviously will pass over the receiving antenna. Similarly, the receiving antenna should be most responsive to waves that arrive horizontally.

In general, the polarization of a space wave remains constant during its travels. Therefore, the receiving antenna should be designed to give maximum response to the polarization set up at the transmitting antenna. Antennas used at v.h.f. are usually either horizontally or vertically polarized, the former being more generally preferred. The principal reason for this preference is that the chief source of radio noise at v.h.f. — that generated by the spark in the ignition systems of automobiles — is predominantly vertically polarized. Thus horizontally-polarized antennas tend to discriminate against such noise and thereby improve the signal-to-noise ratio.

At the present time there is increasing interest in circular polarization (a special case of elliptical polarization). The direction of rotation with this type of polarization depends on the design of the antenna, and may be made either clockwise or counterclockwise. An antenna designed for clockwise rotation will give poor response to a wave that is rotating counterclockwise. This feature can be used to advantage, because the direction of rotation of the ground-reflected wave is reversed on reflection. Hence a properly-designed receiving antenna will respond chiefly to the direct ray and will discriminate against the ground-reflected ray.

PROPAGATION IN THE TROPOSPHERE

Weather conditions in the atmosphere at heights of a few thousand feet to a mile or two at times are responsible for bending waves downward. This **tropospheric refraction** makes communication possible over far greater distances than can be covered with the ordinary space wave. The amount of the bending increases with frequency, so tropospheric communication improves as the frequency is raised. The bending is relatively inconsequential at frequencies below 28 Mc., but provides interesting communication possibilities at 50 Mc. and above.

Refraction in the troposphere takes place when masses of air become stratified into regions having differing dielectric constants. If the boundary between the two masses of air is sharply defined, reflection as well as refraction may take place for waves striking the boundary at grazing angles.

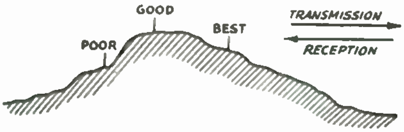


Fig. 1-6 — Propagation conditions are generally best when the antenna is located slightly below the top of a hill on the side that faces the distant station. Communication is poor when there is a sharp rise immediately in front of the antenna in the direction of communication.

The most common cause of tropospheric refraction is the **temperature inversion.** Normally, the temperature of the lower atmosphere decreases at a constant rate of approximately 3 degrees F. per 1000 feet of height. When this rate is decreased for any reason, a temperature inversion is said to exist and greater-than-normal wave bending takes place. Some of the types of temperature inversion are the *dynamic* inversion, resulting when a warm air mass overruns a colder mass; the *subsidence* inversion, caused by the sinking of an air mass heated by compression; the *nocturnal* inversion, brought about by the rapid cooling of surface air after sunset; and the *cloud-layer* inversion, caused by the heating of air above a cloud layer by reflection of the sun's rays from the upper surface of the clouds. Sharp transitions in the water-vapor content of the atmosphere may also bring about refraction and reflection of v.h.f. waves.

Because the atmospheric conditions that produce tropospheric refraction are seldom stable over any considerable period of time, the strength of the received signal usually varies or "fades" over a wide range. Hourly and seasonal variations are observed. Best conditions often occur in the evening and just before sunrise, and conditions are generally poorest at midday when the atmosphere is relatively stable. Tropospheric refraction is generally greatest in the early summer and early fall. It is also more pronounced along the seacoasts.

The tropospheric wave maintains essentially the same polarization throughout its travel, so the transmitting and receiving antennas should have the same type of polarization. Since waves that enter the refracting region at anything other than practically grazing incidence are not bent enough to be useful for communication, the transmitting antenna should be designed for maximum radiation horizontally. The receiving antenna likewise should be a low-angle affair if the received signal is to be most efficiently utilized.

Atmospheric Ducts

In some parts of the world, particularly in the tropics and over large bodies of water, temperature inversions are present practically continuously at heights of the order of a few hundred feet or less. The boundary of the inversion is usually well enough defined so that waves traveling horizontally are "trapped" by the refracting layer of air and continually bent back toward the earth. The air layer and the earth form the upper and lower walls of a "duct" in which waves are guided in much the same fashion as in a metallic wave guide. The waves therefore follow the curvature of the earth for distances (sometimes hundreds of miles) far beyond the optical horizon of the transmitter.

Because the height of an atmospheric duct is relatively small, only waves smaller than a certain limit will be trapped. If the refracting layer is only a few feet above the surface the lowest usable frequency may be as high as a few thousand megacyles, so ultrahigh or superhigh frequencies must be used. Under some conditions, however, the height and dielectric characteristics of the layer may be such that waves in the medium v.h.f. region will be transmitted. The line of distinction, if any, between ducting and ordinary tropospheric propagation is hard to draw in such a case.

A feature of duct transmission is that the antennas, both transmitting and receiving, must be *inside* the duct if communication is to be established. If the duct extends only a few feet above the earth and the transmitting antenna is on a tower or promontory *above* the duct, no signals will be heard at the receiving point. Likewise, a receiving antenna situated above the duct will not pick up energy trapped nearer the earth.

Atmospheric ducts also are formed between two layers of air having suitable characteristics. If the lower layer refracts the waves *upward* while the upper layer refracts them *downward*, waves will be trapped between the two layers and again can travel for great distances. In such a case antennas either below or above the duct will be ineffective. Ducts of this type have been observed from airplanes, where good signals will be received with the 'plane at the optimum height but the signal strength drops off rapidly at either higher or lower altitudes.

Much remains to be learned about the extent of duct transmission at amateur frequencies. There appears to be no significant difference in the signal strength with either horizontal or vertical polarization. Communication via ducts may well turn out to be the most important mode at u.h.f. and s.h.f. when the amateur bands in that frequency range become more thickly populated.

The Sky Wave

At frequencies below 30 Mc. practically all amateur communication — except for "local" work over distances of a few miles — is carried on by means of the **sky wave**. This is a wave that, on leaving the transmitting antenna, would travel on out into empty space if it were not for the fact that under certain conditions it can be sufficiently reflected or refracted, high up in the earth's atmosphere, to reach the earth again at distances varying from zero to about 2500 miles from the transmitter. By successive reflections at the earth's surface and in the upper atmosphere, communication can be established over the maximum possible terrestrial distances.

THE IONOSPHERE

The region in which the waves are bent back to earth is called the **ionosphere**. This is a section of the upper atmosphere in which the air pressure is so low that "free" electrons and ions can exist for a long time without getting close enough to each other to be attracted together and thus recombine into a neutral atom. A wave entering a region in which there are many free electrons will be affected in much the same way as one entering a region of differing dielectric constant; that is, its direction of travel will be shifted. The mechanism is complicated, but in a broad sense is the result of the interaction of electric forces when the free electron is set in motion by the passing wave. In the ionosphere the moving wave tends to be bent back toward the earth.

Ultraviolet light from the sun is the primary cause of ionization in the upper atmosphere. The amount of ionization does not change uniformly

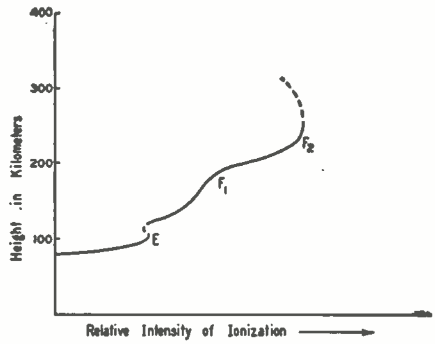


Fig. 1-7 — Typical curve of distribution of ion density with height in latitude of Washington, D. C., near a sunspot minimum. The ion density and the heights at which the maxima occur for each layer will depend on the latitude and the period in the sunspot cycle. The *D* region is not shown because it is not ordinarily measurable by the methods used in ionosphere sounding; its presence is known principally because of the observed absorption of lower-frequency waves during the daytime hours.

with height above the earth, as might be expected at first thought. Instead, it is found that there are relatively dense **layers** of ionization, quite thick vertically, at rather well-defined heights. Nor is the ionization uniform within the layer itself; it is highest at the center of the layer and tapers off gradually both above and below. Fig. 1-7 is a representative plot of the intensity of ionization with height above the earth. Both the height and ionization intensity of any given region vary with the time of day, the season of the year, and the 11-year sunspot cycle. This is because the amount of ultraviolet radiation received from the sun at any given spot depends on these factors.

Layer Characteristics

The ionized layers or regions are designated by letters. The lowest one known, at a height of about 30 to 55 miles, is called the ***D* region.** Because it is in a relatively dense part of the atmosphere the atoms broken up into ions by sunlight quickly recombine, so the amount of ionization depends directly on the amount of sunlight. Thus *D*-region ionization is maximum at local noon and disappears at sundown. When electrons in the *D* region are set in motion by a passing wave the collisions between particles are so frequent, because of the rather high air density, that a substantial proportion of the wave energy may be used up as heat. The probability of collisions depends on the distance an electron can travel under the influence of the wave. This distance depends on the frequency of the wave, because during a long cycle (low frequency) the electron has time to move farther, before the direction of the field reverses and sends it back again, than it does in a short cycle (high frequency). If the frequency is low enough the collisions between particles will be so frequent that practically all of the energy in the wave will be absorbed in the *D* region. This usually happens at frequencies in the 3.5–4.0 Mc. amateur band at the time of maximum *D*-region ionization, particularly for waves that enter the layer at the lower vertical angles and thus have to travel a relatively long distance through it. At times of sunspot maxima even waves entering the layer directly upward will be almost wholly absorbed, in this frequency band, around the middle of the day. The absorption is less in the 7-Mc. band and is quite small at 14 Mc. and higher frequencies. The *D* region is relatively ineffective in bending waves back to earth, and so plays no significant part in long-distance communication except as an absorber of energy. It is a principal reason why amateur communication on the lower frequencies (3.5 and 7 Mc.) is confined to short distances during the daytime.

The lowest layer that affords long-distance communication has a mean height of about 65 miles and is called the *E* layer. It is a region of fairly high atmospheric density and consequently the ionization varies with the height of the sun. The ionization drops rapidly after sundown, when the ions and electrons recombine in the absence of sunlight, and reaches a minimum at about midnight. It again increases rapidly at sunrise and reaches a maximum at about noon local time. As in the case of the *D* region, the *E* layer absorbs energy from low-frequency waves during the time of maximum ionization.

The second important "communication" layer is the F_2 layer. This is the most intensely ionized layer, and its height — of the order of 150 to 250 miles — varies with the time of day, the season, and the sunspot cycle. At these heights the atmosphere is very thin, and so the ions and electrons are slow to recombine. Because of this, the ionization is not so responsive to the height of the sun; it reaches a maximum shortly after noon, local time, but "tails off" gradually thereafter. It continues at a rather high, but gradually decreasing, level throughout the night, reaching a minimum just before sunrise. At sunrise it increases rapidly and attains the daytime level in the course of an hour or two.

During the day the F_2 layer sometimes splits into two layers, the lower (and weaker) one occurring at a height of 120 miles or so and being designated the F_1 layer. The F_1 layer is, in general, of little importance in communication, except for introducing additional energy absorption of waves traveling through it. It disappears at night. After sundown, too, the height of the F_2 layer decreases, the maximum ionization occurring in the neighborhood of 175 miles. (The nighttime F_2 layer was formerly called the *F* layer, but present usage is to drop the separate designation for the night layer.)

Refraction in the Ionosphere

The amount by which the path of a wave is bent in an ionized layer depends on the intensity of ionization and the wavelength. The greater the ionization, the more the bending at any given frequency. Or, to put it another way, at a given degree of ionization the bending will be greater as the frequency of the wave is lowered — in other words, as its wavelength is increased.

Two extremes thus become possible. If the ionization is intense enough and the frequency is low enough, a wave entering the ionized region *perpendicularly* will be turned back to earth. But as the frequency is raised or the ionization is decreased, a condition will eventually be reached where the bending will not be sufficient to return the wave to earth, even though the wave leaves the transmitting antenna at the lowest possible angle and thus requires the least bending in the ionosphere. A typical "in-between" condition is illustrated in Fig. 1-8, a simplified illustration of the paths taken by high-frequency waves and considering only the effect of a single layer.

Fig. 1-8 shows a condition that is frequently typical of the way waves are bent in a single layer. (When several layers are involved, the paths are naturally more complex, since the layers have differing characteristics.) In this case the layer is capable of refracting waves that enter it at low angles. However, as the angle at which the ray strikes the layer is increased, a critical angle is reached at which the ray just manages to be bent back to earth. Rays entering at still greater angles are not bent enough and pass through the layer into empty space. Since such rays are useless for communication, it is obvious that energy radiated at angles above the critical angle is wasted.

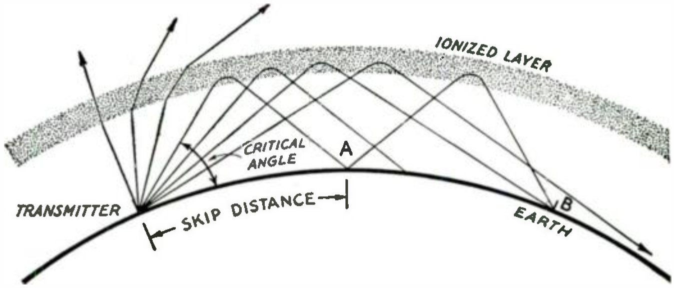


Fig. 1-8 — Behavior of waves on encountering the ionosphere. Waves entering the ionized region at angles higher than the critical angle are not bent enough to be returned to earth. Waves entering below the critical angle reach the earth at increasingly greater distances as the angle approaches the horizontal.

Note also that the point at which a ray reaches the earth on its return journey from the ionosphere depends on the angle at which it left the transmitting antenna. The larger the angle with the surface of the earth the shorter the distance from the transmitter to the point at which the returning ray arrives.

Skip Distance

When the critical angle is less than 90 degrees the highest-angle wave that can be bent back to earth will return at an appreciable distance from the transmitter. For some distance, then, depending on the critical angle, there is a region about the transmitter where the sky-wave signal will not be heard. This "silent" region, extending from the limit of the useful ground wave to the distant point where the sky-wave signal can first be heard, is called the **skip zone**, because all signals skip over it. The skip zone is indicated in Fig. 1-8.

The **skip distance** — the distance from the transmitter to the point where the sky-wave signal is first heard — depends on the critical angle and the layer height. The lower the critical angle the farther the skip distance extends. Since higher frequencies are, in general, bent less than lower frequencies, the skip distance is greater the higher the frequency. For a given critical angle, it is also greater the greater the height of the layer in which the bending takes place. Thus for the same critical angle, the skip distance with F_2-layer bending will be greater than for waves returned to earth from the *E* layer, because the F_2 layer is higher.

When waves at any and all angles are returned to earth from the layer there is, of course, no skip zone. In such instances the sky wave frequently is stronger than the ground wave, at amateur frequencies, even as close as a few miles from the transmitter location. This is because the wave is attenuated less in its travel up to the layer and back again than it is in going a few miles over the ground, surprising as it may seem.

Single- and Multihop Propagation

Fig. 1-8 also shows two of the modes by which the signal can reach a distant receiving point. In one case the wave is bent in the layer at a point about midway between the transmitter and the receiving point, *B*. The wave thus makes the trip in one "hop." However, that is not the only possibility. A ray that is reflected midway between the transmitter and point *A* (which in turn is midway between the transmitter and *B*) will be reflected when it strikes the earth at *A* and will go up to the layer again. Here it is once more reflected, returning to earth, finally, at *B*. This is "two-hop" transmission. More than two hops are readily possible.

The picture of long-distance propagation by means of a series of hops, while satisfying because it is geometrically simple, is seriously open to question. It does not explain all the phenomena observed in practice. For one thing, under some conditions it would be expected to lead to a second skip zone. So far as is known, such a thing has not been observed. For another, it would require that the layer be capable of reflecting the wave all along a path that may be extremely long. In actual practice it has been found that it is only the conditions in the layer at points relatively close to the transmitter and receiver locations that determine whether or not satisfactory communication is possible. (This is discussed further in this chapter in the section on maximum usable frequencies.) In long-distance transmission (distances beyond the theoretical limit for one hop) it seems quite likely that a more accurate theory of propagation will develop as more data on propagation via the ionosphere are collected.

Virtual Height and Critical Frequencies

By using a frequency low enough so that waves entering the ionosphere at the maximum angle of 90 degrees (i.e., waves going vertically from the transmitting antenna to the ionosphere) are returned to earth, it is possible to measure the height of the ionosphere. This is done by measuring the time taken by the wave to go up and back. Knowing the time and velocity of propagation, the distance can be readily calculated. The distance so found is the *virtual height*, or the height from which a pure reflection would give the same effect as the refraction that actually takes place. The method is illustrated in Fig. 1-9. Because a certain amount of time is required for the wave to make the turn at the top of its travel, the virtual height is somewhat higher than the actual height, as the illustration shows.

If the transmitting frequency is gradually increased while height measurements of this type are being made, eventually a frequency range will be encountered where the virtual height seems to increase rapidly, until finally the wave does not come back. The highest frequency that is returned to earth is known as the **critical frequency**. As the frequency is further increased beyond the critical frequency, the wave must enter the ionosphere at progressively smaller angles in order for it to be bent back to earth. At the very lowest angles long-distance transmission is possible at frequencies up to about 2.5 times the critical frequency. Thus, the critical frequency is a measure of the ability of the ionosphere to return high-frequency waves to earth.

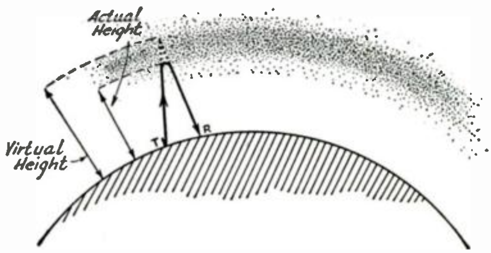


Fig. 1-9 — The "virtual" height of the refracting layer is measured by sending a wave vertically to the layer and measuring the time it takes for it to come back to the receiver. The actual height is somewhat less because of the time required for the wave to "turn around" in the ionized region.

Since the refracted wave acts as though it were reflected from a mirror at the virtual height, it is customary to use the terms "reflection" and "refraction" almost interchangeably in connection with ionospheric propagation. In most cases the actual process is refraction. However, it *is* possible for true reflection to occur if the boundary of the layer is sharply defined and the wave strikes it at a small enough angle.

Virtual heights, of course, depend on the height of the ionized region. The critical frequencies vary with the intensity of ionization in the layers, being greater when the ionization increases. Since the ionization is greatest at the peak of the sunspot cycle, critical frequencies are highest in both the E and F_2 layers during that period. Conversely, they are lowest during a sunspot minimum. The E-layer critical frequency ranges from about 1 to 4 megacycles, depending on the period in the sunspot cycle and the time of day. The F_2 critical frequency varies with the time of day, the season, and the sunspot cycle, ranging from a low of perhaps 2 to 3 Mc. at night in a sunspot minimum to a high of 12 or 13 Mc. in daytime during a sunspot maximum. Whenever the critical frequency is above an amateur band, it is possible to communicate on that band over distances from zero to the maximum possible via that layer, absorption permitting.

Maximum Usable Frequency

Of more interest, from a practical standpoint, than the critical frequency is the frequency *range* over which communication can be carried on via one or the other of the two reflecting layers. In particular, it is useful to know the **maximum usable frequency** (abbreviated **m.u.f.**) for a particular distance at the time of day at which communication is desired. It is always advantageous to use the highest possible frequency because the absorption is less the higher the frequency. Therefore the m.u.f. always gives the greatest signal strength at the receiving point for a given transmitting power.

The m.u.f. depends upon the critical frequency and thus is subject to seasonal variations as well as variations throughout the day. To employ the m.u.f. requires that the antenna system radiate well at very low vertical angles, because at the m.u.f. the critical angle is practically horizontal.

Regular observations of the ionosphere, and correlation with observed signals from various distances over different paths, have made it possible to predict with a high order of reliability the maximum usable frequencies to be expected over periods of several months. These predictions, for both the E and F_2 layers, are issued by the Central Radio Propagation Laboratory of the National Bureau of Standards each month in the form of charts showing the predicted m.u.f. three months in advance. The charts, known as the CRPL-D Series, are available from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D.C., at 10 cents per copy, or on an annual subscription basis (12 issues) for one dollar. The method of using the charts is described in Circular 465, "Instructions for the Use of Basic Radio Propagation Predictions," available from the Supt. of Documents for twenty-five cents. A description of the charts and their uses is outside the scope of this book, but any amateur interested in communication over long distances will find them extremely valuable. They are not at all difficult to use.

As the frequency is decreased below the m.u.f., the signal strength also decreases because of greater absorption. Eventually, as the frequency continues to be lowered, the signal will disappear in the noise background that is always present. Thus there is a low-frequency limit, under a given set of ionosphere conditions, as well as a high-frequency limit to the range of frequencies that can be used for a given distance. The **lowest useful high frequency** (abbreviated **l.u.h.f.**) depends considerably on the transmitter power available, since high power will push the signal through the noise where low power would fail. But when the frequency is near the m.u.f., even low-power signals often will give surprising signal strength at long distances.

In commercial communication it is considered good practice to operate on a frequency about 15 per cent below the m.u.f. This allows for variations in the ionosphere and for the fact that the radiation angle of antennas cannot be made truly horizontal in the high-frequency part of the spectrum. This somewhat lower frequency is known as the **optimum working frequency** (abbreviated **o.w.f.**). Since amateur stations work in fixed bands of frequencies, it is not possible to choose either the m.u.f. or o.w.f. at will. Instead, with the aid of the prediction charts, the *time of day* at which optimum conditions can be expected for a given distance on a particular band must be determined.

Transmission Distance and Layer Height

Consideration of Fig. 1-8 shows that the distance at which a particular ray returns to earth depends upon the angle at which it enters the layer. This angle in turn is determined by the angle (called the **wave angle**) at which the ray leaves the transmitting antenna. Not shown in the drawing, but inherent in the geometry of the situation, is the fact that the distance also depends on the layer height. As the layer height is increased, the distance at which a ray leaving at a fixed wave angle returns to earth also increases. The same wave angle, therefore, will result in transmission over a greater distance when the wave is reflected by the F_2 layer than when it is reflected by the E layer.

The maximum distance that can be covered by one-hop transmission is approximately 1250 miles when the reflection is from the E layer, and approximately 2500 miles when the reflection is from the F_2 layer. These distances are based on average virtual heights, and in both cases a wave angle of zero (ray leaving the antenna tangent to the earth) is required. The actual distance covered by good one-hop transmission is somewhat less, at least at frequencies below 28 Mc., because of ground losses at wave angles below about 3 degrees.

The wave angle required for distances less than the maximum are shown in the chart of Fig. 1-10. The curves are based on average values of virtual height, and are for one-hop transmission. For two or more hops, the distance should be divided by the number of hops and the wave angle read from the chart on the basis of a single hop for the shorter distance. The critical angle must, of course, be greater than the wave angle that is required for the number of hops selected.

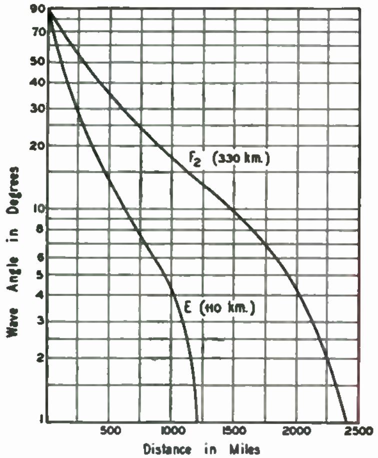


Fig. 1-10 — **Distance plotted against wave angle (one-hop transmission) for average virtual heights of the *E* and F_2 layers.**

One-hop transmission, when possible, will provide the greatest signal strength at the receiving point because there is some energy loss at each reflection, whether in the ionosphere or at the earth. At the longer distances, this requires a small wave angle, or "low-angle radiation" from the antenna. High-angle radiation is most useful for covering short distances. It will be appreciated that "long distance" and "short distance" are relative terms when it is remembered that the distance depends on the layer height as well as the wave angle. At times when the frequency in use is reflected by the *E* layer the distance will be one thing, but at another time of day when the *E* layer is ineffective and the F_2 layer comes into play the same wave angle from the same antenna will cover a much larger distance. That is one reason why it is possible to communicate at longer distances at night on frequencies in the vicinity of 7 Mc. than it is in the daytime.

Long-Distance Transmission

From the discussion in the preceding section, it should be clear that transmission over distances greater than 2500 miles must involve multihop propagation, because 2500 miles is the maximum distance that can be covered by one hop via the highest layer. Since multihop transmission increases the energy loss, it is quite important, for most effective long-distance transmission, that a frequency near the m.u.f. be used, and that the antenna concentrate the radiation at low angles so that the number of reflections will be as small as possible.

The propagation of waves over long paths is complicated by a number of factors. For example, at the particular frequency used the *E* layer may reflect the waves along part or parts of the path while the *F* layer does the reflecting at other parts. This will depend on the time of day, whether the path is generally north-south or east-west, the part of the world over which the path lies — in short, on the state of the ionosphere all along the path. It is also possible that a wave reflected downward from the F_2 layer will be reflected *upward* from the *E* layer instead of being reflected from the earth. However, all these possibilities have but little effect on the primary consideration in DX-antenna design — that the antenna should concentrate the radiation at the lowest possible angle.

Despite the complexity of long-distance propagation, a relatively simple method of determining communication possibilities in advance was developed during the war. This is based on **control points** located 1250 miles from the transmitter and receiver, respectively, along the great-circle path connecting them. If the m.u.f. at the transmitter's control point is, say, 14 Mc., transmission in the direction of the receiver is possible on that frequency. If the m.u.f. at the receiver's control point is 14 Mc. or higher the signal will be heard. On the other hand, if the m.u.f. at the receiving control point is 10 Mc., a 14-Mc. signal from the transmitter will not be heard. The transmitting frequency must be lowered to 10 Mc. before communication is possible. In other words, the *lower* of the m.u.f.s at the two control points is the m.u.f. for the *circuit.* The values of m.u.f. at control points in any part of the world can be determined in advance from the CRPL charts mentioned earlier in this chapter. While communication is possible, in theory, at any frequency below the circuit m.u.f., in practice the absorption becomes too great if the frequency is lowered too much below the m.u.f.

The 1250-mile control point is used for F_2-layer transmission. The control point is 625 miles away when the *E* layer is effective. This may occur at either end of the circuit. If the frequency to be used is below the *E*-layer m.u.f. at that particular time, the *E* layer will control at the end of the circuit at which it is operating. This fact should not be forgotten when using the charts, because it frequently happens that the *E* layer is controlling at one end when the F_2 layer is controlling at the other. Under such circumstances the F_2 m.u.f. may be so high at both ends of the circuit that high absorption would be expected, whereas the actual case is that rather good signals will be received because the operating frequency is near the E-layer m.u.f. at one or both ends of the circuit.

The control-point method of prediction does not explain *how* the waves travel from the transmitter to the receiver. Its justification is that it has been found to be a reliable method for predicting whether or not communication will be possible at a given frequency, or for selecting a frequency that *will* give communication between any two points. One of its puzzling aspects is that the selection of an F_2-layer control point 1250 miles away implies that the wave angle is practically zero, or horizontal. On the other hand, measurements have shown that the amount of radiation from practicable antennas is quite small at angles of less than a few degrees, in the high-frequency range.

The vertical angle at which a wave arrives at the receiving point in long-distance transmission has been found by measurement to vary over a considerable range. For example, measurements on a path from England to the New Jersey coast indicate that on 7 Mc. the wave angle of the received signal at times is as high as 35 degrees, and on 14 Mc. is at times as high as 17 degrees. For 99 per cent of the time it is below those figures on these two frequencies. On the other hand, the same measurements showed that for 99 per cent of the time the angle was above 10 degrees on 7 Mc. and above 6 degrees on 14 Mc. For about half the time the angle was between 22 and 35 degrees on 7 Mc. and between 11 degrees and 17 degrees on 14 Mc. Whether or not there is exact reciprocity between the transmitting and receiving wave angles, these figures indicate the importance of keeping the wave angle low. They also show that the higher the frequency the less useful the higher wave angles become, in transmission over distances.

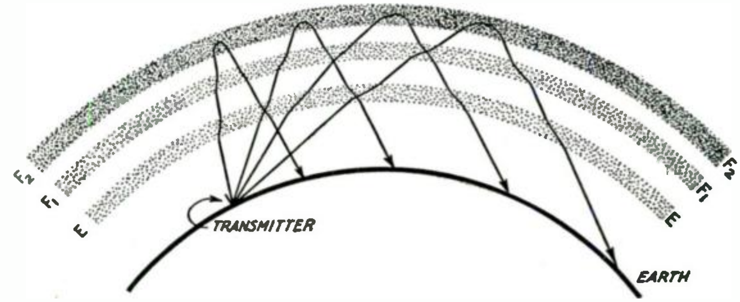


Fig. 1-11 — Typical daytime propagation at high frequencies (14 to 28 Mc.). The waves are partially bent in going through the E and F_1 layers, but not enough to be returned to earth. The actual reflection is from the F_2 layer.

Polarization and Direction of Travel

Because of the nature of refraction in the ionosphere, the polarization of the refracted wave usually is shifted from the direction it had on leaving the transmitting antenna. Because of this, it is not at all necessary to use antennas having the same polarization at the receiving and transmitting points. At the frequencies for which sky waves are useful, most amateurs use horizontal antennas. Depending on the type, such antennas may generate either horizontally-or elliptically-polarized waves.

For the most part, a wave follows the most direct path between the transmitter and receiver. In other words, it follows the great circle connecting the two points. Because of variations in the ionosphere the actual path may vary slightly, and shifts of as much as 5 degrees from the true great-circle path occur at times.

There are always two great-circle paths connecting two points on the earth's surface, one representing the shortest distance between them and the other a path in exactly the opposite direction — around the world the other way. Most communication is via the "short path." However, "long-path" communication is not uncommon, particularly when there is not too much difference between the two distances. At certain times of the day, when the short path would be inoperative, the ionosphere may be able to support communication over the long path.

Occasionally waves arrive from directions that seem to bear no visible relationship to the direction in which the transmitting station lies. While there are well-authenticated cases of this, and reasonable explanations have been worked out on the basis of known behavior of the ionosphere, it is probable that the apparent shift in direction frequently observed by amateurs is a result of "scattering," described in a later section. It is also possible that a combination of the vertical angle at which the wave arrives and minor responses of the antenna system used gives a false direction indication. Accurate direction finding with the sky wave at high frequencies is extremely difficult, requiring highly-specialized care in design and construction of equipment.

MISCELLANEOUS FEATURES OF SKY-WAVE PROPAGATION

Although not having any very direct bearing on antenna design, there are several aspects of sky-wave transmission that are of considerable interest from an operating standpoint. The ability to recognize and appraise unusual propagation effects often will help to explain seeming inconsistencies that may wrongly be blamed on faulty antenna design.

Ionosphere Variations

The daily and seasonal variations in the ionized layers that result from changes in the amount of ultraviolet light received from the sun have already been mentioned. Reference has also been made to the 11-year sunspot cycle, which directly affects propagation conditions because there is a rather direct correlation between sunspot activity and ionization. The 11-year figure for the time between successive peaks of sunspot activity is only an average; any given cycle may vary a few years either way. The peak of the present cycle is considered to have been reached during the winter of 1947–48. This peak was an unusually high one, as sunspot activity goes. On occasions the F_2 m.u.f. rose above 50 Mc., a condition which on the basis of past sunspot peaks would occur only about once in 50 years. The next minimum, when the m.u.f. will drop to low values, should occur about 1953. At a sunspot minimum, there is a period of a year or two when the F_2 m.u.f. does not get as high as 28 Mc.

A small, but regular, variation in sunspot activity occurs over a period of about 28 days. This is the time required for the sun to make one rotation on its axis. The consequent rise and fall of the m.u.f. makes a noticeable change in transmission conditions at frequencies from 14 to 28 Mc.

Fading

Variations in the strength of a received signal are classified under the general term **fading.** Long-period variations are to be expected through the day, on any given frequency, because the absorption changes with the height of the sun. In addition, there is the daily variation of the m.u.f.; when the m.u.f. drops below the frequency in use the signal will "fade out."

In addition, the ionization at any part of the layer is in a continual state of change; there is turbulence in the ionosphere just as there is some turbulence in the atmosphere even on quiet days when the weather seems stable. The amount of absorption is continually varying; waves entering the ionosphere at slightly different angles will be refracted differently; the polarization is continually changing with refraction. The wave reaching the receiving antenna is usually made up of a group of rays each of which has been acted on a little differently by the ionosphere. Sometimes the rays are more or less in phase when they strike the receiving antenna; at other times some of the rays may be out of phase with others. The result is a continual variation in signal strength that may occur at rates varying from several times a second to once every few minutes.

When transmission conditions are not alike for waves of slightly different frequency, the sidebands in voice transmission may have a different fading pattern than the carrier. This is known as **selective fading.** It causes severe distortion of the modulation, especially when the carrier fades down while the sidebands do not. The distortion is in general worse with frequency modulation than with amplitude modulation. Selective fading is more serious at the lower frequencies, such as 4 Mc., where the sideband frequencies represent a larger percentage of the carrier frequency than they do at a frequency such as 28 Mc.

Fading may be entirely different at two receiving points only a short distance apart. By the use of antennas separated by a wavelength or two, feeding separate receivers, it is possible to take advantage of this to overcome the effects of amplitude fading, but not of selective fading. Simultaneous use of inputs from antennas of differing polarization also will often serve the same purpose. Such a receiving arrangement is known as a "diversity" system.

Magnetic Storms

One of the points omitted from the simplified picture of refraction in the ionosphere given earlier was the fact that the process is influenced by the earth's magnetic field. When the earth's magnetism is substantially constant the effect on refraction also is constant. However, on occasions the earth's magnetic field becomes disturbed — presumably because of some effect of the sun, since the disturbances are more frequent during peaks of the sunspot cycle — and a **magnetic storm** is said to be in progress. During magnetic storms the ionosphere is likewise disturbed, with a consequent effect on wave propagation.

The effect of a magnetic storm on the lower frequencies is to decrease the daytime absorption and thus increase the daytime signal strength. At night the signal strength is below normal and is comparable to the daytime levels. On high frequencies communication frequently becomes impossible, as though the refracting layers had disappeared. Magnetic storms vary in intensity and duration. They may last from one to several days.

Aurora

During magnetic storms auroral activity becomes more pronounced and extends farther from the polar regions than is normally the case. During abnormal auroral activity a peculiar form of wave propagation is frequently observed, in which the auroral curtain acts as a reflector. Waves directed toward the polar regions will be reflected back and can be used for communication on frequencies and over distances that normally would be skipped over. When this condition obtains it is necessary, when directive antennas are used, that both the transmitting and receiving antennas be directed toward the polar regions rather than along the great-circle path between the two stations.

Auroral propagation is particularly noticeable on the 28- and 50-Mc. bands. It is easily identified by a very rapid "flutter" fading that in effect modulates the received signal. On v.h.f. the flutter is frequently so pronounced that voice-modulated signals cannot be understood, even though the signals are quite strong. Under these conditions only c.w. signals will provide intelligible communication.

Sudden Fade-Outs

For some reason not completely understood, sky-wave transmission occasionally ceases abruptly and a radio "black-out" occurs, lasting sometimes for a few minutes and sometimes several hours. These sudden fade-outs are always associated with an unusual eruption on the sun, and affect only the part of the earth illuminated by the sun. Although the whole sky-wave spectrum may be affected, the more usual case is that the lower-frequency part is completely blacked out while long-distance propagation continues at the high-frequency end above about 25 Mc.

During one of these sudden fade-outs there is a marked decrease in noise level, since much of the received noise originates at distant points and arrives by way of the ionosphere. The only signals that can be heard during a true fade-out are those from stations within the ground-wave range.

Sporadic-*E* Layer Refraction

In addition to the normal variations in ionization in the *E* layer as previously described, there are also "patches" of relatively intense ionization scattered throughout the layer. These are of varying intensity and size, usually are moving, and appear and disappear apparently at random. They may occur at any time of the day or night, and their cause is unknown.

Sporadic-*E* ionization has no critical frequency. That is, there is no well-defined frequency at which a wave striking the ionized region ceases to be returned to earth. Instead, the intensity of the returning signal simply drops off as the frequency is raised until eventually it is too weak to be usable.

Sporadic-*E* ionization is present all the time, although not at all places nor in sufficient intensity to provide regular communication on high frequencies, such as 28 and 50 Mc., where the normal *E* layer is not operative. However, it will often take part in propagation over long distances at 14 Mc. and below, accounting for communication at times and between points where the F_2 layer alone would not support it. In the 3.5- and 7-Mc. bands it is more of a factor than is generally realized.

At times the intensity of ionization in a sporadic-*E* patch is much higher than in the normal *E* layer at its best. When this occurs sky-wave communication sometimes becomes possible at frequencies as high as the 50-Mc. band over distances as short as 500 miles. Provided the patch is situated midway between two stations, communication is possible to the limit of one-hop *E*-layer transmission, or approximately 1250 miles. There have been a few instances of two-hop 50-Mc. communication between the east and west coasts of the United States, but such instances are relatively rare because *two* sporadic-*E* "clouds" have to be located at just the right spots between the two stations.

Sporadic-*E* transmission is more common on 28 than on 50 Mc. because it can be supported by a lesser degree of ionization on the lower frequency. In other respects the behavior on the two bands is the same, except that when the ionization is high enough to permit 50-Mc. communication, the skip distance will be shorter on 28 Mc. than on 50 Mc.

The comparison between 28 Mc. and 50 Mc. can also be carried to the relationship between 14 Mc. and 28 Mc.; sporadic-*E* is frequently observed on 14 Mc. when it is not operative on 28 Mc., but when it is operative on both bands the skip distance will be shorter on 14 Mc. than it is on 28 Mc.

Maximum utilization of sporadic-*E* ionization at 28 and 50 Mc. requires antennas having low-angle characteristics, because the ionization is not intense enough to return high-angle waves to earth.

Scatter Signals

When a skip zone exists it might be expected, from the simplified explanations of propagation given earlier, that no signals at all would be heard from stations too near to be reached by the sky wave and too far away for the ground wave to be heard. Actually, however, signals from these stations usually can be detected. The strength is low and the signal has a "fluttery" or "warbly" fade that is very characteristic. The signal results from "scattered" radiation that arrives at the receiving point from random directions and in random phase relationships.

Scattering may take place in the ionosphere or may result from repeated reflections between the ionosphere and earth. It may even occur as a result of repeated reflections between the two ionized layers. The signal heard in the skip zone may have traveled thousands of miles, even though the transmitter and receiver are separated by less than 100 miles. One possibility, for example, is that some of the wave energy is reflected back toward the transmitter when it strikes the earth after its first hop, a condition that is easily possible because the earth's surface is not perfectly smooth.

Scattered reflections often will permit communication to be maintained for periods after the normal type of transmission would be expected to die out because of decreasing maximum usable frequency.

Meteor Trails

Meteorites entering the upper atmosphere travel at such high speed that a large amount of energy is released when the meteorite is slowed down by friction with the air. Part of this is used in ionizing the atmosphere along the path followed by the meteorite. Even a very small meteorite can ionize a region 50 or more feet in diameter and a mile or so long. Such a region is large enough to refract the shorter wavelengths back to earth. The ions quickly recombine, however, so the effect of a meteor usually lasts only a short time — from a fraction of a second to a few seconds, in the average case. It is long enough, though, to produce a "burst" of signal from stations not normally heard, or heard only weakly by scattered propagation.

Bursts caused by meteorites can be observed at amateur frequencies from 14 Mc. through 50 Mc. During meteor showers the bursts are so frequent that it is sometimes possible to carry on continuous communication on 28 and 50 Mc. by that means. Just as in the case of sporadic-*E* patches, the ionized meteorite trail must be midway between two stations if a burst is to be heard, and the two stations must be separated by enough distance so that the wave angle will be low enough to be refracted.

CHAPTER 2

Antenna Fundamentals

An antenna is an electric circuit of a special kind. In the ordinary type of circuit the dimensions of coils, condensers and connections usually are small compared with the wavelength that corresponds to the frequency in use. When this is the case most of the electromagnetic energy stays in the circuit itself and either is used up in performing useful work or is converted into heat. But when the dimensions of wiring or components become appreciable compared with the wavelength some of the energy escapes by **radiation** in the form of electromagnetic waves. When the circuit is intentionally designed so that the major portion of the energy is radiated, we have an **antenna.**

Usually, the antenna is a straight section of conductor, or a combination of such conductors. Very frequently the conductor is a wire, although rods and tubing also are used. In this chapter we shall use the term "wire" to mean any type of conductor having a cross section that is small compared with its length.

The strength of the electromagnetic field radiated from a section of wire carrying radio-frequency current depends on the length of the wire and the amount of current flowing. Other things being equal, the field strength will be directly proportional to the current. It is therefore desirable to make the current as large as possible, considering the power available. In any circuit that contains both resistance and reactance, the largest current will flow (for a given amount of power) when the reactance is "tuned out" — in other words, when the circuit is made **resonant** at the operating frequency. So it is with the common type of antenna; the current in it will be largest, and the radiation therefore greatest, when the antenna is resonant.

In an ordinary circuit the inductance is usually concentrated in a coil, the capacitance in a condenser, and the resistance is principally concentrated in resistors, although some may be distributed around the circuit wiring and coil conductors. Such circuits are said to have **lumped constants.** In an antenna, on the other hand, the inductance, capacitance and resistance are distributed along the wire. Such a circuit is said to have **distributed constants.** Circuits with distributed constants are so frequently straight-line conductors that they are customarily called **linear circuits.**

● RESONANCE IN LINEAR CIRCUITS

The shortest length of wire that will resonate to a given frequency is one just long enough to permit an electric charge to travel from one end to the other and then back again in the time of one r.f. cycle. If the speed at which the charge travels is equal to the velocity of light, 300,000,000 meters per second, the distance it will cover in one cycle will be equal to this velocity divided by the frequency in cycles per second, or

$$\lambda = \frac{300{,}000{,}000}{f}$$

in which λ is the wavelength in meters. Since the charge traverses the wire *twice*, the length of wire needed to permit the charge to travel a distance λ in one cycle is $\lambda/2$, or one-half wavelength. Therefore the shortest *resonant* wire will be a half wavelength long.

The reason for this length can be made clear by a simple example. Imagine a trough with barriers at each end. If an elastic ball is started along the trough from one end, it will strike the far barrier, bounce back, travel along to the near barrier, bounce again, and continue until the energy imparted to it originally is all dissipated. If, however, whenever it returns to the near barrier it is given a new push just as it starts away, its back-and-forth motion can be kept up indefinitely. The impulses, however, must be *timed* properly; in other words, the rate or frequency of the impulses must be adjusted to the length of travel and the rate of travel. Or, if the timing of the impulses and the speed of the ball are fixed, the length of the trough must be adjusted to "fit."

In the case of the antenna, the speed is essentially constant, so we have the alternatives of adjusting the frequency to a given length of wire, or the length of wire to a given operating frequency.

The latter is usually the practical condition.

By changing the units in the equation just given, and dividing by 2, the more useful formula

$$l = \frac{492}{f\,(\mathrm{Mc.})}$$

is obtained. In this case l is the length *in feet* of a *half* wavelength for a frequency f, given in megacycles, when the wave travels with the velocity of light. This formula is the basis upon which several significant lengths in antenna work are developed. It represents the length of a half wavelength in space, when no factors that modify the speed of propagation exist.

Current and Voltage Distribution

If the wire in the first illustration had been infinitely long the charge (voltage) and the current (an electric current is simply a charge in motion) would both slowly decrease in amplitude with distance from the source. The slow decrease would result from dissipation of energy in the form of radio waves and in heating the wire because of its resistance. However, when the wire is short the charge is reflected when it reaches the far end, just as the ball bounced back from the barrier. With radio-frequency excitation of a half-wave antenna, there is of course not just a single charge but a continuous supply of energy, varying in voltage according to a sine-wave cycle.

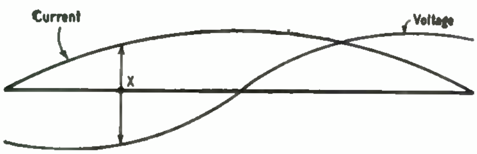


Fig. 2-1 — Current and voltage distribution on a half-wave wire. In this conventional representation the distance at any point (X, for instance) from the wire, represented by the heavy line, to the curve gives the relative intensity of current or voltage at that point. The relative direction of current flow (or polarity of voltage) is indicated by drawing the curve either above or below the line that represents the antenna. The curve above, for example, shows that the instantaneous polarity of the voltage in one half of the antenna is opposite to that in the other half.

We might consider this as a series of charges, each of slightly different amplitude than the preceding one. When a charge reaches the end of the antenna and is reflected, the direction of current flow reverses, since the charge is now traveling in the opposite direction. However, the next charge is just reaching the end of the antenna, so we have two currents of practically the same amplitude flowing in opposite directions. The resultant current at the end of the antenna therefore is zero. As we move farther back from the end of the antenna the magnitudes of the outgoing and returning currents are no longer the same because the charges causing them have been supplied to the antenna at different parts of the r.f. cycle. There is less cancellation, therefore, and a measurable current exists. The greatest difference — that is, the largest resultant current — will be found to exist a quarter wavelength away from the end of the antenna. As we move back still farther from this point the current will decrease until, a half wavelength away from the end of the antenna, it will reach zero again. Thus in a half-wave antenna the current is zero at the ends and maximum at the center.

This current distribution along a half-wave wire is shown in Fig. 2-1. The distance measured vertically from the antenna wire to the curve marked "current," at any point along the wire, represents the relative amplitude of the current as measured by an ammeter at that point. This is called a **standing wave** of current. The *instantaneous* value of current at any point varies sinusoidally at the applied frequency, but its amplitude is different at every point along the wire as shown by the curve. The standing-wave curve itself has the shape of a half sine wave, at least to a good approximation.

The voltage along the wire will behave differently; it is obviously greatest at the end since at this point we have two practically equal charges adding. As we move back along the wire, however, the outgoing and returning charges are not equal and their sum is smaller. At the quarter-wave point the returning charge is of equal magnitude but of opposite sign to the outgoing charge, since at this time the polarity of the voltage wave from the source has reversed (one-half cycle). The two voltages therefore cancel each other and the resultant voltage is zero. Beyond the quarter-wave point, away from the end of the wire, the voltage again increases, but this time with the opposite polarity.

It will be observed, therefore, that the voltage is maximum at every point where the current is minimum, and vice versa. The polarity of the current or voltage reverses every half wavelength along the wire, but the reversals do not occur at the same points for both current and voltage; the respective reversals occur, in fact, at points a quarter wave apart.

A maximum point on a standing wave is called a **loop** (or **antinode**); a minimum point is called a **node.**

Harmonic Operation

If there is reflection from the end of a wire, the number of standing waves on the wire will be equal to the length of the wire divided by a half wavelength. Thus, if the wire is two half waves long there will be two standing waves; if three half waves long, three standing waves, and so on. These longer wires, each multiples of a half wave in length, also will be resonant, therefore, to the same frequency as the single half-wave wire. When an antenna is two or more half waves in length at the operating frequency it is said to be **harmonically resonant**, or to operate at a harmonic. The number of the harmonic is the number of standing waves on the wire. For example, a wire two half-waves long is said to be operating on its **second harmonic**; one three half-waves long on its **third harmonic**, and so on.

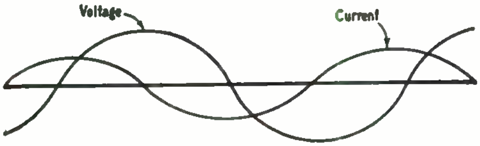


Fig. 2-2 — Harmonic operation of a long wire. The wire is long enough to contain several half waves. The current and voltage curves cross the heavy line representing the wire to indicate that there is a reversal in the direction of the current, and a reversal in the polarity of the voltage, at intervals of a half wavelength. The reversals of current and voltage do not coincide, but occur at points a quarter wavelength apart.

Harmonic operation is often utilized in antenna work because it permits operating the same antenna on several harmonically-related amateur bands. It is also an important principle in the operation of certain types of directive antennas.

Electrical Length

The *electrical* length of a linear circuit such as an antenna wire is not necessarily the same as its *physical* length in wavelengths or fractions of a wavelength. Rather, the electrical length is measured by the *time* taken for the completion of a specified phenomenon.

For instance, we might imagine two linear circuits having such different characteristics that the speed at which a charge travels is not the same in both. Suppose we wish to make both circuits resonant to the same frequency, and for that purpose adjust the physical length of each until a charge started at one end travels to the far end, is reflected and completes its return journey to the near end in exactly the time of one r.f. cycle. Then it will be found that the physical length of the circuit with the lower velocity of propagation is shorter than the *physical* length of the other. The *electrical* lengths, however, are identical, each being a half-wavelength.

In alternating-current circuits the instantaneous values of current or voltage are determined by the instant during the cycle at which the measurement is made (assuming, of course, that such a measurement could be made rapidly enough). If the current and voltage follow a sine curve — which is the usual case — the time, for any instantaneous value, can be specified in terms of an angle. The sine of the angle gives the instantaneous value when multiplied by the *peak* value of the current or voltage. A complete sine curve occupies the 360 degrees of a circle and represents one cycle of a.c. current or voltage. Thus a half cycle is equal to 180 degrees, a quarter cycle to 90 degrees, and so on.

It is often convenient to use this same form of representation for linear circuits. When the electrical length of such a circuit is such that a charge, *traveling in one direction*, takes the time of one cycle to traverse it, the length of the circuit is said to be 360 degrees. This corresponds to one wavelength. On a wire a half-wave in electrical length the charge completes a one-way journey in one-half cycle, and its length is said to be 180 degrees. The angular method of measurement is quite useful for lengths that are not easily-remembered fractions or simple multiples of such fractions. A chart for converting fractions of a wavelength to degrees is given in Fig. 2-3.

Velocity of Propagation

The speed or velocity at which electromagnetic waves travel through a medium depends upon the dielectric constant of the medium. At radio frequencies the dielectric constant of air is practically unity, so the waves travel at essentially the same velocity as light in a vacuum. This is also the velocity of the charge traveling along a wire.

If the dielectric constant is greater than 1, the velocity of propagation is lowered. Thus the introduction, in appreciable quantity, of insulating material having a dielectric constant greater than 1 will cause the wave to slow down. This effect is frequently encountered in practice in connection with both antennas and transmission lines. It causes the electrical length of the line or antenna to be somewhat greater than the actual physical length.

Length of a "Half-Wave" Antenna

Even if the antenna could be supported by insulators that did not cause the electromagnetic fields traveling along the wire to slow down, the

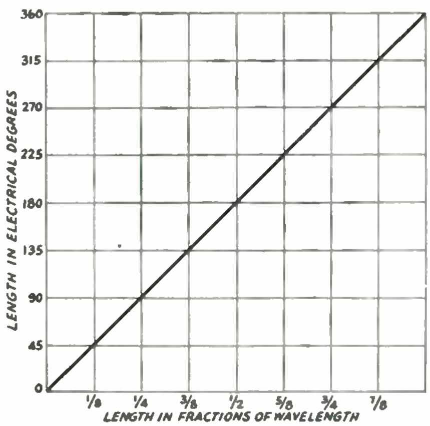


Fig. 2-3 — Chart for converting electrical degrees to fractions of a wavelength.

physical length of a practical antenna always is somewhat less than its electrical length. That is, a "half-wave" antenna is not one having the same length as a half wavelength in space. It is one having an *electrical* length equal to 180 degrees. Or, to put it another way, it is one whose length has been adjusted to "tune out" any reactance, so it is a *resonant* antenna.

The antenna length required to resonate at a given frequency (independently of any dielectric effects) depends on the ratio of the length of the conductor to its diameter. The smaller this ratio, the shorter the antenna for a given electrical length. This effect is shown in Fig. 2-4 as a factor (K) by which a free-space half wavelength must be multiplied to find the resonant length, as a function of the length/diameter ratio. If the antenna conductor could be infinitely small in diameter, K would equal 1 and the antenna length would be equal to a free-space half wavelength. A length/diameter ratio of 10,000 is roughly average for wire antennas (actually, it is approximately the ratio for a 7-Mc. half-wave antenna made of No. 12 wire). In this region K changes rather slowly and a half-wave antenna made of wire is about 2 per cent shorter than a half wavelength in space.

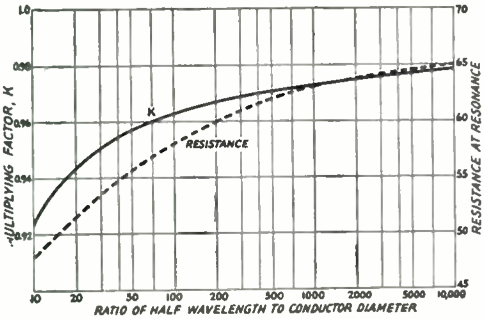


Fig. 2-4 — The solid curve shows the factor, K, by which the length of a half wave in free space should be multiplied to obtain the physical length of a resonant half-wave antenna having the length/diameter ratio shown along the horizontal axis. The broken curve shows how the radiation resistance of a half-wave antenna varies with the length/diameter ratio.

The shortening effect is most pronounced when the length/diameter ratio is 100 or less. An antenna constructed of 1-inch diameter tubing for use on 144 Mc., for example, would have a length/diameter ratio of about 40 and would be almost 5 per cent shorter than a free-space half wavelength.

If the antenna is made of rod or tubing and is not supported near the ends by insulators, the following formula will give the required physical length of a half-wave antenna to a close enough approximation:

$$\textit{Length} \text{ (feet)} = \frac{492 \times K}{f \text{ (Mc.)}}$$

or

$$\textit{Length} \text{ (inches)} = \frac{5905 \times K}{f \text{ (Mc.)}}$$

where K is taken from Fig. 2-4 for the particular length/diameter ratio of the conductor used.

End Effect

If the formulas of the preceding section are used to determine the length of a wire antenna, the antenna will resonate at a somewhat lower frequency than is desired. The reason for this is that there is an additional "loading" effect caused by the insulators that must be used at the ends of the wire for suspending it. These insulators and the wire loop that ties the insulator to the antenna add a small amount of capacitance to the system. This capacitance helps to tune the antenna to a slightly lower frequency, in much the same way that additional capacitance in any tuned circuit will lower the resonant frequency. In an antenna it is called **end effect.**

End effect increases with frequency and varies slightly with different installations. However, at frequencies up to 30 Mc. (the frequency range over which wire antennas are most likely to be used) experience shows that the length of a half-wave antenna is about 5 per cent less than the length of a half wave in space. As an average, then, the physical length of a resonant half-wave antenna may be taken to be

$$l \text{ (feet)} = \frac{492 \times 0.95}{f \text{ (Mc.)}}$$

or

$$l \text{ (feet)} = \frac{468}{f \text{ (Mc.)}}$$

This formula is sufficiently accurate, for all practical purposes, for finding the physical length of a half-wave antenna for a given frequency, but does not apply to antennas longer than a half wave in length.

The current at the ends of the antenna does not quite reach zero because of the end effect, as there is some current flowing into the end capacitance.

● ANTENNA IMPEDANCE

In the simplified description given earlier of voltage and current distribution along an antenna it was stated that the voltage was zero at the center of a half-wave antenna (or at any current loop along a longer antenna). It is more accurate to say that the voltage reaches a *minimum* rather than zero. Zero voltage with a finite value of current would imply that the circuit is entirely without resistance. It would also imply that no energy is radiated by the antenna, since a circuit without resistance would take no real power from the driving source.

Actually, of course, an antenna, like any other circuit, consumes power. The current that flows in it therefore must be supplied at a finite value of voltage. The **impedance** of the antenna is simply equal to the voltage applied to its terminals divided by the current flowing into those terminals. If the current and voltage are exactly in phase the impedance is purely resistive. This is the case when the antenna is resonant. If the antenna is not exactly resonant the current will be somewhat out of phase with the applied voltage and the antenna shows reactance along with resistance.

Most amateur transmitting antennas are operated at or quite close to resonance so that reactive effects are in general comparatively small. They are nevertheless present, and must be taken into account, whenever an antenna is operated at other than the exact frequency for which it is designed.

In the following discussion it is assumed that power is applied to the antenna by opening the conductor at the center and applying the driving voltage across the gap. This is shown in Fig. 2-5. While it is possible to supply power to the antenna by other methods, the selection of different driving points leads to different values of impedance; this can be appreciated after study of Fig. 2-1, which shows that the ratio of voltage to current (which is, by definition, the impedance) is different at every point along the antenna. To avoid confusion, therefore, it is desirable to use the conditions at the center of the antenna as a basis.

The Antenna as a Circuit

If the frequency applied at the center of a half-wave antenna is varied above and below the resonant frequency, the antenna will exhibit much the same characteristics as a conventional series-resonant circuit. Exactly at resonance the current at the input terminals will be in phase with the applied voltage. If the frequency is on the low side of resonance the phase of the current will lead the voltage; that is, the reactance of the antenna is capacitive. When the frequency is on the high side of resonance the opposite occurs; the current lags the applied voltage and the antenna exhibits inductive reactance.

It is not hard to see why this is so. Consider the antennas shown in Fig. 2-6, one resonant, one too long for the applied frequency, and one too short.

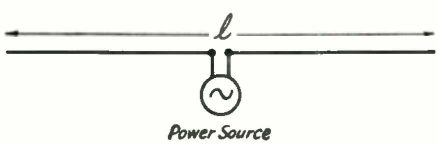


Fig. 2-5 — The center-fed antenna discussed in the text. It is assumed that the leads from the source of power to the antenna have zero length.

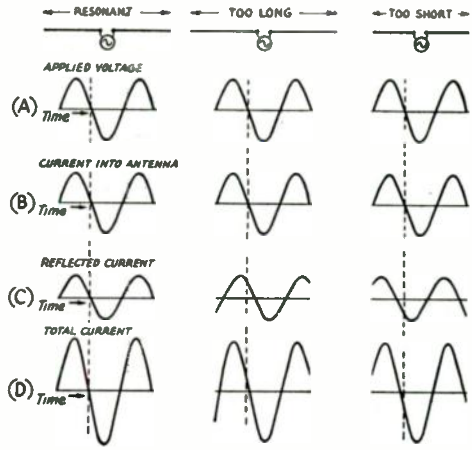


Fig. 2-6 — Current flow in resonant and off-resonant antennas. The initial current flow, B, caused by the source of power, is in phase with the applied voltage, A. This is the outgoing current discussed in the text. The reflected current, C, combines with the outgoing current to form the resultant current, D, at the input terminals of the antenna.

In each case the applied voltage is shown at A, and also in each case the instantaneous current going *into* the antenna because of the applied voltage is shown at B. Note that this current is always in phase with the applied voltage, regardless of the antenna length. For the sake of simplicity only the current flowing in one leg of the antenna is considered; conditions in the other leg are similar.

In the case of the resonant antenna, the current travels out to the end and back to the driving point in one-half cycle, since one leg of the antenna is 90 degrees long and the total path out and back is therefore 180 degrees. This would make the phase of the *reflected* component of current differ from that of the outgoing current by 180 degrees, since the latter current has gone through a half cycle in the meantime. However, it will be remembered that there is a phase shift of 180 degrees at the end of the antenna, because the direction of current reverses at the end. The *total* phase shift between the outgoing and reflected currents, therefore, is 360 degrees. In other words, the reflected component arrives at the driving point exactly in phase with the outgoing component. The reflected component, shown at C, adds to the outgoing component to form the resultant or total current at the driving point. The resultant current is shown at D, and in the case of the resonant antenna it is easily seen that the resultant is exactly in phase with the applied voltage. This being the case, the load seen by the source of power is a pure resistance.

Now consider the antenna that is too short to be resonant. The outgoing component of current is still in phase with the applied voltage, as shown at B. The reflected component, however, gets back to the driving point *too soon,* because it travels over a path less than 180 degrees, out and back. This means that the maximum value of the reflected component occurs at the driving point ahead of (in time) the maximum value of the outgoing component, since that particular charge took less than a half cycle to get back. Including the 180-degree reversal at the end of the antenna, the total phase shift is therefore less than 360 degrees. This is shown at C, and the resultant current is the combination of the outgoing and reflected components as given at D. It can be seen that the resultant current leads the applied voltage, so the antenna looks like a resistance in series with a capacitance. The shorter the antenna the greater the phase shift between voltage and current; that is, the capacitive reactance increases as the antenna is shortened.

When the antenna is too long for the applied frequency the reflected component of current arrives too late to be exactly in phase with the outgoing component, because it must travel over a path more than 180 degrees long. The maximum value of the reflected component therefore occurs later (in time) than the maximum value of the outgoing component, as shown at C. The resultant current at the antenna input terminals therefore lags behind the applied voltage. The phase lag increases as the antenna is made longer. That is, an antenna that is too long shows inductive reactance along with resistance, and this reactance increases with an increase in antenna length over the length required for resonance.

If the antenna length is increased to 180 degrees on each leg the path length for the reflected component becomes 360 degrees. This, plus the 180-degree reversal at the end, makes the total phase shift 540 degrees, which is the same as a 180-degree shift. In this case the reflected current arrives at the input terminals exactly *out* of phase with the outgoing component, so the resultant current is very small. The resultant is in phase with the applied voltage, so the antenna impedance is again purely resistive. The resistance under these conditions is very high, and the antenna has the characteristics of a parallel-resonant circuit. A *voltage* loop, instead of a *current* loop, appears at the input terminals when each leg of the antenna is 180 degrees long.

The amplitude of the reflected component is less than that of the component of current going into the antenna. This is the result of energy loss by radiation as the current travels along the wire. It is perhaps easier to understand if, instead of thinking of the electromagnetic fields as being brought into being by the current flow, we adopt the more fundamental viewpoint that *current flow along a conductor is caused by a moving electromagnetic field.* When some of the energy escapes from the system because the field travels out into space, it is not hard to understand why the current becomes less the farther it travels. There is simply less energy left to cause it. The difference between the outgoing- and reflected-current amplitudes accounts for the fact that the current does not go to zero at a voltage loop, and a similar difference between the applied- and reflected-voltage components explains why the voltage does not go to zero at a current loop.

Resistance

The energy supplied to an antenna is dissipated in the form of radio waves and in heat losses in the wire and near-by dielectrics. The radiated energy is the useful part, but so far as the antenna is concerned it represents a loss just as much as the energy lost in heating the wire is a loss. In either case the dissipated power is equal to I^2R: in the case of heat losses, the R is a real resistance, but in the case of radiation R is an assumed resistance, which, if present, would dissipate the power that actually disappears by radiation. This fictitious resistance is called the **radiation resistance.** The total power loss in the antenna is therefore equal to $I^2(R_0 + R)$, where R_0 is the radiation resistance and R the real resistance, or **ohmic** resistance.

In the ordinary half-wave antenna operated at amateur frequencies the power lost as heat in the conductor does not exceed a few per cent of the total power supplied to the antenna. This is because the r.f. resistance of copper wire even as small as No. 14 is very low compared with the radiation resistance of an antenna that is reasonably clear of surrounding objects and is not too close to the ground. Therefore it can be assumed that the ohmic loss in a reasonably well-located antenna is negligible, and that all of the resistance shown by the antenna is radiation resistance. As a radiator of electromagnetic waves such an antenna is a highly-efficient device.

The value of radiation resistance, as measured at the center of a half-wave antenna, depends on a number of factors. One is the location of the antenna with respect to other objects, particularly the earth. Another is the length/diameter ratio of the conductor used. In "free space" — with the antenna remote from everything else — the radiation resistance of a resonant antenna made of an infinitely-thin conductor is approximately 73 ohms. While there is no such thing as having an antenna in free space, it is a convenient basis for calculation because the modifying effect of the ground can be taken into account separately. If the antenna is at least several wavelengths away from ground and other objects, it can be considered to be in free space for all practical purposes. This condition can be met with antennas in the v.h.f. and u.h.f. range.

The way in which the free-space radiation resistance varies with the length/diameter ratio of a half-wave antenna is shown by the broken curve in Fig. 2-4. As the antenna is made thicker the radiation resistance decreases. For most wire antennas it is close to 65 ohms. It will usually lie between 55 and 60 ohms for antennas constructed of rod or tubing.

The actual value of the radiation resistance — at least so long as it is 50 ohms or more — has no appreciable effect on the radiation efficiency of the antenna. This is because the ohmic resistance is only of the order of 1 ohm with the conductors used for thick antennas. The ohmic resistance does not become important until the radiation resistance drops to very low values — say less than 10 ohms — as may be the case when several antennas are coupled together.

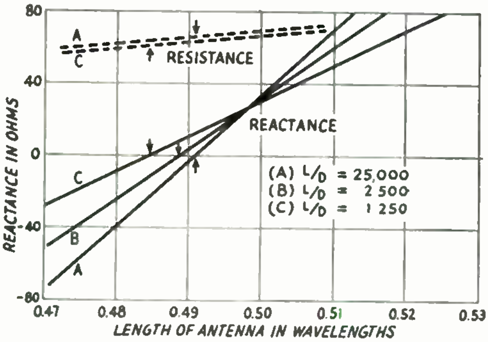


Fig. 2-7 — **Resistance and reactance at input terminals of a center-fed antenna as a function of its length near a half wavelength. As shown by curves *A*, *B* and *C*, the reactance is affected more by the length/diameter ratio of the conductor than is the radiation resistance.**

The radiation resistance of a resonant antenna is the "load" for the transmitter or for the r.f. transmission line connecting the transmitter and antenna. Its value is important, therefore, in determining the way in which the antenna and transmitter or line are coupled together. The power must be supplied at the right voltage if the system as a whole is to be efficient.

The resistance of an antenna varies with its length as well as with the ratio of its length to its diameter. When the antenna is approximately a half wave long, the resistance changes rather slowly with length. This is shown by the curves of Fig. 2-7, where the change in resistance as the length is varied a few per cent on either side of resonance is shown by the broken curves. The resistance decreases somewhat when the antenna is slightly short, and increases when it is slightly long. These curves also illustrate the effect of changing the frequency applied to an antenna of fixed length, since increasing the frequency above resonance is the same thing as having an antenna that is too long, and vice versa.

The range covered by the curves in Fig. 2-7 is representative of the frequency range over which a fixed antenna is operated between the limits of an amateur band. At greater departures from the resonant length the resistance continues to decrease about uniformly as the antenna is shortened, but tends to increase rapidly as the antenna is made longer. The resistance increases very rapidly when the length of a leg exceeds about 135 degrees, or about ⅜ wavelength, and reaches a maximum when the length of one side is 180 degrees. This is considered in more detail in a later chapter.

Reactance

The rate at which the reactance of the antenna increases as the length is varied from resonance depends on the length/diameter ratio of the conductor. The thicker the conductor the smaller the reactance change for a given change in length. This is shown by the reactance curves in Fig. 2-7. Curves for three values of length/diameter ratio are shown; *L* represents the length of a half wave in space, approximately, and *D* is the diameter of the conductor in the same units as the length. The point where each curve crosses the zero axis (indicated by an arrow in each case) is the length at which an antenna of that particular length/diameter ratio is resonant. The effect of *L/D* on the resonant length also is illustrated by these curves: the smaller the ratio, the shorter the length at which the reactance is zero.

It will be observed that the reactance changes about twice as rapidly in the antenna with the smallest diameter (*A*) as it does in the antenna with the largest diameter (*C*). With still larger diameters the rate at which the reactance changes would be even smaller. As a practical matter it is advantageous to keep the reactance change with a given change in length as small as possible. This means that the reactance is comparatively low, when the antenna is operated over a small band of frequencies centered on its resonant frequency, and that the impedance change with frequency is small. This simplifies the problem of supplying power to the antenna when it must be worked at frequencies somewhat different from its resonant frequency.

At lengths considerably different from the resonant length the reactance changes more rapidly than in the curves shown in Fig. 2-7. As in the case of the resistance change, the change is most rapid when the length exceeds 135 degrees (⅜ wavelength) and approaches 180 degrees (½ wavelength) on a side. In this case the reactance is inductive and reaches a maximum at a length somewhat less than 180 degrees. Between this maximum point and 180 degrees of electrical length the reactance decreases very rapidly, becoming zero when the length is such as to be parallel-resonant.

Very short antennas have a large capacitive reactance. It was pointed out in the preceding section that with antennas shorter than 90 degrees on a side the resistance decreases at a fairly uniform rate, but this is not true of the reactance. It increases rather rapidly when the length of a side is shortened below about 45 degrees.

The behavior of antennas with different length/diameter ratios corresponds with the behavior of ordinary resonant circuits having different Qs. When the Q of a circuit is low, the reactance is small and changes rather slowly as the applied frequency is varied on either side of the resonant frequency. If the Q is high, the converse is true. The response curve of the low-Q circuit is "broad"; that of the high-Q circuit "sharp". So it is with antennas; a thick antenna works well over a comparatively wide band of frequencies while a thin antenna is rather sharp in tuning. The Q of the thick antenna is low; the Q of the thin antenna is high, assuming essentially the same value of radiation resistance in both cases.

Coupled Antennas

A conventional tuned circuit far enough away from all other circuits so that no external coupling exists can be likened to an antenna in free space, in one sense. That is, its characteristics are unaffected by its surroundings. It will have a Q and resonant impedance determined by the inductance, capacitance and resistance of which it is composed, and those quantities alone. But as soon as it is coupled to another circuit its Q and impedance will change, depending on the characteristics of the other circuit and the degree of coupling.

A similar situation arises when two or more "elementary" antennas — half-wave antennas, frequently called **half-wave dipoles** — are coupled together. This coupling takes place merely by having the two antennas in proximity to each other. The sharpness of resonance and the radiation resistance of each "element" of the system are affected by the mutual interchange of energy between the coupled elements. The exact effect depends on the degree of coupling (that is, how close the antennas are to each other in terms of wavelength, and whether the wires are parallel or not) and the tuning condition (whether tuned to resonance or slightly off resonance) of each element. Analysis is extremely difficult and even then has to be based on some simplifying assumptions that may not be true in practice. Only a few relatively simple cases have been analyzed. Such data as are available for even moderately complicated systems of coupled antennas are confined to a few types and are based on experimental measurements. They are, therefore, subject to the inaccuracies that accompany any measurements in a field where measurement is difficult at best.

Antenna systems consisting of coupled elements will be taken up in later chapters. At this point it is sufficient to appreciate that the free-space values that have been discussed in this chapter may be modified drastically when more than one antenna element is involved in the system. It has already been pointed out, also, that the presence of the ground, as well as near-by conductors or dielectrics, also will modify the free-space values. The free-space characteristics of the elementary half-wave dipole are only the point of departure for a practical antenna system. In other words, they give the basis for understanding antenna principles but cannot be applied too literally in the practical case.

It is of interest to note that the comparison between an isolated tuned circuit and an antenna in free space is likewise not to be taken too literally. In one sense the comparison is wholly misleading. The tuned circuit is usually so small, physically, in comparison with the wavelength, that practically no energy escapes from it by radiation. An antenna, to be worthy of the name, is always so large in comparison with the wavelength that practically *all* the energy supplied to it escapes by radiation. Thus the antenna can be said to be very tightly coupled to space, while the tuned circuit is not coupled to anything. This very fundamental difference is one reason why antenna systems cannot be analyzed as readily, and with as satisfactory results in the shape of simple formulas, as ordinary electrical circuits.

● HARMONICALLY-OPERATED ANTENNAS

An antenna operated at a harmonic of its fundamental frequency has considerably different properties than the half-wave dipole previously discussed. It must first be emphasized that harmonic operation implies that there is a reversal of the direction of current flow in alternate half-wave sections of the antenna, as shown in Fig. 2-2 and again at A in Fig. 2-8. In the latter figure, the curve shows the standing wave of current intensity along the wire; the curve is above the line to indicate current flow in one direction (assumed to be to the right, in the direction of the arrow) and below the line to indicate current flow in the opposite direction in the other half-wave section. (During the next radio-frequency half cycle the current flow in the left-hand half-wave section would be toward the left, and in the right-hand half-wave section to the right; this alternation in direction takes place in each succeeding half cycle. However, the direction of current flow in adjacent half-wave sections is at all times opposite.) The antenna in this drawing is one wavelength long and is operating on its second harmonic.

Now consider the half-wave antenna shown at Fig. 2-8B. It is opened in the center and fed by a source of r.f. power through leads that are assumed to have zero length. Since one terminal of the generator is positive at the same instant that the other terminal is negative, current flows *into* one side of the generator while it is flowing *out* at the other terminal. Consequently the current flows in the same direction in both sections of the half-wave antenna. It has the amplitude distribution shown by the curve over the antenna wire.

If we now increase the length of the wire on each side of the generator in Fig. 2-8B to one-half wavelength, we have the situation shown in Fig. 2-8C. At the instant shown, current flows into the generator from the left-hand half-wave section, and out of the generator into the right-hand half-wave section. Thus the currents in the two sections are in the same direction, just as they were in Fig. 2-8B. The current distribution in this case obviously is not the same as in Fig. 2-8A. Although the over-all lengths of the antennas shown at A and C is the same, the antenna at A is operating on a harmonic but the one in C is not.

For true harmonic operation it is necessary that the power be fed into the antenna at the proper point. Two methods that result in the proper current distribution are shown at D and E in Fig. 2-8. If the source of power is connected to the antenna at one end, as in D, the direction of current flow will be reversed in alternate half-wave sections. Or if the power is inserted at the center of a half-wave section, as in E, there will be a similar reversal of current in the next half-wave section. For harmonic operation, therefore, the antenna should be fed either at the end or at a current loop. If the feed point is at a current node the current distribution will not be that expected on a harmonic antenna.

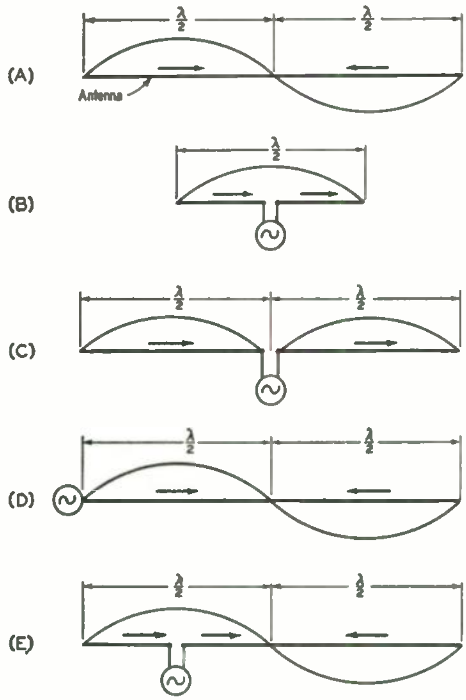


Fig. 2-8 — How the feed point makes a difference in current distribution along the antenna. With center feed, increasing the length of each side of the antenna keeps the current flowing in the same direction in the two halves, up to the point where each side is a half wavelength long. For harmonic operation, the antenna must be fed in such a way that the current direction reverses in alternate half-wavelength sections. Suitable methods are shown at D and E.

Length of a Harmonic Wire

The physical length of a harmonic antenna is not exactly the same as its electrical length, for the same reasons discussed earlier in connection with the half-wave antenna. The physical length is somewhat shorter than the length of the same number of half waves in space because of the length/diameter ratio of the conductor and the end effects. Since the latter are appreciable only where insulators introduce additional capacitance at a high-voltage point along the wire, and since a harmonic antenna usually has such insulation only at the ends, the end-effect shortening affects only the half-wave sections at each end of the antenna. It has been found that the following formula for the length of a harmonic antenna of the usual wire sizes works out very well in practice:

$$\text{Length (feet)} = \frac{492\,(N-0.05)}{\text{Freq. (Mc.)}}$$

where N is the number of *half* waves on the antenna.

Because the number of half waves varies with the harmonic on which the antenna is operated, consideration of this formula as well as that for the half-wave antenna (the fundamental frequency) will show that the relationship between the fundamental frequency and the harmonics of an antenna is not exactly integral. That is, the "second-harmonic" frequency to which a given length of wire is resonant is not exactly twice the fundamental frequency; the "third-harmonic" resonance is not exactly three times the fundamental, and so on. The actual resonant frequency on a harmonic always is a little higher than the exact multiple of the fundamental. A second-harmonic or full-wave antenna, for example, must be a little longer than twice the length of a half-wave antenna cut for the same frequency.

Impedance of Harmonic Antennas

A harmonic antenna can be looked upon as a series of half-wave sections placed end to end (collinear) and supplied with power in such a way that the currents in alternate sections are out of phase. There is a certain amount of coupling between adjacent half-wave sections. Because of this coupling the impedance as measured at a current loop in a half-wave section is not the same as the impedance at the center of a half-wave antenna.

Just as in the case of a half-wave antenna, the impedance consists of two main components, radiation resistance and reactance. The ohmic or loss resistance is low enough to be ignored in the practical case. If the antenna is exactly resonant there will be no reactance at the input terminals and the impedance consists only of the radiation resistance. The value of the radiation resistance depends on the number of half waves on the wire and, as in the case of the half-wave antenna, is modified by the presence of near-by conductors and dielectrics, particularly the earth. As a point of departure, however, it is of interest to know the order of magnitude of the radiation resistance of a theoretical harmonic antenna consisting of an infinitely-thin conductor in free space, with its length adjusted to exact harmonic resonance. The radiation resistance of such an antenna having a length of one wavelength is approximately 90 ohms, and as the antenna length is increased the resistance also increases. At ten wavelengths it is approximately 160 ohms, for example. The way in which the radiation resistance of a theoretical harmonic wire varies with length is shown by curve *A* in Fig. 2-23, page 44. It is to be understood that the radiation resistance always is measured at a current loop.

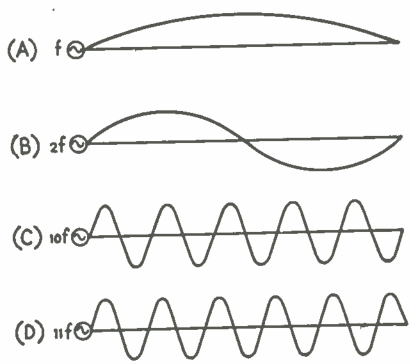


Fig. 2-9 **— The percentage frequency change from one high-order harmonic to the next (for example, between the 10th and 11th harmonics shown at C and D) is much smaller than between the fundamental and second harmonic (A and B). This makes impedance variations more rapid as the wire becomes longer in terms of wavelength.**

When the antenna is operated at a frequency slightly off its exact resonant frequency, reactance as well as resistance will appear at its input terminals. In a general way, the reactance varies with applied frequency in much the same fashion as in the case of the half-wave antenna already described. However, the reactance varies at a more rapid *rate* as the applied frequency is varied; on a harmonic antenna a given percentage change in applied frequency causes a greater change in the phase of the reflected current as related to the outgoing current than is the case with a half-wave antenna. This is because, in traveling the greater length of wire in a harmonic antenna, the reflected current gains the same amount of time in *each* half-wave section, if the antenna is too short for resonance, and these gains add up as the current travels back to the driving point. When the antenna is too long, the reverse occurs and the reflected current progressively drops behind in phase as it travels back to the point at which the voltage is applied. This effect, obviously, increases with the length of the antenna, and the change of phase can be quite rapid when the frequency applied to an antenna operated on a high-order harmonic is varied.

Another way of looking at it is this: Consider the antenna of Fig. 2-9A, driven at the end by a source of power having a frequency *f*. When *f* is equal to the fundamental resonant frequency of the antenna there is one half wavelength on the wire, with the current distribution as shown. At this frequency the antenna is exactly resonant and it appears as a pure resistance to the source of power.

If the frequency is now increased slightly the antenna will be too long and the resultant current at the input terminals will lag behind the applied voltage (as explained by Fig. 2-6) and the antenna will have inductive reactance along with resistance. As we continue to raise the frequency the value of reactance increases to a maximum and then decreases, reaching zero when the frequency is such that the wire is an odd number of *quarter* wavelengths long. On further increasing the frequency the reactance becomes capacitive, increases to a maximum and then decreases to zero again. At this second zero-reactance point there are two complete standing waves of current (two half waves or one wavelength) and the wire is exactly resonant on its second harmonic. This last condition is shown in Fig. 2-9B; the frequency is now 2*f*, twice the original figure. In varying the frequency from *f* to 2*f* the resistance seen by the source of power also varies, decreasing as the frequency is raised above *f* and reaching a minimum when the wire is an odd number of quarter wavelengths long, then rising again with increasing frequency until it reaches a new maximum when the frequency is 2*f*.

This behavior of reactance and resistance is shown in Fig. 2-10. A similar change in reactance and resistance occurs when the frequency is moved from *any* harmonic to the *next adjacent* one, as well as between the fundamental and second harmonic shown in the drawing. That is, the impedance goes through a cycle, starting with a high value of pure resistance, then becoming inductive and decreasing, passing through a low value of pure resistance, and then becoming capacitive and increasing until it again reaches a high value of pure resistance. This cycle occurs as the frequency is continuously varied from any harmonic to the next higher one.

Looking now at Fig. 2-9C, the frequency has been increased to 10*f*, ten times its original value, so the antenna is operated on its tenth harmonic. Raising the frequency to 11*f*, the eleventh harmonic, causes the impedance of the antenna to go through the complete cycle described above. But 11*f* is only 10% higher than 10*f*, so a 10% change in frequency has caused a complete impedance cycle. In contrast, changing from *f* to 2*f* is a 100% increase in frequency — for the same impedance cycle. The impedance changes ten times as fast when the frequency is varied about the 10th harmonic as it does when the frequency is varied the same percentage about the fundamental.

To offset this, the actual impedance change — that is, the ratio of the maximum to the minimum impedance through the impedance cycle — is not as great at the higher harmonics as it is near the fundamental. This is because the radiation resistance increases with the order of the harmonic, raising the minimum point on the resistance curve and also lowering the maximum point. In practice, also, the reflected current returning to the input end of a long harmonic wire is not as great as the outgoing current because energy has been lost by radiation; this is not taken into account in the theoretical pictures of current distribution so far discussed.

The over-all result, then, is that the *magnitude* of the impedance variations becomes less as the wire is operated at increasingly higher harmonics. Nevertheless, the impedance is maximum at each adjacent harmonic and minimum halfway between, independently of the actual values of impedance.

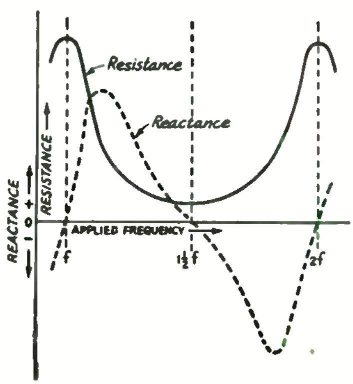


Fig. 2-10 — This drawing shows qualitatively the way in which the reactance and resistance of an end-fed antenna vary as the frequency is increased from the fundamental (*f*) to the second harmonic (2*f*). Actual values of resistance and reactance and the frequencies at which the reactances are maximum will depend on the size of the conductor and the height of the antenna above ground.

OTHER ANTENNA PROPERTIES

Polarization

The polarization of a half-wave dipole is the same as the direction of its axis. That is, at distances far enough from the antenna for the waves to be considered as plane waves (see Chapter One) the direction of the electric component of the field is the same as the direction of the antenna wire. Vertical and horizontal polarization, the two most commonly used for antennas, are indicated in Fig. 2-11.

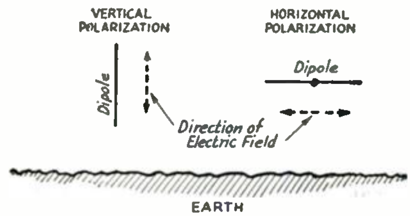


Fig. 2-11 — Vertical and horizontal polarization of a dipole. The direction of polarization is the direction of the electric field with respect to earth.

Antennas composed of a number of half-wave elements with all arranged so that their axes lie in the same or parallel directions will have the same polarization as that of any one of the elements. A system composed of a group of horizontal dipoles, for example, will be horizontally polarized. If both horizontal and vertical elements are used, the polarization will be the *resultant* of the contributions made by each set of elements to the total electromagnetic field at a distance. In such a case the resultant polarization will be tilted between horizontal and vertical.

Harmonic antennas also are polarized in the direction of the wire axis. However, in some combinations of harmonic wires such as the "V" and rhombic antennas described in a later chapter the polarization becomes elliptical in most directions with respect to the antenna.

As pointed out in Chapter One, sky-wave transmission usually changes the polarization of the traveling waves. The polarization of receiving and transmitting antennas in the 3-to-30 Mc. range, where almost all communication is by means of the sky wave (except for distances of a few miles), is therefore relatively unimportant. In this range the choice of polarization for the antenna is usually determined by other factors such as the height of available antenna supports, the polarization of man-made r.f. noise from near-by sources, probable energy losses in near-by houses, wiring, etc., and the likelihood of interfering with neighborhood broadcast reception. Generally speaking, the most favorable solution to these problems is achieved when the antenna is horizontally polarized.

Reciprocity in Receiving and Transmitting

The basic conditions existing when an antenna is used for radiating power are not the same as when it is used for receiving a distant signal. In the former case the electromagnetic field originates with the antenna and the waves are not plane-polarized in the immediate vicinity. In the receiving case the antenna is always far enough away from the transmitter so that the waves that the antenna intercepts are plane-polarized. This causes the current distribution in a receiving antenna to be different than in a transmitting antenna except in a few special cases. These, however, are the cases of most interest in amateur practice, since they occur when the antenna is resonant and is delivering power to a receiver.

For all practical purposes, then, the properties of a resonant antenna used for reception are the same as its properties in transmission. It has the same directive pattern in both cases, and so will deliver maximum signal to the receiver when the signal comes from the direction in which the antenna transmits best. The impedance of the antenna is the same, at the same point of measurement, in receiving as in transmitting.

In the receiving case, the antenna is to be considered as the *source* of power delivered to the receiver, rather than as the *load* for a source of power as in transmitting. Maximum output from the receiving antenna is secured when the load to which the antenna is connected is equal to the impedance of the antenna. Under these conditions half of the total power picked up by the antenna from the passing waves is delivered to the receiver and half is reradiated into space. "Impedance matching" in the case of a receiving antenna does not have quite the same meaning as in the transmitting case; this is considered in later chapters.

The power gain (defined later in this chapter) in receiving is the same as the gain in transmitting, assuming that certain conditions are met. One such condition is that both the antenna under test and the comparison antenna (usually a half-wave antenna) work into load impedances equal to their own impedances so that maximum power is delivered in both cases. In addition, of course, the comparison antenna should be oriented so that it gives maximum response to the signal used in the test; that is, it should have the same polarization as the incoming signal and should be placed so that its direction of maximum gain is toward the signal source.

In long-distance transmission and reception via the ionosphere the relationship between receiving and transmitting may not be exactly reciprocal. This is because the waves do not take exactly the same paths at all times and so may show considerable variation in alternate transmission and reception. Also, when more than one layer is involved in the wave travel it is sometimes possible for transmission to be good in one direction and reception to be poor in the other, over the same path. In addition, the polarization of the waves is shifted in the ionosphere, as pointed out in Chapter One. Since the tendency is for the arriving wave to be more or less horizontally-polarized, regardless of the polarization of the transmitting antenna, a vertically-polarized antenna can be expected to show more difference between transmission and reception than a horizontal antenna. On the average, however, an antenna that transmits well in a certain direction will also give favorable reception from the same direction, despite possible ionosphere variations.

Pick-Up Efficiency

Although the transmitting and receiving properties of an antenna are, in general, reciprocal, there is another fundamental difference between the two cases that is of very great practical importance. In the transmitting case all the power supplied to the antenna is radiated (assuming negligible ohmic resistance) regardless of the physical size of the antenna system. For example, a 300-Mc. half-wave radiator, which is only about 19 *inches* long, radiates every bit as efficiently as a 3.5-Mc. half-wave antenna, which is about 134 *feet* long. But, in receiving, the 300-Mc. antenna does not abstract anything like the amount of energy from passing waves that the 3.5-Mc. antenna does.

This is because the section of wave front from which the antenna can draw energy extends only about a quarter wavelength from the conductor. At 3.5 Mc. this represents an area roughly ½ wavelength or 134 feet in diameter, but at 300 Mc. the diameter of the area is only about 3 feet. Since the energy is evenly distributed throughout the wave front regardless of the wavelength, the effective area that the receiving antenna can utilize varies directly with the *square* of the wavelength. A 3.5-Mc. half-wave antenna therefore picks up something like 2500 times as much energy as a 300-Mc. half-wave antenna, the field strength of the signal being the same in both cases.

The higher the frequency, consequently, the less energy a receiving antenna has to work with. This, it should be noted, does not affect the *gain* of the antenna. In making gain measurements, both the antenna under test and the comparison antenna are working at the same frequency. Both therefore are under the same handicap with respect to the amount of energy that can be intercepted. Although the pick-up efficiency decreases rapidly with increasing frequency, the smaller dimensions of antenna systems in the v.h.f. and u.h.f. regions make it relatively easy to obtain high gain. This helps to overcome the loss of received signal energy.

The Induction Field

Throughout this chapter the fields we have been discussing are those forming the traveling electromagnetic waves — the waves that go long distances from the antenna. These are the **radiation fields.** They are distinguished by the fact that their intensity is inversely proportional to the distance and that the electric and magnetic components, although perpendicular to each other in the wave front, are in phase in time. At a distance of several wavelengths from the antenna these are the only fields that need to be considered.

Close to the antenna, however, the fields are much more complicated. In an ordinary electric circuit containing inductance or capacitance the magnetic field is a quarter cycle out of phase (in time) with the electric field. The intensity of these fields decreases in a complex way with distance from the source. These are the **induction fields.** The induction field exists about an antenna along with the radiation field, but dies away with much greater rapidity as the distance from the antenna is increased. At a distance equal to the wavelength divided by 2π, or slightly less than 1/6 wavelength, the two types of field have equal intensity.

Although the induction field is of no importance insofar as effects at a distance are concerned, it *is* important when antenna elements are coupled together, particularly when the spacing between elements is small. Also, its existence must be kept in mind in making field-strength measurements about an antenna. The measurements will not be reliable if the measuring equipment is set up too close to the antenna system.

Radiation Patterns and Directivity

The radiation from a practical antenna never has the same intensity in all directions. The intensity may even be zero in some directions from the antenna; in others it may be greater than one would expect from an antenna that *did* radiate equally well in all directions. But even though no actual antenna radiates with equal intensity in all directions, it is nevertheless convenient to assume that such an antenna exists. It can be used as a sort of "measuring stick" for comparing the properties of actual antenna systems. Such a hypothetical antenna is called an **isotropic radiator.**

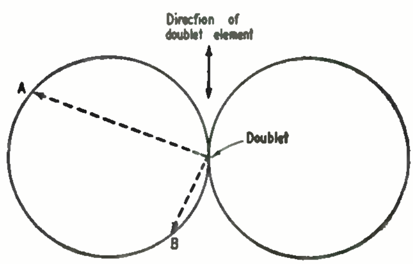


Fig. 2-12 — Directive diagram of an elementary doublet in the plane containing the wire axis. The same diagram applies in the case of any antenna that is considerably less than a half wavelength long. The length of each arrow represents the relative field strength in that direction.

A graph showing the actual or relative intensity, at a fixed distance, as a function of the direction from the antenna system is called a **radiation pattern.** At the outset it must be realized that such a pattern is a three-dimensional affair and therefore cannot be represented in a plane drawing. The "solid" radiation pattern of an antenna in free space would be found by measuring the field strength at every point on the surface of an imaginary sphere having the antenna at its center. The information so obtained is then used to construct a solid figure such that the distance from a fixed point (representing the antenna) to the surface, in any direction, is proportional to the field strength from the antenna in that direction. The solid pattern of an isotropic radiator, therefore, would be a sphere, since the field strength is the same in all directions. An infinite variety of pattern shapes, some quite complicated, is possible with actual antenna systems.

● RADIATION FROM DIPOLES

In the analysis of antenna systems it is convenient to make use of another fictitious type of antenna called an **elementary doublet or elementary dipole.** This is just a very short length of conductor, so short that it can be assumed that the current is the same throughout its length. (In an actual antenna, it will be remembered, the current is different all along its length.) The radiation intensity from an elementary doublet is greatest at right angles to the line of the conductor, and decreases as the direction becomes more nearly in line with the conductor until, right off the ends, the intensity is zero. The directive pattern in a single plane, one containing the conductor, is shown in Fig. 2-12. It consists of two tangent circles. The *solid* pattern is the figure described when the plane shown in the drawing is rotated about the conductor as an axis.

The radiation from an elementary doublet is not uniform in all directions because there is a definite direction to the current flow along the conductor. It will be recalled that a similar condition exists in the ordinary electric and magnetic fields set up when current flows along any conductor; the field strength near a coil, for ex-

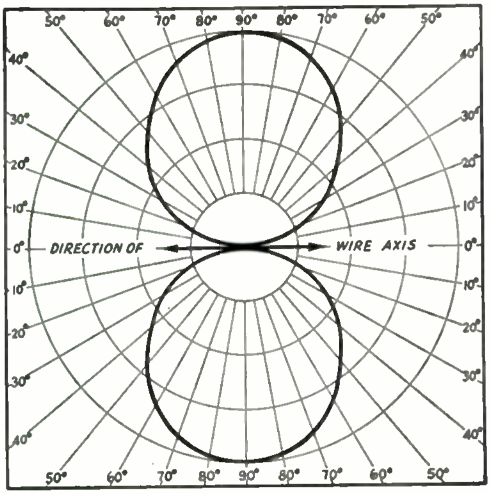


Fig. 2-13 — Plane directive diagram of a half-wave antenna. The solid line shows the direction of the wire, although the antenna itself is considered to be merely a point at the center of the diagram. As explained in the text, a diagram such as this is simply a cross section of the solid figure that describes the relative radiation in all possible directions. The radial scale is purely arbitrary and is proportional to the field strength (voltage). This is also true of the diagrams in Figs. 2-16, 2-17 and 2-18.

ample, is greatest at the ends and least on the outside of the coil near the middle of its length. There is nothing strange, therefore, in the idea that the field strength should depend on the direction in which it is measured from the radiator.

When the antenna has appreciable length, so that the current in every part is not the same at any given instant, the shape of the radiation pattern changes. In this case the pattern is the summation of the fields from *each* elementary doublet of which the antenna may be assumed to consist, strung together in chain fashion. If the antenna is short compared with a half wavelength there is very little change in the pattern, but at a half wavelength the pattern takes the shape shown in cross section in Fig. 2-13. The intensity decreases somewhat more rapidly, as the angle with the wire is made smaller, than in the case of the elementary doublet. This tendency continues as the wire length is increased, the shape of one half of the diagram becoming more and more compressed as the antenna length approaches a full wavelength. (The antenna is assumed to be driven at the center, as in Fig. 2-8B and 2-8C.)

The solid pattern from a half-wave wire is formed, just as in the case of the doublet, by rotating the plane diagram shown in Fig. 2-13 about the wire as an axis. If there is trouble in visualizing the solid pattern, it will help if the diagrams shown in Figs. 2-12 and 2-13 are copied on cardboard, cut out to shape, and then fastened at the center to a piece of stiff wire lying in the direction indicated for the antennas in the drawings. If the wire is rotated rapidly in the fingers the cardboard will appear to form a solid figure that has the same shape as the solid radiation pattern.

The directive diagrams just discussed are actually cross sections of the solid pattern, cut by planes in which the axis of the antenna lies. If the solid pattern is cut by any other plane the diagram will be different. For instance, imagine a plane passing through the center of the wire at right angles to the axis. The cross section of the pattern for either the elementary doublet or the half-wave antenna will simply be a circle in that case. This is shown in Fig. 2-14, where the dot at the center represents the antenna as viewed "end-on"; in other words, the antenna is perpendicular to the page. This means that in any direction in a plane at right angles to the wire, the field intensity is exactly the same at the same distance from the antenna. At right angles to the wire, then, an antenna a half wave or less in length is *nondirectional*. Also, at every point on such a circle the field is in the *same phase*.

Although the solid pattern of an antenna cannot be shown adequately on a flat sheet of paper, the cross-sectional or **plane diagrams** are very useful. Two such diagrams, one in the plane containing the axis of the antenna and one in the plane perpendicular to the axis, can give a great deal of information. After a little practice, and with the exercise of a little imagination, the complete solid pattern can be visualized with fair accuracy from inspection of just the two diagrams. Plane diagrams are plotted on polar-coordinate paper — graph paper with radial lines marking the 360 degrees of a circle, and having a linear scale along the radii for marking amplitudes.

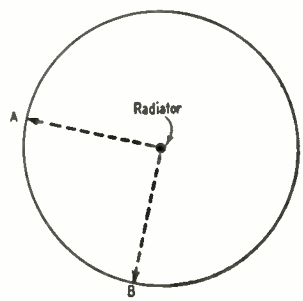


Fig. 2-14 — Directive diagram of a doublet or dipole in the plane perpendicular to the wire axis. The direction of the wire is into or out of the page.

The points on the pattern where the radiation is zero are called **nulls**, and the curved section from one null to the next on the plane diagram, or the corresponding section on the solid pattern, is called a **lobe or ear**.

HARMONIC-ANTENNA PATTERNS

In view of the change in radiation pattern as the length of the antenna is increased from the elementary doublet to the half-wave dipole, it is to be expected that further pattern changes will occur as the antenna is made still longer. We find, as a matter of fact, that the patterns of harmonic antennas differ very considerably from the pattern of the half-wave dipole.

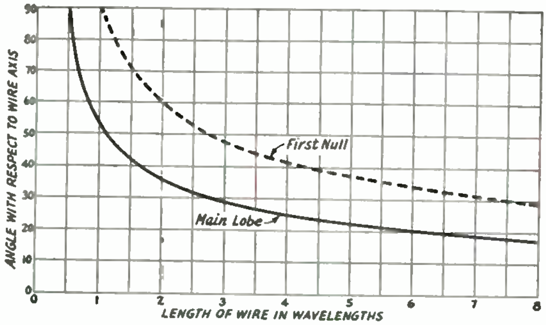


Fig. 2-15 — Angle at which the field intensity from the main lobe of a harmonic antenna is maximum, as a function of the wire length in wavelengths. The curve labeled "First Null" locates the angle at which the intensity of the main lobe decreases to zero. The null marking the other boundary of the main lobe always is at zero degree with the wire axis.

As explained earlier in this chapter, a harmonic antenna consists of a series of half-wave sections with the currents in adjacent sections always flowing in opposite directions. This type of current flow causes the pattern to be split up into a number of lobes. If there is an *even* number of half waves in the harmonic antenna there is always a null in the plane at right angles to the wire; this is because the radiation from one half-wave section cancels the radiation from the next, in that particular direction. If there is an *odd* number of half waves in the antenna, the radiation from all but one of the sections cancels itself out in the plane perpendicular to the wire. The "left-over" section radiates like a half-wave dipole, and so a harmonic antenna with an odd number of half wavelengths does have some radiation at right angles to its axis.

The greater the number of half waves in a harmonic antenna, the larger the number of lobes into which the pattern splits. A feature of all such patterns is the fact that the "main" lobe — the one that gives the largest field strength at a given distance — always is the one that makes the smallest angle with the antenna wire. Furthermore, this angle becomes smaller as the length of the antenna is increased. Fig. 2-15 shows how the angle which the main lobe makes with the axis of the antenna varies with the antenna length in wavelengths. The angle shown by the solid curve is the maximum point of the lobe; that is, the direction in which the field strength is greatest. The broken curve shows the angle at which the first null (the one that occurs at the smallest angle with the wire) appears. There is also a null in the direction of the wire itself (0 degrees) and so the total width of the main lobe is the angle between the wire and the first null. It can be seen from Fig. 2-15 that the width of the lobe decreases as the wire becomes longer. At 1 wavelength, for example, it has a width of 90 degrees, but at 8 wavelengths the width is slightly less than 30 degrees.

Plane diagrams of the radiation patterns of several harmonic wires are shown in Figs. 2-16 to 2-18, inclusive. These are free-space diagrams in the plane containing the wire axis, corresponding to the diagrams for the elementary doublet and half-wave dipole shown in Figs. 2-12 and 2-13. They are based on an infinitely-thin antenna conductor with ideal current distribution, and in a practical antenna system will be modified by the presence of the earth and other effects that will be considered later. However, these few diagrams do show the tendency for the pattern to break up into more and more lobes as the antenna is made longer, and also show that these "secondary" lobes have progressively smaller amplitude as compared with the main lobe.

HOW PATTERNS ARE FORMED

In a multielement antenna system the over-all radiation pattern is determined by the way in which the fields at a distant point from the separate antenna elements combine. With two antenna elements, for example, the field strength at a given point depends on the amplitudes and

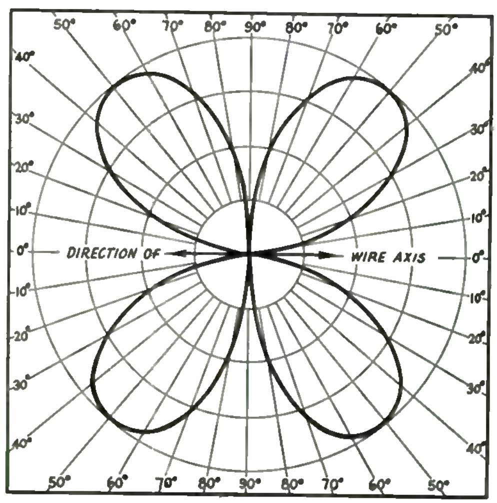


Fig. 2-16 — Free-space directive diagram of a 1-wavelength harmonic antenna in the plane containing the wire axis.

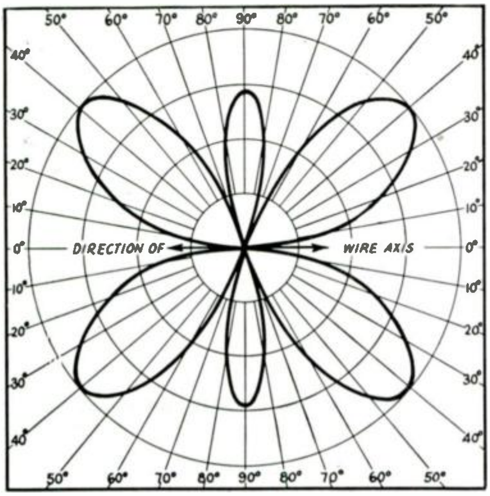


Fig. 2-17 — Free-space directive diagram of a 1½-wavelength harmonic antenna in the plane containing the wire axis.

phase relationship of the fields from each antenna. A requirement in working out a radiation pattern is that the field strength be measured or calculated at a *distant* point — distant enough so that, if the elements carry equal currents, the field strength from each is exactly the same even though the size of the antenna system may be such that one antenna element is a little nearer the measuring point than another. On the other hand, this slight difference in distance, even though it may be only a small fraction of a wavelength in many miles, is very important in determining the *phase* relationships between the fields from the various elements.

The principle on which the radiated fields combine to produce the directive pattern, in the case of multielement antennas, is illustrated in the simple example shown in Fig. 2-19. In this case it is assumed that there are two antenna elements, each having a circular directive pattern. The two elements therefore could be half-wave dipoles oriented perpendicular to the page (which gives the plane pattern shown in Fig. 2-14). The separation between the two elements is assumed to be a half-wavelength, and the currents in them are assumed to be equal. Furthermore, the two currents are in phase; that is, they reach their maximum values in the same polarity at the same instant.

Under these conditions the fields from the two antennas will be in the same phase at any point that is equally distant from both antenna elements. At the instant of time selected for the drawing of Fig. 2-19 the solid circles having the upper antenna at their centers represent, let us say, the location of all points at which the field intensity is maximum and has the direction indicated by the arrowheads. The *distance* between each pair of concentric solid circles, measured along a radius, is equal to one wavelength because, as described earlier in this chapter, it is only at intervals of this distance that the fields are in phase. The broken circle locates the points at which the field intensity is the same as in the case of the solid circles, but is *oppositely* directed. It is therefore 180 degrees out of phase with the field denoted by the solid circles, and the distance between the solid and broken circles is therefore one-half wavelength.

Similarly, the solid circles centered on the lower antenna locate all points at which the field intensity from that antenna is maximum and has the same direction as the solid circles about the upper antenna. In other words, these circles represent points in the same phase as the solid circles around the upper antenna. The broken circle having the lower antenna at its center likewise locates the points of opposite phase.

Considering now the fields from both antennas, it can be seen that along the line *AB* the fields from the two always are exactly in phase, because every point along *AB* is equally distant from both antenna elements. However, along the line *XY* the field from one antenna always is *out* of phase with the other, because every point along *XY* is a half-wavelength nearer one element than the other. It takes one-half cycle longer, therefore, for the field from the more distant element to reach the same point as the field from the nearer antenna, and thus the one field arrives 180 degrees out of phase with the other. Since we have assumed that the points considered are sufficiently distant so that the amplitudes of the fields from the two antennas are the same, the *resultant* field at any point

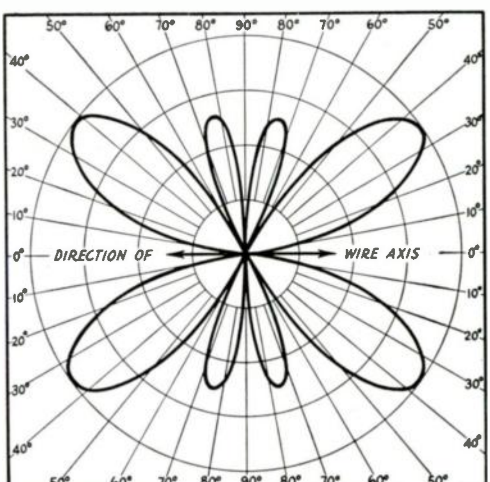


Fig. 2-18 — Free-space directive diagram of a 2-wavelength harmonic antenna in the plane containing the wire axis.

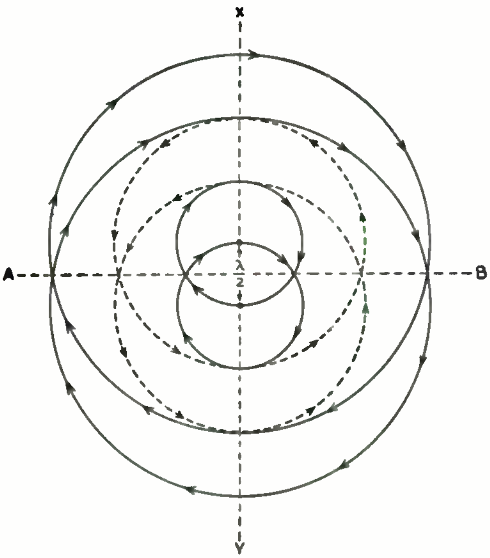


Fig. 2-19 — Interference between waves from two separated radiators causes the resultant directional effects to differ from those of either radiator alone. The two radiators shown here are separated one-half wavelength. The radiation fields of the two cancel along the line XY but, at distances which are large compared with the separation between the radiators, add together along line AB. The resultant field decreases uniformly as the line is swung through intermediate positions from AB to XY.

along XY is zero and the antenna combination shown will have a null in that direction. However, the two fields add together along AB, and the field strength in that direction will be twice the amplitude of the field from either antenna alone.

The drawing of Fig. 2-19 is not quite accurate because it cannot be made large enough. Actually, the two fields along AB do not have exactly the same direction until the distance to the measuring point is large enough, compared with the dimensions of the antenna system, so that the waves become plane. In a drawing of limited size the waves are necessarily represented as circles — that is, as representations of a *spherical* wave. The reader, therefore, should imagine Fig. 2-19 as being so much enlarged that the circles crossing AB are substantially straight lines in the region under discussion.

Pattern Construction

The drawing of Fig. 2-19 does not tell us much about what happens to the field strength at points that do not lie on either AB or XY, although we could make the reasonable guess that the field strength at intermediate points probably would decrease as the point was moved along the arc of a circle farther away from AB and nearer to XY. To construct an actual pattern it is necessary to use a different method. It is simple in principle and can be done with a ruler, protractor and pencil.

In Fig. 2-20 the two antennas, A and B, are assumed to have circular radiation patterns, and to carry equal currents in the same phase. (In other words, the conditions are the same as in Fig. 2-19.) The relative field strength at a distant point P is to be determined. Here again the limitations of the printed page make it necessary to use the imagination, because we assume that P is far enough from A and B so that the lines AP and BP are, for all practical purposes, parallel. When this is so, the distance d, between B and a perpendicular dropped to BP from A, will be equal to the difference in length between the distance from A to P and the distance from B to P. The distance d thus measures the difference in the distances the waves from A and B have to cover to reach P; d is also, therefore, a measure of the difference in the time of arrival or *phase* of the waves at P.

Under the assumed conditions, the relative field strengths easily can be combined graphically. The phase angle in degrees between the two fields at P is equal to

$$\frac{d}{\lambda} \times 360$$

where λ is the wavelength and d is found by constructing a figure similar to that shown in Fig. 2-20 for P in any desired direction. The angle θ is the angle between a line to P and the line drawn between the two antenna elements, and is used simply to identify the direction of P from the antenna system. λ and d must be expressed in the same length units.

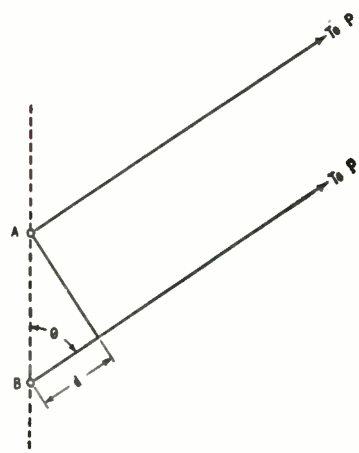


Fig. 2-20 — Graphical construction to determine the relative phase, at a distant point, of waves originating at two antennas, A and B. The phase is determined by the additional distance, d, that the wave from B has to travel to reach the distant point. This distance will vary with the angle that the direction to P makes with the axis of the antenna system.

For example, let us assume that θ is 40 degrees. We then arbitrarily choose a scale such that four inches is equal to one wavelength — a scale large enough for reasonable accuracy but not too large to be unwieldy. Since the two antenna elements are assumed to be a half-wavelength apart, we start the drawing by placing two points two inches apart and connecting them by a line, as shown in Fig. 2-21. Then, using B as a center and employing the protractor, we lay off an angle of

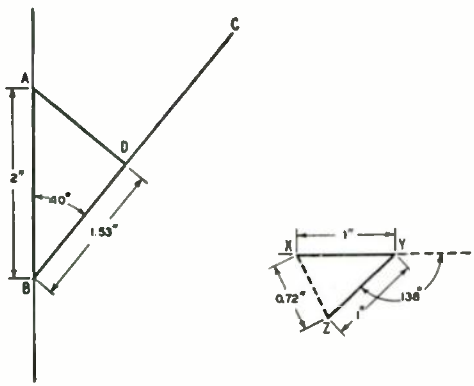


Fig. 2-21 — **Graphical construction in the example discussed in the text.**

40 degrees and draw the line BC. The next step is to drop a perpendicular from A to BC; this may be done with the 90-degree mark on the protractor but the corner of an ordinary sheet of paper will do just about as well. The distance BD is then measured, preferably with a ruler graduated in tenths of inches rather than the more usual eighths. By actual measurement distance BD is found to be 1.53 inches. The phase difference is therefore $d/\lambda \times 360 = 1.53/4 \times 360 = 138$ degrees.

The relative field strength in the direction given by θ (40 degrees in this example) is found by arbitrarily selecting a line length to represent the strength of the field from each antenna, and then combining them "vectorially." One inch is a convenient length to select. XY, Fig. 2-21, is such a line, representing the strength of the field from antenna element A. We then measure off an angle of 138 degrees from XY, using Y as a center, and draw YZ one inch long to represent the strength and phase of the field from antenna element B. The angle is measured off clockwise from XY because the field from B lags behind that from A. The distance from X to Z then represents the relative field strength resulting from the combination of the separate fields from the two antennas, and measurement shows it to be approximately 0.72 inch. In the direction θ, therefore, the field strength is 72% as great as the field from either antenna alone.

By selecting different values for θ and proceeding as above in each case, the complete pattern can be determined. When θ is 90 degrees, the phase difference is zero and YZ and XY are simply end-to-end along the same line. The maximum field strength is therefore twice that of either antenna alone. When θ is zero, YZ lies on top of XY (phase difference 180 degrees) and the distance XZ is therefore zero; in other words, the radiation from B cancels that from A at such an angle.

The patterns of more complex antenna systems can readily be worked out by this method, although more labor is required if the number of elements is increased. But whether or not actual patterns are worked out, an understanding of the method will do much to make it plain why certain combinations of antenna elements result in specific directive patterns.

The illustration above is a very simple case, but it is only a short step to systems in which the antenna elements do not carry equal currents or currents in the same phase. A difference in current amplitude is easily handled by making the lengths of lines XY and YZ proportional to the current in the respective elements; if the current in B is one-half that in A, for example, YZ would be drawn one-half as long as XY. If B's current leads the current in A by 25 degrees, then after the angle determined by the distance d is found the line YZ is simply rotated 25 degrees in the *counterclockwise* direction before measuring the distance XZ. The rotation would be *clockwise* for any line representing a lagging current. The lead or lag of current always has to be referred to the current in *one* element of the system, but any desired element can be chosen as the reference.

DIRECTIVITY AND GAIN

It has been stated that all antennas, even the simplest types, exhibit directive effects in that the intensity of radiation is not the same in all directions from the antenna. Let us assume that a certain amount of power is supplied to an isotropic radiator and that it produces a certain field strength at a given distance from the antenna. Now if an actual antenna is substituted for the hypothetical one and the same amount of power is supplied, the actual antenna will radiate just the same power as was radiated by the isotropic antenna. But in the second case we know that the intensity of radiation in certain directions is zero and in many other directions will be smaller than the field strength produced by the isotropic radiator at the same distance. This power cannot just disappear. Actually, it appears as *increased* field strength, over that produced by the isotropic radiator, in the most favored directions of the actual antenna.

If we call the field strength at a given distance from the actual antenna, in its most favored direction, F_1, and the field strength from an isotropic radiator at the same distance F_2, then the ratio

$$\frac{F_1}{F_2}$$

is called the **gain** of the actual antenna. The gain is thus the ratio of the *voltages* produced at a given point by the actual and hypothetical antennas. More frequently, the gain is expressed as a *power* ratio. Since power varies as the square of the voltage, the power gain is

$$\frac{P_1}{P_2} = \frac{F_1^2}{F_2^2} = \left(\frac{F_1}{F_2}\right)^2$$

Thus if the field produced by an actual antenna is twice as great as the field produced by the isotropic antenna the gain ratio is 2 and the power gain is 2^2, or 4. This means that to produce the same field strength at the same distance, four times as much power would have to be supplied to an isotropic radiator as to the actual antenna under consideration; or, conversely, that one-fourth as much power in the actual antenna would produce the same field strength as the isotropic radiator.

The more the directive diagram is compressed — or, in common terminology, the "sharper" the lobes — the greater the power gain of the antenna. This is a natural consequence of the fact that as power is taken away from a larger and larger portion of the sphere surrounding the radiator it is added to the smaller and smaller volume represented by the lobes. The power is therefore concentrated in some directions at the expense of others. In a general way, the smaller the volume of the solid radiation pattern, compared with the volume of a sphere having the same radius as the length of the largest lobe in the actual pattern, the greater the power gain.

Gain referred to an isotropic radiator is necessarily theoretical; that is, it has to be calculated rather than measured because the isotropic radiator has no existence. In practice, measurements on the antenna being tested usually are compared with measurements made on a half-wave dipole. The latter should be at the same height and have the same polarization as the antenna under test, and the reference field — that from the half-wave dipole comparison antenna — should be measured in the most favored direction of the dipole. The data can be secured either by measuring the field strengths produced at the same distance from both antennas when the same power is supplied to each, or by measuring the power required in each antenna to produce the same field strength at the same distance. The latter method is preferable in amateur work, because it is easier to measure the relative powers supplied to the antennas than it is to measure field strengths accurately.

The Decibel

As a convenience, the power gain of an antenna system is usually expressed in **decibels.** The decibel is an excellent practical unit for measuring power ratios because it is more closely related to the actual effect produced than the power ratio itself. One decibel represents a just-detectable *change* in signal strength, regardless of the actual value of the signal voltage. A 20-decibel (20-db.) increase in signal, for example, represents 20 observable "steps" in increased signal. The power ratio (100 to 1) corresponding to 20 db. would give an entirely exaggerated idea of the improvement in communication to be expected. The number of decibels corresponding to any power ratio is equal to 10 times the common logarithm of the power ratio, or

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

If the *voltage* ratio is given, the number of decibels is equal to 20 times the common logarithm of the ratio. That is,

$$\text{db.} = 20 \log \frac{E_1}{E_2}$$

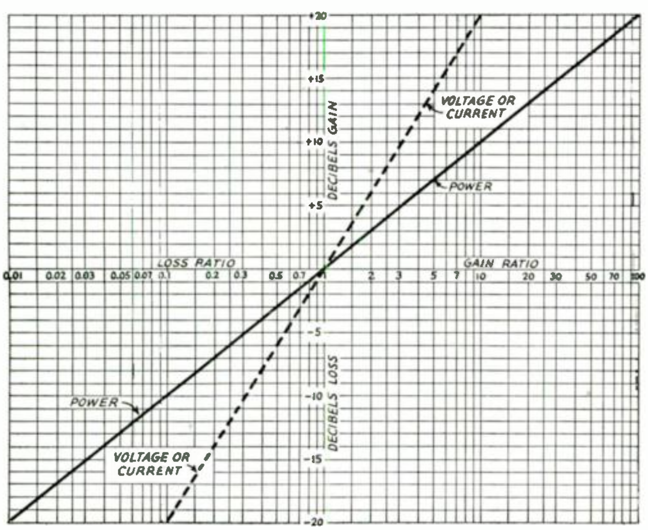


Fig. 2-22 — Chart of decibels *vs.* power or voltage gain or loss. When the voltage curve is used the voltages must be measured across identical impedances. The range of the chart can be extended by adding (or subtracting, if a loss) 10 db. each time the power ratio is multiplied or divided by 10, or 20 db. each time the voltage is multiplied or divided by 10.

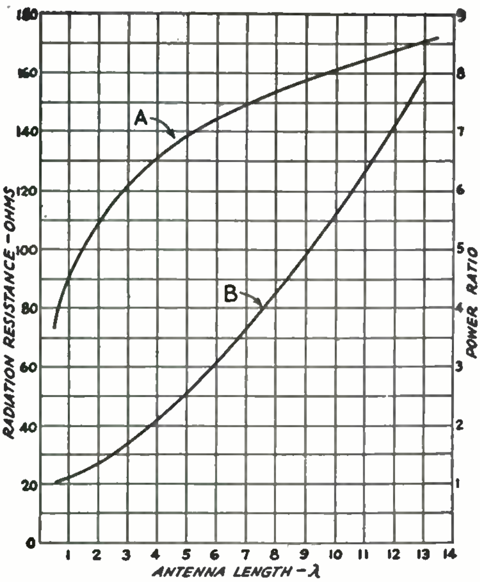


Fig. 2-23 — The variation in radiation resistance and power in the major lobe of long-wire antennas. Curve *A* shows the change in radiation resistance with antenna length, while Curve *B* shows the power in the lobes of maximum radiation for long-wire antennas as a ratio to the maximum of a half-wave antenna.

When a voltage ratio is used both voltages must be measured in the same value of impedance. Unless this is done the decibel figure is meaningless, because it is fundamentally a measure of a *power* ratio.

The chart of Fig. 2-22 shows the number of decibels corresponding to various power and voltage ratios. One advantage of the decibel is that successive power gains expressed in decibels may simply be added together. Thus a gain of 3 db. followed by a gain of 6 db. gives a total gain of 9 db. In ordinary power ratios, the ratios would have to be multiplied together to find the total gain. Furthermore, a *reduction* in power is handled simply by subtracting the requisite number of decibels. Thus reducing the power to ½ is the same as *subtracting* 3 decibels. We might, for example, have a power gain of 4 in one part of a system and a reduction of ½ in another part, so that the total power gain is $4 \times \frac{1}{2} = 2$. In decibels, this would be $6 - 3 = 3$ db. A power reduction or "loss" is simply indicated by putting a negative sign in front of the appropriate number of decibels.

Power Gains of Harmonic Antennas

In splitting off into a series of lobes, the solid radiation pattern of a harmonic antenna is compressed into a smaller volume as compared with the single-lobed pattern of the half-wave dipole. This means that there is a concentration of power in certain directions with a harmonic antenna, particularly in the main lobe. The result is that a harmonic antenna will produce an increase in field strength, in its most favored direction, over a half-wave dipole in *its* most favored direction, when both antennas are supplied with the same amount of power.

The power gain from harmonic operation is small when the antenna is small in terms of wavelengths, but is quite appreciable when the antenna is fairly long. The theoretical power gain of harmonic antennas or "long wires" is shown by curve *B* in Fig. 2-23, using the half-wave dipole as a base. A 1-wavelength or "second-harmonic" antenna has only a slight power gain, but an antenna 9 wavelengths long will show a power increase of 5 times over the dipole. This gain is secured in one direction by eliminating the power radiated in other directions; thus the longer the wire the more directive the antenna becomes.

Curve *A* in Fig. 2-23 shows how the radiation resistance, as measured at a current loop, varies with the length of a harmonic antenna.

Ground Effects

● REFLECTION FROM THE GROUND

The performance of an antenna, particularly with respect to its directive properties, is considerably modified by the presence of the earth underneath it. The earth acts like a huge reflector for those waves that are radiated from the antenna at angles lower than the horizontal. These downcoming waves strike the surface and are reflected by a process very similar to that by which light waves are reflected from a mirror. As in the case of light waves, the angle of reflection is the same as the angle of incidence, so that a wave striking the surface at an angle of, for instance, 15 degrees, is reflected upward from the surface at the same angle.

The reflected waves combine with the direct waves (those radiated at angles above the horizontal) in various ways, depending upon the orientation of the antenna with respect to earth, the height of the antenna, its length, and the character of the ground. At some vertical angles above the horizontal the direct and reflected waves may be exactly in phase — that is, the maximum field strengths of both waves are reached at the same time at the same spot, and the directions of the fields are the same. In such a case the resultant field strength is simply equal to the sum of the two. At other vertical angles the two waves may be completely out of phase — that is, the fields are maximum at the same instant and the directions are opposite, at the same spot. The resultant field strength in that case is the *difference* between the two. At still other angles the resultant field will have intermediate values. Thus the effect of the ground is to increase the intensity of radiation at some vertical angles and to decrease it at others.

The effect of reflection from the ground is shown graphically in Fig. 2-24. At a sufficiently large distance, two rays converging at the distant point can be considered to be parallel. However, the reflected ray travels a greater distance in reaching P than the direct ray does, and this difference in path length accounts for the effect described in the preceding paragraph. If the ground were a perfect conductor for electric currents reflection would take place without a change in phase when the waves are vertically polarized. Under similar conditions there would be a complete reversal (180 degrees) of phase when a horizontally-polarized wave is reflected. The actual earth is of course not a perfect conductor, but is usually assumed to be one for purposes of calculating the vertical pattern of an antenna. The error is small except at very low vertical angles.

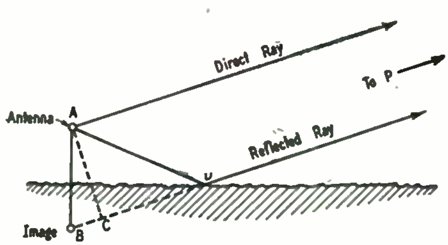


Fig. 2-24 — At any distant point, P, the field strength will be the resultant of two rays, one direct from the antenna, the other reflected from the ground. The reflected ray travels farther than the direct ray by the distance BC, where the reflected ray is considered to originate at the "image" antenna.

As an example, when the path of the reflected ray is exactly a half wave longer than the path of the direct ray, the two waves will arrive out of phase if the polarization is vertical. This corresponds to the condition illustrated in Fig. 2-19 along the line XY. However, if the path of the reflected ray is just a wavelength longer than that of the direct ray, the two rays arrive in phase.

Image Antennas

It is often convenient to use the concept of an image antenna to show the effect of reflection. As Fig. 2-24 shows, the reflected ray has the same path length (AD equals BD) that it would if it originated at a second antenna, of the same characteristics as the real antenna, but situated below the ground just as far as the actual antenna is above it. Like an image in a mirror, this image antenna is "in reverse," as shown in Fig. 2-25.

If the real antenna is horizontal, and is instantaneously charged so that one end is positive and the other negative, then the image antenna, also horizontal, is oppositely poled; the end under the positively-charged end of the real antenna is negative, and vice versa. Likewise, if the lower end of a half-wave vertical antenna is instantaneously positive, the end of the vertical image antenna nearest the surface is negative. Now if we look at the antenna and its image from a remote point on the surface of the ground, it will be obvious that the currents in the horizontal antenna and its image are flowing in opposite directions, or are 180 degrees *out* of phase, but the currents in the vertical antenna and its image are flowing in the *same* direction, or are *in* phase. The effect of ground reflection, or the image antenna, is therefore different for horizontal and vertical half-wave antennas. The physical reason for this difference is the fact that vertically-polarized waves are reflected from a perfectly-conducting earth with no change in phase, but that horizontally-polarized waves have their phase shifted by 180 degrees on reflection.

Reflection Factor

The effect of reflection can be expressed as a factor which, when multiplied by the free-space figure for relative intensity of radiation at a given vertical angle from an antenna, gives the resultant relative radiation intensity at that same angle. The limiting conditions are those represented by the direct ray and reflected ray being exactly in phase and exactly out of phase when both, assuming there are no ground losses, have exactly equal amplitudes. Thus the resultant field strength at a distant point may be either twice the field strength from the antenna alone, or zero.

The way in which the reflection factor (based on perfectly-conducting ground) varies with antenna height is shown in the series of graphs, Figs. 2-26 to 2-43. Figs. 2-26 to 2-37 apply to horizontal antennas of any length, and to vertical antennas an *even* number of half waves long. Figs. 2-38 to 2-43 apply to vertical antennas an *odd* number of half waves long. Comparing the two sets, it is seen that the positions of nulls

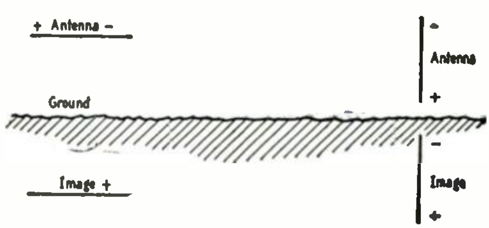


Fig. 2-25 — Horizontal and vertical half-wave antennas and their images.

(multiplying factor zero) and maxima (multiplying factor 2) are interchanged for the two sets of conditions.

It must be remembered that these graphs are not plots of vertical patterns of antennas, but represent simply multiplying factors representing the result of reflection from the ground. With the distinction between vertical and horizontal antennas noted, the graphs apply equally well to *all* antennas. Also, it should be understood that they apply at *vertical* angles only. The ground, if of uniform characteristics, makes no distinction between geographical directions — that is, horizontal angles from the antenna — in reflecting waves.

Fig. 2-44 shows the angles at which nulls and maxima occur as a function of the height of the antenna. This chart gives a rough idea of the ground-reflection pattern for heights intermediate to those shown in detail in Figs. 2-26 to 2-43. It also facilitates picking the right height for any desired angle of radiation.

● GROUND CHARACTERISTICS

As already explained, the charts of Figs. 2-26 to 2-43 are based on the assumption that the earth is a perfect conductor. The actual ground is far from being "perfect" as a conductor of electricity. Actually, its behavior depends considerably on the transmitted frequency. At low frequencies — through the standard broadcast band, for example — most types of ground do act very much like a good conductor. At these frequencies the waves can penetrate for quite a distance and thus find a large cross section in which to cause current flow along their paths. The resistance of even a moderately-good conductor will be low if its cross section is large enough. The ground acts as a fairly good conductor even at frequencies as high as the 3.5-Mc.

Factors by which the free-space radiation pattern of a horizontal antenna should be multiplied to include the effect of reflection from perfectly-conducting ground. These factors affect only the *vertical* angle of radiation (wave angle).

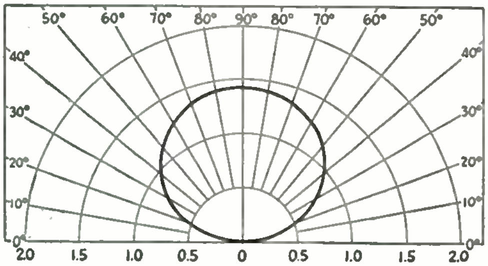


Fig. 2-26 — Horizontal antennas 1/8 wavelength high.

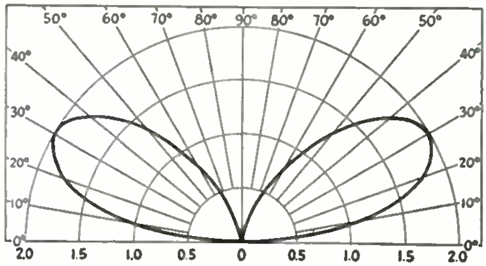


Fig. 2-29 — Horizontal antennas 1/2 wavelength high.

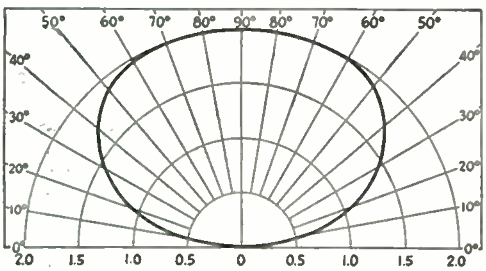


Fig. 2-27 — Horizontal antennas 1/4 wavelength high.

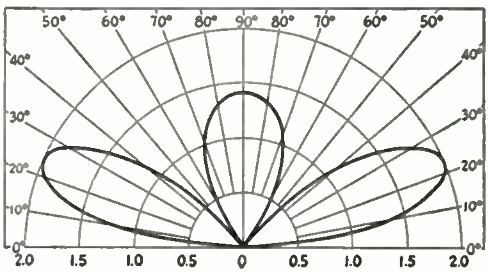


Fig. 2-30 — Horizontal antennas 5/8 wavelength high.

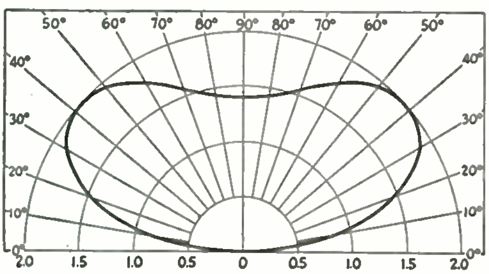


Fig. 2-28 — Horizontal antennas 3/8 wavelength high.

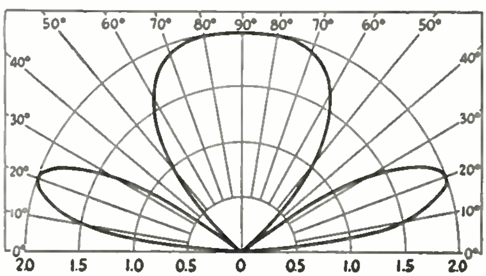


Fig. 2-31 — Horizontal antennas 3/4 wavelength high.

band, and so the charts give a rather good approximation of the effect of the ground at this frequency.

In the higher-frequency region the penetration decreases and the ground may even take on the characteristics of a lossy dielectric rather than a good conductor. The chief effect of this change is to absorb most of the energy radiated at the very low angles, in the frequency region from about 7 to 21 Mc. In general, the reflection factor will be lower than given by the charts at angles of less than about 10 degrees, and it is generally considered that the radiation below about 3 degrees is very small compared with the radiation at higher angles. This applies to both vertical and horizontal antennas, so that the "zero-angle" reflection factor with a vertical half-wave antenna, theoretically 2 as shown by the charts, actually is a small fraction. Thus the apparent advantage of the vertical antenna for very low-angle radiation is not realized in practice in this frequency range.

The "effective reflecting plane" of the ground — that is, the surface from which the reflection is considered to take place at the heights given in the charts — seldom coincides with the actual surface of the ground. Usually it will be found that this plane appears to be a few feet below the surface; in other words, the height of the antenna taken for purposes of estimating reflection is a few feet more than the actual height of the antenna. A great deal depends upon the character of the ground, and in some cases the reflecting plane may be "buried" a surprising distance. Thus in some instances the charts will not give an accurate indication of the effect of reflection. On the average, however, they will give a reasonably satisfactory representation of reflection effects, with the qualifications with respect to high frequencies and low angles mentioned above.

Factors by which the free-space radiation pattern of a horizontal antenna should be multiplied to include the effect of reflection from perfectly-conducting ground. These factors affect only the *vertical* angle of radiation (wave angle).

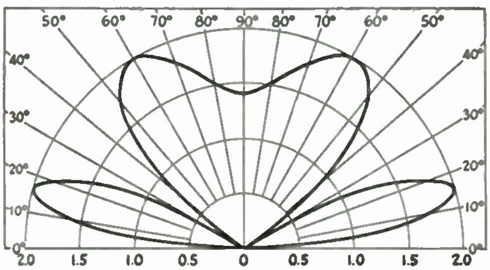


Fig. 2-32 — Horizontal antennas ⅞ wavelength high.

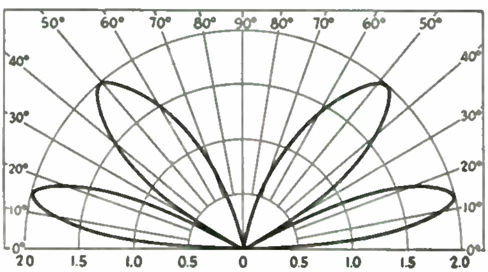


Fig. 2-33 — Horizontal antennas 1 wavelength high.

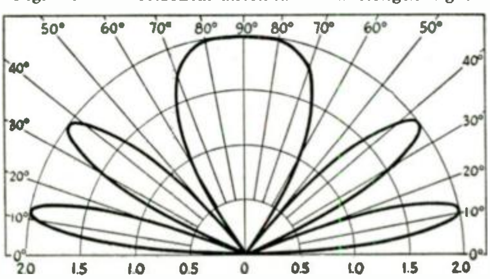


Fig. 2-34 — Horizontal antennas 1¼ wavelengths high.

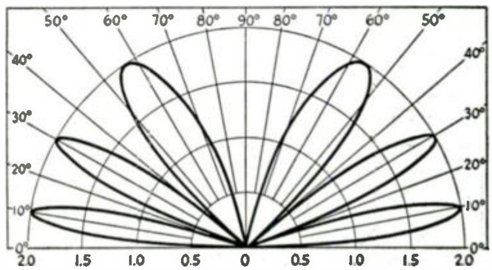


Fig. 2-35 — Horizontal antennas 1½ wavelengths high.

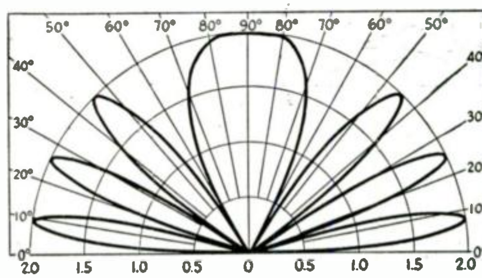


Fig. 2-36 — Horizontal antennas 1¾ wavelengths high.

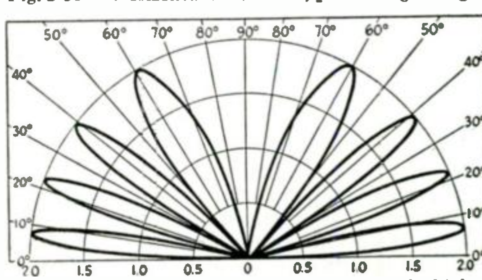


Fig. 2-37 — Horizontal antennas 2 wavelengths high.

Factors by which the free-space radiation pattern of a half-wave vertical antenna should be multiplied to include the effect of reflection from perfectly-conducting ground. These factors affect only the *vertical* angle of radiation (wave angle).

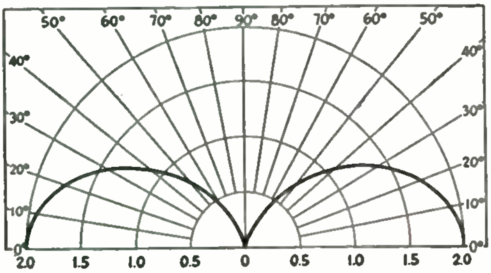


Fig. 2-38 — Vertical dipole antenna with center ¼ wavelength high.

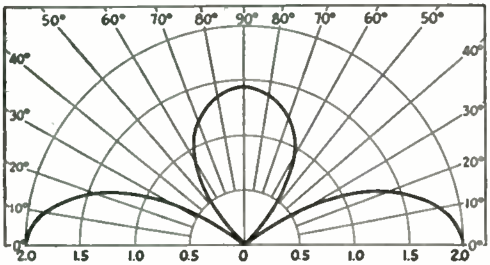


Fig. 2-39 — Vertical dipole antenna with center ⅜ wavelength high.

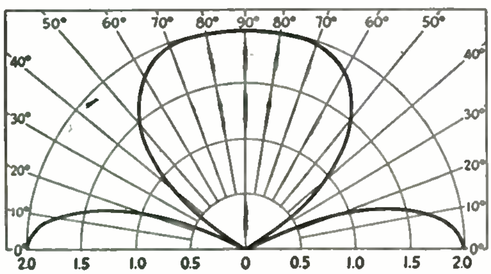


Fig. 2-40 — Vertical dipole antenna with center ½ wavelength high.

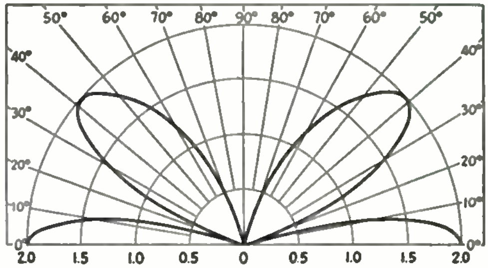


Fig. 2-41 — Vertical dipole antenna with center ¾ wavelength high.

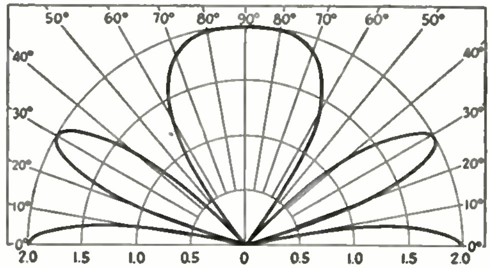


Fig. 2-42 — Vertical dipole antenna with center 1 wavelength high.

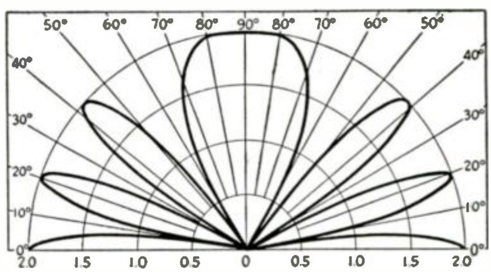


Fig. 2-43 — Vertical dipole antenna with center 1½ wavelengths high.

In the v.h.f. and u.h.f. region (starting in the vicinity of the 28-Mc. band) a different situation exists. At these frequencies little, if any, use is made of the part of the wave that travels in contact with the ground. The antennas, both transmitting and receiving, usually are rather high in terms of wavelength. The wave that is actually used — at least for line-of-sight communication — is in most cases several wavelengths above the surface of the ground. At such a height there is no consequential loss of energy; the direct ray travels from the transmitter to the receiver with only the normal attenuation caused by spreading, as explained in Chapter One. The loss of energy in the reflected ray is beneficial rather than otherwise, as also explained in that chapter. The net result is that radiation at very low angles is quite practicable in this frequency region. Also, there is little practical difference between horizontal and vertical polarization.

GROUND REFLECTION AND RADIATION RESISTANCE

Waves radiated from the antenna directly downward reflect vertically from the ground and, in passing the antenna on their upward journey, induce a current in it. The magnitude and phase of this induced current depends upon the height of the antenna above the reflecting surface.

The total current in the antenna thus consists of two components. The amplitude of the first is determined by the power supplied by the transmitter and the free-space radiation resistance of

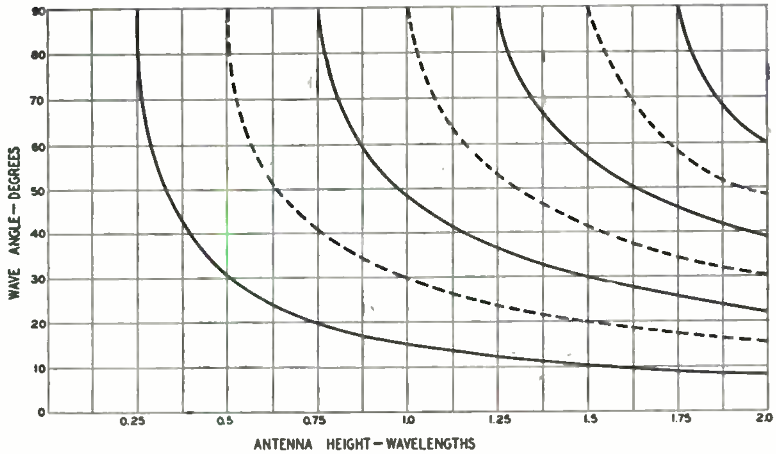


Fig. 2-44 — Angles at which nulls and maxima (factor = 2) in the ground-reflection factor appear for antenna heights up to two wavelengths. The solid lines are maxima, dashed lines nulls, for all horizontal antennas and for vertical antennas having a length equal to an even multiple of one-half wavelength. For vertical antennas an odd number of half waves long, the dashed lines are maxima and the solid lines nulls. For example, if it is desired to have the ground reflection give maximum reinforcement of the direct ray at a 20-degree wave angle (angle of radiation) the antenna height should be 0.75 wavelength. The same height will give a null at 42 degrees and a second maximum at 90 degrees.

the antenna. The second component is induced in the antenna by the wave reflected from the ground. The second component, while considerably smaller than the first at most useful antenna heights, is by no means inappreciable. At some heights the two components will be more or less in phase, so the total current is larger than would be expected from the free-space radiation resistance. At other heights the two components are out of phase, and at such heights the total current is the difference between the two components.

Thus merely changing the height of the antenna above ground will change the amount of current flow, assuming that the power input to the antenna is held constant. A higher current at the same value of power means that the effective resistance of the antenna is lower, and vice versa. In other words, the radiation resistance of the antenna is affected by the height of the antenna above ground. Fig. 2-45 shows the way in which the radiation resistance of a horizontal half-wave antenna varies with height, in terms of wavelengths, over perfectly-conducting ground. Over actual ground the variations will be somewhat lower, but the chart shows the approximate magnitude of the change to be expected. The antenna on which this chart is based is assumed to have an infinitely-thin conductor, and thus has a somewhat higher free-space value of radiation resistance (73 ohms) than an antenna constructed of wire or tubing. (See Fig. 2-7.)

Ground Screens

The effect of a perfectly-conducting ground can be simulated, in the vicinity of the antenna, by installing a metal screen or mesh such as chicken wire underneath the antenna near or on the surface of the ground. Such a screen often will improve the performance of the antenna by reducing losses in the ground near the antenna. It should preferably extend at least a half wavelength in every direction from the antenna, although good results have been reported with

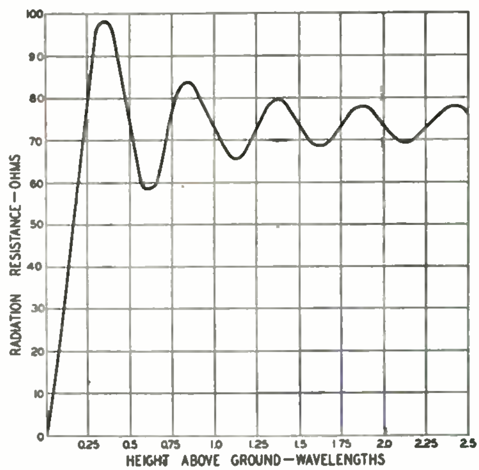


Fig. 2-45 — Variation in radiation resistance of a horizontal half-wave antenna with height above perfectly-conducting ground.

ground screens having 25 per cent smaller dimensions.

Besides reducing losses, a ground screen rather effectively establishes the height of the antenna insofar as the radiation resistance is concerned. For this purpose, the height will be the actual height of the antenna above the screen. However, reflection from a screen of reasonable dimensions takes place only at the higher angles. The presence of the ground screen therefore will not appreciably modify the effect of the actual ground at the lower angles, because the low-angle waves are reflected at considerable distances from the antenna.

Directive Diagrams and the Wave Angle

In the discussion of radiation patterns or directive diagrams of antennas it was brought out that such patterns always are three-dimensional affairs, but that it is difficult to show, on a plane sheet of paper, more than a cross section of the solid pattern at a time. The cross sections usually selected are those cut by a plane that contains the wire axis, and those cut by a plane perpendicular to the wire axis.

If the antenna is horizontal, the former pattern (the cross section cut by a plane containing the wire) represents the radiation pattern of the antenna when the wave leaves the antenna (or arrives at it) at zero angle of elevation above the earth. The angle of elevation — the "vertical" angle referred to in the discussion of ground effects — is usually called the **wave angle.** In the case of the vertical antenna the radiation pattern at zero wave angle is given by the cross section cut by the plane perpendicular to the wire axis. (In the latter case it must be remembered that the antenna itself is assumed to be merely a point at the center of the pattern, so the plane must pass through this point.)

Now with two exceptions — surface waves at low frequencies and space waves at v.h.f. and higher — the wave angle used for communication is not zero. In ionosphere transmission waves sent *directly upward* can be reflected back to earth, if the frequency is low enough; on the other hand, as pointed out earlier in this chapter, in most of the frequency range useful for ionospheric communication waves leaving at an angle of less than about 3 degrees are largely absorbed by ground losses. What we are interested in at these frequencies, then, is the directive pattern of the antenna at a wave angle that is of value in communication.

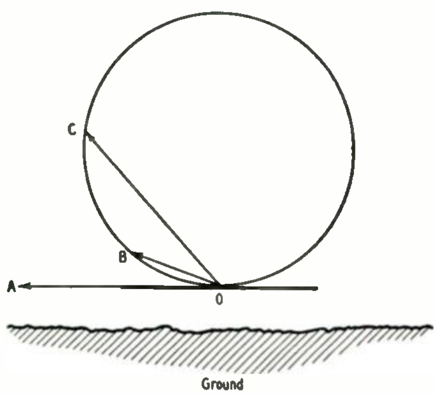


Fig. 2-46 — The effective directive pattern of the antenna depends upon the angle of radiation considered. As shown by the arrows, the field strength in a given compass direction will be quite different at different vertical angles.

● EFFECTIVE DIRECTIVE DIAGRAMS

The directive diagram for a wave angle of zero elevation (purely horizontal radiation) does not give an accurate indication of the directive properties of a horizontal antenna at wave angles above zero. For example, consider the half-wave dipole pattern in Fig. 2-13. It shows that there is no radiation directly in line with the antenna itself, and this is true at zero wave angle. However, if the antenna is horizontal and some wave angle other than zero is considered, it is not true at all.

The reason why will become clear on inspection of Fig. 2-46, which shows a horizontal half-wave antenna with a cross section of its free-space radiation pattern, cut by a plane that is vertical with respect to the earth and which contains the axis of the antenna conductor. (For the moment, reflections from the ground are neglected.) The lines *OA*, *OB* and *OC* all point in the same *geographical* direction (the direction in which the wire itself points), but make different angles, in the vertical plane, with the antenna. In other words, they correspond to different wave angles or angles of radiation, with all three rays aimed along the same line on the earth's surface. So far as compass directions are concerned, *all three waves are leaving the end of the antenna.*

The purely horizontal wave *OA* has zero amplitude, but at a somewhat higher angle corresponding to the line *OB* the field strength is appreciable. At a still higher angle corresponding to the line *OC* the field strength is still greater. In this particular pattern, the higher the wave angle the greater the field strength in the same compass direction. It should be obvious that it is necessary, in plotting a directive diagram that purports to show the behavior of the antenna in different compass directions, to specify the angle of radiation for which the diagram applies. When the antenna is horizontal the shape of the diagram will be altered considerably as the wave angle is changed.

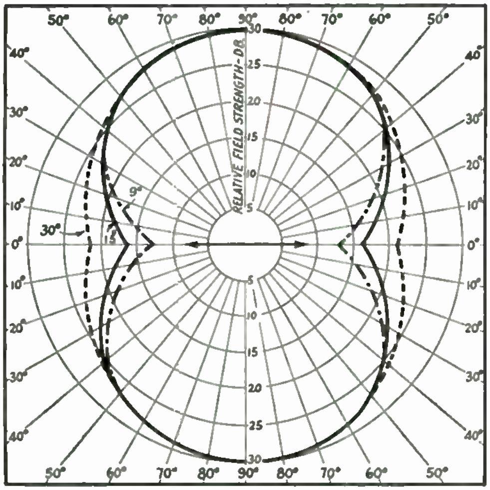


Fig. 2-47 — Directive patterns of a horizontal half-wave antenna at three radiation angles, 9, 15 (solid line) and 30 degrees. The direction of the antenna itself is shown by the arrow. These patterns are plotted to a 30-db. scale, which is about proportional to signal strength as determined by ear. If 30 db. represents an S9 signal, 0 on the scale will be about S1. All three patterns are plotted to the same maximum, but the actual amplitudes at the various angles will depend upon the antenna height, as described in the text. The patterns shown here show only the *shape* of the directive diagram as the angle is varied.

As described in Chapter One, the wave angles that are useful depend on two things — the distance over which communication is to be carried on, and the height of the ionosphere layer that does the reflecting. Whether the *E* or F_2 layer (or a combination of the two) will be used depends on the operating frequency, the time of day, season, and the sunspot cycle. The same half-wave antenna, operating on the same frequency, may be almost nondirectional for distances of a few hundred miles but will give substantially better results broadside than off its ends at distances of the order of 1000–1200 miles during the day when transmission is by the *E* layer. In the evening, when the *F* layer takes over, the directivity may be fairly well marked at long distances and not at all pronounced at 1000 miles or less. From this it might seem that it would be impossible to predict the directivity of an antenna without all sorts of qualifications. However, it is possible to get a very good idea of the directivity by choosing a few angles that, on the average, are representative for different types of work. With patterns for such angles available it is fairly simple to interpolate for intermediate angles. Combined with some knowledge of the behavior of the ionosphere, a fairly good estimate of the directive characteristics of a particular antenna can be made for the particular time of day and distance of interest.

In the directive patterns given in Figs. 2-47 to 2-53, inclusive, the wave angles considered are 9, 15 and 30 degrees. These represent, respectively, the median values of a range of angles that have been found to be effective for communication at 28, 14 and 7 Mc. Because of the variable nature of ionosphere propagation the patterns should not be considered to be more than general guides to the sort of directivity to be expected.

In the directive patterns of Figs. 2-47 to 2-53 the relative field strength has been plotted in decibels. This makes the patterns more representative of the effect produced than is the case when the relative intensity is plotted in either voltage or power. Since one "S" point on the signal-strength scale is roughly 5 or 6 db., it is easy to get an approximate idea of the operation of the antenna. For example, off the ends of a half-wave antenna the signal can be expected to be "down" between 2 and 3 "S" points compared with its strength at right angles or broadside to the antenna, at a wave angle of 15 degrees. This would be fairly representative of its performance on 14 Mc. at distances of 500 miles or more. With a wave angle of 30 degrees the signal off the ends would be down only 1 to 2 "S" points, while at an angle of 9 degrees it would be down 3 to 4 "S" points. Since high wave angles become less useful as the frequency is increased, this illustrates the importance of running the antenna wire in the proper direction at the higher frequencies, if best results are wanted in a particular direction.

Height Above Ground

The *shapes* of the directive patterns given in Figs. 2-47 to 2-53 are not affected by the height of the antenna above the ground. However, the *amplitude* relationships between the patterns of

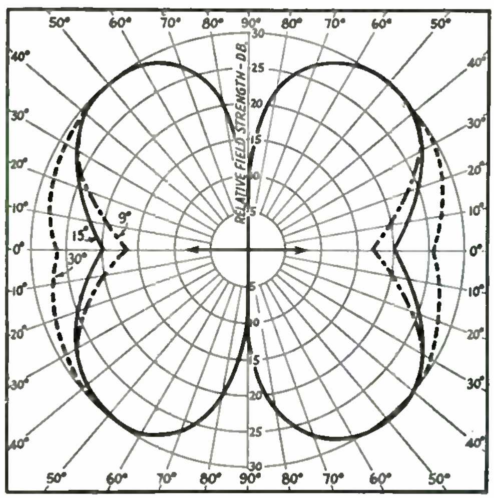


Fig. 2-48 — The horizontal patterns for a one-wave-length antenna at vertical angles of 9, 15 and 30 degrees.

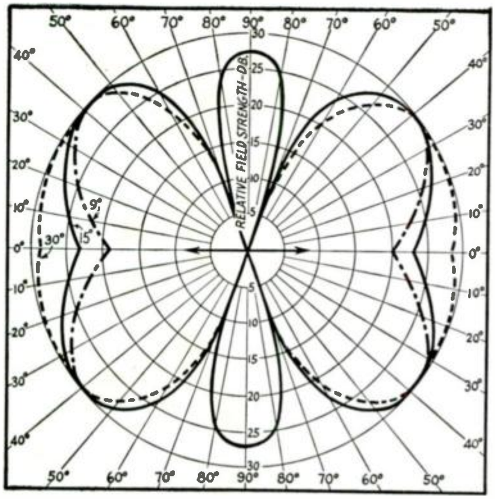


Fig. 2-49 — The horizontal patterns for a 1½-wavelength antenna at vertical angles of 9, 15 and 30 degrees.

a given antenna for various wave angles are modified by the height. In the figures as given, the scale is such that the same field intensity is assumed in the direction of maximum radiation, regardless of the wave angle. To make best use of the patterns the effect of the ground-reflection factor should be included.

Take the horizontal half-wave antenna shown in Fig. 2-47 as an example, and assume that the antenna is a half wavelength above perfectly-conducting ground. The graph of the ground-reflection factor for this height is given in Fig. 2-29. For angles of 9, 15 and 30 degrees the values of the factor as read from the curve are 1.0, 1.5 and 2.0, respectively. These factors are applied to field strength. For convenience, take the 9-degree angle as a reference. Then at a wave angle of 15 degrees the field strength will be 1.5 times the field strength at 9 degrees, *in any compass direction*, and at a wave angle of 30 degrees will be 2.0 times the field strength at 9 degrees, in any compass direction. Since the factors apply to field strength (voltage) the ratios just obtained may be converted to decibels by using the voltage curve in Fig. 2-22. A voltage ratio of 1.5 corresponds to 3.5 db., and a ratio of 2.0 corresponds to 6 db. Hence at an angle of 15 degrees the radiation in any direction is 3.5 db. above the radiation at 9 degrees, and at a wave angle of 30 degrees it is 6 db. above. To put it another way, at a wave angle of 30 degrees the antenna is about an "S" point better than it is at 9 degrees. There is about a half "S"- point difference between 9 and 15 degrees, and between 15 and 30 degrees. If we wanted, we could add 3.5 db. to every point on the 15-degree graph in Fig. 2-47, and 6 db. to every point on the 30-degree graph, and thus show graphically the comparison in amplitude as well as shape of the directive pattern at the three angles. This has been done in the graph of Fig. 2-54. However, it is generally unnecessary to take the trouble to draw separate graphs because it is so easy to add or subtract the requisite number of decibels as based on the appropriate ground-reflection factor.

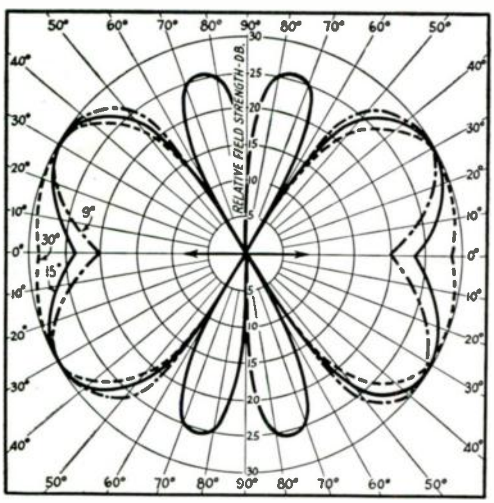


Fig. 2-50 — The horizontal patterns for a 2-wavelength antenna at vertical angles of 9, 15 and 30 degrees.

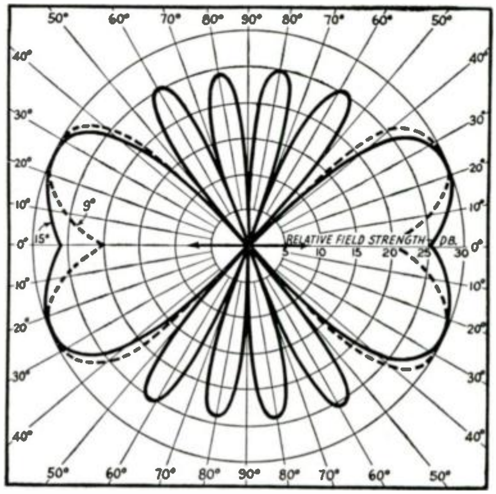


Fig. 2-51 — The horizontal patterns for a 3-wavelength antenna at vertical angles of 9 and 15 degrees.

It should be emphasized again that the patterns are based on idealized conditions not realized over actual ground. Nevertheless, they are useful in indicating about what *order* of effect to expect.

Using the Patterns

Directive patterns can be of considerable help in solving practical problems in the choice and

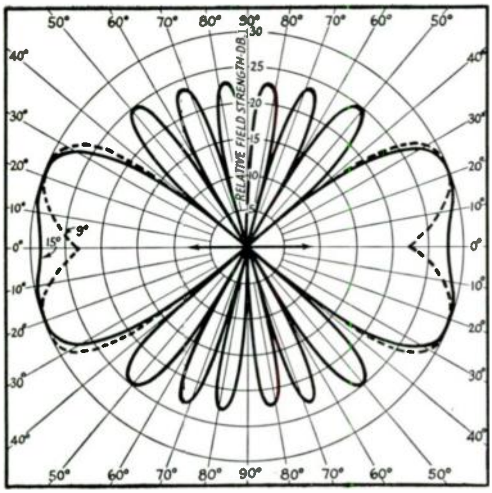


Fig. 2-52 — The horizontal patterns for a 4-wavelength antenna at vertical angles of 9 and 15 degrees.

location of antennas, particularly in cases where a simple type of antenna (such as the antennas discussed in this chapter) has to be used. While it is not to be expected that antenna performance will conform to the theoretical with the same exactness that you would expect Ohm's Law calculations to work out, the results averaged over a period of time will be sufficiently close to the predictions to make a little preliminary estimating worth while.

Here is one example: Suppose that a clear space of about 70 feet is available between two supports that will hold the antenna about 35 feet above ground. The operating frequency is to be 28 Mc. and the positions of the supports is such that the antenna will run west of north by 10 degrees. The principal direction of transmission is to be 35 degrees east of north, but there is another area in the general direction 15 degrees south of west that it is also hoped to cover as well as possible. The situation is shown in Fig. 2-55 (in this figure the last direction is shown with reference to the north-south line).

In the available space, it is possible to erect antennas ½, 1, 1½ or 2 wavelengths long. Since the supports are at fixed height, the ground-reflection factor will be the same for all the possible antennas and so may be left out of the estimates. The principal direction is 45 degrees off the line of the antenna and the secondary direction is 85 degrees off. For simple antennas such as these the directive patterns are symmetrical about the wire axis and so we do not have to worry about whether the angles lie east or west of the antenna.

Since the frequency is 28 Mc. the 9-degree patterns should be used. From Fig. 2-47 we see that the relative amplitudes at 45 and 85 degrees are 27 and 30 db., respectively, for the half-wave antenna. From Fig. 2-48 the corresponding amplitudes are 30 and 17 db.; from Fig. 2-49 the amplitudes are 28 and 27 db.; and from Fig. 2-50 the amplitudes are 28 and 20 db. To these amplitudes we should add the gains realized by harmonic operation as given in Fig. 2-23. These are, for the ½-, 1-, 1½- and 2-wavelength antennas, respectively, 1, 1.1, 1.2 and 1.3 in power. Converted to decibels by using Fig. 2-22 they are 0, 0.5, 0.8 and 1 db. respectively. They are small enough to be less important than the probable error in reading the charts, but will be included for the sake of completeness. Arranging the information in table form gives

	Antenna Length in Wavelengths			
	0.5	*1*	*1.5*	*2*
Relative intensity at 45 degrees, db.	27	30	28	28
Gain from harmonic operation, db.	0	0.5	0.8	1
Total	*27*	*30.5*	*28.8*	*29*
Relative intensity at 85 degrees, db.	30	17	27	20
Gain from harmonic operation, db.	0	0.5	0.8	1
Total	*30*	*17.5*	*27.8*	*21*

It is seen that either a 1-wavelength or 1½-wavelength antenna will give the best results in the principal direction, but that neither is as good as a half-wave antenna in the secondary direction. In a case such as this, the best all-around results would be obtained by using *two* antennas, since there is room to string them end to end. A good combination, for example, would be a 1-wavelength and ½-wavelength antenna, arranged with a little space between the ends so the cou-

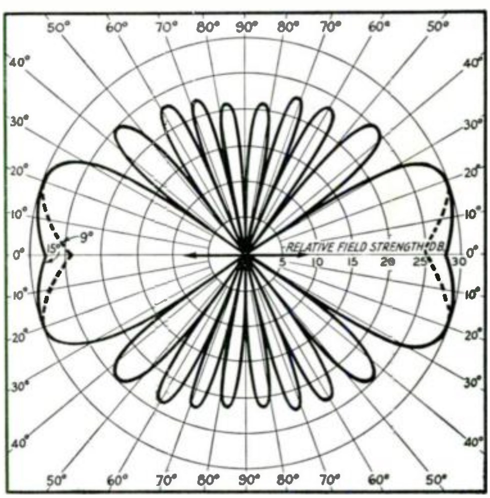


Fig. 2-53 — The horizontal patterns for a 5-wavelength antenna at vertical angles of 9 and 15 degrees.

pling between the pair of antennas is substantially reduced.

Another example: Space is available to erect a 4-wavelength antenna at a height of 30 feet, for operation on 28 Mc., and it is possible to orient the antenna so that its major lobe will point in the direction of transmission desired. Alternatively, a self-supporting half-wave dipole could be erected at a height of 45 feet and oriented so that its maximum radiation would be in the desired direction. Which antenna is likely to be the better one?

From Fig. 2-23, the power gain of a 4-wavelength antenna over a dipole is 2.1, and from Fig. 2-22 this power ratio corresponds to a gain of 3 db. At 28 Mc. the length of a half wavelength is 492/28 = 17.6 feet, so one wavelength is 35 feet, near enough. At 30 feet the height in wavelengths is 30/35 = 0.86λ, and at 45 feet it is 45/35 = 1.3λ. (It is not necessary to carry out the calculations to more than two significant figures because the height above the effective ground plane is not known. As explained earlier, the "effective" height would tend to be higher than the actual height.) A height of 0.85λ is near enough to ⅞λ to permit us to use Fig. 2-32, in which the ground-reflection factor is shown as 1.5 at a wave angle of 9 degrees. From Fig. 2-22 the corresponding figure is 3.5 db. A height of 1.3λ is slightly over 1¼λ, and inspecting Figs. 2-34 and 2-35 shows that the reflection factor therefore will lie in the vicinity of 1.9, or about 5.5 db. Putting these figures into table form, we have

	Dipole	*4λ Antenna*
Relative intensity of main lobe, db.	30	30
Gain from harmonic operation, db.	0	3
Ground-reflection factor, db.	5.5	3.5
Total	*35.5*	*36.5*

The difference, 1 db., is in favor of the 4λ antenna, but for all practical purposes the two antennas could be considered equally good in the desired direction; the additional height of the dipole is just about sufficient to overcome the gain of the harmonic antenna. The choice therefore could be based purely on other considerations such as convenience in erection, the fact that the dipole antenna will be effective over a wider horizontal angle than the harmonic antenna, and so on.

Once again it must be emphasized that calculations such as these should not be taken *too* literally. Too many factors, particularly the behavior of the ground, are unknown. The calculations are useful principally as a guide to determining the type of antenna that, in all probability, will best meet the required working conditions.

RADIATION RESISTANCE AND GAIN

The field strength produced at a distant point by a given antenna system is directly proportional to the current flowing in the antenna. In turn, the amount of current that will flow, when a fixed amount of power is applied, will depend on the radiation resistance. Since power is equal to current squared multiplied by resistance, the current for fixed power input is inversely proportional to the square root of the radiation resistance. Lowering the radiation resistance will increase the field strength and raising the radiation resistance will decrease it.

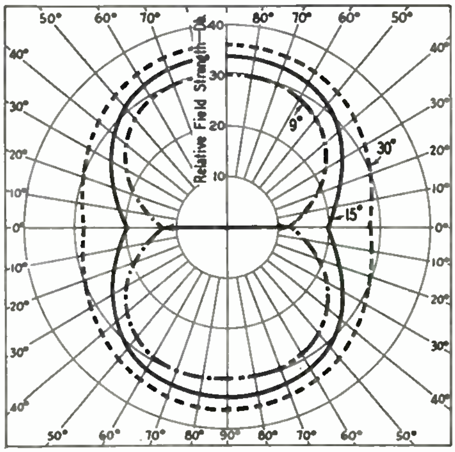


Fig. 2-54 — These diagrams compare the amplitudes of radiation at wave angles of 9, 15 and 30 degrees from a horizontal half-wave antenna when the height is ½ wavelength.

The radiation resistance can be lowered by distorting the radiation pattern in such a way that the directivity of the antenna system is increased. This principle is used in certain types of directive systems described in detail in a later chapter.

The shape of the directional pattern in the vertical plane is, as previously described, modified by the height of the antenna above ground. The effect of height on radiation resistance has been shown in Fig. 2-45 for the horizontal half-wave dipole. Because the radiation pattern of a half-wave antenna in the plane perpendicular to its axis is a circle (Fig. 2-14) the plots of ground-reflection factors shown in Figs. 2-26 to 2-37, inclusive, show the actual shape of the pattern of such an antenna in the vertical plane at right angles to the wire. That is, they show the variation in intensity with wave angle in the direction broadside to the antenna. In an approximate way, the radiation resistance is larger as the area of the pattern is less, as can be seen

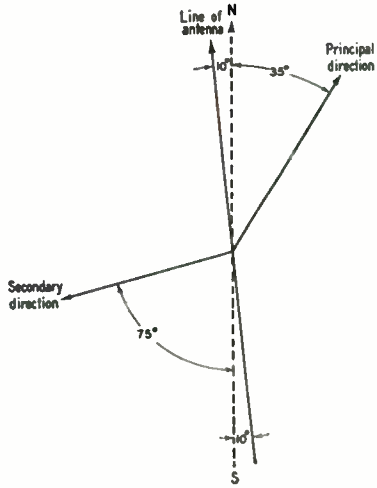


Fig. 2-55 — Example discussed in the text.

by comparing the ground-reflection patterns with the curve of Fig. 2-45.

Varying the height of a horizontal half-wave antenna while the power input is held constant will cause the current in the antenna to vary as its radiation resistance changes. Under the idealized conditions represented in Fig. 2-45 (an infinitely-thin conductor over perfectly-conducting ground) the field intensity at a distant point will vary as shown in Fig. 2-56. In this figure the relative field intensity is expressed in decibels, using the field when the radiation resistance is 73 ohms as a reference (0 db.). From this cause alone, there is a gain of about 1 db. when the antenna height is $\frac{5}{8}$ wavelength as compared with either $\frac{1}{2}$ or $\frac{3}{4}$ wavelength. The gain or loss from the change in radiation resistance should be combined with the reflection factor for the particular wave angle and antenna height considered, in judging the over-all effect of height on performance.

For example, Fig. 2-57 shows the reflection factor, plotted in decibels, for a wave angle of 15 degrees (solid curve). This curve is based on data

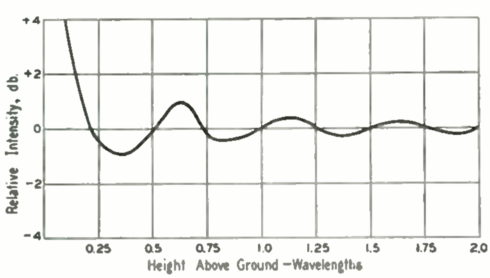


Fig. 2-56 — Gain or loss in decibels because of change in antenna current with radiation resistance, for fixed power input. Perfectly-conducting ground is assumed.

from Figs. 2-26 to 2-37, inclusive. Taken alone, it would indicate that a height of slightly less than 1 wavelength is optimum for this wave angle. However, when the values taken from the curve of Fig. 2-56 are added the broken curve results. Because of the change in radiation resistance, there is a maximum near a height of $\frac{5}{8}$ wavelength that is very nearly as good as the next maximum at a height of 1 to $1\frac{1}{4}$ wavelength. The change in radiation resistance also has the effect of steepening the curve at the lower heights and flattening it in the optimum region. Thus it would be expected that, for this wave angle, increasing the height of a half-wave dipole is very much worth while up to about $\frac{5}{8}$ wavelength, but that further increases would not result in any material improvement. At 14 Mc., where a 15-degree wave angle is taken to be average, $\frac{5}{8}$ wavelength is about 45 feet.

There is, of course, some difficulty in applying the information obtained in this fashion because of the uncertainty as to just where the ground plane is. One possibility, if the antenna can be raised and lowered conveniently, is to measure the current in it while changing its height, keeping the power input constant. Starting with low heights, the current should first go through a

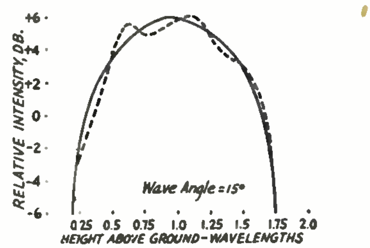


Fig. 2-57 — Solid curve, relative intensity *vs.* height at a wave angle of 15 degrees, because of reflection from perfect ground. Broken curve, height and effect of change in radiation resistance (Fig. 2-56) combined.

minimum (at a theoretical height of about $\frac{3}{8}$ wavelength) and then increase to a maximum as the height is increased. The height at which this maximum is obtained is the optimum.

It should be kept in mind that no one wave angle does all the work. Designing for optimum results under average conditions does not mean that best results will be secured for all types of work and under all conditions. For long-distance work, for example, it is best to try for the lowest possible angle — 10 degrees or less is better for multihop propagation at 14 Mc., for example. However, an antenna that radiates well at such low angles may not be as good for work over shorter distances as one having a broader lobe in the vertical plane.

The effect of radiation resistance is somewhat more marked at the lower frequencies. To cover a distance of 200 miles at night (*F*-layer propagation) requires a wave angle of 60 degrees. As shown by the patterns of Figs. 2-26 to 2-28, the optimum antenna height for this wave angle is ¼ wavelength. However, it is in the region below ¼ wavelength that the radiation resistance decreases most rapidly. At a height of ⅛ wavelength there is a gain of 3.5 db. over a height of ¼ wavelength because of lowered radiation resistance. To offset this, the ground-reflection factor for a wave angle of 60 degrees is about 1.25 at ⅛ wavelength (Fig. 2-26) as compared with 2.0 for ¼ wavelength; this is a loss of 4 db. There is thus a difference of only ½ db., which is not observable, between ⅛ and ¼ wavelength. At 3.5 Mc. this is a considerable difference in actual height, since $\frac{1}{8}\lambda$ is about 35 feet and $\frac{1}{4}\lambda$ is about 70 feet. For *short*-distance work the cost of the supports required for the greater height would not be justified.

Information on the variation in radiation resistance with height for other types of antenna than the half-wave dipole is not readily available. A harmonic antenna can be expected to show such variations, but in general an antenna system that tends to minimize the radiation directly toward the ground under the antenna can be expected to have a lesser order of variation in radiation resistance with height than is the case with the half-wave dipole.

VERTICAL DIRECTIVITY PATTERNS

It was explained in the preceding section that the directive patterns of Figs. 2-47 to 2-53, inclusive, show the relative intensity of radiation in different compass directions for each wave angle selected, but do not attempt to show the amplitude relationship between the wave angles. This is because the intensity at different wave angles varies with the height of the antenna above ground, and an extremely large number

Vertical-Plane Radiation Patterns of Horizontal Half-Wave Antennas Above Perfectly-Conducting Ground

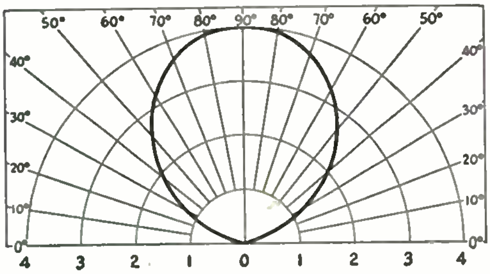


Fig. 2-58 — In direction of wire; height ¼ wavelength.

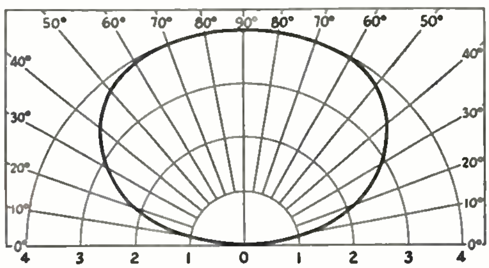


Fig. 2-59 — At right angles to wire; height ¼ wavelength.

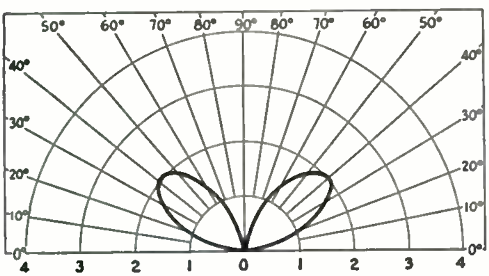


Fig. 2-60 — In direction of wire; height ½ wavelength.

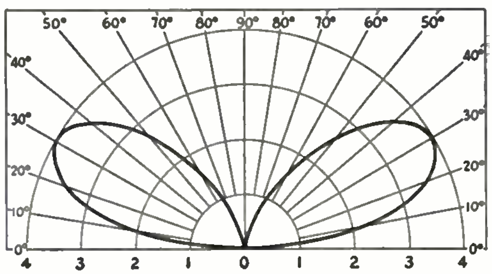


Fig. 2-61 — At right angles to wire; height ½ wavelength.

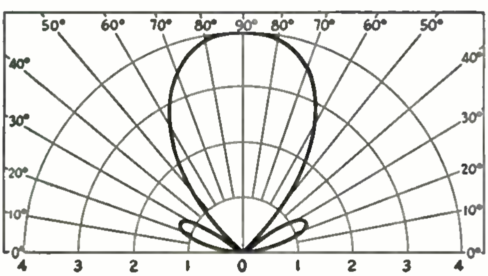


Fig. 2-62 — In direction of wire; height ¾ wavelength.

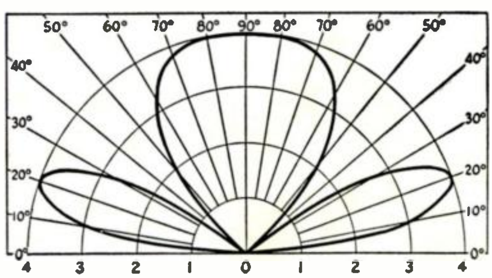


Fig. 2-63 — At right angles to wire; height ¾ wavelength.

of diagrams would be needed to represent the range of heights and lengths of antennas encountered in practice. The information on relative intensity at different wave angles is easily secured from the ground-reflection charts.

However, it is helpful in forming a picture of the operation of antennas to use a form of representation in which the *vertical* directional characteristic is shown for different heights. Inasmuch as we are still confronted by a three-dimensional pattern, it is only possible to do this for selected vertical planes oriented in various directions with respect to the antenna axis. In the case of the horizontal half-wave dipole a plane running in a direction along the axis and another broadside to the antenna will give a good deal of information.

A series of such patterns for a horizontal half-wave dipole at different heights is given in Figs. 2-58 to 2-69, inclusive. The scale is simply an arbitrary one in which the length of a radius drawn from the origin to any point on the graph is proportional to field strength (voltage). The reduction in field strength off the ends of the wire at the lower angles, as compared with the broadside field strength, is quite apparent. It is also apparent that, at some heights, the high-angle radiation off the ends is nearly as great as the broadside radiation.

In vertical planes making some angle intermediate between 0 and 90 degrees with the wire axis the pattern will have a shape intermediate between the two planes shown. By visualizing a smooth transition from the end-on pattern to the broadside pattern as the horizontal angle is varied from 0 to 90 degrees a fairly-good mental picture of the actual solid pattern can be formed.

In the case of a vertical half-wave dipole, the horizontal directional pattern is simply a circle at any wave angle (although the actual field strength

Vertical-Plane Radiation Patterns of Horizontal Half-Wave Antennas Above Perfectly-Conducting Ground

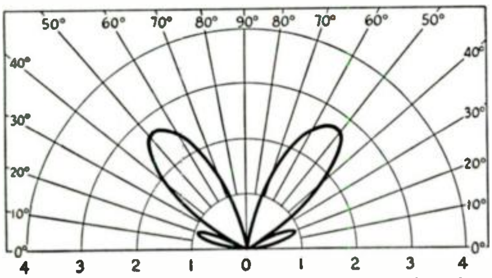


Fig. 2-64 — In direction of wire; height 1 wavelength.

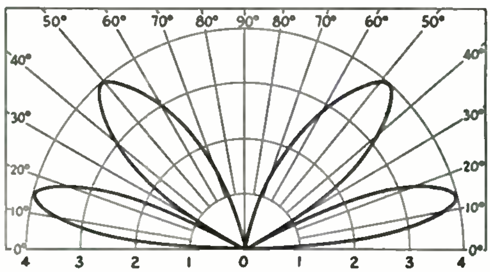


Fig. 2-65 — At right angles to wire; height 1 wavelength.

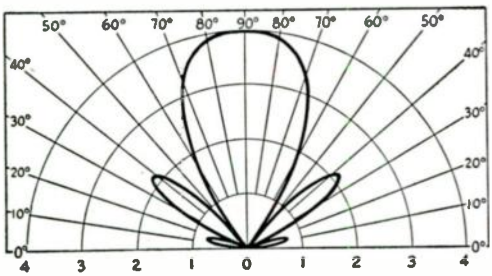


Fig. 2-66 — In direction of wire; height 1¼ wavelength.

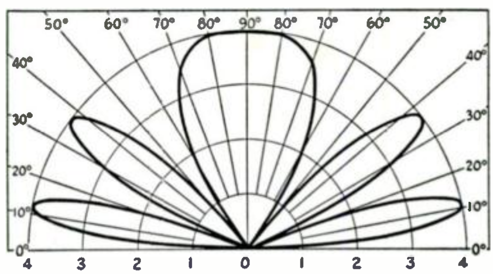


Fig. 2-67 — At right angles to wire; height 1¼ wavelength.

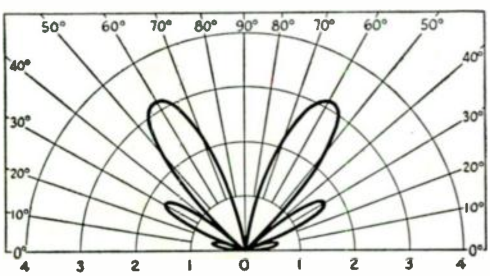


Fig. 2-68 — In direction of wire; height 1½ wavelength.

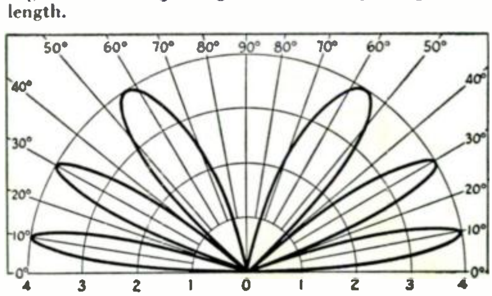


Fig. 2-69 — At right angles to wire; height 1½ wavelength.

will vary, at the different wave angles, with the height above ground). Hence one vertical pattern is sufficient to give complete information, for a selected antenna height, about the antenna in any direction with respect to the wire. A series of such patterns is given in Figs. 2-70 to 2-73, inclusive. These patterns are formed by multiplying one lobe of the free-space pattern of a half-wave dipole by the ground-reflection factor that applies at each wave angle for the antenna height selected, to obtain the resultant relative field strength at each wave angle. The solid pattern in each case is formed by rotating the plane pattern about the 90-degree axis of the graph.

The effect of ground losses at high frequencies is simulated by the broken curves at the very low wave angles. In other respects the curves are based on the assumption that the antenna is erected over perfectly-conducting ground.

SOME PRACTICAL CONSIDERATIONS

At the risk of being repetitious, we must state again that the results from a practical antenna cannot be expected to be exactly according to the theoretical performance outlined in this chapter. The theory that leads to the impedances, radiation patterns, and power-gain figures discussed is necessarily based on idealized assumptions that cannot be exactly realized, although they may be approached reasonably well, in practice.

The effect of imperfectly-conducting earth has been mentioned several times. It will cause the actual radiation resistance of an antenna to differ somewhat from the theoretical figure at a given height. In addition, there is the effect of the length/diameter ratio of the conductor to be considered. Nevertheless, the theoretical figure will approximate the actual radiation resistance closely enough for most practical work. The value of radiation resistance is of most importance in feeding power to the antenna through a transmission line, and a variation of 10 or even 20 per cent will not be serious. In any event, adjustments can be made on the ground to compensate for the discrepancy between practice and theory.

So far as radiation patterns are concerned, the effect of imperfect earth is to decrease the amplitude of the reflected ray and to introduce some phase shift on reflection. The phase shift is generally small with horizontal polarization. Both effects combine to make the maximum reflection factor somewhat less than 2, and to prevent complete cancellation of radiation in the nulls in the theoretical patterns. There may also be a slight change in the wave angle at which maximum reinforcement occurs, as a result of the phase shift. The effect of ground losses on very low angles already has been emphasized.

Aside from ground effects, the theoretical patterns of the antennas discussed are developed on the basis of sinusoidal distribution of current along the antenna, and on the assumption (in

Vertical-Plane Radiation Patterns of Vertical Half-Wave Antennas Above Perfectly-Conducting Ground
The height is that of the center of the antenna. Dotted lines indicate approximate effect of attenuation of the very low-angle radiation because of ground losses.

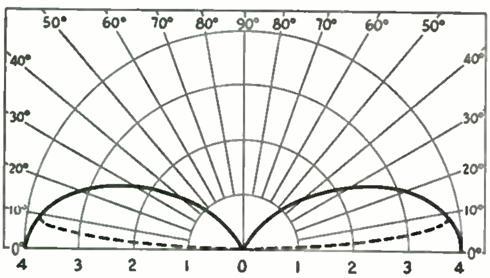


Fig. 2-70 — Height ¼ wavelength.

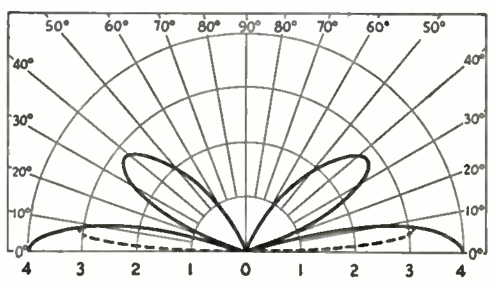


Fig. 2-72 — Height ¾ wavelength.

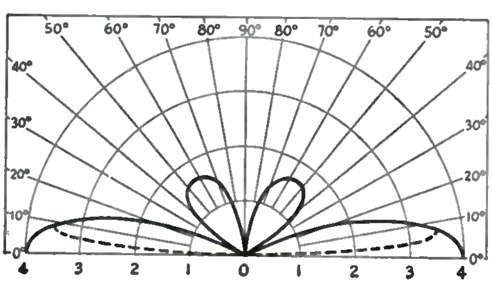


Fig. 2-71 — Height ½ wavelength.

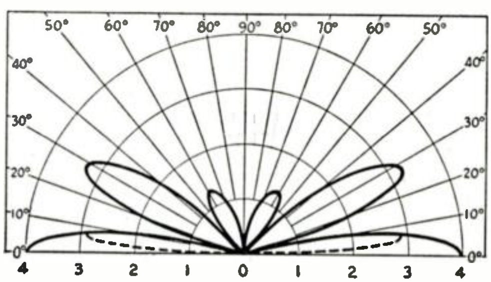


Fig. 2-73 — Height 1 wavelength.

harmonic antennas) that the value of the current is the same at every current loop. Neither is strictly true. In particular, the current in a long harmonic antenna is not the same at every loop because some energy is lost all along the antenna by radiation. This affects both the current going out and the current returning after reflection at the far end of the antenna. The result is that the radiation pattern does not have the perfect symmetry, as indicated in the drawings in this chapter. The lobes pointing away from the end at which the antenna is fed are tilted somewhat toward the direction of the antenna wire, and the lobes pointing toward the fed end are tilted away from the wire. The latter also have less amplitude than the former. Typical measured patterns are shown in Fig. 2-74. There is even a tilt to the pattern of a half-wave antenna when it is fed at one end; however, when such an antenna is fed at the center the pattern is symmetrical.

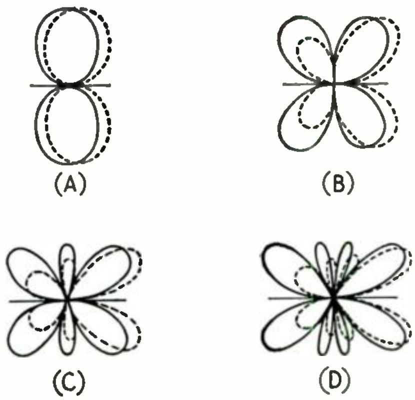


Fig. 2-74 — The effect of feeding an antenna at the end is to cause a tilt to the directional pattern, as shown by these experimentally-determined patterns. A, half wave; B, 1 wavelength; C, 1½ wavelengths; D, 2 wavelengths. Solid patterns are theoretical, dotted patterns experimental. In each case the antenna is fed from the left-hand end.

Finally, the effect of near-by conductors and dielectrics cannot, of necessity, be included in the theoretical patterns. Conductors such as power and telephone lines, house wiring, piping, etc., close to the antenna can cause considerable distortion of the pattern if currents of appreciable magnitude are induced in them. Under similar conditions they can also have a marked effect on the radiation resistance. Poor dielectrics such as green foliage near the antenna can introduce loss, and may make a noticeable difference between summer and winter performance.

The directional effects of an antenna will conform more closely to theory if the antenna is located in a clear space, at least a half wavelength from anything that might affect its properties. In cities, it may be difficult to find such a space at low frequencies. The worst condition arises when near-by wires or piping happen to be resonant, or nearly so, at the operating frequency. Such resonances often can be destroyed by bonding pipes or BX coverings at trial points, checking with a crystal-detector wavemeter to determine the measures necessary to reduce the induced current. Metal masts or guy wires can cause distortion of the pattern unless detuned by grounding or by breaking up the wires with insulators. However, masts and guy wires usually have relatively little effect on the performance of horizontal antennas because, being vertical or nearly so, they do not pick up much energy from a horizontally-polarized wave. In considering near-by conductors, too, the transmission line that feeds the antenna should not be overlooked. Under some conditions that are rather typical with amateur antennas currents will be induced in line by antenna, leading to some undesirable effects. This is considered in detail in Chapter Three.

Special Antenna Types

The underlying principles of antenna operation have been discussed in this chapter in terms of the half-wave dipole, which is the elementary form from which more elaborate antenna systems are built. However, there are other types of antennas that find some application in amateur work, particularly when space limitations do not permit using a full-sized dipole. These include, principally, grounded antennas and loops.

● THE GROUNDED ANTENNA

In cases where vertical polarization is required — for example, when a low wave angle is desired at frequencies below 4 Mc. — the antenna must be vertical. At low frequencies the height of a vertical half-wave antenna would be beyond the constructional reach of most amateurs. A 3.5-Mc. vertical half wave would be 133 feet high, for instance.

However, if the lower end of the antenna is grounded it need be only a quarter wave high to resonate at the same frequency as an ungrounded half-wave antenna. The operation can be understood when it is remembered that ground having high conductivity acts as an electrical mirror, and the missing half of the antenna is supplied by the mirror image. This is shown in Fig. 2-75.

The directional characteristic of a grounded quarter-wave antenna will be the same as that of a half-wave antenna in free space. Thus a vertical grounded quarter-wave antenna will have a circular radiation pattern in the horizontal plane. In the vertical plane the radiation will decrease from maximum along the ground to zero directly overhead.

The grounded antenna may be much smaller than a quarter-wave and still be made resonant by "loading" it with inductance at the base, as in Fig. 2-76 at B and C. By adjusting the inductance of the loading coil even very short wires can be tuned to resonance. However, the efficiency of the wire as a radiator is decreased considerably by decreasing its length. This is because the

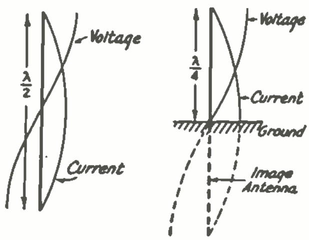


Fig. 2-75—The half-wave antenna and its grounded quarter-wave counterpart. The missing quarter-wavelength can be considered to be supplied by the image in ground of good conductivity.

current at the top of a simple vertical wire such as is indicated in the figures is necessarily zero, so that as the length is reduced less and less of the wire is carrying the high current which produces the greatest radiation.

Current and Voltage Distribution

The current along a grounded quarter-wave vertical wire varies practically sinusoidally, as is the case with a half-wave wire, and is highest at the ground connection. The r.f. voltage, however, is highest at the open end and minimum at the ground. The current and voltage distribution are shown in Fig. 2-76A. When the antenna is shorter than a quarter wave but is loaded to resonance, the current and voltage distribution are part sine waves along the antenna wire. If the loading coil is substantially free from distributed capacity, the voltage across it will increase uniformly from minimum at the ground, as shown at B and C, while the current will be the same throughout.

The radiation resistance of a grounded quarter-wave vertical antenna is approximately 36 ohms. With shorter antennas the radiation resistance decreases. The ratio of the radiation resistance to the resistance of the ground contact system determines the portion of the input power that is actually radiated.

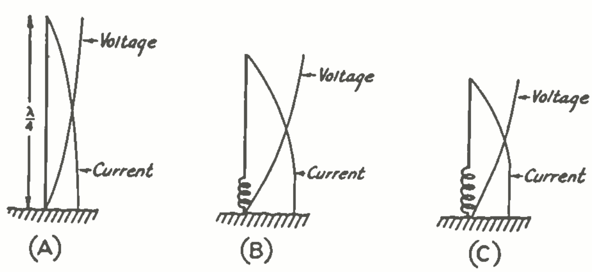


Fig. 2-76 — Current and voltage distribution on a grounded quarter-wave antenna (A) and on successively shorter antennas loaded to resonate at the same frequency.

If the grounding resistance cannot be reduced, the ratio of radiated power to power lost in the ground connection can be increased by increasing the radiation resistance of the antenna. The radiation resistance as measured at the base of the antenna can be increased by making the antenna longer than a quarter wave. The current distribution then becomes as shown in Fig. 2-77. The highest value is secured when the length becomes a half wave, since this length brings a current node at the ground connection.

Note that as the length increases beyond a quarter-wave the maximum current point on the antenna is no longer at the base, but has moved up on the wire. When the antenna height is a half wave the current is maximum halfway up, or one-quarter wavelength above the ground. The upward shift in the current loop is beneficial in two respects: a greater length of wire is carrying high current, thus giving greater effective radiation, and the high-angle radiation is decreased.

Top Loading

The heights required for realization of high radiation resistance usually are impracticable for amateur work. The object of design of vertical

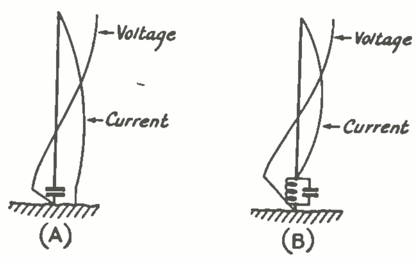


Fig. 2-77 — Current and voltage distribution on grounded antennas longer than ¼ wavelength. (A), between ¼ and ⅜ wave, approximately; (B) half-wave.

grounded antennas which are necessarily ¼ wavelength or less high is to make the current loop come near the top of the antenna, and to keep the current as large as possible throughout the length of the vertical wire. This requires "top loading" — which, in essence, means replacing the missing height by some form of electrical circuit having the same characteristics as the missing part of the antenna, so far as energy traveling up to the end of the antenna is concerned.

One method of top loading is to take a length of wire equal to the missing length in the antenna and fold it into a "flat-top" in such a way that radiation from it is minimized. Several ways of folding are shown in Fig. 2-78. The currents in adjacent folds flow in opposite directions and, providing the lengths are not too large a fraction of a wavelength, the radiation

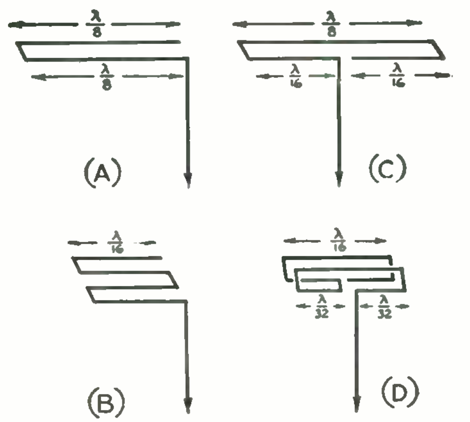


Fig. 2-78 — Several ways in which the top quarter-wave section may be folded to reduce radiation. Spacing between folded wires is not critical but the lengths of the connecting pieces at the folds should be counted in the total length of the quarter-wave top section.

will be substantially canceled because the amplitudes of the currents will not be greatly different in short sections of wire. The dimensions given in Fig. 2-78 are for antennas having the vertical section ¼ wave high. This brings the maximum current at the top of the antenna and makes the impedance between the base and ground of the order of a few thousand ohms. The antenna thus simulates one a half wave high. By dividing all the dimensions by 2, the antenna height can be reduced to ⅛ wavelength, in which case the current will be maximum at the ground connection and the impedance will be considerably less than the 36-ohm value for a grounded quarter-wave antenna. The top-loaded antenna having an actual height of ¼ wavelength is the more desirable type, but the top-loaded ⅛-wave antenna may be used when the greater height is not available.

Instead of a folded top, it is possible to use a simple vertical wire with concentrated capacity and/or inductance at its top to simulate the effect of the missing length. The capacity used is not the usual type of condenser, which would be ineffective since the connection is one-sided, but consists of a metallic structure which exhibits the necessary capacity to space. Practically any sufficiently-large metallic structure can be used for the purpose, but simple geometric forms such as the sphere, cylinder and disc are preferred because of the relative ease with which their capacity can be calculated. The inductance may be the usual type of r.f. coil, with suitable protection from the weather.

The minimum value of capacitive reactance required depends principally upon the ground resistance. Fig. 2-80 is a set of curves giving the reactances required under representative conditions. These curves are based on obtaining 75 per cent of the maximum possible increase in field strength over an antenna of the same height without top loading, and apply with sufficient accuracy to all antenna heights. An inductance

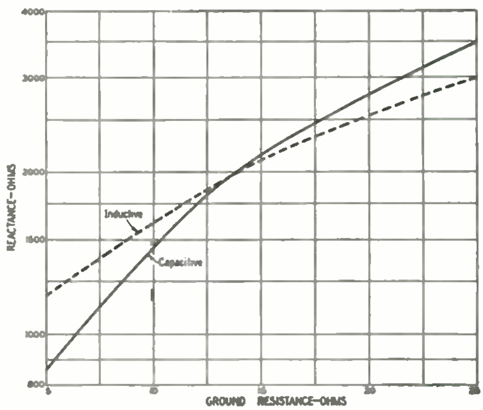


Fig. 2-80 — Inductive and capacitive reactance required for top loading a grounded antenna by the method shown in Fig. 2-79. The reactance values should be converted to inductance and capacitance, using the ordinary formulas, at the operating frequency.

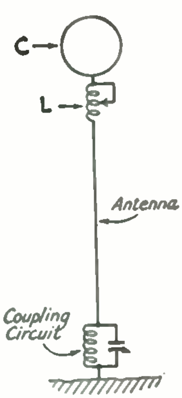


Fig. 2-79 — Top loading with lumped constants. The inductance, *L*, should be adjusted to give maximum field strength with constant power input to the antenna. A parallel-tuned circuit, independently resonant at the operating frequency, is required for coupling to the transmitter when the top loading is adjusted to bring a current node at the lower end of the antenna.

coil of reasonably low-loss construction is assumed. The general rule is to use as large a capacity (low capacitive reactance) as the circumstances will permit, since an increase in capacity will cause an improvement in the field strength. It is particularly important to do this when, as is usually the case, the ground resistance is not known and cannot be measured.

The capacitance of three geometric forms is shown by the curves of Fig. 2-81 as a function of their size. For the cylinder, the length is specified equal to the diameter. The sphere, disc and cylinder can be constructed from sheet metal, if such construction is feasible, but the capacitance will be practically the same in each case if a "skeleton" type of construction, using screening or wire networks, is used.

Ground-Plane Antennas

Instead of being actually grounded, a ¼-wave antenna can work against a simulated ground called a **ground plane.** Such a simulated ground can be formed from wires at least ¼ wavelength long radiating from the base of the antenna as shown in Fig. 2-82. It is obvious that with ¼-wave radials the antenna and any one radial have a total length of ½ wavelength and therefore will be a resonant system. However, with only one radial the directive pattern would be that of a half-wave antenna bent into a right angle at the center; if one section is vertical and the other horizontal this would result in equal components of horizontal and vertical polarization and a nonuniform pattern in the horizontal plane. This can be overcome by using a ground plane in the shape of a disc with a radius of ¼ wavelength. The effect of the disc can be simulated, with simpler construction, by using at least four straight radials equally spaced around the circle, as indicated in the drawing.

The lengths of the radials is not particularly critical when several of them are used, but should be at least ¼ wavelength. A multiradial system can be looked upon as being equivalent to the top loading just described (except, of course, that it is used at the bottom of the antenna) for the special case where the antenna height is approximately ¼ wavelength and the capacitive reactance of the "top" is essentially zero.

The ground-plane antenna is used principally at v.h.f., for the purpose of establishing a "ground" for a vertical antenna mounted many wavelengths above actual ground. This prevents a metallic antenna support from carrying currents that tend to turn the system into the equivalent of a long-wire antenna and raising the wave angle. At the lower frequencies a ground plane can be used beneficially with a vertical antenna as a substitute for an actual ground connection. It will eliminate the resistance loss in the ground

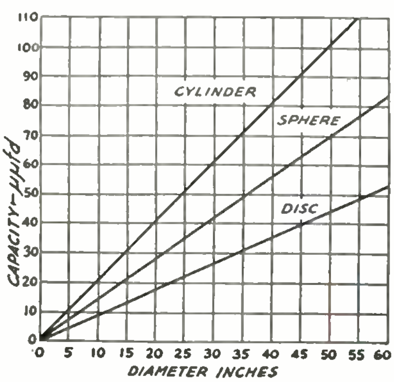


Fig. 2-81 — Capacitance of sphere, disc and cylinder as a function of their diameters. The cylinder length is assumed equal to its diameter.

connection, although not losses in the ground itself in the vicinity of the antenna. The lengths of wires and the configuration used are not especially critical in such a case, particularly when the ground plane is close (in terms of wavelength) to the actual ground. The ground plane is usually called a **counterpoise** when so used.

LOOP ANTENNAS

A loop antenna is a closed-circuit antenna — that is, one in which a conductor is formed into one or more turns so that its two ends are close together. Loops can be divided into two general

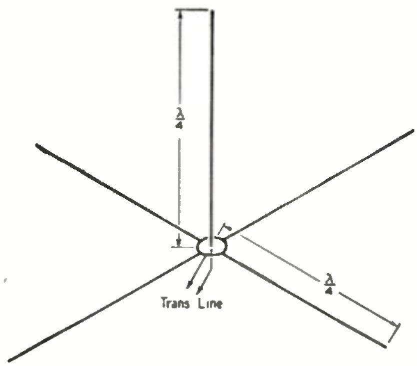


Fig. 2-82 — The ground-plane antenna. Power is applied between the base of the antenna and the center of the ground plane, as indicated in the drawing. The impedance at this point is approximately 36 ohms.

classes, those in which both the total conductor length and the maximum linear dimension of a turn are both very small compared with the wavelength, and those in which both conductor length and loop dimensions begin to be comparable with the wavelength.

A "small" loop can be considered to be simply a rather large coil, and the current distribution in such a loop is the same as in a coil. That is, the current is in the same phase and has the same amplitude in every part of the loop. To meet this condition the total length of conductor in the loop must not exceed about 0.08 wavelength.

A "large" loop is one in which the current is not the same either in amplitude or phase in every part of the loop. This change in current distribution gives rise to entirely different properties as compared with a small loop.

Small Loops

Small loops can be made in the form of a circle, triangle, rectangle, etc., with little or no change in properties. The most convenient form, generally, is a square such as is shown in Fig. 2-83. So long as the total length of the conductor is very small compared with the wavelength the loop acts like a simple inductance and can be tuned to resonance at the desired frequency by a condenser, *C*. The directive pattern of such a loop is given by the small drawing, and is the same as that of an elementary doublet (Fig. 2-12).

Loops of this type do not have much application in amateur work, although they are widely used at frequencies below the standard broadcast band for direction finding. They are not very useful for this purpose at high frequencies because waves arriving at a receiving point via the ionosphere have random polarization and wave angles, and this introduces large errors.

The radiation resistance of a small loop is extremely low. For this reason most of the power supplied to the loop is wasted in conductor resistance loss, when the loop is used for transmitting. A similar situation exists when the loop is used for receiving; because of its small size only a very small amount of energy is absorbed from passing waves. Consequently the loop is very inefficient as compared with a half-wave dipole in both receiving and transmitting. However, when resonated by a condenser it can be markedly better than the few feet of wire often used for reception on the lower frequencies, and it is sometimes possible to take advantage of its directional effects to reduce interference.

Half-Wave Loops

The smallest size of "large" loop generally used is one having a conductor length of ½ wavelength. The conductor is generally formed into a square, as shown in Fig. 2-84, making each side ⅛ wavelength long. When fed at the center of one side the current flows in a closed loop as shown at A. The current distribution is approximately the same as on a half-wave wire, and so is maximum at the center of the side opposite the terminals *X–Y*, and minimum at the terminals themselves. This current distribution causes the field strength to be maximum in the plane of the loop and in the direction looking from the low-current side to the high-current side. If the side opposite the terminals is opened at the center as shown at B (strictly speaking, it is then no longer a loop because it is no longer a closed circuit) the direction of current flow remains unchanged but the maximum current occurs at the terminals. This reverses the direction of maximum radiation.

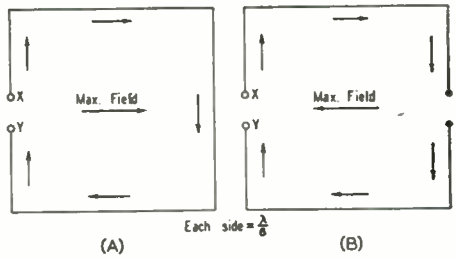


Fig. 2-84 — **Half-wave loops, consisting of a single turn having a total length of ½ wavelength.**

The radiation resistance at a current loop (which is also the resistance at *X–Y* in Fig. 2-84B) is of the order of 50 ohms. The impedance at the terminals in A is a few thousand ohms. This can be reduced by using two identical loops side by side with a few inches spacing between them and applying power between terminal *X* on one loop and terminal *Y* on the other.

Unlike a half-wave dipole or a small loop, there is no direction in which the radiation from a loop of the type shown in Fig. 2-84 is zero. There is appreciable radiation in the direction perpendicular to the plane of the loop, as well as to the "rear" — the opposite direction to the arrows shown. The front-to-back ratio is of the order of 4 to 6 db. The small size and the shape of the directive pattern result in a loss of about 1 db. when the field strength in the optimum direction from such a loop is compared with the field from a half-wave dipole in its optimum direction.

The ratio of the forward radiation to the backward radiation can be increased, and the field strength likewise increased at the same time to give a gain of about 1 db. over a dipole, by using inductive reactances to "load" the sides joining the front and back of the loop. This is shown in Fig. 2-85. The reactances, which should have a value of approximately 360 ohms, decrease the current in the sides in which they are inserted and increase it in the side having the terminals. This increases the directivity and thus increases the efficiency of the loop as a radiator.

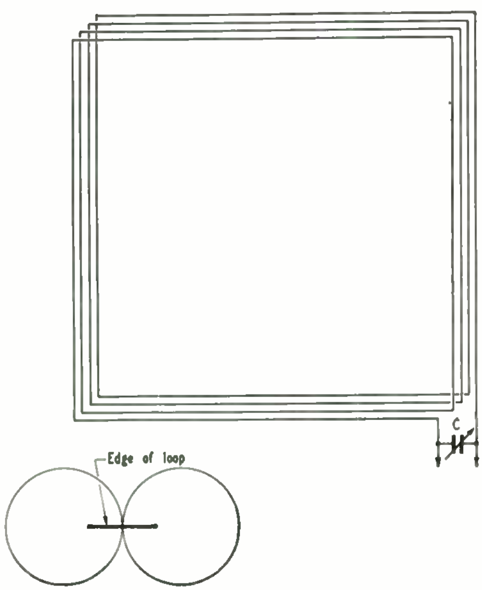


Fig. 2-83 — **Small loop, consisting of several turns of wire having a total length very much less than a wavelength. The directional pattern of such a loop is as shown in the small drawing, with maximum response in the plane of the loop.**

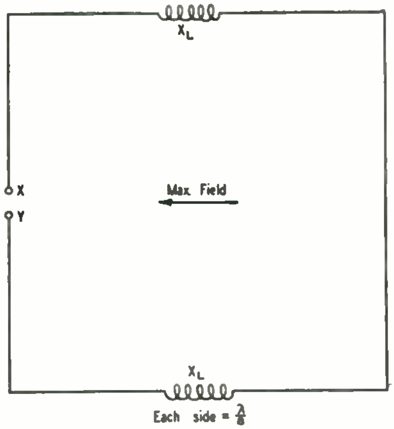


Fig. 2-85 — Inductive loading in the sides of a half-wave loop to increase the directivity and gain. Maximum radiation or response is in the plane of the loop in the direction shown by the arrow.

One-Wavelength Loops

Loops in which the conductor length is one wavelength (sides of square equal to ¼ wavelength) have different characteristics than half-wave loops. Two forms of one-wavelength loops are shown in Fig. 2-86, the difference being in the point at which the terminals are inserted. The relative direction of current flow is as shown in the drawings. This direction reverses halfway around the perimeter of the loop, since such reversals always occur at the junction of each half-wave section of wire.

The directional characteristics of loops of this type are opposite in sense to those of a small loop. That is, the radiation is maximum perpendicular to the plane of the loop and is minimum in any direction in the plane containing the loop. If the two loops shown in Fig. 2-86 are mounted in a vertical plane with the terminals at the bottom, the radiation is horizontally polarized. When the terminals are moved to the center of one vertical side in A, or to a side corner in B, the radiation is vertically polarized.

The dimensions of loops of this type are comparable with those of a half-wave dipole and the efficiency tends to be high. In the optimum direction of radiation either type will show a small gain (less than 1 db.) over a half-wave dipole. The resistance at the input terminals has not been measured accurately but general considerations indicate that it should be of the order of 50 ohms.

● OTHER TYPES OF ANTENNAS

The half-wave dipole and the few special types of antennas described in this chapter form the basis for practically all antenna systems in use at frequencies from the v.h.f. region down. Other fundamental types of radiators are applicable at microwaves, but they are not used at lower frequencies because the dimensions are such as to be wholly impracticable when the wavelength is measured in meters rather than centimeters.

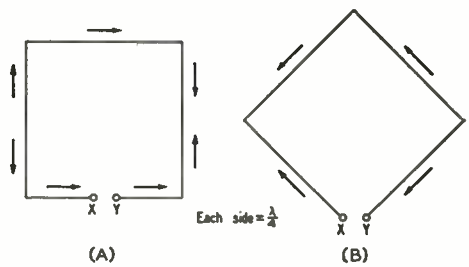


Fig. 2-86 — Loops having sides ¼ wavelength long (total conductor length 1 wavelength). The polarization depends on the orientation of the loop and the point at which the terminals *X–Y* are located.

CHAPTER 3

Transmission Lines

The desirability of installing an antenna in a clear space, not too near buildings or power and telephone lines, has been emphasized in the preceding chapter. On the other hand, the transmitter that generates the r.f. power for driving the antenna is usually, as a matter of necessity, located at some distance from the antenna terminals. The connecting link between the two is the **r.f. transmission line or feeder.** Its sole purpose is to carry r.f. power from one place to another, and to do it as efficiently as possible. In other words, the power *transferred* by the line, as compared with the power *lost* in it, should be as large as the circumstances will permit.

Meeting this requirement at radio frequencies is not as simple as it is at commercial power-line frequencies. At 50 or 60 cycles there is no problem, in the ordinary case, beyond providing for adequate insulation and selecting a conductor size that will carry the load current without excessive voltage drop. These points must be given consideration at radio frequencies, too. But, in addition, at radio frequencies every conductor that has appreciable length compared with the wavelength in use will *radiate* power. That is, every conductor becomes an antenna. Special care must be used, therefore, to minimize radiation from the conductors used in r.f. transmission lines. Without such care, the power radiated by the line may be much larger than that which is lost in the resistance of conductors and dielectrics. Power loss in resistance is inescapable, at least to a degree, but loss by radiation is largely avoidable.

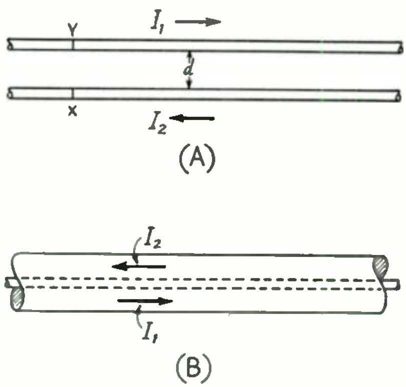


Fig. 3-1 — The two basic types of transmission lines.

Preventing Radiation

Radiation loss from transmission lines can be prevented by using two conductors so arranged and operated that the electromagnetic field from one is everywhere balanced by an equal and opposite field from the other. In such a case the resultant field is zero everywhere in space; in other words, there is no radiation.

For example, Fig. 3-1A shows two parallel conductors having currents I_1 and I_2 flowing in opposite directions. If the current I_1 at point Y on the upper conductor has the same amplitude as the current I_2 at the corresponding point X on the lower conductor, the fields set up by the two currents will be equal in magnitude. Because the two currents are flowing in opposite directions, the field from I_1 at Y will be 180 degrees out of phase with the field from I_2 at X. This does *not* mean that the two fields are 180 degrees out of phase at every point in space. It takes a measurable interval of time for the field from X to travel to Y. If I_1 and I_2 are alternating currents, the phase of the field from I_1 at Y will have changed in such a time interval, and so at the instant the field from X reaches Y the two fields at Y are not exactly 180 degrees out of phase. The two fields will be exactly 180 degrees out of phase at every point in space only when the two conductors occupy the same space — an obviously impossible condition if they are to remain separate conductors.

The best that can be done is to make the two fields cancel each other as completely as possible. This can be accomplished by making the distance, d, between the two conductors small enough so that the time interval during which the field from X is moving to Y is a very small part of a cycle. When this is the case the phase difference between the two fields at any given point will be so close to 180 degrees that the cancellation is practically complete.

Practicable values of d, the separation between the two conductors, are determined by the physical limitations of line construction. A separation that meets the condition of being "very small" at one frequency may be quite large at another. For example, if d is six inches (a commonly-used value) the phase difference between the two fields at Y will be only a fraction of a degree if the frequency is 3500 kc. This is because a distance of six inches is such a small fraction of a wavelength (one wavelength = 360 degrees) at

3500 kc. But at 144 Mc. the phase difference would be 26 degrees, and at 420 Mc. it would be 73 degrees. In neither of these cases could the two fields be considered to "cancel" each other. The separation must be very small in comparison with the wavelength used; it should never exceed 1 per cent of the wavelength and smaller separations are desirable.

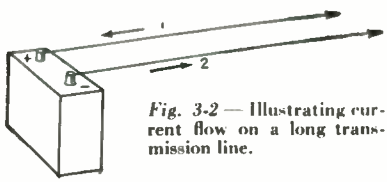


Fig. 3-2 — Illustrating current flow on a long transmission line.

Transmission lines consisting of two parallel conductors as in Fig. 3-1A are called **parallel-conductor lines,** or **open-wire lines,** or **two-wire lines.**

A second general type of line construction is shown in Fig. 3-1B. In this case one of the conductors is tube-shaped and encloses the other conductor. This is called a **coaxial line** ("coax") or **concentric line.** The current flowing on the inner conductor is balanced by an equal current flowing in the opposite direction on the *inside* surface of the outer conductor. Because of skin effect the current on the inner surface of the tube does not penetrate far enough to appear on the outer surface. In fact, the total electromagnetic field outside the coaxial line, as a result of currents flowing on the conductors inside, always is zero because the tube acts as a shield at radio frequencies. (This is because of the extremely low velocity of the field in good conductors, as mentioned in Chapter One.) The separation between the inner conductor and the outer conductor is therefore unimportant from the standpoint of reducing radiation.

The characteristics and applications of both types of lines are covered in detail later in this chapter. Before they can be used intelligently it is necessary to become familiar with some inherent features common to all types of transmission lines. These features arise chiefly as a result of the fact that *time* becomes a dominating factor. In dealing with antennas and r.f. transmission lines it must never be forgotten that electrical and magnetic effects do not take place instantaneously. The maximum speed at which a field can travel is 300,000,000 meters per second — a tremendous speed judged by the standards of human perception, but not so rapid when the frequencies considered also are measured in millions of cycles per second. At high radio frequencies one cycle can be over and a new one beginning by the time the effect of the first has traveled only a few feet. The behavior of a circuit in which this is so is quite different from that of one in which the electromagnetic effects reach the most distant point in a small fraction of a cycle.

CURRENT FLOW IN LONG LINES

Suppose we have a battery connected to a pair of parallel wires that extends to a very great distance, as in Fig. 3-2. At the moment the battery is connected to the wires, electrons in wire No. 1 near the positive terminal of the battery will be attracted to the battery, and the same number of electrons in wire No. 2, near the battery terminal, will be repelled outward along the wire. The directions are shown by the arrows. Thus a current flows in both wires at the instant the battery is connected. These currents do *not* flow throughout the entire length of both wires simultaneously. They start instantaneously *in both wires* at the battery terminals, but a definite time interval will elapse before they are evident at a distance from the battery.

This time interval may be very small. For example, one-millionth of a second (one microsecond) after the connection is made the currents in the wires will have traveled 300 meters, or nearly 1000 feet, from the battery terminals. Note that they flow in both wires simultaneously, even though there may be no connection between the two wires at the end (which is infinitely far away) to form what we ordinarily think of as a closed circuit.

The current is in the nature of a charging current, flowing to charge the capacitance between the two wires. But unlike an ordinary condenser, the conductors of this "linear" condenser have appreciable inductance. In fact, we may think of the line as being composed of a whole series of

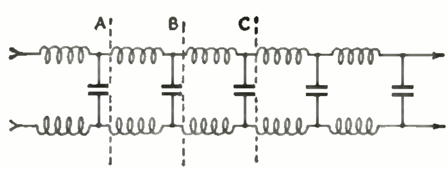


Fig. 3-3 — Equivalent of a transmission line in terms of ordinary circuit constants. The values of *L* and *C* depend on the line construction.

small inductances and capacitances connected as shown in Fig. 3-3, where each coil is the inductance of a very short section of one wire and each condenser is the capacitance between two such short sections.

Characteristic Impedance

An indefinitely-long chain of coils and condensers connected as in Fig. 3-3, where each *L* is the same as all others and all the *C*'s have the same value, has an interesting and important peculiarity. If a voltage is suddenly applied to the terminals at the left-hand end, the condenser in section A cannot charge instantly to the applied voltage, because the inductances in series prevent the current from rising too suddenly. The current flowing into the condenser at the instant the voltage is applied is not, therefore, infinitely large — as it would be, theoretically, if the voltage source had zero resistance and no inductance was present. The current flow is held to an amplitude determined by the value of the inductance in relation to the capacitance. This same relationship holds in section B, and in section C, and so on in each section. As the condenser in section A charges, the voltage across it constitutes the voltage applied to section B, charging the condenser in that section. This goes on, chain fashion, throughout the network.

Since there is a definite limit to the amount of current that can flow for a given applied voltage, the combination of coils and condensers shown in Fig. 3-3 (or the transmission line that it simulates) appears to have a definite value of impedance. The value of this impedance is equal to $\sqrt{L/C}$ (if the coils have zero resistance and there is no leakage across the condensers). Furthermore, if the transmission line or chain of coils and condensers is infinitely long the energy flow is all in one direction — outward from the source — because the process of charging successive sections of the line is never finished. Since there is no energy coming back to the source, there are no reactive effects. Consequently the impedance of the line is a *pure resistance*. (The line itself will "look like" such an impedance only when it is infinitely long. However, even a short line can be made to act as though it were infinitely long by a simple means to be discussed a little later.)

This inherent line impedance is called the **characteristic impedance** or **surge impedance** of the line. As stated, its value is determined by the inductance and capacitance per unit length. These quantities in turn depend upon the size of the line conductors and the spacing between them. The closer the two conductors of the line and the greater their diameter, the higher the capacitance and the lower the inductance. A line with large conductors closely spaced will have low impedance, while one with small conductors widely spaced will have relatively high impedance. The characteristic impedances of typical lines range from a low of about 50 ohms in the coaxial type to a high of somewhat more than 600 ohms for the open-wire type.

The characteristic impedance of the line is a very important property. As we have seen, it determines the amount of current that can flow when a voltage is applied to the line. When a line is infinitely long, the current is simply equal to E/Z_0, where E is the voltage applied to the line and Z_0 is the characteristic impedance. This has nothing to do with the *resistance* of the conductors; in fact, in this simplified picture of a transmission line we have tacitly assumed that the conductors do not have any resistance. The line is an impedance (like any circuit composed of L and C, without any R) that does not consume power. Actually, of course, the conductors do have resistance, so power cannot be transmitted along the line without some loss. But if the line is properly constructed and operated, this loss will be small compared with the amount of power carried to the load to do useful work.

R.F. on Lines

Bearing in mind that *time* must elapse before the currents initiated at the "input" end of the line — that is, the end to which the source of power is connected — can appear some distance away, consider now what happens when a radio-frequency voltage is applied to a transmission line. Suppose an r.f. generator is connected to a long line as shown in Fig. 3-4. To make the figures easy, assume that the frequency is 10 Mc., or 10,000,000 cycles per second. Then each cycle will occupy 0.1 microsecond, as shown by the

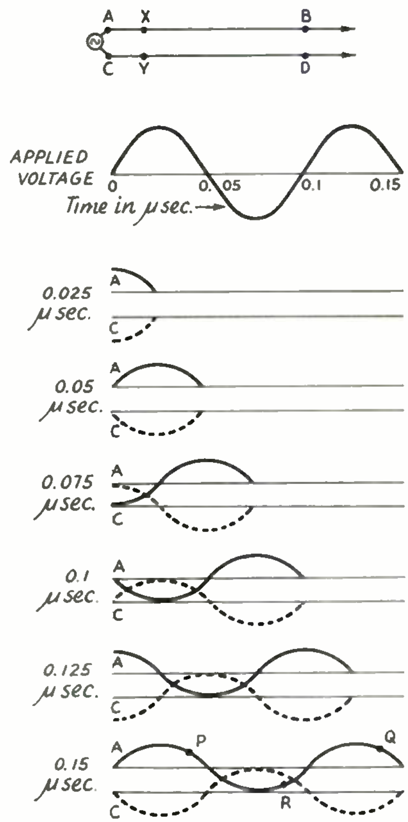


Fig. 3-4 — **Instantaneous current along a transmission line at successive time intervals. The frequency is such that the time of one cycle is 0.1 microsecond.**

drawing of the applied voltage. Suppose that the points B and D along the line are 30 meters away from A and C, respectively. If the current travels with the velocity of light, in 0.1 microsecond (one cycle) it will move 30 meters (300,000,000 meters divided by 10,000,000 cycles) along the line. This is a distance of one wavelength. Thus any currents observed at B and D occur just one cycle later in time than the currents at A and C. To put it another way, the currents initiated at A and C do not appear at B and D, one wavelength away, until the applied voltage has had time to go through a complete cycle.

Since the applied voltage is always changing, the currents at A and C are changing in proportion. The current a short distance away from A and C — for instance, at X and Y — is not the same as the current at A and C because the current at X and Y was caused by a value of voltage that occurred slightly earlier in the cycle. This is true all along the line; at any instant the current anywhere along the line from A to B and C to D is different from the current at every other point in that same distance. The series of drawings shows how the instantaneous currents might be distributed if we could snapshot them at intervals of one-quarter cycle. The current travels out from the input end of the line in waves.

At any selected point on the line the current goes through its complete range of a.c. values in the time of one cycle just as it does at the input end. Therefore (if there are no losses) an ammeter inserted in either conductor would read exactly the same current at any point along the line, because the ammeter averages the current over a whole cycle. The *phases* of the currents at any two separated points would be different, but an ammeter cannot show phase.

"Matched" Lines

In this picture of current traveling along a transmission line we have assumed that the line was infinitely long. Lines have a definite length, of course, and they are connected to or **terminated** in a load at the "output" end, or end to which the power is delivered. If the load is a pure resistance of a value equal to the characteristic impedance of the line, the current traveling along the line to the load does not find conditions changed in the least when it meets the load; in fact, the load just looks like still more transmission line of the same characteristic impedance. Connecting such a load to a short transmission line allows the current to travel in exactly the same fashion as it would on an infinitely-long line.

The reason for this can perhaps be made a little clearer by considering it from another viewpoint. In flowing along a transmission line, the power is constantly being handed from one of the elementary sections in Fig. 3-3 to the next. When the line is infinitely long this power transfer goes on in one direction indefinitely. From the standpoint of section B, Fig. 3-3, for instance, the power it has handed over to section C has simply disappeared in C. So far as section B is concerned, it makes no difference whether C has absorbed the power itself or has in turn handed it along to more line. Consequently, if we substitute something for section C that has the same electrical characteristics, section B will not know the difference. A pure resistance equal to the characteristic impedance of C, which is also the characteristic impedance of the line, meets this condition. It absorbs all the power just as the infinitely-long line absorbs all the power transferred by section B.

A short line terminated in a purely-resistive load equal to the characteristic impedance of the line is said to be **matched.** In a matched transmission line, power travels outward along the line from the source until it reaches the load, where it is completely absorbed.

With either the infinitely-long line or its matched counterpart the impedance presented to the source of power by the line (the **input impedance**) is the same regardless of the line length. It is simply equal to the characteristic impedance of the line. The current in such a line is equal to the applied voltage divided by the characteristic impedance.

STANDING WAVES

Now suppose that the line is terminated in a load that is not equal to the line's characteristic impedance. To take an extreme case, suppose that the output end of the line is short-circuited, as in Fig. 3-5. The impedance of a resistanceless short-circuit would be zero; and whatever the value of power traveling along the line, on reaching the short-circuit it would cause an infinitely-large current to flow. At the same time, the voltage across zero resistance, even with infinitely-large current, would be zero. Furthermore, no power would be consumed in the short-circuit Of course, there is no such thing as a short with no resistance at all, but the actual resistance can be extremely small compared with the characteristic impedance of the line. In the practical case, the current in the short circuit will be very large compared with the current that would flow in the same line when terminated in its characteristic impedance. Also, the voltage across the short circuit will be extremely small and the power consumption likewise small.

What happens is that the outgoing power, on meeting the short-circuit, simply reverses its direction of flow and goes back along the transmission line toward the input end. It has nowhere else to go. The reflected power has essentially the same value as the outgoing power, because the power lost in the short-circuit is quite small if the resistance of the short is low.

The power going into the short-circuit from the line (which we term the "outgoing" power) can be represented by a voltage and current in phase, as indicated at A in Fig. 3-6. The reflected power also can be represented by a voltage and current. The reflected voltage is 180 degrees out of phase with the outgoing voltage, because their sum must be essentially zero across the short-circuit. However, the reflected power is traveling in the opposite direction to the outgoing power, and to represent this the reflected current must be shown 180 degrees out of phase with the reflected voltage; this "negative" power is comparable with the reactive power in an inductance or capacitance and represents power returned to the source. As shown in Fig. 3-6B, this puts the reflected current in phase with the outgoing current. The total current in the short-circuit is the numerical sum of the outgoing and reflected currents, as indicated in Fig. 3-6C, but when the two out-of-phase voltages are added together their sum is very small.

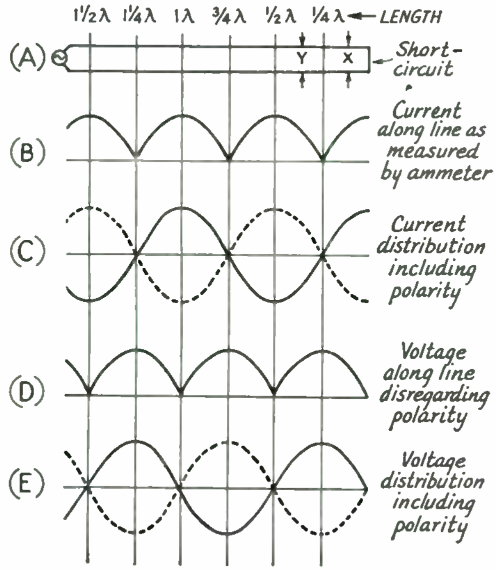


Fig. 3-5 — Standing waves of current and voltage on a short-circuited transmission line.

In discussing Fig. 3-4 it was pointed out that the instantaneous value of a traveling current is different at every point along a conductor in the space of one wavelength, because the current travels that far in one cycle. This is true in each successive one-wavelength portion of the line. At any two points exactly one wavelength apart along the conductor, such as *P* and *Q* in Fig. 3-4, the currents will be in phase. At any two points exactly *one-half* wavelength apart, such as *P* and *R* or *R* and *Q*, the currents will be 180 degrees out of phase, because such a distance represents one-half cycle. The same relationships hold for the phases of the instantaneous voltages along the line.

Now consider the outgoing and reflected currents at some distance from the end of the short-circuited line. The reflected current travels at the same speed as the outgoing current, so its instantaneous value also will be different at every point along the line in the space of one wavelength. This means that, because the currents are traveling in opposite directions, there are bound to be points along the line where they will be in phase and points where they will be completely out of phase. At all *other* points the two currents will be neither completely in nor completely out of phase.

The phase relationship between the outgoing and reflected currents is dependent only on the distance to and from the short-circuit, since this determines the *time* relationship between the two currents. The points at which the two are in phase are therefore at fixed distances from the short; so, likewise, are the points at which they are out of phase. At the in-phase points the two currents go through the usual sequence of instantaneous values during a cycle, but always adding. Similarly, at the out-of-phase points the two go through the normal cyclic variations, but always canceling each other. The result is that the effective current, as measured by an ammeter, will be zero at an out-of-phase point, but will be higher at an in-phase point than at any other place along the line.

We have shown that the outgoing and reflected currents are exactly in phase in the short-circuit. They will therefore also be in phase at a point along the line that represents a total distance of one wavelength. Since one current is traveling toward the short while the other is traveling away from it, a total distance of one wavelength is equal to a *half* wavelength of travel for each current. Consequently the two currents are in phase at a point exactly ½ wavelength from the short. Likewise, they will be in phase at every point along the line that is an integral multiple of a half-wavelength from the short. At such points, therefore, an ammeter will read maximum current.

The two currents will be *out* of phase at a point that represents a total distance of one-half wavelength. Again because the currents are traveling in opposite directions, each current has to travel only one-quarter wavelength to make the total distance equal to one-half wavelength. As a result, the currents are out of phase at a point exactly ¼ wavelength from the short-circuit. They will again be out of phase a half-wavelength from that point — that is, ¾ wavelength from the short — and at every point an odd integral multiple of a quarter-wavelength. The ammeter will read zero at such points.

If the current along the line is measured at successive points it will be found to vary about as shown in Fig. 3-5B. The same result would be obtained by measuring the current in either wire, since the ammeter cannot measure phase. However, if the phase could be checked it would be found that in each successive half-wavelength section of the line the currents at any given instant are flowing in opposite directions, as indicated by the solid line in Fig. 3-5C. Furthermore, the current in the second wire is flowing in the opposite direction to the current in the adjacent section of the first. This is indicated by the broken curve in Fig. 3-5C.

Thus a short-circuited transmission line has **standing waves** of current along its length. Just as in the case of the antennas discussed in Chapter Two, a point of maximum current is called a **current loop** and a point of minimum current is called a **current node.**

Voltage Relationships

The 180-degree reversal of phase in the voltage reflected at the short-circuit, already described, is equivalent to a half cycle or half wavelength of travel. Remembering again that when one voltage travels a quarter wavelength the other covers the same distance (so the relative distance covered by both in one quarter wavelength along the line totals to a *half* wavelength) the phase reversal in the reflected voltage causes the two components to be in phase at a distance of *one-quarter* wavelength from the short-circuit. At a distance of *one-half* wavelength from the short they are again out of phase and cancel.

The standing waves of voltage, shown at D in Fig. 3-5, are therefore displaced by one-quarter wavelength from the standing waves of current. The drawing at E shows the voltages on both wires when phase is taken into account. The polarity of the voltage on each wire reverses

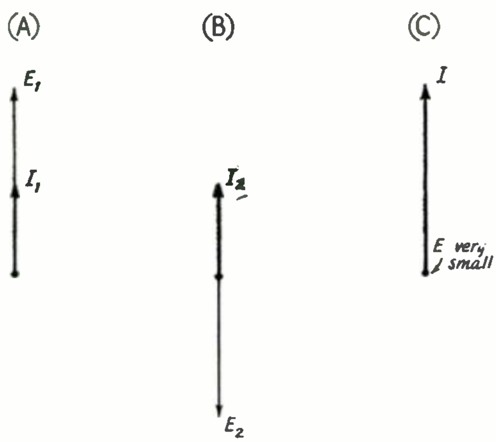


Fig. 3-6 — Voltage and current at the short-circuit on a short-circuited line. These vectors show how the outgoing voltage and current (A) combine with the reflected voltage and current (B) to result in high current and very low voltage in the short-circuit.

in each half-wavelength section of transmission line. A voltage maximum on the line is called a **voltage loop** and a voltage minimum is called a **voltage node.**

Input Impedance

It is apparent, from examination of B and D in Fig. 3-5, that at points that are a multiple of a half wavelength — i.e., ½, 1, 1½ wavelengths, etc. — from the short-circuited end of the line the current and voltage have the same values that they do at the short-circuit. In other words, if the line were an exact multiple of a half wavelength long the generator or source of power would "look into" a short-circuit. On the other hand, at points that are an odd multiple of a quarter wavelength — i.e., ¼, ¾, 1¼, etc. — from the short-circuit the voltage is maximum and the current is zero. Since $Z = E/I$, the impedance at the input terminals **(input impedance)** of lines having such lengths is theoretically infinite. In practice it is very high, but not infinite. This is because the current does not actually go to zero when there are losses in the line. While losses are always present, they are usually small enough on short-circuited lines so that the impedance is of the order of tens or hundreds of thousands of ohms.

At either the odd or even multiples of a quarter wavelength the impedance is a pure resistance, because at these points the current and voltage in the transmission line are exactly in phase.

A detailed study of the outgoing and reflected components of voltage and current will show that at a point such as X in Fig. 3-5, lying anywhere in the section of line between the short-circuit and the first quarter-wavelength point, the current lags behind the voltage. (The reasoning is similar to that used in Chapter Two in connection with Fig. 2-6, although in reverse because an antenna corresponds to the open-circuited line described in the next section.) This is exactly what happens in an inductance, so it can be said that a short-circuited transmission line less than a quarter-wavelength long has inductive reactance. The line also has inductive reactance when its length is between ½ and ¾ wavelength, between 1 and 1¼ wavelengths, and so on.

On the other hand, in the section of line between ¼ and ½ wavelength from the short-circuit the current leads the voltage, so a short-circuited line having a length between these two limits "looks like" a capacitive reactance to the generator to which it is connected. This is also the case when the line length is between ¾ and 1 wavelength, between 1¼ and 1½ wavelengths, and so on.

Fig. 3-7 shows the general way in which the reactance appearing at the input end of a short-circuited line having no losses varies with the length of the line. The reactance is low at lengths that are near a multiple of a half-wavelength, and

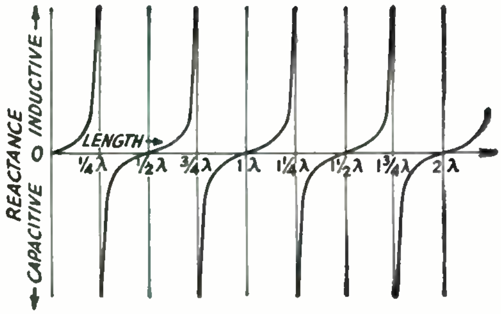


Fig. 3-7 — **Reactance at the input terminals as a function of line length in wavelengths, short-circuited line. These curves show the general way in which the input reactance varies. Specific values are determined by the characteristic impedance of the line.**

very high at lengths that are near an odd multiple of a quarter wavelength. At the exact odd quarter wavelengths both the inductive and capacitive reactances are infinitely large in this theoretical case of no losses — a condition that means, physically, that the line impedance is purely resistive. In a practical line having some loss, the reactance curves do not become infinite at such line lengths. The effect of power dissipation is considered a little later in this chapter.

Open-Circuited Line

If the end of the line is open-circuited instead of short-circuited, there can be no current at the end of the line but a large voltage can exist. Again the outgoing power is reflected back toward the source because it has nowhere else to go. As shown in Fig. 3-8, the outgoing voltage and current are in phase, just as they were in the short-circuited case, because their product represents power delivered to the end of the line. The reflected power, traveling in the opposite direction, is represented by a reflected current out of phase with the reflected voltage. In this case the outgoing and reflected *currents* are out of phase because the total current in an open circuit must be zero. This brings the reflected voltage in phase with the outgoing voltage, so the total voltage at the end of the line is the sum of the two.

The result is that we again have standing waves, but the conditions are reversed as compared with the short-circuited line. Fig. 3-9 shows the open-circuited line case. It may be compared directly with Fig. 3-5. The impedance looking into the line toward the open end is purely resistive for line lengths that are multiples of one-quarter wavelength. It is very low at odd multiples of one-quarter wavelength, and very high at even multiples. In fact, an open-circuited line and short-circuited line behave just alike *if* the length of one differs by one-quarter wavelength from the length of the other.

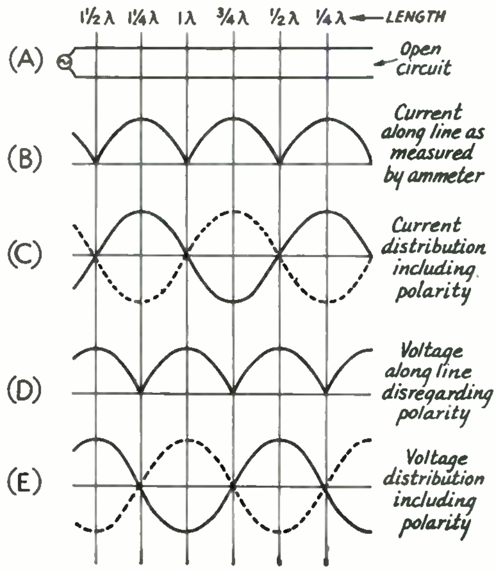


Fig. 3-9 — **Standing waves of voltage and current on an open-circuited transmission line.**

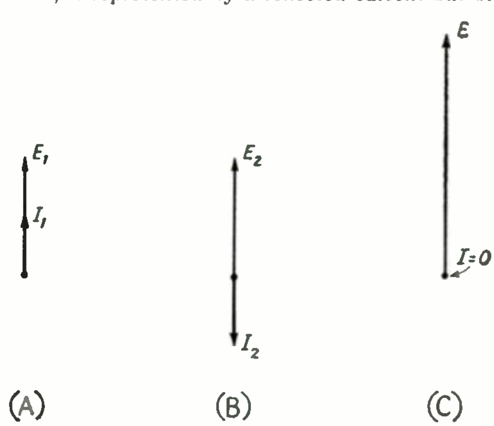


Fig. 3-8 — **Voltage and current at the end of an open-circuited line. A — outgoing voltage and current; B — reflected voltage and current; C — resultant.**

Fig. 3-10 shows how the reactance varies with line length for the open-circuited line. Comparing this with Fig. 3-7 shows that the reactance of any given length of line is of the opposite type to that obtained with a short-circuited line of the same length.

Lines as Circuit Elements

An open- or short-circuited line does not deliver any power to a load, and for that reason is not, strictly speaking, a "transmission" line. However, the fact that a line of the proper length has inductive reactance makes it possible to substitute the line for a coil in an ordinary circuit; likewise, another line of appropriate length hav-

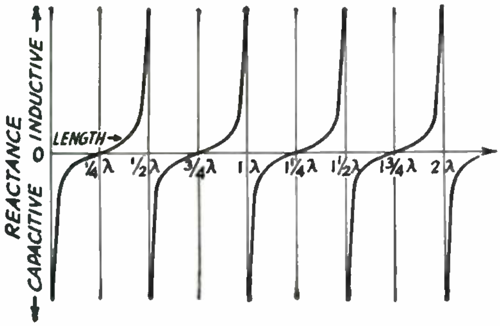


Fig. 3-10 — Reactance at the input terminals as a function of line length in wavelengths, open-circuited line.

ing capacitive reactance can be substituted for a condenser.

Furthermore, at lengths that are exact multiples of a quarter wavelength such lines have the properties of resonant circuits. At lengths where the input reactance passes through zero the line acts like a series-resonant circuit; at lengths for which the reactances are theoretically infinite the line simulates a parallel-resonant circuit. The effective Q of such linear resonant circuits is very high if the line losses, both in resistance and by radiation, are kept down. This can be done without much difficulty, particularly in coaxial lines, if air insulation is used between the conductors. Air-insulated open-wire lines are likewise very good at frequencies for which the conductor spacing is very small in terms of wavelength.

Sections of lines used as circuit elements are usually a quarter wavelength or less long. The desired type of reactance (inductive or capacitive) or the desired type of resonance (series or parallel) is obtained by shorting or opening the far end of the line. The circuit equivalents of various types of line sections are shown in Fig. 3-11.

When a line section is used as a reactance, the amount of reactance obtained is determined by the characteristic impedance and the length of the line. In the case of a line having no losses, and to a close approximation when the losses are small, the inductive reactance of a short-circuited line less than a quarter wave in length is

$$X_L \text{ (ohms)} = Z_0 \tan l$$

where l is the length of the line in electrical degrees (see Chapter Two for explanation of the degree method of measuring length) and Z_0 is the characteristic impedance of the line. The capacitive reactance of an open-circuited line less than a quarter wave in length is

$$X_C \text{ (ohms)} = Z_0 \cot l$$

Fig. 3-12 is a graph of the quantity X/Z_0 for both cases. To find the actual reactance at a given length, multiply the value of X/Z_0 by the characteristic impedance of the line used. For example, a section of 600-ohm line 60 degrees long and short-circuited at the far end will have an inductive reactance of 1.73 × 600 = 1040 ohms. The same line open-circuited would have a capacitive reactance of 0.57 × 600 = 340 ohms. The equivalent inductance and capacitance can be found by substituting these values in the formulas relating inductance and capacitance to reactance, or by using the various charts and calculators available. The frequency for which the line length is 60 degrees must be used, of course. In this example, if the frequency is 14 Mc. the equivalent inductance and capacitance in the two cases are 12.1 μh. and 33.4 μμfd., respectively. Note that when the line length is 45 degrees (⅛ wavelength) the reactance in either case is numerically equal to the characteristic impedance of the line.

Applications of line sections as circuit elements in connection with antenna and transmission-line systems are discussed later in this chapter. In using the graphs in Fig. 3-12 it should be kept in mind that electrical length and physical length

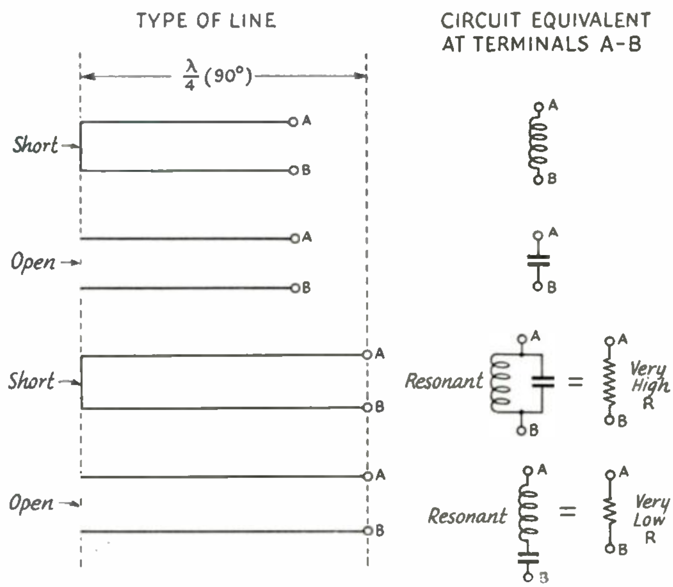


Fig. 3-11 — Lumped-constant circuit equivalents of open- and short-circuited transmission lines.

are not necessarily the same. As pointed out in Chapter Two, the electrical length depends on the velocity at which the electromagnetic fields travel. Data on velocity of propagation along lines of various types of construction are given in this chapter in the section on the practical characteristics of transmission lines.

TERMINATED LINES

Fig. 3-13 shows a line terminated in a resistive load. In such a case at least part of the outgoing power is absorbed in the load, and so is not available to be reflected back toward the source. Because only part of the power is reflected, the reflected voltage and current do not have the same magnitude as the outgoing voltage and current. Therefore there is no such thing as complete cancellation of either voltage or current at any point along the line. However, the *speed* at which the outgoing and reflected components travel is not affected by their amplitude, so the phase relationships between them are similar to those in open- or short-circuited lines.

It was pointed out earlier that if the load resistance, which we will call Z_R, is equal to the characteristic impedance, Z_0, of the line all the power is absorbed in the load. In such a case there is no reflected power and therefore no standing waves of current and voltage. This is a special case that represents the changeover point between "short-circuited" and "open-circuited" lines. If Z_R is less than Z_0, the current is largest

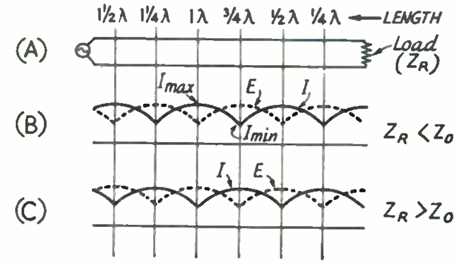


Fig. 3-13 — **Standing waves on a transmission line terminated in a pure resistance.**

at the load and the reflected voltage is out of phase with the outgoing voltage at the load. If Z_R is greater than Z_0, the voltage is largest at the load and the reflected current is out of phase with the outgoing current. The relationships between outgoing and reflected voltages and currents at the load are shown for the two cases in Fig. 3-14.

In either case the reflected power causes standing waves to be formed along the line. In the one case (Z_R less than Z_0) the positions of the minima and maxima are the same as for the short-circuited line; in the other (Z_R greater than Z_0) the positions of the minima and maxima correspond to their positions on the open-circuited line. But in neither case is there a complete null at a minimum point; although the phase relationships are right for cancellation at these points the amplitudes are not. The reflected current or voltage, being smaller, cannot completely oppose the outgoing current or voltage. Typical conditions are shown at B and C in Fig. 3-13.

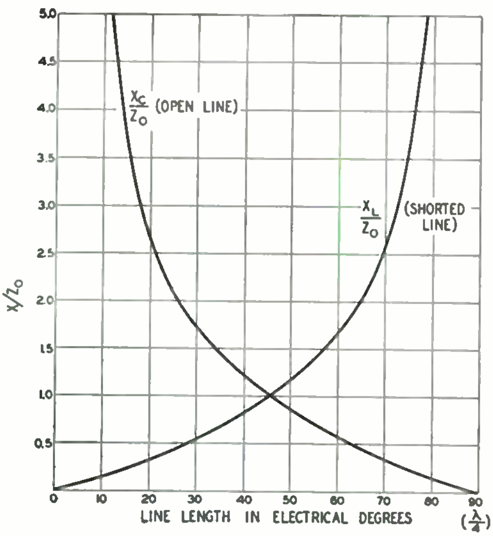


Fig. 3-12 — **Universal reactance curves for open- and short-circuited low-loss lines. The quantity X/Z_0 multiplied by the characteristic impedance of the line is equal to the value of reactance at the input terminals. In lines of greater length the type of reactance varies according to the following table:**

Length	*Short-Circuited Line*	*Open-Circuited Line*
90 to 180 deg.	Capacitive	Inductive
180 to 270 deg.	Inductive	Capacitive
270 to 360 deg.	Capacitive	Inductive

The sign of the reactance reverses in each consecutive 90-degree (¼-wavelength) section.

Input Impedance of Terminated Lines

A transmission line terminated in a purely-resistive load is an important practical case. The load is usually an antenna — which, as we have seen in Chapter Two, "looks like" a pure resistance when its length is adjusted to resonance at the operating frequency. If the antenna resistance is the same as the characteristic impedance of the transmission line the line will be matched just as well as though an actual resistor were used for the termination. In such a case all the power going along the line is absorbed by the antenna and there are no standing waves.

However, more often than not the antenna resistance does not equal the characteristic impedance of the line. We then have standing waves and the input impedance depends on the line length. This input impedance is of first importance because it determines the way in which the source of power must be coupled to the line. We would not expect to get good results by connecting a 115-volt lamp to the 6-volt secondary of a transformer, because the lamp impedance is too high for such a power source. On the other hand, we would not expect to operate a 6-volt lamp directly from 115 volts; the impedance is too low. The same problem arises in coupling a transmitter to a transmission line, with this additional complication: Not only must the voltage be right, but we must also take care of the phase relationship between current and voltage. This phase relationship is determined by the line itself.

It is difficult to give a descriptive picture of what happens along a terminated line without going into detail not easily followed without a good background of a.c.-circuit theory and vector diagrams. It can be said that, in general, the resistive part of any impedance represents power used up or "dissipated" in the circuit, while the reactive part represents power at first accepted by the circuit but then returned to the source at some later part of the a.c. cycle. In a transmission line, then, we would expect that the smaller the amount of power reflected back along the line in comparison with that going out, the smaller the reactive effects. This is true. Also, we would expect that the closer the load comes to being a match for the characteristic impedance of the line, the smaller the amount of power reflected compared with that absorbed in the load. That is also true. Hence the closer Z_R comes to being equal to Z_0, the smaller the reactive effects. It is, in fact, the *ratio* of Z_R to Z_0 that determines the relationship between resistance and reactance at the input terminals of a line of given length.

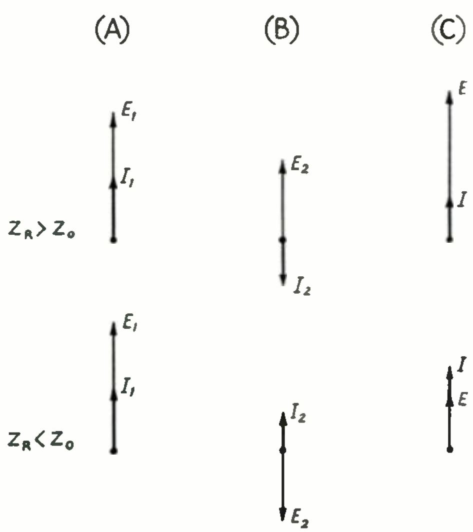


Fig. 3-14 — **Outgoing voltage and current (A), reflected voltage and current (B) and resultants (C) for lines terminated in pure resistance. The amplitudes of the reflected voltage and current depend on the magnitude of the ratio Z_R/Z_0.**

Resistance, reactance and impedance are usually visualized in terms of coils, condensers and resistors, because they are the things dealt with in ordinary circuits. In a broader sense, impedance is simply a ratio of voltage to current, but is not completely specified unless the phase relationship

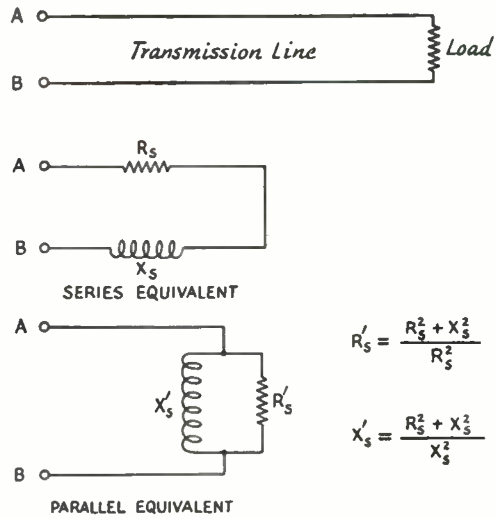


Fig. 3-15 — **Input impedance of a line terminated in a pure resistance. The input impedance can be represented either by a resistance and reactance in series or a resistance and reactance in parallel. The relationship between the series R and X values in the series and parallel equivalents is given by the formulas. X may be either inductive or capacitive, depending on the line length and the ratio Z_R/Z_0.**

between the voltage and current also is given. In the transmission-line case, applying a given voltage to the input terminals will cause a definite current to flow, but the current will not, in general, be in phase with the voltage. No actual coils, condensers or resistors are involved, but we can nevertheless form from such components a circuit that will result in exactly the same current flow for the same applied voltage, and in which the phase relationship between current and voltage will be exactly the same as in the transmission line. In fact, it is possible to form two such "equivalent" circuits and select whichever is most convenient for our purpose. One of these, as shown in Fig. 3-15, consists of a resistance, R_S, and a reactance, X_S (shown here as an inductance, but it can be a capacitance equally well), in series. The other consists of a different value of resistance, R_S, in parallel with a different value of reactance, X_S. But when the values are properly chosen, the impedance between terminals A-B will be exactly the same in the two equivalent circuits as in the transmission line itself.

The type of the reactance — whether inductive or capacitive — at the input terminals of a line is the same as for the short-circuited line when Z_R is less than Z_0, and the same as for the open-circuited line when Z_R is greater than Z_0. The *magnitude* of the reactance is not the same as given in Fig. 3-12, however. The nature of the input impedance of terminated lines of various lengths is shown in Fig. 3-16. Comparing this with Fig. 3-11, it is seen that the difference lies in the fact that there is resistance as well as reactance in the equivalent circuit when the line length is not an exact multiple of a quarter wavelength. At such multiples the input impedance is a pure resistance having the value shown on the drawing. The input impedance of a line having a length of exactly one-half wavelength is equal to the terminating resistance, regardless of the characteristic impedance of the line.

Figs. 3-17 and 3-18 show the resistance and reactance in the series-circuit equivalent of lines having several ratios of load resistance (Z_R) to line characteristic impedance (Z_0). The values are plotted in terms of the ratio of the resistive part of the input impedance (R_S) to the characteristic impedance, and the ratio of the reactive part of the input impedance (X_S) to the characteristic impedance. This makes the curves applicable to all types of lines.

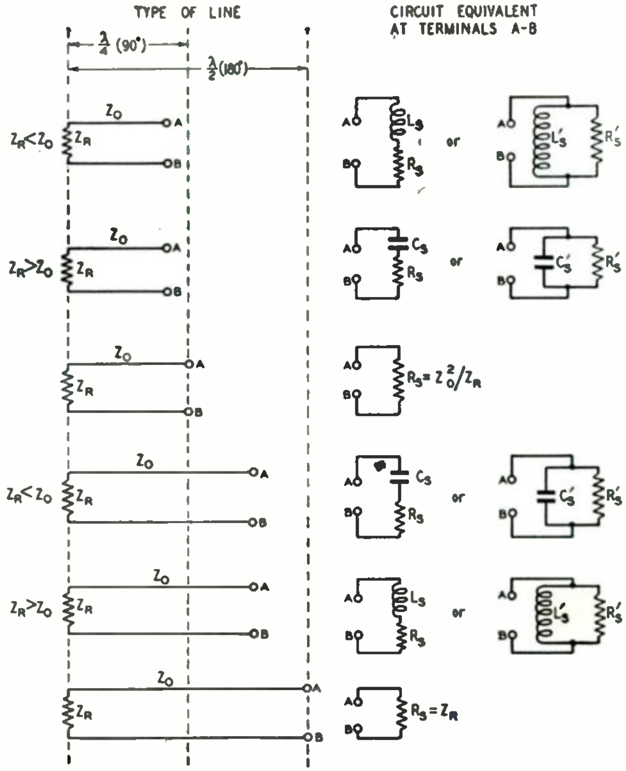


Fig. 3-16 — Equivalent circuits of the input impedance of lines terminated in pure resistance.

Several features of these curves are noteworthy. When Z_R is close to the same value as Z_0 the resistance changes rather slowly with line length and varies over only a relatively small range. Under the same conditions the reactance is never large, and reaches a maximum value at lengths intermediate to exact multiples of a quarter wavelength. When Z_R is considerably different than Z_0, the resistance varies between widely different values in the course of a quarter wavelength and shows a rather sharp peak of the type characteristic of parallel-resonant circuits. The reactance variation is likewise large, and the point of maximum reactance moves closer to a quarter-wave point as the ratio Z_R/Z_0 becomes either very large or very small. The maximum value of reactance approaches one-half the maximum value of resistance, at the extreme Z_R/Z_0 ratios.

For line lengths between 90 and 180 degrees, 270 and 360 degrees, and so on, use the curves for the *opposite* case. For instance, if Z_R is less than Z_0 and the line length is between 90 and 180 degrees, Fig. 3-18 applies. Simply subtract 90 from the actual line length, invert the ratio Z_R/Z_0, and read off the values for the new length. Likewise, if Z_R is greater than Z_0 and the line length is between 90 and 180 degrees (or between 270 and 360 degrees, etc.) subtract 90 from the actual length and use the curves in Fig. 3-17 with the Z_R/Z_0 ratio inverted.

The practical applications of curves of this kind are covered later in this chapter in the section on coupling to transmission lines.

Standing-Wave Ratio

The ratio of maximum current to minimum current along a line, as indicated in Fig. 3-19, is called the **standing-wave ratio.** It is a measure of the mismatch between the load and the line, and is equal to 1 when the line is perfectly matched. (In that case the "maximum" and "minimum" current are the same, since the current does not vary along the line.) When the line is terminated in a purely-resistive load, the standing-wave ratio is

$$S.W.R. = \frac{Z_R}{Z_0} \text{ or } \frac{Z_0}{Z_R}$$

Where $S.W.R.$ = Standing-wave ratio
Z_R = Impedance of load (must be pure resistance)
Z_0 = Characteristic impedance of line

It is customary to put the larger of the two quantities, Z_R or Z_0, in the numerator of the fraction so that the s.w.r. will be expressed by a number larger than 1.

It is somewhat easier to measure the standing-wave ratio than many of the other quantities that enter into transmission-line computations. Consequently, the s.w.r. offers a convenient basis for work with lines. The higher the s.w.r., the greater the mismatch between line and load. Also, the higher the s.w.r. the more marked are the reactive effects. This is shown by the curves of Figs. 3-17 and 3-18, where the s.w.r. in each case is as follows: A, 10; B, 5; C, 3; and D, 2.

Since the voltages along a line are proportional to the amplitude of the current, the s.w.r. may be determined by measuring the maximum and minimum voltages and taking the ratio. That is, the *voltage* s.w.r. and *current* s.w.r. are identical on a given line. However, it is generally more convenient to measure current than voltage.

Impedance Transformation

As shown in Fig. 3-16, the input impedance of a line an exact multiple of a quarter wave in length is a pure resistance when the line is terminated in a purely-resistive load. At even multiples (i.e., multiples of a half wavelength) the input resistance, R_S, is equal to the load resistance, R_L. As a matter of fact, a line an exact multiple of a half wave in length simply repeats, at its input or sending end, whatever impedance exists at its output or receiving end; it does not matter whether the impedance at the receiving end is resistive, reactive, or a combination of both. Sections of line having such length can be cut in or out without changing any of the operating conditions, at least when the losses in the line itself are negligible.

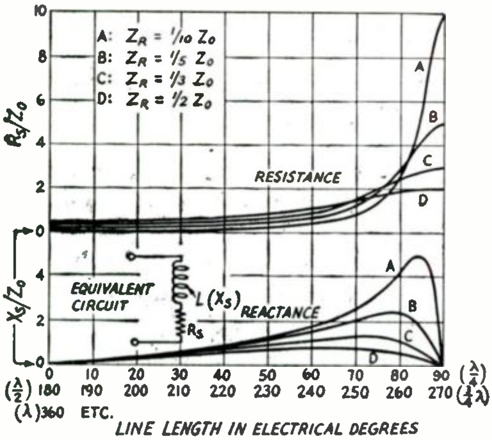


Fig. 3-17 — Universal curves of resistance and reactance *vs.* line length, for various Z_R/Z_0 ratios with Z_R less than Z_0. Actual values of resistance and reactance are found by multiplying the quantity R_S/Z_0 or X_S/Z_0 by the characteristic impedance of the line.

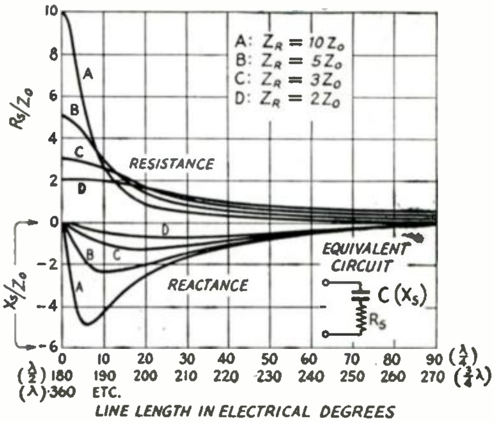


Fig. 3-18 — Universal resistance-reactance curves for Z_R greater than Z_0. These curves are used in the same way as those of Fig. 3-17.

The input impedance of a line an odd multiple of a quarter wavelength long is

$$Z_S = \frac{Z_0^2}{Z_R}$$

If Z_R is a pure resistance, Z_S also will be a pure resistance. Rearranging this equation gives

$$Z_0 = \sqrt{Z_S Z_R}$$

This means that if we have two values of impedance that we wish to "match," we can do so if we connect them together by a quarter-wave transmission line having a characteristic impedance equal to the square root of their product. A quarter-wave line is, in effect, a transformer. It is frequently used as such in antenna work when it is desired, for example, to transform the impedance of an antenna to a new value that will match a given transmission line. This subject is considered in greater detail in a later section of this chapter.

Reactive Terminations

In most amateur applications of transmission lines the load is — or should be — a pure resistance. At least, every attempt is made to make it

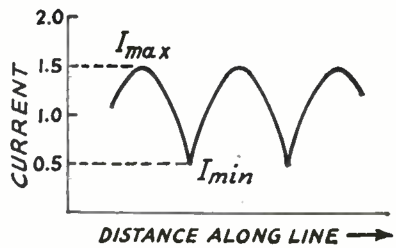


Fig. 3-19 — Measurement of standing-wave ratio. In this drawing I_{max} is 1.5 and I_{min} is 0.5, so the s.w.r. is I_{max}/I_{min} = 1.5/0.5, or 3 to 1.

so. However, there are cases where the load has reactance as well as resistance, and recognizing the symptoms of reactance in the load is of value in indicating what steps should be taken to convert the load to a pure resistance.

The situation is easier to visualize if a line terminated in a pure reactance is considered first. For example, suppose the line is terminated in a capacitive reactance as shown in Fig. 3-20A. It does not matter to the line what physical form the reactance takes; the important thing is that in it the phase relationships are such that the current leads the voltage. The reactance might be a condenser, for example — or it might simply be an additional section of transmission line that exhibits capacitive reactance at its input end, as indicated in Fig. 3-20B.

From Fig. 3-11, a section of open-circuited transmission line less than one-quarter wavelength long will have capacitive reactance. By proper choice of line length, any desired value of reactance can be obtained, as shown in Fig. 3-12. Consequently, any "lumped" reactance, such as a condenser, connected to the end of the transmission line can be replaced by a section of open-circuited line of appropriate length. In other words, a condenser connected across the end of the transmission line lengthens the line, electrically. The amount of effective lengthening depends on the capacitance of the condenser. If the capacitance is small, the reactance is high and the line is only slightly lengthened, electrically. If the capacitance is large and its reactance therefore low, the electrical lengthening may amount to nearly a quarter wavelength.

Once the equivalent lengthening is determined, we can simply look upon the line as one having the new length and apply all that has been said previously about open-circuited lines. In the case just considered, this would mean that the point of maximum current, instead of appearing exactly a quarter wavelength from the end of the line, would appear at something less than a quarter wavelength from the end. This is shown in Fig. 3-20C. The larger the capacitance of the terminating condenser the closer the current loop comes to the physical end of the line. All the other loops and nodes of both current and voltage would be shifted accordingly.

If the line is terminated in an inductance, we can substitute a short-circuited section of line less than one-quarter wavelength long for the lumped inductance. Thus, terminating a line in an inductance is equivalent to extending its length by something less than one-quarter wavelength and short-circuiting it. This is shown at D, E and F in Fig. 3-20. The smaller the inductance, the greater the length of line, up to one-quarter wavelength, that is added electrically. When the equivalent section of line is substituted for the inductance, all that has been said about shorted lines applies, based on the new equivalent length.

When the load has both resistance and reactance the apparent length of the line is again increased. The amount of the apparent increase is affected by the resistance component of the load as well as by the reactive component. However, inductive reactance still will cause the first voltage maximum to appear less than one-quarter wavelength from the load, just as in Fig. 3-20F. Capacitive reactance will cause the first current maximum to appear less than one-quarter wavelength from the load, as in Fig. 3-20C. The positions of the maxima will not be exactly the same, when the load has both resistance and reactance, as they are when the load is purely reactive. Also, the voltage and current nodes do not reach zero because not all the outgoing power is reflected. The actual standing waves would be more like those shown in Fig. 3-13, but with the positions of the nodes and loops shifted in the directions indicated in Fig. 3-20.

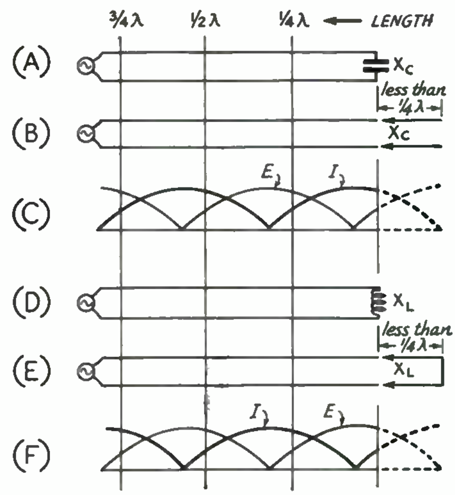


Fig. 3-20 — **Lines terminated in pure reactance. A reactive load is equivalent to a change in the length of the line.**

The presence of reactance in the load, along with resistance, increases the standing-wave ratio over the value it would have with the same load resistance without reactance. Thus in the case of an antenna terminating the transmission line, the s.w.r. will have its lowest value when the antenna is exactly resonant at the operating frequency. At either higher or lower frequencies the s.w.r. will increase because the antenna exhibits reactance at its input terminals when operated off exact resonance, as described in Chapter Two. This effect is modified somewhat by the fact that the radiation resistance also changes as the frequency is varied, but in general the reactance changes will be greater than the resistance changes.

Knowledge of the effect of reactance in the load on the positions of current and voltage maxima can be used to good advantage in adjusting the length of an antenna to resonance as described later.

ATTENUATION

The discussion in the preceding part of this chapter applies to all types of transmission lines, regardless of their physical construction. It is, however, based on the assumption that there is no power loss in the line. Every actual line will have some inherent loss, partly because of the resistance of the conductors, partly because of the fact that power is consumed in every dielectric used for insulating the conductors, and partly because in many cases power escapes from the line by radiation.

Losses in a line modify its characteristic impedance slightly, but usually not to a sufficient extent to be significant. They will also affect the input impedance; in this case the theoretical values will be modified only slightly if the line is short and has only a small loss, but may be changed considerably if an appreciable proportion of the power input to the line is dissipated by the line itself. A large loss may exist because the line is long, because it has inherently high loss per unit length, because the standing-wave ratio is high, or because of a combination of two or all three of these factors.

The reflected power returning to the input terminals of the line is less when the line has losses than it would be if there were none. The over-all effect is that the s.w.r. changes along the line, being highest at the load and smallest at the input terminals. A long, high-loss line therefore tends to act, so far as its input impedance is concerned, as though the impedance match at the load end were better than is actually the case.

Line Losses

Of the three types of loss in a line — resistance loss, dielectric loss, and radiation loss — the radiation loss is likely to be the most variable. Theoretically, radiation loss is negligible in a coaxial line or in a parallel-conductor line having conductor spacing that is very small compared with the wavelength. Practically, it is very low if the line is operating under good conditions — that is, properly balanced. Under poor conditions the power radiated by the line may be comparable with the power radiated by the antenna. But this is a fault of the installation, not the line itself. In comparing lines on a loss basis, therefore, it is customary to ignore radiation and consider only the losses in the conductors and dielectric.

Conductor loss and dielectric loss both increase as the operating frequency is increased, but not in the same way. This, together with the fact that the relative amount of each type of loss depends on the actual construction of the line, makes it impossible to give a specific relationship between loss and frequency that will apply to all types of lines. Each line has to be considered individually.

The power lost in a transmission line is not directly proportional to the line length but varies logarithmically with the length. That is, if 10% of the input power is lost in a section of line of certain length, 10% of the remaining power will be lost in the next section of the same length, and so on. For this reason it is customary to express line losses in terms of decibels per unit length, since the decibel is a logarithmic unit. Calculations are very simple because the total loss in a line is found by multiplying the db. loss per unit length by the total length of the line. Line loss is usually expressed in decibels per 100 feet. However, it is necessary to specify the frequency for which the loss applies, since the loss varies with frequency. In an approximate way, the losses in some types of lines increase directly with the frequency, so the loss for such lines is sometimes expressed in decibels per wavelength of line. In such a case, of course, there are twice as many wavelengths in a given physical length of line when the frequency is doubled, and so the actual loss is doubled.

Effect of S.W.R.

The power lost in a line is least when the line is terminated in a resistance equal to its characteristic impedance, and increases with an increase in the standing-wave ratio. This is because the effective values of both current and voltage become larger as the s.w.r. becomes greater. The increase in effective current raises the ohmic losses in the conductors, and the increase in effective voltage increases the losses in the dielectric.

The increased loss caused by an s.w.r. greater than 1 may or may not be serious. If the s.w.r. is not greater than 2, the *additional* loss caused by the standing waves, as compared with the loss when the line is perfectly matched, does not amount to more than about ½ db. even on very long lines. Since ½ db. is an undetectable change in signal strength, it can be said that from a practical standpoint an s.w.r. of 2 or less is, so far as losses are concerned, every bit as good as a perfect match.

The effect of s.w.r. on line loss is shown in Fig. 3-21. The horizontal axis is the attenuation, in decibels, of the line when perfectly matched. The vertical axis gives the *additional* attenuation, in decibels, caused by standing waves. For example, if the loss in a certain line is 4 db. when perfectly matched, an s.w.r. of 3 on that same line will cause an additional loss of 1.1 db., approximately. The total loss on the poorly-matched line is therefore $4 + 1.1 = 5.1$ db. If the s.w.r. had been 10 instead of 3, the additional loss would be 4.3 db. and the total loss $4 + 4.3 = 8.3$ db.

It is of interest to note that when the line loss is high with perfect matching, the *additional* loss in db. caused by the s.w.r. tends to be constant regardless of the matched line loss. The reason for this is that the amount of power available to be reflected from the load is reduced, because relatively little power reaches the load in the first place. For example, if the line loss with perfect matching is 6 db., only 25% of the power originally put into the line reaches the load. If the mismatch at the load (the s.w.r. at the load) is 4 to 1, 36% of the power reaching the load will be reflected. Of the power originally put into the line, then, $0.25 \times 0.36 = 0.09$ or 9% will be reflected. This in turn will be attenuated 6 db. traveling back to the input end of the line, so that only $0.09 \times 0.25 = 0.0225$ or slightly over 2% of the original power actually gets back to the input terminals. With such a small proportion of power returning to the input terminals the s.w.r. measured at the *input* end of the line would be only about 1.35 to 1 — although it is 4 to 1 at the load. In the presence of line losses the s.w.r. always decreases along the line going from the load to the input end.

On lines having low losses when perfectly matched, a high standing-wave ratio may increase the power loss by a large factor. However, in this case the *total* loss may still be inconsequential in comparison with the power delivered to the load. An s.w.r. of 10 on a line having only 0.3 db. loss when perfectly matched will cause an additional loss of 1 db., as shown by the curves. This loss would produce a just-detectable difference in signal strength.

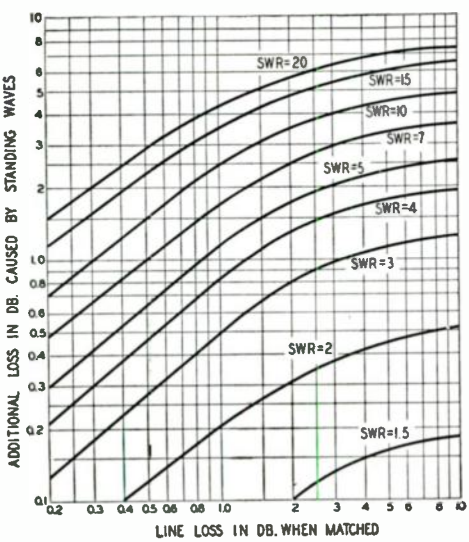


Fig. 3-21 — Increase in line loss because of standing waves. To determine the total loss in decibels in a line having an s.w.r. greater than 1, first determine the loss for the particular type of line, length and frequency, on the assumption that the line is perfectly matched (Tables 3-I and 3-II). Locate this point on the horizontal axis and move up to the curve corresponding to the actual s.w.r. The corresponding value on the vertical axis gives the *additional* loss in decibels caused by the standing waves.

Voltages and Currents on Lines

The power reflected from a mismatched load does not represent an actual loss, except as it is attenuated in traveling back to the input end of the line. It merely represents power returned, and the actual effect is to reduce the power taken from the source. That is, it reduces the coupling between the power source and the line. This is easily overcome by readjusting the coupling until the actual power put into the line is the same as it would be with a matched load. In doing this, of course, the voltages and currents at loops along the line are increased.

As an example, suppose that a line having a characteristic impedance of 600 ohms is matched by a resistive load of 600 ohms and that 100 watts of power goes into the input terminals. The line simply looks like a 600-ohm resistance to the source of power. By Ohm's Law the current and voltage in such a matched line are

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

$$E = \sqrt{PR}$$

Substituting 100 watts for P and 600 ohms for R, the current is 0.408 ampere and the voltage is 245 volts. Assuming for the moment that the line has no losses, all the power will reach the load so the voltage and current at the load will be the same as at the input terminals.

Now suppose that the load is 60 ohms instead of 600 ohms. The s.w.r. is 10, therefore. The reflection coefficient, or ratio of the reflected voltage or current to the voltage or current arriving at the load, is

$$k = \frac{S.W.R. - 1}{S.W.R. + 1}$$

In this case the reflection coefficient is $(10 - 1)/(10 + 1) = 9/11 = 0.818$, so that the reflected voltage and current are both equal to 81.8% of the outgoing voltage and current. The reflected *power* is proportional to the square of either the current or voltage, and so is equal to $(0.818)^2 = 0.67$ times the outgoing power, or 67 watts. Since we have assumed that the line has no losses, this amount of power arrives back at the input terminals and subtracts from the original 100 watts, leaving only 33 watts as the amount of power actually taken from the source.

In order to put 100 watts into the 60-ohm load the coupling to the source must be increased so that the outgoing power minus the reflected power equals 100 watts, and since the power absorbed by the load is only 33% of that reaching it, the outgoing power must equal 100/0.33 = 303 watts. In a perfectly-matched line, the current and voltage with 303 watts input would be 0.71 ampere and 426 volts, respectively. The reflected current and voltage are 0.818 times these values, or 0.581 ampere and 349 volts. At current maxima or loops the current will therefore be 0.71 + 0.58 = 1.29 amp., and at a minimum point will be 0.71 − 0.58 = 0.13 amp. The voltage maxima and minima will be 426 + 349 = 775 volts and 426 − 349 = 77 volts. (Because of rounding off figures in the calculation process the s.w.r. does not work out to be exactly 10 in either the voltage or current case, but the error is very small.)

In the interests of simplicity this example has been based on a line with no losses, but the approximate effect of line attenuation could be included without much difficulty. If the matched-line loss were 3 db., for instance, only half the input power would reach the load, so new values of current and voltage at the load would be computed accordingly. The reflected power would then be based on the attenuated figure, and then itself attenuated 3 db. to find the power arriving back at the input terminals. The over-all result would be, as stated before, a reduction in the s.w.r. at the input terminals as compared with that at the load, along with less actual power delivered to the load for the same power input to the line.

Fig. 3-22 shows the ratio of current or voltage at a loop, in the presence of standing waves, to the current or voltage that would exist with the same power in a perfectly-matched line. Strictly speaking, the curve applies only near the load in the case of lines with appreciable losses. However, the curve shows the maximum possible value of current or voltage that can exist along the line whether there are line losses or not, and so is useful in determining whether or not a particular line can operate safely with a given s.w.r.

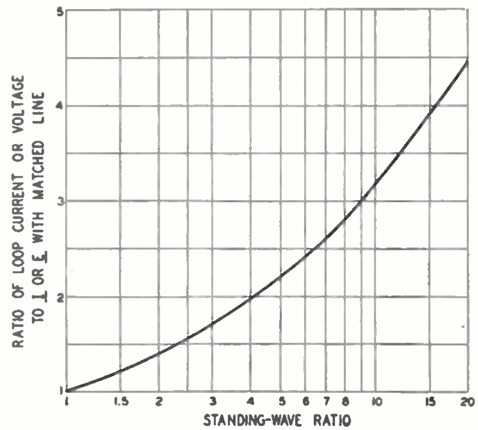


Fig. 3-22 — **Increase in maximum value of current or voltage on a line with standing waves, as referred to the current or voltage on a perfectly-matched line, for the same power delivered to the load. Voltage and current at null points are given by the reciprocals of the values along the vertical axis.**

Line Construction and Operating Characteristics

The two basic types of transmission lines, parallel-conductor and coaxial, can be constructed in a variety of forms. Both types can be divided into two classes: those in which the majority of the insulation between the conductors is air, only the minimum of solid dielectric necessary for mechanical support being used; and those in which the conductors are imbedded in and separated by a solid dielectric. The former class **(air-insulated)** has the lowest loss per unit length because there is no power loss in dry air so long as the voltage between conductors is below the value at which corona forms. At the maximum power permitted in amateur transmitters it is seldom necessary to consider corona unless the s.w.r. on the line is very high.

AIR-INSULATED LINES

A typical type of construction used for parallel-conductor or "two-wire" air-insulated transmission lines is shown in Fig. 3-23. The two line wires, ordinarily No. 12 or No. 14 wire, are supported a fixed distance apart by means of insulating rods called **spacers.** Spacers may be made from insulating material, such as bakelite, or can be purchased ready-made. Materials commonly used in manufactured spacers are isolantite, Lucite, and polystyrene. The spacings used vary from two to six inches, the smaller spacings being necessary at the higher frequencies (28 Mc.) so that radiation will be minimized. The number of spacers required on "open-wire" lines of this type depends on the separation between conductors; it is necessary to use them at

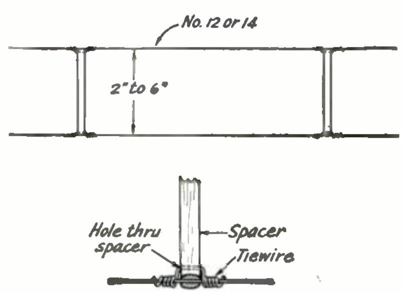


Fig. 3-23 — **Typical open-wire line construction. Commercial spacers are usually provided with grooved ends for the line conductors. The conductor is held in place by a tie wire anchored in a hole near the groove.**

small-enough intervals along the line to prevent the two wires from swinging appreciably with respect to each other in a wind.

The characteristic impedance of open-wire lines of the most useful conductor sizes and spacings is approximately 400 to 600 ohms. When an air-insulated line having lower characteristic impedance is needed, metal tubing having a diameter from ¼ to ½ inch is frequently used. With the larger conductor diameter and relatively close spacing it is possible to build a line having a characteristic impedance as low as about 200 ohms. This type of construction is principally for quarter-wave matching transformers at the higher frequencies.

The characteristic impedance of an air-insulated parallel-conductor line is given by:

$$Z_0 = 276 \log \frac{b}{a}$$

where Z_0 = Characteristic impedance
b = Center-to-center distance between conductors
a = Radius of conductor (in same units as b)

It does not matter what units are used for a and b so long as they are the *same* units. Both quantities may be measured in centimeters, inches, etc. Since it is necessary to have a table of common logarithms to solve practical problems, the solution is given in graphical form in Fig. 3-24 for a number of common conductor sizes.

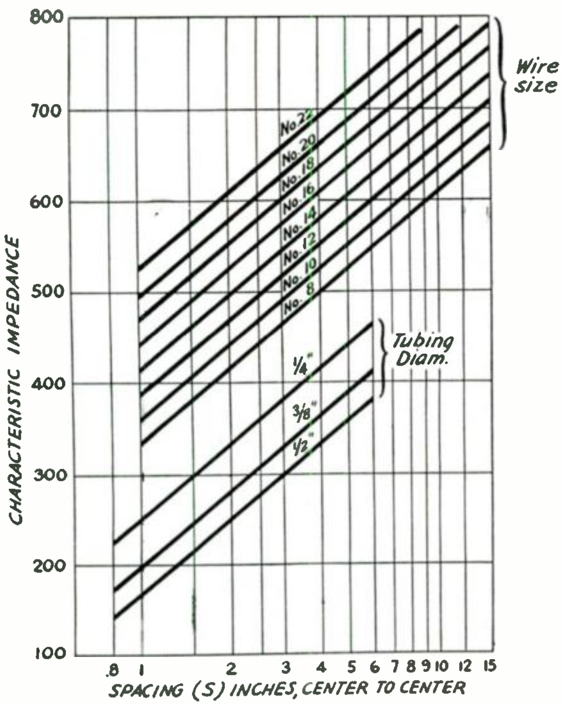


Fig. 3-24 — Characteristic impedance *vs.* conductor size and spacing for parallel-conductor lines.

Another type of parallel-conductor line that is useful in some special applications is the **four-wire** line. In cross-section, the conductors of the four-wire line are at the corners of a square, the spacings being of the same order as those used in two-wire lines. The conductors at *opposite* corners of the square are connected together to operate in parallel. This type of line has a lower characteristic impedance than the simple two-wire type. Also, because of the more symmetrical construction it is better balanced, electrically, to ground and other objects that may be close to the line. The spacers for a four-wire line may be disks of insulating material, "X"-shaped members, etc. The characteristic impedance of four-wire lines is shown graphically in Fig. 3-25.

In coaxial lines of the air-insulated type a considerable proportion of the insulation between conductors may actually be a solid dielectric, because of the necessity for maintaining constant separation between the inner and outer conductors. This is particularly likely to be true in small-diameter lines, typical construction of which is shown in Fig. 3-26. The inner conductor, usually a solid copper wire, is supported by insulating beads at the center of the copper-tubing outer conductor. The beads usually are isolantite and the wire is generally crimped on each side of each bead to prevent the beads from sliding. The material of which the beads are made, and the number of them per unit length of line, will affect the characteristic impedance of the line. The greater the number of beads in a given length the lower the characteristic impedance compared with the value that would be obtained with air insulation only. The presence of the solid dielectric also increases the losses in the line. On the whole, however, a coaxial line of this type tends to have lower actual loss, at frequencies up to about 100 Mc., than any other line construction, provided the air inside the line can be kept dry. This usually means that air-tight seals must be used at the ends of the line and at every joint.

The characteristic impedance of an air-insulated coaxial line is given by the formula

$$Z_0 = 138 \log \frac{b}{a}$$

where Z_0 = Characteristic impedance
b = Inside diameter of outer conductor
a = Outside diameter of inner conductor (in same units as b)

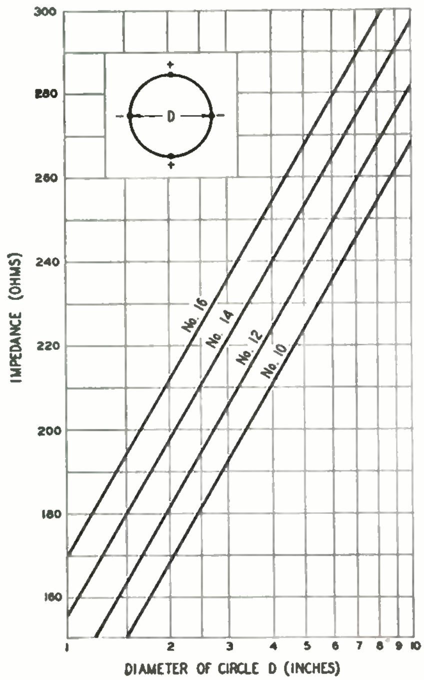


Fig. 3-25 — **Characteristic impedance *vs.* conductor size and spacing for four-wire lines. Opposite wires (not adjacent ones) are connected together at each end of the line.**

Again it does not matter what units are used for *b* and *a*, so long as they are the same. Curves for typical conductor sizes are given in Fig. 3-27.

The formula and curves for coaxial lines are approximately correct for lines in which bead spacers are used, provided the beads are not too closely spaced.

SOLID-DIELECTRIC LINES

Lines in which the conductors are separated by a flexible solid dielectric have a number of advantages over the air-insulated type. They are less bulky, weigh less in comparable types, maintain more uniform spacing between conductors, are generally easier to install, and are neater in appearance. Both parallel-conductor and coaxial lines are available with this type of insulation. The flexibility of the solid-dielectric cable is practically impossible to duplicate in an air-insulated line, and since the solid cable is completely weatherproof there is no danger of excessive power loss or voltage breakdown because of moisture. The flexibility of the parallel-conductor type is also a marked advantage, greatly simplifying installation when the line has to go around corners as compared with the open-wire type.

The chief disadvantage of the solid dielectric lines is that the power loss per unit length is greater than in air-insulated lines. Also, the

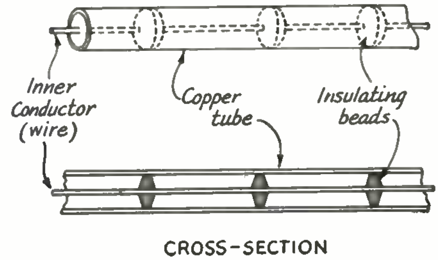


Fig. 3-26 — **Construction of air-insulated coaxial lines.**

power-handling capability is somewhat lower when the line is operated with an appreciable standing-wave ratio. The dielectric used (polyethylene) has quite low losses compared with most solid dielectrics, but still the losses are considerably higher than in air. The power loss causes heating of the dielectric, and when there are standing waves on the line the heating is greatest at points where the voltage maxima occur. Also, if the conductors are small the large currents flowing at current maxima with a high s.w.r. may cause an excessive rise in conductor temperature. Polyethylene softens at fairly low temper-

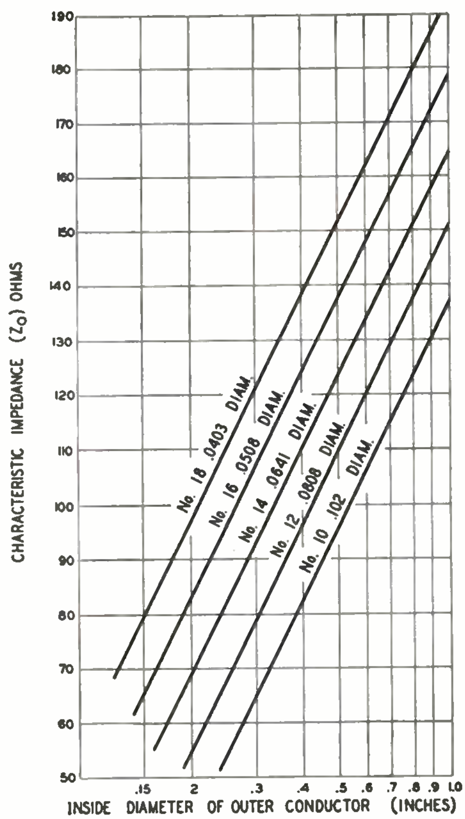


Fig. 3-27 — **Characteristic impedance of typical air-insulated coaxial lines.**

atures, and if the heating is great enough — as it may well be with high power and a high standing-wave ratio — the line will break down both mechanically and electrically.

The construction of a number of types of solid-dielectric lines is shown in Fig. 3-28. In the 300-ohm type ("Twin-Lead" is one trade name) the conductors are stranded wire equivalent to No. 20 in cross-sectional area and are molded in the edges of a polyethylene ribbon about a half inch wide. The effective dielectric is partly solid and partly air. The presence of the solid dielectric lowers the characteristic impedance of the line as compared with the same conductors in air, the result being that the impedance is approximately 300 ohms. The fact that part of the field between the conductors exists outside the solid dielectric leads to an operating disadvantage in that dirt or

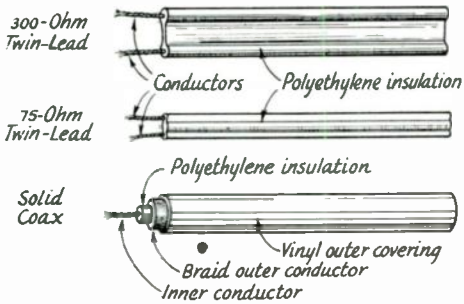


Fig. 3-28 — Construction of parallel-conductor and coaxial lines with solid dielectric.

moisture on the surface of the ribbon tends to change the characteristic impedance. The operation of the line is therefore affected by weather conditions. The effect will not be very serious in a line terminated in its characteristic impedance, but if there is a considerable standing-wave ratio a small change in Z_0 may cause wide fluctuations of the input impedance. Weather effects can be minimized by cleaning the line occasionally and giving it a thin coating of a water-repellent material such as silicone grease or automobile wax.

The transmitting-type 75-ohm Twin-Lead uses stranded conductors about equivalent to solid No. 12 wire, with quite close spacing between conductors. Because of the close spacing most of the field is confined to the solid dielectric, very little existing in the surrounding air. This makes the 75-ohm line much less susceptible to weather effects than the 300-ohm type.

Solid flexible "coax" is available in a rather large number of different types. However, the basic construction is the same in all, and is typified in the drawing in Fig. 3-28. The over-all diameter varies from a little less than ¼ inch to somewhat over an inch, depending chiefly on the power requirements for which the cable was designed. In some cables the inner conductor is stranded; in others, solid wire is used. In some the outer conductor is a single braid; in others it is double. The outer jacket, usually vinyl plastic, plays no part in the electrical performance of the cable but is simply a waterproof covering.

A whole series of fittings for making detachable connections to flexible coaxial cable is available. These include general-purpose connectors, some of which are quite inexpensive, and "constant-impedance" units especially designed so that lengths of cable can be spliced together or terminated without causing a change in the characteristic impedance. Such impedance "bumps" along a line correspond in a general way to having a load that is not matched to the line; that is, they will cause some of the outgoing power to be reflected back toward the input end. In most amateur applications it is not necessary to worry about such impedance discontinuities because most installations will use a single length of cable with connectors only at the load and at the transmitter. In such a case it is a simple matter to compensate for the extra reactance introduced by the connector.

Solid coaxial cables are available in two characteristic impedances, approximately 50 and 75 ohms. In addition to the two parallel-conductor types described above, there are also lightweight two-wire lines of 150 and 75 ohms. These are useful for receiving antennas, but are not heavy enough to carry very much power.

The attenuation and other characteristics of the various types of lines commonly used by amateurs are shown in Table 3-I. Table 3-II is a more comprehensive listing of the standard types of solid-dielectric coaxial cable.

OTHER TYPES OF LINES

There are two types of lines, in addition to those already described, that deserve some mention since they are still used to a limited extent. One is the **single-wire line,** consisting simply of a single conductor running from the transmitter to the antenna. The "return" circuit for such a line is the earth; in fact, the second conductor of the line can be considered to be the image of the actual conductor in the same way that an antenna strung above the earth has an image (see Chapter Two). The characteristic impedance of the single-wire line depends on the conductor size and the height of the wire above ground, ranging from 500 to 600 ohms for No. 12 or No. 14 conductors at heights of 10 to 30 feet. By connecting the line to the antenna at a point that represents a resistive impedance of 500 to 600 ohms the line can be matched and will operate without standing waves.

Although the single-wire line is very simple to install, it has at least two outstanding disadvantages that have made it practically obsolete. Since the return circuit is through the earth, the behavior of the system depends on the kind of

TABLE 3-I

Transmission-Line Velocity Factors and Attenuation

Type of Line	*Velocity Factor V*	*Attenuation ** db./100 ft.; Mc.*						*Capacitance per Foot μμfd.*
		3.5	*7*	*14*	*28*	*50*	*144*	
Open-wire, 400 to 600 ohms	0.975*	0.03	0.05	0.07	0.1	0.13	0.25	
Parallel-tubing	0.95*	***						
Coaxial, air-insulated	0.85*	0.2	0.28	0.42	0.55	0.7	1.4	
RG-8/U (53 ohms)	0.66	0.28	0.42	0.64	1.0	1.4	2.6	29.5
RG-58/U (53 ohms)	0.66	0.53	0.8	1.2	1.9	2.7	5.1	28.5
RG-11/U (75 ohms)	0.66	0.27	0.41	0.61	0.92	1.3	2.4	20.5
RG-59/U (75 ohms)	0.66	0.56	0.82	1.4	1.8	2.5	4.6	21.0
Twin-Lead, 300 ohms	0.82	0.18	0.3	0.5	0.84	1.3	2.8	5.8
Twin-Lead, 150 ohms	0.77	0.2	0.35	0.6	1.0	1.6	3.5	10
Twin-Lead, 75 ohms	0.68	0.37	0.64	1.1	1.9	3.0	6.8	19
Transmitting Twin-Lead, 300 ohms	0.84	0.19	0.32	0.55	0.9	1.4	3.0	
Transmitting Twin-Lead, 75 ohms	0.71	0.29	0.49	0.82	1.4	2.1	4.8	
Rubber-insulated twisted-pair or coaxial ****	0.56 to 0.65	0.96	1.6	2.5	4.2	6.2	13	

* Average figures for air-insulated lines taking into account effect of insulating spacers.
** For lines terminated in characteristic impedance.
*** Losses between open-wire line and air-insulated coaxial cable. Actual loss with both open-wire and parallel-tubing lines is higher than listed because of radiation, especially at higher frequencies.
**** Approximate figures for good-quality rubber insulation.

ground over which the antenna and transmission line are erected. In practice, it may not be possible to get the necessary good connection to actual ground that is required at the transmitter. Second, the line always radiates since there is no near-by second conductor to cancel the fields. The radiation will be minimum when the line is properly terminated because the line current is lowest under those conditions. However, the line is always a part of the radiating antenna system to a greater or lesser extent.

The second type of line is the twisted pair. This is a two-conductor line of the twisted-lampcord type, usually with rubber insulation. It was rather popular some years ago before the introduction of the polyethylene-insulated lines. Its chief disadvantage is the high power loss per unit length (see Table 3-I) together with the fact that the losses increase markedly when the line is damp or wet. The characteristic impedance of most lampcords and special twisted lines for radio use is in the region from 70 to 140 ohms. In view of the much lower losses of the polyethylene lines now available, twisted-pair lines are not recommended — although in an emergency ordinary lampcord can be pressed into service if nothing better is available.

ELECTRICAL LENGTH

In the discussion of line operation earlier in this chapter it was assumed that currents traveled along the conductors at the speed of light. Actually, the velocity is somewhat less, the reason being that electromagnetic fields

TABLE 3-II — STANDARD COAXIAL CABLES

Class of Cables		*Army-Navy Type Number*	*Inner Conductor*	*Nominal Diameter of Dielectric (Inches)*	*Nominal Over-all Diameter (Inches)*	*Weight Pounds Foot*	*Nominal Impedance Ohms*	*Nominal Capacitance µµfd./ft*	*Attenuation in db. per 100 ft.* 30 Mc.	*Attenuation in db. per 100 ft.* 100 Mc.	*Max. Operating Voltage R.M.S.*	*Remarks*
50–55 Ohms	Single Braid	RG-58/U [1]	20 A.W.G. copper	0.116	0.195	0.025	53.5	28.5	2.0	4.1	1900	General-purpose small-size flexible cable
		RG-8/U	7/21 A.W.G. copper	0.285	0.405	0.106	52.0	29.5	1.0	2.1	4000	General-purpose medium-size flexible cable
		RG-17/U	0.188 copper	0.680	0.870	0.460	52.0	27.5	0.37	0.85	11,000	Large high-power low-attenuation transmission cable
		RG-19/U	0.250 copper	0.910	1.120	0.740	52.0	29.5	0.31	0.70	14,000	Very large high-power low-attenuation transmission cable
	Double Braid	RG-55/U [2]	20 A.W.G. copper	0.116	Max. 0.206	0.034	53.5	28.5	2.0	4.1	1900	Small-size flexible cable
		RG-5/U	16 A.W.G. copper	0.185	0.332	0.087	53.5	28.5	1.4	2.7	2000	Small microwave cable
		RG-14/U	10 A.W.G. copper	0.370	0.545	0.216	52.0	29.5	0.66	1.4	5500	General-purpose semiflexible power transmission cable
70–80 Ohms	Single Braid	RG-59/U	22 A.W.G. copperweld	0.146	0.242	0.032	73.0	21.0	1.9	3.7	2300	General-purpose small-size video cable
		RG-11/U	7/26 A.W.G. tinned copper	0.285	0.405	0.096	75.0	20.5	0.93	1.9	4000	Medium-size flexible video and communication cable
	Double Braid	RG-6/U [3]	21 A.W.G. copperweld	0.185	0.332	0.082	76.0	20.0	1.4	2.7	2700	Small-size video and i.f. cable
		RG-13/U	7/26 A.W.G. tinned copper	0.280	0.420	0.126	74.0	20.5	0.93	1.9	4000	I.f. cable
Low Capacitance	Single Braid	RG-62/U	A.W.G. copperweld	0.146	0.242	0.0382	93.0	13.5 Max. 14.5	1.6	3.0	750	Small-size low-capacitance air-spaced cable
		RG-63/U	22 A.W.G. copperweld	0.285	0.405	0.0832	125	10.0 Max. 11.0	1.1	2.0	1000	Medium-size low-capacitance air-spaced cable
	Double Braid	RG-71/U [4]	22 A.W.G. copperweld	0.146	Max. 0.250	0.0457	93.0	13.5 Max. 14.5	1.6	3.0	750	Small-size low-capacitance air-spaced cable for i.f. purposes
Twisting Application	Single Braid	RG-41/U [5]	16/30 A.W.G. tinned copper	0.250	0.425	0.150	67.5	27.0	4.6	10.0	3000	Special twist cable

All cables use copper braid and vinyl protective covering unless otherwise noted.

[1] Tinned-copper shielding braid.

[2] Tinned-copper shielding braid and polyethylene protective covering.

[3] Shielding braid: inner — silver-coated copper; outer — copper.

[4] Shielding braid: inner — plain copper; outer — tinned copper. Polyethylene protective covering.

[5] Tinned-copper shielding braid and neoprene protective covering.

travel more slowly in dielectric materials than they do in free space. In air the velocity is practically the same as in empty space, but a practical line always has to be supported in some fashion by solid insulating materials. The result is that the fields are slowed down; the currents travel a shorter distance in the time of one cycle than they do in space, and so the wavelength along the line is less than the wavelength would be in free space at the same frequency.

Whenever reference is made to a line as being so many wavelengths (such as a "half wavelength" or "quarter wavelength") long, it is to be understood that the *electrical* length of the line is meant. Its actual physical length as measured by a tape always will be somewhat less. The physical length corresponding to an electrical wavelength is given by

$$\textit{Length}\ \text{(feet)} = \frac{984}{f}V$$

where f = Frequency in megacycles
V = Velocity factor

The velocity factor is the ratio of the actual velocity along the line to the velocity in free space. Values of V for several common types of lines are given in Table 3-I.

Because a quarter-wavelength line is frequently used as an impedance transformer, it is convenient to calculate the length of a quarter-wave line directly. The formula is

$$\textit{Length}\ \text{(feet)} = \frac{246}{f}V$$

where the symbols have the same meaning as above.

LINE INSTALLATION

One great advantage of coaxial line, particularly the flexible solid-dielectric type, is that it can be installed with almost no regard for its surroundings. It requires no insulation, can be run on or in the ground or in piping, can be bent around corners with a reasonable radius, and can be "snaked" through places such as the space between walls where it would be impracticable to use other types of lines. However, coax lines always should be operated in systems that permit a low standing-wave ratio, and precautions must be taken to prevent r.f. currents from flowing on the *outside* of the line. This point is discussed later in this chapter.

In installing parallel-wire lines care must be used to prevent the line from being affected by moisture, snow and ice. Spacers that are impervious to moisture and are unaffected by sunlight and the weather should be used on air-insulated lines. Steatite spacers meet this requirement adequately, although they are somewhat heavy. The wider the line spacing the longer the leakage path across the spacers, but this cannot be carried too far without running into line radiation, particularly at the higher frequencies. Where an open-wire line must be anchored to a building or other structure, stand-off insulators of a height comparable with the line spacing should be used if mounted in a spot that is open to the weather. Lead-in bushings for bringing the line into a building also should have a long leakage path.

The line should be kept away from other conductors, including downspouting, metal window frames, flashing, etc., by a distance equal to two or three times the line spacing. Conductors that are very close to the line will be coupled to it in greater or lesser degree, and the effect is that of placing an additional load across the line at the point where the coupling occurs. Reflections take place from this coupled "load," raising the standing-wave ratio. The effect is at its worst when one line wire is closer than the other to the external conductor. In such a case one wire carries a heavier load than the other, with the result that the line currents are no longer equal. The line then becomes "unbalanced."

Solid-dielectric two-wire lines have a relatively small external field because of the small spacing, and can be mounted within a few inches of other conductors without much danger of coupling between the line and such conductors. Clip-type stand-off insulators are available for supporting the line when it must run along walls or similar structures.

Sharp bends should be avoided in any type of transmission line, because such bends cause a change in the characteristic impedance of the line. The result is that reflections take place from each bend. This is of less importance when the s.w.r. is high than when an attempt is being made to match the load to the line's characteristic impedance. It may be impossible to get the s.w.r. down to a desired figure until the necessary bends in the line are made more gradual.

Coupling the Line to the Antenna

Throughout the discussion of transmission-line principles in the first part of this chapter the operation of the line has been described in terms of an abstract "load." This load had the electrical properties of resistance and, sometimes, reactance. It did not, however, have any physical attributes that associated it with a particular electrical device. That is, it could be anything at all that exhibits electrical resistance and/or reactance.

The fact is that, so far as the line is concerned, it does not matter what the load *is*, just so long as it will accept power. It is important to keep this in mind. Many amateurs make the mistake of confusing transmission lines with antennas, believing that because two identical antennas have different kinds of lines feeding them, or the same kind of line with different methods of coupling to the antenna, the "antennas" are different. Nothing could be farther from the truth. There may be practical reasons why one **system** (including antenna, transmission line, and coupling method) may be preferred over another in a particular application. But to the transmission line an antenna is just a load that terminates it, and the important thing is what that load looks like to the line in terms of resistance and reactance. *Any* kind of transmission line can be used with *any* kind of antenna, if the proper measures are taken to couple the two together.

Resonant and Nonresonant Lines

Probably the principal factor that determines the way a transmission line is operated is the frequency range over which the antenna is to work. Very few types of antennas will present essentially the same load impedance to the line on harmonically-related frequencies. As a result, the builder is usually faced with the necessity of choosing between an antenna system that will permit operating the transmission line with a low standing-wave ratio, but is confined to one operating frequency or a narrow band of frequencies, and a system that will permit operation in several harmonically-related bands but with a large s.w.r. on the line.

Methods of coupling the line to the antenna

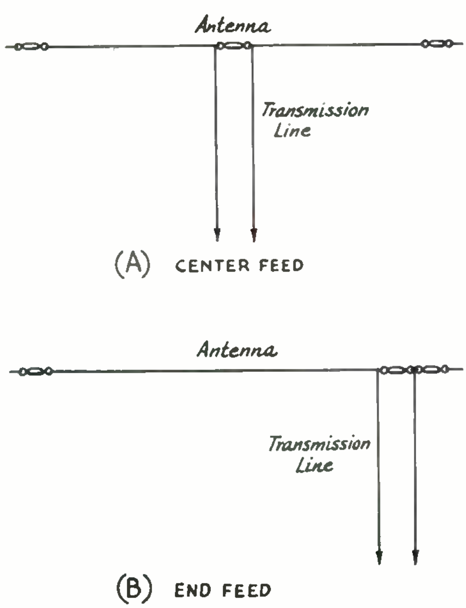


Fig. 3-29 — Center and end feed as used in simple antenna systems.

therefore divide, from a practical standpoint, into two classes. In the first, operation on several amateur bands is the prime consideration and the standing-wave ratio is secondary. The s.w.r. is normally rather large and the input impedance of the line depends rather critically on the line length and the operating frequency. Lines operated in this way are called **resonant** or **tuned** lines, because it is necessary to "tune out" the reactive part of the input impedance in order to get adequate power transfer between the transmitter and the line.

In the second class, a conscious attempt is made, when necessary, to transform the antenna impedance to a value that matches the characteristic impedance of the line. When this is done the line operates with a very low standing-wave ratio and its input impedance is essentially a pure resistance, regardless of the line length. The line in such a case is **nonresonant,** or **flat** — "flat" because a graph of the current along the line is a straight line. A transmission line is considered to be flat, within practical limits, if the s.w.r. is not more than about 1.5 to 1.

A principal reason for matching the antenna to the line impedance is that a flat line operates with the least power loss. While it is always desirable to reduce losses and thus increase efficiency, the effect of standing waves in this connection can be overemphasized. This is particularly true at the lower amateur frequencies, where the inherent loss in most types of lines is quite low even for runs that, in the average amateur installation, are rather long. For example, 100 feet of 300-ohm receiving-type Twin-Lead has a loss of only 0.18 db. at 3.5 Mc., as shown by Table 3-I. Even with an s.w.r. as high as 10 to 1 the additional loss caused by standing waves is less than 0.7 db., from Fig. 3-21. Since 1 db. represents the minimum detectable change in signal strength, it does not matter in this case whether the line is flat or not. But at 144 Mc. the loss in the same length of line perfectly matched is 2.8 db., and an s.w.r. of 10 would mean an *additional* loss of 4 db. At the higher frequency, then, it is worth while to match the antenna and line as closely as possible.

Another reason for matching is that certain types of lines, particularly those with solid dielectric, have definite voltage and current limitations. At the lower frequencies this is a far more compelling reason than power loss for at least approximate matching. Where the voltage and current must not exceed definite maximum values, the amount of power that the line can handle is inversely proportional to the standing-wave ratio. If the safe rating on the 300-ohm line in the example above is 500 watts when perfectly matched, the line can handle only 50 watts with equal safety when the s.w.r. is 10 to 1. Thus, despite the fact that the line losses are low enough to make no difference in the signal

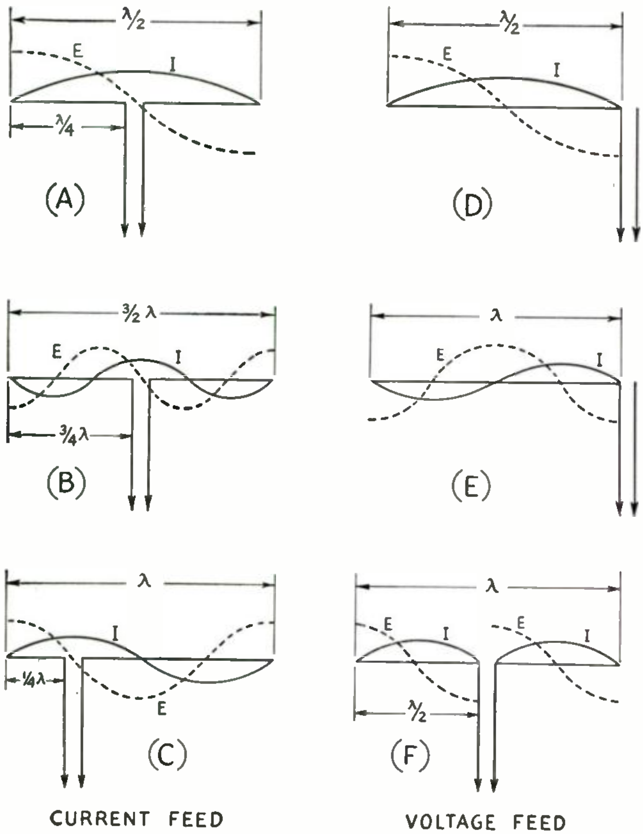


Fig. 3-30 — **Current and voltage feed, in antennas operated at the fundamental frequency, 2 times the fundamental and 3 times the fundamental. The current and voltage distribution on the antenna are identical with both methods only at the fundamental frequency.**

strength the high s.w.r. could be tolerated only with low-power transmitters.

Aside from power considerations, there is a more-or-less common belief that a flat line "does not radiate" while one with a high s.w.r. does radiate. This impression is quite unfounded. It is true that the radiation from a parallel-conductor line increases with the current in the line, and that the effective line current increases with the s.w.r. However, the loss by radiation from a properly-balanced line is so small in the first place (and is, furthermore, independent of the line length) that multiplying it several times still does not bring it out of the "negligible" classification. Whenever a line radiates it is because of faulty installation. Radiation from this cause can take place from either resonant or nonresonant lines, parallel-conductor or coaxial.

RESONANT LINES

Since no attempt is made to match the antenna impedance to the characteristic impedance of the line in resonant-line systems, there is no real problem in coupling the line to the antenna. Customary practice is to connect the line either to the center of the antenna **(center feed)** as indicated in Fig. 3-29A, or to one end **(end feed)** as shown in Fig. 3-29B.

Because the line operates at a rather high standing-wave ratio, the best type to use for resonant operation is the open-wire line. Twin-Lead of the 300-ohm variety can also be used, but the power limitations discussed in the preceding section should be kept in mind. Although the manufacturers have placed no power rating on receiving-type 300-ohm line it seems reasonable to make the assumption, based on the conductor size, that the current should not exceed 1.5 amp. in a line installed so that there is free air circulation about it. This corresponds to a power of 675 watts in a perfectly-matched 300-ohm line. When there are standing waves, the safe power can be found by dividing 675 by the s.w.r. In the case of a center-fed antenna as in Fig. 3-29A, the s.w.r. in the average case should not exceed about 5 to 1, so using receiving-type 300-ohm Twin-Lead as a resonant line would appear to be safe for transmitter power outputs up to 100 watts or so. Transmitting-type Twin-Lead rated at 1 kw., matched, should be able to handle a power of 250 watts at a s.w.r. of about 5 to 1.

Since there is no point in using a resonant line to feed an antenna that is to operate on one amateur band only, the discussion to follow will be based on the assumption that the antenna is to be operated on its harmonics for multiband work.

"Current" and "Voltage" Feed

Usual practice is to connect the transmission line to the antenna at a point where either a current or voltage loop occurs. If the feed point is at a current loop the antenna is said to be **current fed**; if at a voltage loop the antenna is **voltage fed.**

These terms should not be confused with center feed and end feed, because they do not necessarily have corresponding meanings. There is always a voltage loop at the end of a resonant antenna, no matter what the number of half-wavelengths, so a resonant end-fed antenna is always voltage fed. This is illustrated at D and E in Fig. 3-30 for end-fed antennas a half wavelength long (antenna fundamental frequency) and one wavelength long (second harmonic). It would continue to be true for an end-fed antenna operated on any harmonic. However, Fig. 3-30F shows voltage feed at the *center* of the antenna; in this case the antenna has a total length of two half wavelengths each of which is voltage fed by one side of the transmission line. Voltage feed is determined not by the physical position of the transmission line on the antenna, but by the fact that a voltage loop occurs on the antenna at the feed point. Since voltage loops always occur at integral multiples of a half wavelength from either end of a resonant antenna, feeding the antenna at any half-wavelength point constitutes voltage feed.

Typical cases of current feed are shown at A, B and C in Fig. 3-30. The feed point is at a current loop, which always occurs at the midpoint of a half-wave section of the antenna. In order to feed at a current loop the transmission line must be connected in at a point that is an odd multiple of one-quarter wavelength from either end of the resonant antenna. A center-fed antenna is also current fed *only* when the antenna length is an *odd* multiple of one-half wavelength. Thus the antenna in Fig. 3-30B is both center fed and current fed since it is three half wavelengths long. It would also be center fed and current fed when five, seven, etc., half wavelengths long. To current feed a one-wavelength antenna, or any resonant antenna having a length that is an *even* multiple of one-half wavelength, it is necessary to shift the feed point from the center of the antenna (where a voltage loop always occurs in such a case) to the middle of one of the half-wave sections. This is indicated in Fig. 3-30C in the case of a one-wavelength antenna; current feed can be used if the line is connected to the antenna at a point ¼ wavelength from either end of the antenna.

In the usual case of an antenna operated on several bands, the point at which the transmission line is attached is of course fixed. The antenna length is such that it is resonant at some frequency in the lowest-frequency band to be used and the transmission line is connected either to the center or the end. The current and voltage distribution along antennas fed at both points is shown in Fig. 3-31. With end feed, A to F inclusive, there is always a voltage loop at the feed point. Also, the current distribution is such that in every case the antenna is a true harmonic radiator of the type described in Chapter Two.

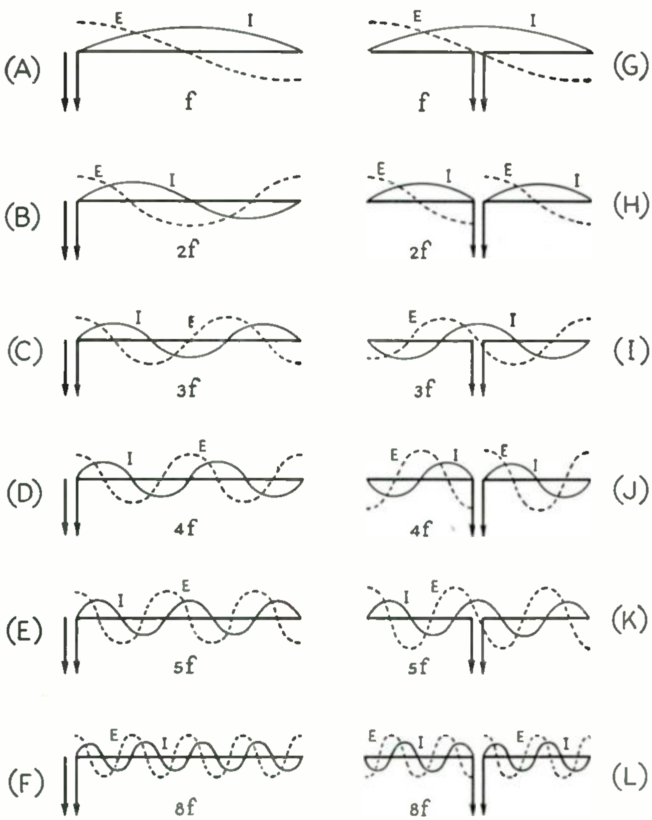


Fig. 3-31 — **Current and voltage distribution at the fundamental frequency and various multiples, with both end feed and center feed. The distributions are the same with both types of feed only when the frequency is an odd multiple of the fundamental.**

With center feed, the feed point is always at a current loop on the fundamental frequency and all *odd* multiples of the fundamental. In these cases the current and voltage distribution are identical with the distribution on an end-fed antenna. This can be seen by comparing A and G, C and I, and E and K, Fig. 3-31. (In I, the phase is reversed as compared with C, but this is merely for convenience in drawing; the actual phases of the currents in each half-wave section reverse each half cycle so it does not matter whether the current curve is drawn above or below the line, so long as the *relative* phases are properly shown in the same antenna.) On odd multiples of the fundamental frequency, therefore, the antenna operates as a true harmonic antenna.

On *even* multiples of the fundamental frequency the feed point with center feed is always at a voltage loop. This is shown at H, J and L in Fig. 3-31. Comparing B and H, it can be seen that the current distribution is different with center feed than with end feed. With center feed the current in both half-wave sections of the antenna are in the same phase, but with end feed the current in one half-wave section is in reverse phase to the current in the other. This does not mean that one antenna is a better radiator than the other, but simply that the two will have different directional characteristics. The center-fed arrangement is commonly known as "two half-waves in phase," while the end-fed system is a "one-wavelength antenna" or "second-harmonic" antenna.

Similarly, the system at J has a different current and voltage distribution than the system at D, although both resonate at four times the fundamental frequency. A similar comparison can be made between F and L. The center-fed arrangement at J really consists of two one-wavelength antennas, while the arrangement at L has two 2-wavelength antennas. These have different directional characteristics than the 2-wavelength and 4-wavelength antennas (D and F) that resonate to the same multiple, respectively, of the fundamental frequency.

The reason for this difference between odd and even multiples of the fundamental frequency in the case of the center-fed antenna can be explained with the aid of Fig. 3-32. It will be recalled from Chapter Two that the direction of current flow reverses in each half wavelength of wire. Also, in any transmission line the currents in the two wires always must be equal and flowing in opposite directions at any point along the line. Starting from the end of the antenna, the current must be flowing in one direction throughout the first half-wavelength section, whether this section is entirely antenna or partly antenna and partly one wire of the transmission line. Thus, in A, Fig. 3-32, the current flows in the same direction from P to Q, since this is all the same conductor. However, one quarter wave is in the antenna and one in the transmission line. The current in the other line wire, starting from R, must flow in the opposite direction in order to balance the current in the first wire, as shown by the arrow. And since the distance from R to S is ½ wavelength, the current continues to flow in the same direction all the way to S. The currents in the two halves of the *antenna* are therefore flowing in the same direction. Furthermore, the current is maximum ¼ wavelength from the ends of the antenna, as previously explained, and so both currents are maximum at the junction of the antenna and transmission line. This makes the current distribution along the antenna exactly the same as with end feed.

Fig. 3-32B shows the case where the over-all length of the antenna is one wavelength, making a half wave on each side. A half wavelength along the transmission line also is shown. If we assume that the current is flowing downward in the line conductor from Q to R, it must be flowing upward from S to T if the line currents are to balance. However, the distance from Q to P is ½ wavelength, and so the current in this section of the antenna must flow in the opposite direction to the current flowing in the section from Q to R. The current in section PQ is therefore flowing *away* from Q. Also, the current in section TU must be flowing in the opposite direction to the current in ST, and so is flowing *toward* T. The

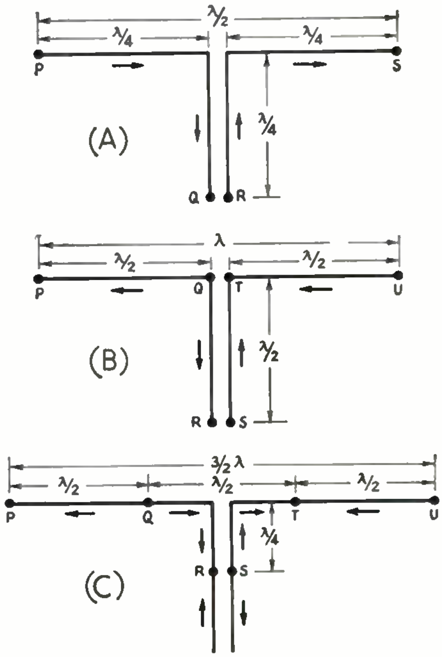


Fig. 3-32 — **Showing how the type of feed changes from current to voltage, with a center-fed antenna, on twice the fundamental frequency, and back to current feed on three times the fundamental. The same change occurs between all even and odd multiples.**

currents in the two half-wave sections of the antenna are therefore flowing in the same direction. That is, they are in the same phase.

With the above in mind, the direction of current flow in a 1½-wavelength antenna, Fig. 3-32C, should be easy to follow. The center half-wave section QT corresponds to the half-wave antenna in A. The currents in the end sections, PQ and TU, simply flow in the opposite direction to the current in QT. Thus the currents are out of phase in alternate half-wave sections.

The shift in voltage distribution between odd and even multiples of the fundamental frequency can be demonstrated by a similar method, making allowance for the fact that the voltage is maximum where the current is minimum, and vice versa. On all even multiples of the fundamental frequency there is a current minimum at the junction of the line and antenna, with center feed, because there is an integral number of half wavelengths in each side of the antenna. The voltage is maximum at the junction in such a case and we have voltage feed. Where the multiple of the fundamental is odd, there is always a current maximum at the junction of the transmission line and antenna, as demonstrated by A and C in Fig. 3-32. At these points the voltage is minimum and we therefore have current feed.

"Zepp" or End Feed

In the early days of short-wave communication an antenna consisting of a half-wave dipole end-fed through a ¼-wavelength resonant transmission line was developed as a trailing antenna for Zeppelin airships. In its utilization by amateurs, over the years, it has become popularly known as the "Zeppelin" or "Zepp" antenna. The term is now applied to practically any resonant antenna fed at the end by a two-wire transmission line.

The mechanism of end feed is perhaps somewhat difficult to visualize, since only one of the two wires of the transmission line is connected to the antenna while the other is simply left "free." The difficulty lies in the natural tendency to think in terms of current flow in ordinary electrical circuits, where it is necessary to have a complete loop between both terminals of the power source before any current can flow at all. But as, explained earlier, this limitation applies only to circuits in which the electromagnetic fields reach the most distant part of the circuit in a time interval that is negligible in comparison with the time of one cycle. When the circuit dimensions are comparable with the wavelength no such complete loop is necessary. The antenna itself is an example of an "open" circuit in which large currents can flow.

One way of looking at end feed is to consider the entire length of wire, including both antenna and feeder, as a single unit. For example, suppose

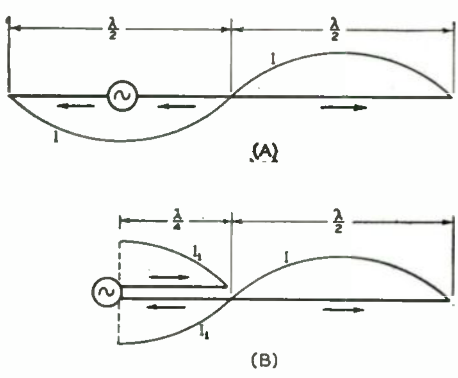


Fig. 3-33 — **Folded-antenna analogy of an end-fed antenna.**

we have a wire one wavelength long, as in Fig. 3-33A, fed at a current loop by a source of r.f. power. The current distribution will be as shown by the curves, with the assumed directions indicated by the arrows. If we now fold back the ¼-wavelength section to the left of the power source, as shown at B, the over-all current distribution will be similar, but the currents in the two wires of the folded section will be flowing in opposite directions. The amplitudes of the currents at any point along the folded-back portion will be equal in the two wires. The folded section therefore has become a ¼-wavelength transmission line, since the fields from the equal and opposite currents cancel. There is, however, nothing to prevent current from continuing to flow in the right-hand half-wavelength section, since there was current there before the left-hand section was folded.

This picture, although showing how power can flow from the transmission line to an antenna through end feed, lacks completeness. It does not take into account the fact that the current I_1 in the transmission line is greatly different from the current I in the antenna. A more basic viewpoint is the one already mentioned in Chapter Two: the current is caused by electromagnetic fields traveling along the wire and simply constitutes a measurable manifestation of those fields; the current does not cause the fields. From this standpoint the transmission-line conductors merely serve as "guides" for the fields so the electromagnetic energy will go where we want it to go. When the energy reaches the end of the transmission line it meets another guide, in the form of the antenna, and continues along it. However, the antenna is a different form of guide: it has a single conductor while the line has two; it has no provision for preventing radiation while the line is designed for that very purpose. This is simply another way of saying that the impedance of the antenna differs from that of the transmission line, so there will be reflection when the energy traveling along the line arrives at the antenna. We are then back on familiar ground, in that we have a transmission line terminated in an impedance different from its characteristic impedance.

Feeder Unbalance

With end feed, the currents in the two line wires do not balance exactly and there is therefore some radiation from the line. The reason for this is that the current at the end of the free wire is zero (neglecting a small charging current in the insulator at the end) while the current does not go to zero at the junction of the "active" line wire and the antenna. This is because not all the energy going into the antenna is reflected back from the far end, some being radiated; hence the outgoing and reflected currents at a node cannot completely cancel.

In addition to this unavoidable line radiation a further unbalance will occur if the antenna is not exactly resonant at the operating frequency. If the frequency is too high (antenna too long) the current node does not occur at the junction of the antenna and "live" feeder but moves out on the antenna. When the frequency is too low the node moves down the active feeder. Since the node on the free feeder has to occur at the end, either case is equivalent to shifting the position of the standing wave along one feeder wire but not the other. Representative cases are shown in Fig. 3-34. The farther off resonance the antenna is operated the greater the unbalance and the greater the line radiation. With center feed this unbalance does not occur, as is also shown in Fig. 3-34, because the system is symmetrical with respect to the line.

To avoid line radiation it is always best to feed the antenna at its center of symmetry. In the case of simple antennas for operation in several bands, this means that center feed should be used. End feed is required only when the antenna is operated on an even harmonic to obtain a desired directional characteristic, and then only when it must be used on more than one band. For single-band operation it is always possible to feed an even-harmonic antenna at a current loop in one of the half-wave sections nearest the center.

S.W.R. with Resonant Lines

When a resonant line is connected to the antenna at a current loop the standing-wave ratio can be estimated with good-enough accuracy with the aid of the curve in Fig. 2-23. Although the actual value of the radiation resistance, as measured at a current loop, will vary with the height of the antenna above ground, the theoretical values given in Fig. 2-23 will at least serve to establish whether the s.w.r. will be high or low.

With center feed the line will connect to the antenna at a current loop on the fundamental frequency and all odd multiples, as shown by Fig 3-31. At the fundamental frequency and usual antenna heights, the antenna resistance should lie between 50 and 100 ohms, so with a line having a characteristic impedance of 600 ohms the s.w.r. will be $Z_0/R_L = 600/50 = 12$ as one limit and $600/100 = 6$ as the other. On the third harmonic the theoretical resistance as given by Fig. 2-23 is 100 ohms, so the s.w.r. should be about 6. For 300-ohm line the s.w.r. can be expected to be between 3 and 6 on the antenna fundamental and about 3 on the third harmonic.

Considerably less is known about the impedances to be expected at voltage loops. Theoretical values are in the neighborhood of 5000 to 8000 ohms, depending on the antenna-conductor size. Such experimental figures as are available indicate a much lower order of resistance, with measurements and estimates running from 1000 to 4000 ohms. Something of the order of 2000 to 3000 ohms is probably near enough for estimating purposes. In any event, there will be some difference between end feed and center feed, since the current distribution on the antenna is different in these two cases at any given even multiple of the fundamental frequency. Also, the higher the multiple the lower the resistance at a voltage loop, so the s.w.r. can be expected to decrease when an antenna is operated at a high multiple of its fundamental frequency. Assuming 2400 ohms as a mean value, the s.w.r. would be 4 with a 600-ohm line and 8 with a 300-ohm line. However, considerable variation is to be expected.

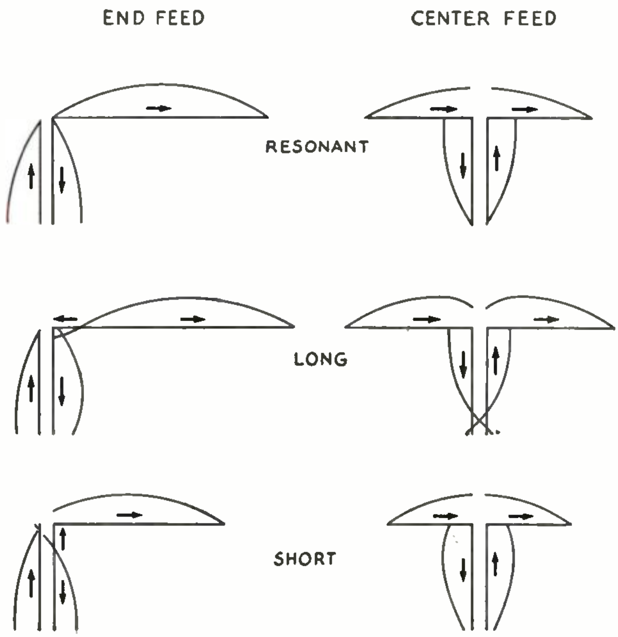


Fig. 3-34 — Effect of departure from exact resonance in antenna with end-fed and center-fed systems.

ANTENNA CURRENTS ON TRANSMISSION LINES

In any discussion of transmission-line operation it is always assumed that the two conductors carry equal and opposite currents throughout their length. This is an ideal condition that may or may not be realized in practice. In the average case the chances are rather good that the currents will *not* be balanced unless special precautions are taken. Whether the line is resonant or nonresonant has little to do with the situation.

Consider the half-wave antenna shown in Fig. 3-35 and assume that it is somehow fed by a source of power at its center, and that the instantaneous direction of current flow is as indicated by the arrows. In the neighborhood of the antenna is a group of conductors disposed in various ways with respect to the antenna itself. All of these conductors are in the field of the antenna and are therefore coupled to it. Consequently, when current flows in the antenna a voltage will be induced in each conductor. This causes a current flow determined by the induced voltage and the impedance of the conductor.

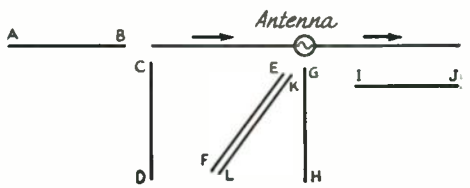


Fig. 3-35 — Coupling between antenna and conductors in the antenna's field.

The degree of coupling depends on the position of the conductor with respect to the antenna, assuming that all the conductors in the figure are the same length. The coupling between the antenna and conductor *IJ* is greater than in any other case, because *IJ* is close to and parallel with the antenna. Ideally, the coupling between conductor *GH* and the antenna is zero, because the voltage induced by current flowing in the left-hand side of the antenna is exactly balanced by a voltage of opposite polarity induced by the current flowing in the right-hand side. This is because the two currents are flowing in opposite directions with respect to *GH*. Complete cancellation of the induced voltages can occur, of course, only if the currents in the two halves of the antenna are symmetrically distributed with respect to the center of the antenna, and also only if every point along *GH* is equidistant from any two points along the antenna that are likewise equidistant from the center. This cannot be true of any of the other conductors shown, so a finite voltage will be induced in any conductor in the vicinity of the antenna except one perpendicular to the antenna at its center.

Now consider the two conductors *EF* and *KL*, which are parallel and very close together. Except for the negligible spacing between them, the two conductors lie in the same position with respect to the antenna. Therefore, identical voltages will be induced in both, and the resulting currents will be flowing *in the same direction in both conductors.* It is only a short step to visualizing conductors *EF* and *KL* as the two conductors of a section of transmission line in the vicinity of the antenna. Because of coupling to the antenna it is not only possible but *certain* that a voltage will be induced in the two conductors of the transmission line in parallel. The resulting current flow is in the same direction in both conductors, whereas the true transmission-line currents are always flowing in opposite directions at each point along the line. These "parallel" currents are of the same nature as the current in the antenna itself, and hence are called "antenna" currents on the line. They are responsible for a great deal of the radiation that takes place from transmission lines, regardless of whether the line is resonant or not.

Unbalanced Currents

Fig. 3-36 shows how the line and antenna currents can combine to produce resultant unbalanced currents in the line conductors. Although the "antenna" current in this example is smaller than the line current, it reduces the total current in one wire and increases it in the other, with the result that the current in Conductor No. 1 is only about half as large as the current in Conductor No. 2. The relative current amplitudes in the two conductors will be affected by the phase relationship between the antenna and line currents as well as their amplitudes.

The currents in Fig. 3-36 are shown on an instantaneous basis, but the standing waves of current along the line are also affected. In general, when there is an antenna current of appreciable amplitude on the line it will be found that not only are the line currents unbalanced but the apparent s.w.r. is different in each conductor, and that the loops and nodes of current in one wire do not occur at the same points as in the other wire. Under these conditions it is impossible to measure the true s.w.r.

It should be obvious from Fig. 3-35 that only in the case of a center-fed antenna can the coupling between the line and antenna be reduced to zero. There is always some such coupling when the antenna is end fed, so there is always the possibility that antenna currents of appreciable ampli-

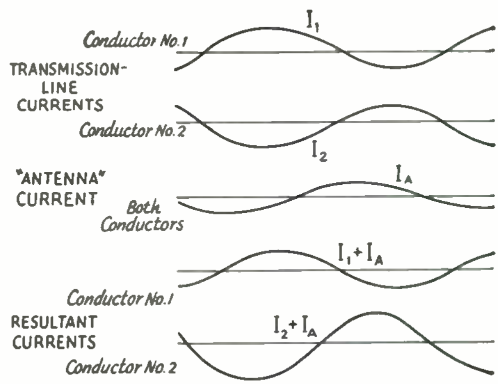


Fig. 3-36 — Instantaneous currents in line conductors when both "line" and "antenna" currents are present.

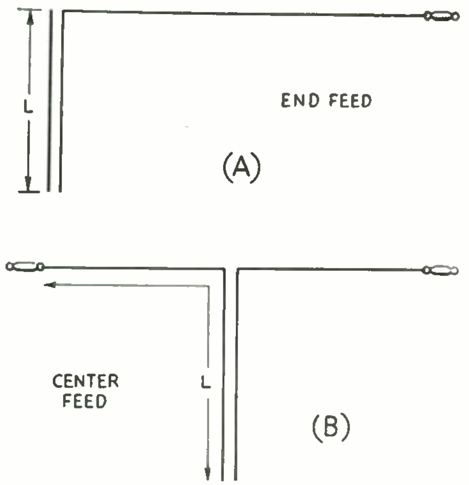


Fig. 3-37 — The important length for resonance to antenna currents coupled from the antenna to the line. In the center-fed system one side of the antenna is part of the "parallel"-resonant system.

tude will exist on the line, contributing further to the inherent line unbalance in the end-fed arrangement. But the center-fed system is also likely to have antenna-to-line coupling if the line is not brought off at right angles to the antenna for a distance of at least a half wavelength.

Detuning the Line

The antenna current flowing in the line as a result of voltage induced from the antenna will be small if the over-all circuit, considering the line simply as a single conductor, is not resonant at the operating frequency. The frequency (or frequencies) at which the system is resonant depends on the total length and whether the transmission line is grounded or not at the transmitter end.

If the line is connected to a coupling circuit that is not grounded, either directly or through a capacitance of more than a few $\mu\mu$fd., it is necessary to consider only the length of the antenna and line. In the end-fed arrangement, shown at A in Fig. 3-37, the line length, *L*, should not be an integral multiple or close to such a multiple of a half wavelength. In the center-fed system, Fig. 3-37B, the length of the line plus *one side* of the antenna should not be a multiple of a half wavelength. In this case the two halves of the antenna are simply in parallel so far as resonance for the induced "antenna" current on the line is concerned, because the line conductors themselves act in parallel. When the antenna is to be used in several bands, resonances of this type should be avoided at all frequencies to be used. Fig. 3-38 shows, as solid lines along the length scale, the lengths that avoid exact resonance on frequencies from 3.5 to 29.7 Mc. These are based on the usual antenna-length formulas, since the velocity factor of the line plays no part in establishing such resonances; it applies only to true transmission-line currents.

Whenever possible, it is best to choose line lengths, such as those indicated by the arrows, that fall midway in the nonresonant range. This is because the resonances are not extremely sharp. Working close to resonance, although not exactly on it, will allow an appreciable "antenna" current to flow even though it may not be as large as it would if the line were exactly resonant. For the same reason the line length should be chosen to fall in a range where there is a considerable distance between resonances. A length of 76 feet, for example, would be definitely less susceptible to resonance effects than a length of 96 feet.

The lengths shown in Fig. 3-38 are subject to some modification in practice. Transmission lines usually have bends, are at varying heights above ground, etc., all of which will modify the resonant frequency. It is advisable to check the system for resonance at and near all operating frequencies

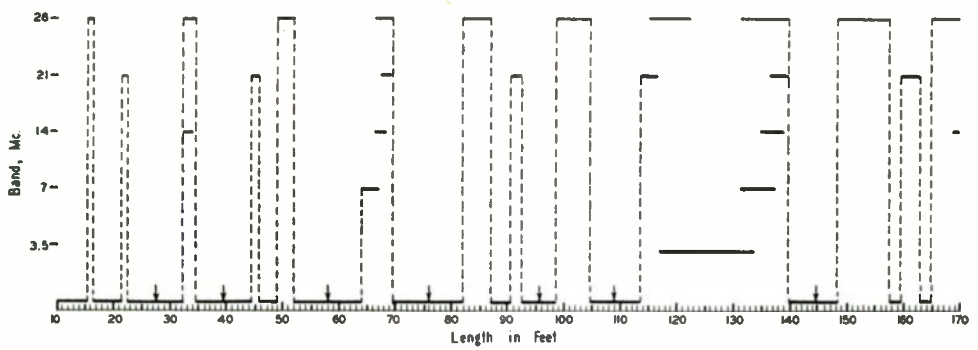


Fig. 3-38 — Lengths shown by solid lines along the horizontal axis avoid exact resonance at frequencies in all amateur bands from 3.5 to 29.7 Mc., in systems where the coupling apparatus is not grounded. Best operating lengths are at the center of the wider ranges, as shown by the arrows. These lengths correspond to *L* in Fig. 3-37.

before assuming that the line is safely detuned for antenna currents. This can be done by temporarily connecting the ends of the line together and coupling them through a small capacitance (not more than a few μμfd.) to a resonance indicator such as a grid-dip meter. Very short leads should be used between the meter and antenna. Fig. 3-39 shows the method. Once the resonance points are known it is a simple matter to prune the feeders to get as far away as possible from resonance at any frequency to be used.

Resonances in systems in which the coupling apparatus is grounded at the transmitter are not so easily predicted. The "ground" in such a case is usually the metal chassis of the transmitter itself, not actual ground. In the average amateur station it is not possible to get a connection to real ground without having a lead that is an appreciable fraction of a wavelength long. At the higher frequencies, and particularly in the v.h.f. region, the distance from the transmitter to ground may be a wavelength or more. Probably the best plan in such cases is to make the length *L* in Fig. 3-37 equal to a multiple of a half-wavelength. If the transmitter has fairly large capacitance to ground, a system of this length will be effectively detuned for the fundamental and all even harmonics when grounded to the transmitter at the coupling apparatus. However, the resonance frequencies will depend on the arrangement and constants of the coupling system even in such a case, and preferably should be checked by means of the grid-dip meter. If this test shows resonance at or near the operating frequency, alternative grounds (to a heating radiator, for example) should be tried until a combination is found that detunes the whole system.

As a general rule, it appears to be preferable to avoid grounding at the coupling circuit whenever the circumstances permit. The lengths given in Fig. 3-38 can then be applied, provided care is used to keep the capacitance between the coupling circuit and the transmitter chassis small. It is seldom necessary to ground an open-wire system. On the other hand, it is highly desirable to ground coaxial line at the transmitter to ensure that the power fed to the line will travel on the inside. With this type of line the presence of antenna currents can be checked by going along the line with a crystal-detector wavemeter, and then trying grounds as described above to reduce the current on the outside of the cable to the lowest possible value.

Unbalance in center-fed systems can arise even when the line is brought away at right angles to the antenna for a considerable distance. If both halves of the antenna are not symmetrically placed with respect to near-by conductors (such as power and telephone wires, downspouting, etc.) the antenna itself becomes unbalanced and the current distribution is different in the two halves. Because of this unbalance a voltage will be induced in the line even if the line is symmetrical with respect to the antenna.

It should be quite clear, from the mechanism that produces antenna currents on a transmission line, that such currents are entirely independent of the normal operation as a true transmission line. It does not matter whether the line is perfectly matched or is operated with a high standing-wave ratio. Nor does it matter what kind of line is used, air-insulated or solid-dielectric,

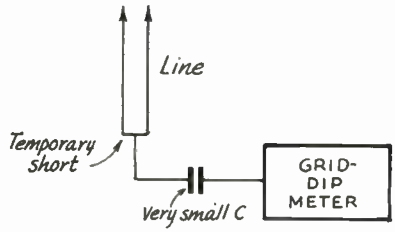


Fig. 3-39 — Using a grid-dip meter to check resonance of the antenna system for antenna currents on the transmission line.

parallel-conductor or coax. With coax the antenna current flows on the *outside* of the outer conductor, while the transmission-line currents flow only *inside* the line. This makes it possible to separate the two in the coax case, whereas no such separation is possible with parallel-wire lines. In every case, the antenna currents should be minimized by detuning the line if the line is to fulfill only its primary purpose of transferring power to the antenna.

NONRESONANT LINES

Operating the transmission line at a low standing-wave ratio requires that the line be terminated, at its output end, in a resistive load matching the characteristic impedance of the line as closely as possible. The problem can be approached from two standpoints: selecting a transmission line having a characteristic impedance that matches the antenna resistance at the point of connection; or transforming the antenna resistance to a value that matches the Z_0 of the line selected. The first approach is simple and direct, but its application is limited because the antenna impedance and line impedance are alike only in a few special cases. The second approach provides a good deal of freedom in that the antenna and line can be selected independently. Its disadvantage is that it is more complicated constructionally. Also, it frequently calls for a somewhat tedious routine of measurement and adjustment before the desired match is achieved.

Operating Considerations

As pointed out earlier in this chapter, most antenna systems show a marked change in resistance when going from the fundamental to multiples of the fundamental frequency. For this reason it is usually possible to match the line impedance only on one frequency. A matched antenna system is consequently a one-band affair, in most cases. It can, however, usually be operated over a fair frequency range in a given band. The frequency range over which the standing-wave ratio is low is determined by the impedance-*vs.*-frequency characteristic of the antenna. If the change in impedance is small for a given change in frequency, the s.w.r. will be low over a fairly

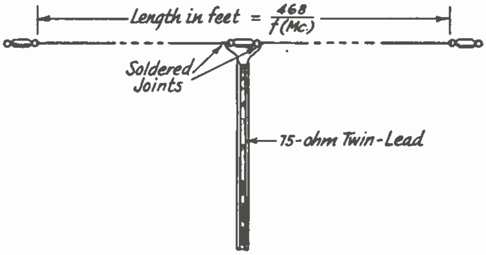


Fig. 3-40 — Half-wave dipole fed with 75-ohm Twin-Lead, giving a close match between antenna and line impedance. The leads in the "Y" from the end of the line to the ends of the center insulator should be as short as possible.

wide band of frequencies. However, if the impedance change is rapid (a sharply-resonant or high-Q antenna) the s.w.r. will also rise rapidly as the operating frequency is shifted to one side or the other of the frequency for which the antenna is resonant and for which the line is matched.

It must be emphasized that the line unbalance discussed in the preceding section is every bit as likely to occur with matched lines as with tuned lines. In fact, the presence of antenna currents on the line is likely to be more noticeable when the line is nonresonant, simply because a matched line operates with minimum line current for a given power input. This does not mean that the radiation from the line is any greater or any less. Actually, if the line is otherwise balanced, the radiation is entirely caused by the antenna currents on the line in either the resonant or nonresonant case. As much care should be used in selecting the length of a nonresonant line, to prevent resonance for parallel currents, as is used in choosing a length for a resonant line. The chart of Fig. 3-38 applies equally well. Equal care should also be used to bring the line away from the antenna at the angle with respect to the antenna conductors that results in the least coupling between antenna and line.

Antenna Resonance

Another point that needs emphasis in connection with matching the antenna to the line is that the impedance at the point where the line is connected must be a *pure resistance.* This means that the antenna system must be resonant at the frequency for which the line is to be matched. (Some types of long-wire antennas are exceptions, in that their input impedances are resistive over a wide band of frequencies. Such systems are themselves essentially nonresonant.) The higher the Q of the antenna system the more essential it is that exact resonance be established before an attempt is made to match the line. This is particularly true of the close-spaced parasitic arrays described in a later chapter. With simple dipole and harmonic antennas the tuning is not so critical and it is usually sufficient to cut the antenna to the length given by the appropriate formula in Chapter Two. The frequency should be selected to be at the center of the range of frequencies (which may be the entire width of an amateur band) over which the antenna is to be used.

Matching with 75-Ohm Twin-Lead

As discussed in Chapter Two, the impedance at the center of a resonant half-wave antenna at heights of the order of ¼ wavelength and more is resistive and is in the neighborhood of 70 ohms. This is rather closely matched by transmitting-type Twin-Lead having a characteristic impedance of 75 ohms. It is possible, therefore, to operate with a low s.w.r. using the arrangement shown in Fig. 3-40. No precautions are necessary beyond those already described in connection with antenna-to-line coupling.

This system is badly mismatched on *even* multiples of the fundamental frequency, since the feed in such cases is at a high-impedance point. However, it is reasonably well matched at *odd* multiples of the fundamental. For example, an antenna resonant near the low-frequency end of the 7-Mc. band will operate with a low s.w.r. over the 21-Mc. band (three times the fundamental).

The same method may be used to feed a harmonic antenna at any current loop along the wire. For lengths up to three or four wavelengths the s.w.r. should not exceed 2 to 1 if the antenna is ¼ or ½ wavelength above ground.

At the fundamental frequency the s.w.r. should not exceed 2 to 1 when the applied frequency is varied plus or minus 2% from the frequency of exact resonance. Such a variation corresponds approximately to the entire width of the 7-Mc. band, if the antenna is resonant at the center of the band. A wire antenna is assumed. Antennas having a greater ratio of diameter to length will have a lower change in s.w.r. with frequency.

Matching with 75-Ohm Coaxial Cable

Instead of using Twin-Lead as just described, the center of a half-wave dipole may be fed through 75-ohm coaxial cable such as RG-11/U, as shown in Fig. 3-41. The principle is exactly the same as with Twin-Lead, and the same remarks as to s.w.r. apply. However, there is a considerable practical difference between the two types of line.

With the parallel-conductor line the system is symmetrical, but with coaxial line it is inherently unbalanced.

Stated broadly, the unbalance with coaxial line is caused by the fact that the outer conductor is not coupled to the antenna in the same way as the inner conductor. The over-all result is that current will flow on the outside of the outer conductor in the simple arrangement shown in Fig. 3-41. The unbalance is rather small if the line diameter is very small compared with the length of the antenna, a condition that is met fairly well at the lower amateur frequencies. However, it is not negligible in the v.h.f. and u.h.f. range, nor should it be ignored at 28 Mc. The current that flows on the outside of the line because of this unbalance, it should be noted, is not the "antenna" current previously discussed. More exactly, it does not arise from the same type of coupling. The coupling pictured in Fig. 3-35 can still occur, *in addition.* However, the remedy is the same in both cases — the system must be detuned for currents on the outside of the line. This can be done by choosing one of the recommended lengths in Fig. 3-38 or by an actual resonance check by the method shown in Fig. 3-39.

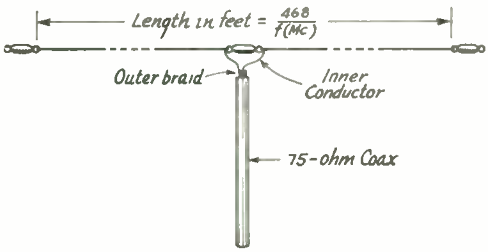


Fig. 3-41 — **Half-wave antenna fed with 75-ohm coaxial cable. The outside of the outer conductor of the line may be grounded for lightning protection.**

Other detuning and line-balancing methods particularly adapted to coaxial lines are shown in Fig. 3-42. The arrangement at A is a ¼-wave section of transmission line, shorted at one end and with its open end connected across the terminals of the antenna and regular transmission line. The connections to the inner and outer conductors are, however, reversed. This helps to restore physical symmetry at the antenna terminals and thus overcomes the unbalance that exists when only one cable is used. The shorted section of line is a very high impedance at the operating frequency and therefore has no effect on the coupling between the regular line and antenna. The length of the shorted section is dependent on the propagation factor of the particular type of line used. There should be no connection between the two cables except at the antenna terminals.

"Bazookas"

Two forms of "bazooka" are shown at B and C in Fig. 3-42. These are quarter-wave sections whose purpose is to detune the outside of the transmission line for the currents caused by lack of symmetry. A parallel-conductor section is shown at B and a coaxial section or **detuning sleeve** at C. In both cases the principal dielectric in the detuning section is air, so the length is calculated on the basis of a velocity factor of 0.95. In B, the second conductor is a piece of coax simply for convenience, since the inner and outer conductors are connected together; a length of tubing of the same diameter as the outer braid could be used with equal effect. The two conductors should be separated sufficiently so that the vinyl covering on the cable represents only a small proportion of the dielectric between them.

The diameter of the coaxial detuning sleeve in C should be fairly large compared with the diameter of the cable it surrounds. A diameter of two inches or so is satisfactory with half-inch cable. The sleeve should be symmetrically placed with respect to the center of the antenna so that it will be equally coupled to both sides. Otherwise a current will be induced from the antenna to the outside of the sleeve. This is particularly important at v.h.f. and u.h.f.

The detuning sections shown at B and C also prevent "antenna" currents of the type described earlier, provided the sections are symmetrical with respect to the antenna and provided the line beyond the detuning section does not pick up energy from the antenna's field. To avoid such pick-up it is well to use a line length, measured from the top of the sleeve to the transmitter, that is safely far away from resonance in the band on which the antenna is to be operated.

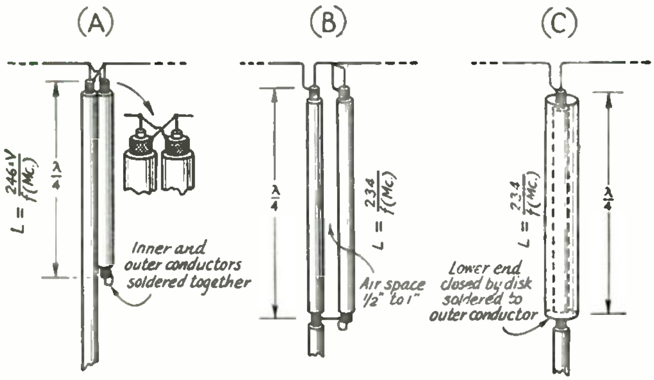


Fig. 3-42 — **Methods of balancing the termination when a coaxial cable is connected to a balanced antenna. The systems at B and C are two forms of "bazooka."**

In all of the balancing methods shown in Fig. 3-42 the quarter-wave section should be cut to be resonant at exactly the same frequency as the antenna itself. These sections tend to have a beneficial effect on the impedance-frequency characteristic of the system, because their reactance varies in the opposite direction to that of the antenna. For instance, if the operating frequency is slightly below resonance the antenna has capacitive reactance, but the quarter-wave sections or stubs have inductive reactance. Thus the reactances tend to cancel, which prevents the impedance from changing rapidly and helps maintain a low standing-wave ratio on the line over a band of frequencies.

Quarter-Wave Transformers

The impedance-transforming properties of a quarter-wave transmission line can be used to good advantage in matching the antenna impedance to the characteristic impedance of the line. As described earlier, the input impedance of a quarter-wave line terminated in a resistive impedance Z_R is

$$Z_S = \frac{Z_0^2}{Z_R}$$

Rearranging this equation gives

$$Z_0 = \sqrt{Z_R Z_S}$$

This means that any value of load impedance Z_R can be transformed into any desired value of impedance Z_S at the input terminals of a quarter-wave line, provided the line can be constructed to have a characteristic impedance Z_0 equal to the square root of the product of the two impedances. The factor that limits the range of impedances that can be matched by this method is the range of values for Z_0 that is physically realizable. The latter range is approximately 50 to 600 ohms.

Practically any type of line can be used for the matching section, including both air-insulated and solid-dielectric lines.

A common application of this type of matching section is in matching a half-wave antenna to a 600-ohm line, as shown in Fig. 3-43. Assuming that the antenna has a resistive impedance in the vicinity of 65 to 70 ohms, the required Z_0 of the matching section is approximately 200 ohms. A section of this type can be constructed of parallel tubing, from the data in Fig. 3-24. Using tubing of ½-inch diameter, the center-to-center spacing is approximately 1.3 inches for this characteristic impedance. In practice, provision should be made for changing the spacing between the matching-section conductors over a small range. The spacing can be varied, while checking the s.w.r. on the main transmission line, until the match is as nearly perfect as possible. It is to be expected that some adjustment will be necessary because the actual antenna resistance will be dependent on the height of the antenna.

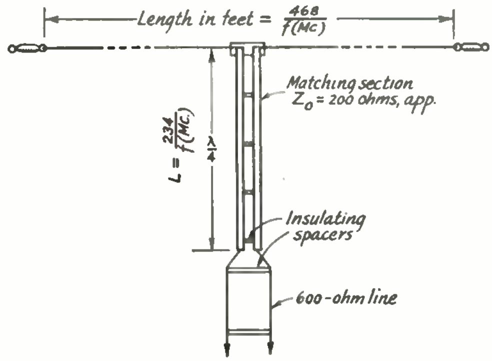


Fig. 3-43 — **Matching a half-wave antenna to a 600-ohm line through a quarter-wave linear transformer. This arrangement is popularly known as the "Q" matching system.**

Antennas operated on harmonics can be matched to a transmission line by this method provided the matching section is connected into the antenna at a current loop. Fig. 3-44 shows the required Z_0 of the quarter-wave matching section for a number of different open-wire transmission line impedances. This graph is based on the theoretical values of radiation resistance for long-wire antennas, and some modification must be anticipated in practice because of ground effects and the effect of radiation of energy from the antenna. Discrepancies of this sort between theory and practice can be compensated by changing the conductor spacing in the matching section to make the s.w.r. as low as possible.

Another application for the quarter-wave "linear transformer" is in matching the very low antenna impedances encountered in close-spaced directional arrays to a transmission line having a characteristic impedance of 300 to 600 ohms. The observed impedances at the antenna feed point in such cases range from about 8 to 20 ohms. A matching section having a Z_0 of 75 ohms is useful

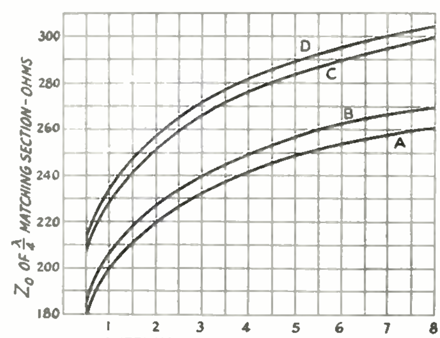


Fig. 3-44 — **Required Z_0 of quarter-wave linear transformers for matching at a current loop in harmonic antennas of various length. The curves are for the following main transmission-line impedances: A, 440 ohms; B, 470 ohms; C, 580 ohms; D, 600 ohms.**

with such arrays. The impedance at its input terminals will vary from approximately 700 ohms with an 8-ohm load to 280 ohms with a 20-ohm load. If the main transmission line has a characteristic impedance of 600 ohms, the s.w.r. will range from less than 1.2 with the 8-ohm load to 2.1 with the 20-ohm load. Since the additional line loss is practically negligible when the s.w.r. is 2 to 1, the 75-ohm section gives a quite adequate match over the range of impedances encountered in most such antenna arrays. Transmitting Twin-Lead is suitable for this application; such a short length is required that the loss in the matching

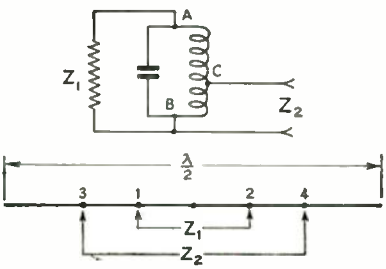


Fig. 3-45 — Impedance transformation with a resonant circuit, together with antenna analogy.

section should not exceed about 0.6 db. even though the s.w.r. in the matching section may be almost 10 to 1 in the extreme case.

Delta Match

Among the properties of a coil-and-condenser resonant circuit is that of transforming impedances. If a resistive impedance, Z_1 in Fig. 3-45, is connected across the outer terminals AB of a resonant LC circuit, the impedance Z_2 as viewed looking into another pair of terminals such as BC will also be resistive, but will have a different value depending on the mutual coupling between the parts of the coil associated with each pair of terminals. Z_2 will be less than Z_1 in the circuit shown, but this relationship will be reversed if Z_1 is connected across terminals BC and Z_2 is viewed from terminals AB.

A resonant antenna has similar properties. The impedance presented between any two points symmetrically placed with respect to the center of a half-wave antenna will depend on the distance between the points. The greater the separation the higher the value of impedance, up to the limiting value that exists between the open ends of the antenna. This is also suggested in Fig. 3-45. The impedance Z_1 between terminals *1* and *2* is lower than the impedance Z_2 between terminals *3* and *4*. Both impedances, however, are purely resistive if the antenna is resonant.

This principle, plus the fact that a "fanned" transmission line — i.e., one with continuously-increasing spacing — has a continuously-increasing value of characteristic impedance, is utilized in the delta matching system shown in Fig. 3-46. The center impedance of a half-wave dipole as a series-resonant circuit is too low to be matched directly by any practicable type of air-insulated parallel-conductor line. However, if the antenna is considered as a parallel-resonant circuit it is possible to find between two points a value of impedance that can be matched to such a line when a "fanned" section or delta is used to couple the line and antenna.

Available information on the delta match is based on experimental data for the case of a simple half-wave antenna coupled to a 600-ohm transmission line. The antenna length, L, should be based on the formula in Chapter Two, using the appropriate factor for the length/diameter ratio. The ends of the delta or "Y" should be attached at points equidistant from the center of the antenna, and the total distance, A, between them is given by

$$A \text{ (feet)} = \frac{118}{f \text{ (Mc.)}}$$

for frequencies up to and including the 28-Mc. band, and

$$A \text{ (feet)} = \frac{113}{f \text{ (Mc.)}}$$

for frequencies above 30 Mc. The length of the delta, B, is given by

$$B \text{ (feet)} = \frac{148}{f \text{ (Mc.)}}$$

These formulas are based on the assumption that the center impedance of the antenna is approximately 70 ohms, and will require modification if the actual impedance is very much different.

The delta match can be used for matching the driven element of a directive array to a transmission line, but if the impedance of the element

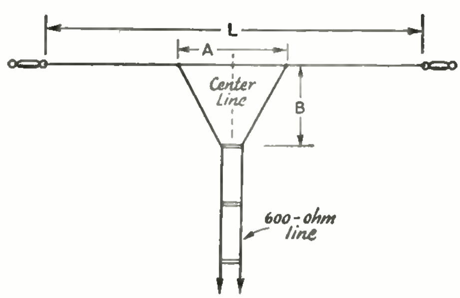


Fig. 3-46 — The "delta" matching system, applied to a half-wave antenna and 600-ohm line.

is low — as is frequently the case — the proper dimensions for A and B must be found by experiment. In general, the points at which the delta conductors are attached to the antenna will be closer together, for proper matching, as the antenna impedance becomes lower.

The delta match is somewhat awkward to adjust when the proper dimensions are unknown, because both the length and width of the delta must be varied. An additional disadvantage is that there is always some radiation from the delta. This is because the conductors are not close enough together throughout their length to meet the requirement that line spacing should be negligibly small in comparison with the wavelength.

The "T"-Match

A matching system that, like the delta, uses the principle of tapping along the antenna to obtain a desired impedance is the "T"-match shown in Fig. 3-47. However, in the "T"-match the conductors in the matching section are parallel to the antenna, a more convenient arrangement constructionally. The system has sufficient flexibility to be used with either a simple half-wave dipole or the driven element of a directive array, and can be adjusted to match the characteristic impedance of most of the commonly-used lines.

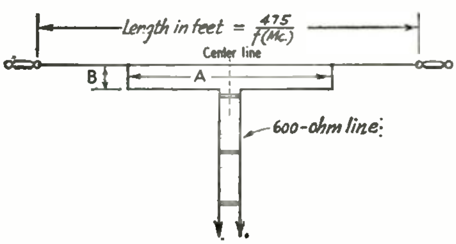


Fig. 3-47 — The "T" matching system, applied to a half-wave antenna and 600-ohm line.

For a dipole having an approximate impedance of 70 ohms, the matching-section dimensions for matching a 600-ohm line are given by the following formulas:

$$A \text{ (feet)} = \frac{180.5}{f \text{ (Mc.)}}$$

$$B \text{ (inches)} = \frac{114}{f \text{ (Mc.)}}$$

With an antenna element of different impedance, or for matching a line having a Z_0 other than 600 ohms, the matching-section dimensions can be determined experimentally. Typical dimensions for directional arrays are given in a later chapter.

This type of matching system is especially adaptable to antennas constructed of tubing. In such case the matching section also can be tubing of the same size and the end pieces can be metal straps whose positions can easily be changed for adjustment. With wire antennas the end of the transmission line and the center of the matching section can be supported from the center of the antenna by strips of insulating material having the same length as dimension B.

Folded Dipoles

In the diagram shown in Fig. 3-48, suppose for the moment that the upper conductor between points B and C is disconnected and removed. The system is then a simple center-fed dipole, and the direction of current flow along the antenna and line at a given instant is as given by the arrows. Then if the upper conductor between B and C is restored, the current in it will flow away from B and toward C, in accordance with the rule for reversal of direction in alternate half-wave sections along a wire. However, the fact that the second wire is "folded" makes the currents in the two conductors of the antenna flow in the *same* direction. Although the antenna physically resembles a transmission line, it is not actually a line but is merely two conductors in parallel. The connections at the ends of the two are assumed to be of negligible length.

For all practical purposes a half-wave dipole formed in this way has exactly the same properties as an ordinary dipole. Its directional properties and radiation resistance will be the same as those of an ordinary dipole. However, the transmission line is connected to only *one* of the conductors. It is therefore to be expected that the antenna will "look" different, in respect to its input impedance, as viewed by the line.

The effect on the impedance at the antenna input terminals can be visualized quite readily. The center impedance of the dipole *as a whole* is the same as the impedance of a single-conductor dipole — that is, approximately 70 ohms. A given amount of power will therefore cause a definite value of current, I. In the ordinary half-wave dipole this current flows at the junction of the line and antenna. In the folded dipole the same current also flows, but is equally divided between *two* conductors in parallel. The current in each conductor is therefore $I/2$. Consequently, the line "sees" a higher impedance because it is delivering the same power at only half the current. It is easy to show that the new value of impedance is equal to four times the impedance of a simple dipole. If more wires are added in parallel the current continues to divide equally between

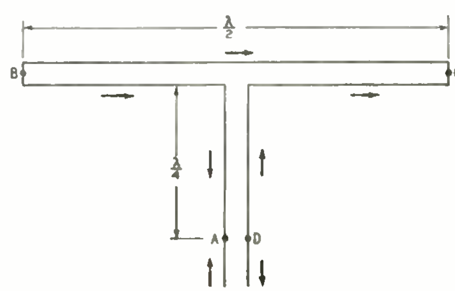


Fig. 3-48 — Direction of current flow in a folded dipole and associated transmission line.

them and the original impedance is raised by a factor N^2, where N is the total number of conductors in parallel.

This explanation is a simplified one based on the assumption that the conductors are close together and that all have the same diameter.

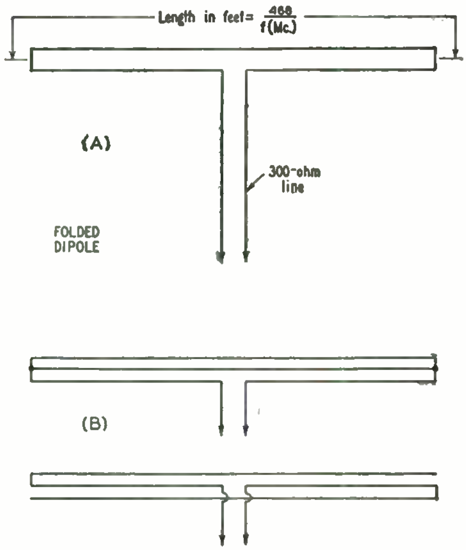


Fig. 3-49 — Two- and three-wire folded dipoles.

The folded dipole is frequently (but not necessarily) constructed with identical conductors paralleled. Fig. 3-49 shows a few common arrangements. The two-wire system at *A* is an especially useful one because the input impedance is so close to 300 ohms that it can be fed directly with 300-ohm Twin-Lead, without any other matching arrangement, and the line will operate with a very low standing-wave ratio. The antenna itself can be built like an open-wire line; that is, the two conductors can be held apart by regular feeder spreaders, the 6-inch size being quite suitable. Alternatively, it is possible to use 300-ohm ribbon for the antenna, in addition to using it for the transmission line. The length of the antenna does not change on this account, since the antenna section does not operate as a transmission line but simply as two wires in parallel.

Fig. 3-49B shows two ways of forming a 3-wire folded dipole. The input impedance of a 3-wire folded dipole is theoretically 9 times the dipole radiation resistance, and therefore offers a very good match for a 600-ohm open-wire line. The parallel- and series-connected arrangements shown in Fig. 3-49B are equivalent, and the choice between them is simply a matter of convenience. The parallel system is usually simpler, from the insulation standpoint.

The folded dipole has a much "flatter" impedance-*vs.*-frequency characteristic than a simple dipole. That is, the reactance varies rather slowly as the frequency is varied on either side of resonance. The result is that it is possible to operate over a considerably wider band of frequencies, while maintaining a low s.w.r. on the line, than with a simple dipole. This can be explained by the fact that, as seen by the line, the system is not only an antenna but also consists of two short-circuited quarter-wave transmission lines. As pointed out in connection with the "bazooka," the reactance of a quarter-wave shorted section varies in the opposite direction to the reactance of an antenna as the frequency is varied. Thus the two tend to cancel. The actual current in the folded antenna consists of two components; the first is the antenna current already described, and the second is a transmission-line current flowing in each quarter-wave section. The transmission-line current is substantially zero at the center of the antenna but is large at the ends. It does not cause radiation if the spacing between the two conductors is small in comparison with the wavelength.

Folded dipoles frequently are made of 300-ohm Twin-Lead as a matter of structural convenience. In this case the velocity factor for the material as an antenna is the same as for wire of any other type, since the antenna currents flow in the same direction on the two conductors. However, the velocity factor of the line does apply in the case of the transmission-line currents just described. As a result, a Twin-Lead folded dipole cut to the correct length as an antenna is not the correct length to give "quarter-wave" transmission-line sections on each side. Because of this the Twin-Lead folded dipole does not have quite as good an impedance-frequency characteristic as one made from separate wires.

This condition can be corrected by changing the electrical length of the line with the condensers shown in Fig. 3-50. The condensers make each side of the antenna equivalent to an open-circuited half-wavelength line for transmission-line type currents, reflecting the desired high impedance at the feed point. The capacitance required is easily calculated on the basis of 6.9 μμfd. per meter of operating wavelength. For

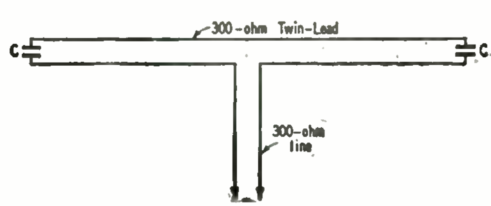


Fig. 3-50 — Condenser loading at ends of a Twin-Lead folded dipole to resonate the antenna for transmission-line currents. The capacitance required at *C* is 6.9 μμfd. per meter of operating wavelength.

7 Mc., for example (40 meters in round numbers), the capacitance should be 6.9 × 40 = 276 μμfd. The nearest available standard size of fixed condenser would be satisfactory. Mica condensers can be used at most amateur powers. The capacitance at the ends has no effect on the operation of the system as an antenna, because the impedance at the ends is high and even a small capacitance is a short-circuit for the antenna currents.

A folded dipole will not accept power at twice the fundamental frequency, or any even multiples of the fundamental. At such multiples the folded section simply acts like a continuation of the transmission line. No other current distribution is possible if the currents in the two conductors of the actual transmission line are to flow in opposite directions. It should be pointed out, however, that the antenna can actually be excited at the second harmonic if there is any capacitive coupling between the transmitter tank circuit and the line, and if the over-all system — considering the line conductors as being in parallel — is at or near resonance at the second harmonic. The significant length is L in Fig. 3-37B. To avoid radiation of even harmonics, therefore, the system should be detuned by proper choice of over-all length.

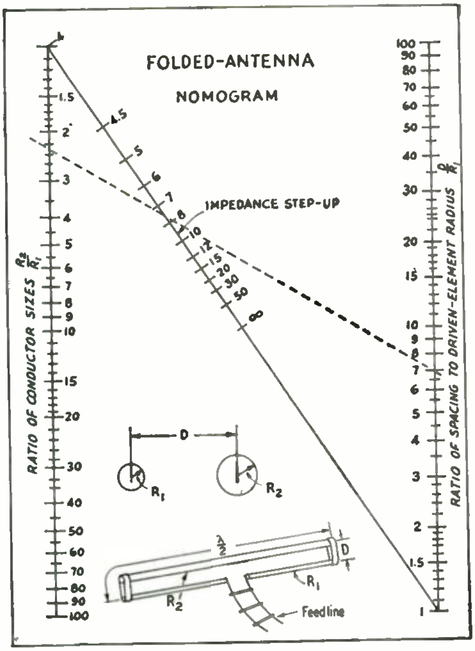


Fig. 3-51 — Nomogram for computing impedance step-up in a folded dipole having conductors of unequal diameter. The scale at the left is the ratio of conductor radii or diameters (driven element smaller). The right-hand scale is the ratio of center-to-center conductor spacing to driven-element radius. The diagonal scale is the impedance step-up ratio. A straightedge connecting two known quantities on the appropriate scales will intersect the third scale at the corresponding value.

On the third and other odd multiples of the fundamental the current distribution is correct for operation of the system as a folded antenna. Since the radiation resistance of a 3/2-wave antenna is not greatly different from that of a half-wave antenna, a folded dipole can be operated on its third harmonic with a low s.w.r. in 300-ohm line. A 7-Mc. folded dipole consequently can be used for the 21-Mc. band as well.

Multiwire and Unequal-Conductor Folded Dipoles

Two- and 3-wire folded dipoles take care of the most popular types of transmission lines (300- and 600-ohm lines) when the antenna is a half-wave radiator. Larger impedance ratios are frequently desirable when the folded dipole is used as the driven element in a directive array because the radiation resistance is frequently quite low. In such a case the number of conductors in the dipole can be increased to give the desired impedance step-up. Measurements have shown that the step-up is in quite good agreement with the N^2 rule up to at least four conductors in parallel, but there is evidence that it does not increase quite as rapidly if a large number of conductors is used. The application of this principle to a directional array is shown in a later chapter. A multiwire dipole should be constructed in the form of a cage, with the conductors equally spaced around the circumference of a circle.

Analysis of the impedance step-up when the conductors of a two-conductor folded dipole have different diameters is rather complicated. The nomogram of Fig. 3-51 is based on assumptions that appear to be borne out satisfactorily in practice. Note that both conductor diameter and spacing are involved. The broken line shows an example of the use of the chart. In this example R_2 is 2.3 times R_1, and the spacing, D, is 7 times R_1; this gives an impedance step-up of 7.5 times. The input impedance of an antenna element having a radiation resistance of 10 ohms, for instance, would be 75 ohms with this particular set of dimensions. R_1, R_2 and D may be measured in any units, since the significant quantities are simply ratios, so long as the same units are used throughout.

Matching Stubs

As explained earlier in this chapter, a terminated transmission line less than ¼ wavelength long has an input impedance that is both resistive and reactive. The equivalent circuit of the line input impedance can be formed either of resistance and reactance in series or resistance and reactance in parallel. Fig. 3-52A shows such a line terminated in a resistance less than its characteristic impedance, together with the equivalent circuits of the input impedance.

Depending on the line length, the resistive component, R_S, can have any value between

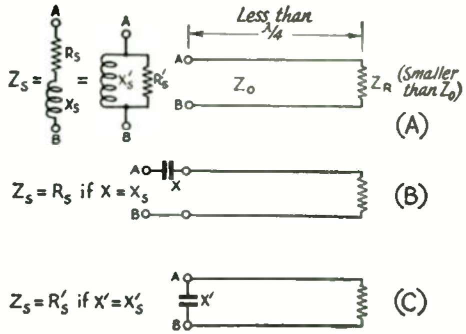


Fig. 3-52 — Circuit equivalent of a transmission line less than ¼ wavelength long terminated in a load resistance lower than Z_0, and methods of compensating for the reactive component of the input impedance.

Z_R (when the line has zero length) and Z_0^2/Z_R (when the line is exactly ¼ wave long). The same thing is true of R'_S. (R_S and R'_S do not, however, have the same values at the same line length, other than zero and ¼ wavelength.) With either equivalent there is some line length that will give a value of R_S or R'_S equal to the characteristic impedance of the line. In either case, also, reactance will accompany the resistance. If provision is made for canceling or "tuning out" this reactive part of the input impedance only the resistance will remain. Since this resistance is equal to Z_0 a transmission line of the same characteristic impedance connected to terminals *AB* will be properly matched.

Tuning out the reactance in the equivalent series circuit requires that a reactance of the same value as X_S, but of opposite kind, be inserted in series with the line. This is shown at B in Fig. 3-52. Tuning out the reactance in the equivalent parallel circuit requires that a reactance of the same value as X'_S but of the opposite kind be connected across terminals *AB*. This is shown at Fig. 3-52 C.

Corresponding circuits and relationships for the case where the line is terminated in a resistive

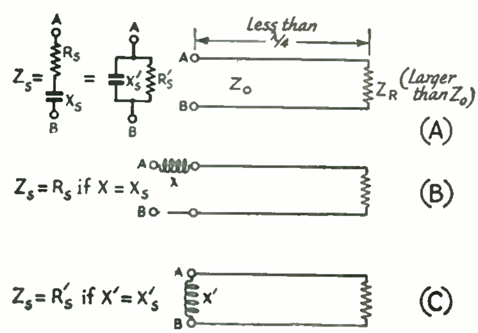


Fig. 3-53 — Circuit equivalent of a transmission line less than ¼ wavelength long terminated in a load resistance greater than Z_0, and methods of compensating for the reactive component of the input impedance.

load greater than its characteristic impedance are shown in Fig. 3-53. Aside from the fact that the reactances are of the opposite kind to those in Fig. 3-52, all of the foregoing remarks apply equally well.

In practice it is convenient to use the parallel equivalent circuit. The transmission line is simply connected to the load (which of course is usually a resonant antenna) and then a reactance of the proper value is connected across the line at the proper distance from the load. From this point back to the transmitter there are no standing waves on the line. A convenient type of reactance to use is a section of transmission line less than ¼ wave long, either open- or short-circuited depending on whether capacitive or inductive reactance is called for. Reactances formed from sections of transmission line are called **matching stubs**, and are designated as **open** or **closed** depending on whether the free end is open- or short-circuited. The two types are shown in Fig. 3-54.

The distance from the load to the stub (dimension *A* in Fig. 3-54) and the length of the stub, *B*, depend on the characteristic impedances of the

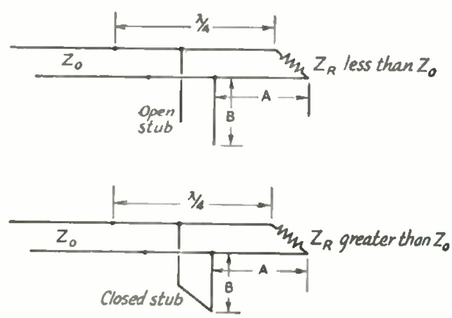


Fig. 3-54 — Use of open or closed stubs for canceling the parallel reactive component of input impedance.

line and stub and on the ratio of Z_R to Z_0. Since the ratio of Z_R to Z_0 is also the standing-wave ratio in the absence of matching, the dimensions are a function of the s.w.r. If the line and stub have the same Z_0, dimensions *A* and *B* are dependent on the s.w.r. only. Consequently, if the s.w.r. can be measured before the stub is installed, the stub can be properly located and its length determined even though the actual value of load impedance is not known.

Typical applications of matching stubs are shown in Fig. 3-55. From inspection of these drawings it will be recognized that when an antenna is fed at a current loop, as in Fig. 3-55A, Z_R is less than Z_0 (in the average case) and therefore calls for an open stub installed within the first quarter wavelength of line measured from the antenna. Voltage feed, as at B and C, corresponds to Z_R greater than Z_0 and therefore requires a closed stub.

Curves showing the length of the stub and its distance from the load, measured in terms of wavelength, are given in Figs. 3-56 and 3-57. These curves are based on the assumption that the load is a pure resistance, and that the characteristic impedances of both the line and stub are identical. The equations from which the curves are constructed are

$$\tan A = \sqrt{S.W.R.}$$

$$\cot B = \frac{S.W.R. - 1}{\sqrt{S.W.R.}}$$

for the closed stub when Z_R is greater than Z_0 and

$$\cot A = \sqrt{S.W.R.}$$

$$\tan B = \frac{S.W.R. - 1}{\sqrt{S.W.R.}}$$

for the open stub when Z_R is less than Z_0. In these equations the lengths A and B must be expressed in electrical degrees. The formula for converting length in wavelengths to electrical degrees is

$$\text{Length (degrees)} = 360 \times \text{Length in wavelengths}$$

In using the curves or the above equations it must be remembered that the wavelength along the line is not the same as in free space. If an open-wire line is used the velocity factor of 0.975 will apply. When solid-dielectric line such as Twin-Lead is used the free-space wavelength as given by the curves must be multiplied by the appropriate velocity factor to obtain the actual length of A and B. The formula for the actual length is

$$A\text{ (feet)} = \frac{985 \times V}{f\text{ (Mc.)}} \times A\text{ (wavelengths)}$$

$$B\text{ (feet)} = \frac{985 \times V}{f\text{ (Mc.)}} \times B\text{ (wavelengths)}$$

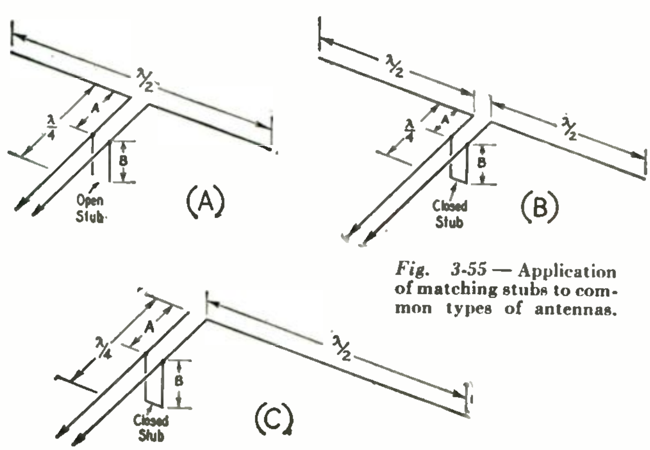


Fig. 3-55 — Application of matching stubs to common types of antennas.

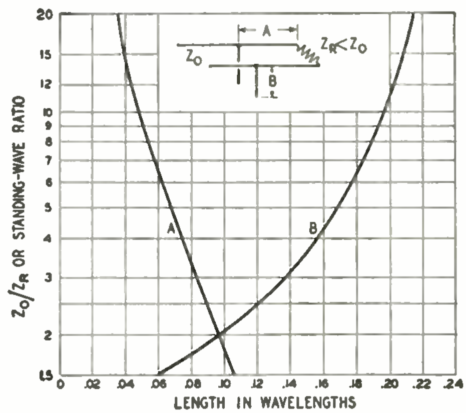


Fig. 3-56 — Position and length of open stub as a function of Z_0/Z_R or standing-wave ratio.

where V is the velocity factor of the line of which sections A and B are constructed.

Although the curves and equations do not apply when the characteristic impedances of the line and stub are not the same, this does not mean that the line cannot be matched under such conditions. With a few exceptions, it is readily possible to match even if the characteristic impedances of section A, the stub, and the line all are different. The exceptions arise when the Z_0 of section A is considerably different from that of the line, and when, at the same time, Z_R is not greatly different from the Z_0 of section A. In such a case it may be impossible to obtain a resistive component of the input impedance of section A that will match the Z_0 of the main transmission line. This cannot happen if section A is simply a continuation of the main transmission line to the load.

On the other hand, the stub itself can have any desired characteristic impedance if its length is chosen so that it has the proper value of reactance. Once the length is determined for a particular case from Figs. 3-56 and 3-57 the corresponding value of reactance can be found from Fig. 3-12 and a new length found from the same figure for a stub of the desired Z_0. For example, suppose that it is determined from Fig. 3-56 that the length of an open stub is 0.16 wavelength when the stub and line have the same characteristic impedance. Assume that this Z_0 is 600 ohms. From Fig. 3-12, X/Z_0 for an open-circuited transmission line 0.16 wavelength long (57.6 degrees) is 0.65, so $X = 0.65 \times 600 = 390$ ohms. If we want to make the stub from 300-ohm line, $X/Z_0 = 390/300 = 1.3$. The value 1.3 on the open-circuited line curve corresponds to a length of 37.5 degrees or 0.104 wavelength. The *actual* length of the 300-ohm Twin-Lead stub would be $0.104 \times V = 0.104 \times 0.82 = 0.085$ wavelength.

Other Stub Positions

As a general rule, matching stubs will be installed less than ¼ wavelength from the load, since this reduces the standing-wave ratio along the greatest length of transmission line. However, the stub may be connected farther from the load, if necessary. The curves of Figs. 3-56 and 3-57 may be used equally well. It is only necessary to remember that at every quarter-wave point along the line the input impedance is the reciprocal of the impedance at a point ¼ wavelength nearer the load. Referring to Fig. 3-54, if Z_R is less than Z_0 the impedance ¼ wave from the load will be *higher* than Z_0 and will equal Z_0^2/Z_R. It is therefore possible to consider this ¼-wave point as the end of a line terminated in a load Z_0^2/Z_R, and then use the curves of Fig. 3-57, adding 0.25 to the value found for dimension A. The length of the open stub is given directly by Fig. 3-57. Similarly, if Z_R is greater than Z_0 Fig. 3-56 can be used, adding 0.25 wavelength to dimension A as given by the curve.

Whether Z_R is larger or smaller than Z_0 can be determined by measuring the current in the line (before the stub is installed) over a length of ¼ wavelength from the load. If the current at the load is higher than at a point ¼ wavelength away, Z_R is less than Z_0; if the load current is smaller, Z_R is greater than Z_0.

In all of the discussion of matching stubs it has been assumed that the load is a pure resistance. This is the most desirable condition, since the antenna that represents the load preferably should be tuned to resonance before any attempt is made to match the line. Nevertheless, matching stubs can be used even when the load is considerably reactive. A reactive load simply means that the loops and nodes of the standing waves of voltage and current along the line do not occur at integral multiples of ¼ wavelength from the load. To use the curves of Figs. 3-56 and 3-57 it is only necessary to find a point along the line at which a current loop or node occurs. Then Fig. 3-56 gives the stub length and distance *toward the transmitter* from a current loop. Fig. 3-57 gives the stub length and distance *toward the transmitter* from a current node.

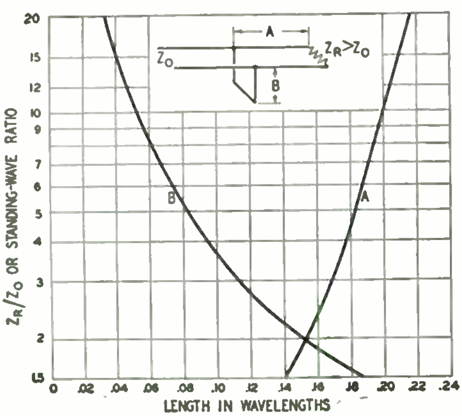


Fig. 3-57 — Position and length of closed stub as a function of Z_R/Z_0 or standing-wave ratio.

In using matching stubs it should be noted that the length and location of the stub should be based on the s.w.r. *at the load.* If the line is long and has fairly high losses, measuring the s.w.r. at the input end will not give the true value of the s.w.r. at the load. This point was discussed earlier in this chapter in the section on attenuation.

Stubs on Coaxial Lines

The principles outlined in the preceding section apply with equal force to coaxial lines. The construction is, of course, different. Also, it is not feasible to measure s.w.r. along a solid-dielectric coaxial cable and still keep the cable in one piece. The ratio Z_R/Z_0 or Z_0/Z_R must be determined by other means, such as direct measurement of Z_R.

The coaxial cases corresponding to the open-wire cases shown in Fig. 3-54 are given in Fig. 3-58. The curves of Figs. 3-56 and 3-57 may be used to determine the dimensions A and B. In a practical installation the junction of the transmission line and stub would be a Tee connector.

If it is impossible to measure either the load resistance or s.w.r., an impedance match can be brought about by using two closed stubs, spaced about ¼ wavelength apart, provided the lengths of the stubs are adjustable. The latter is hardly feasible with solid-dielectric lines used as stubs, but is practicable at u.h.f. where a shorting piston can be moved along inside an air-insulated line. Double-stub impedance matching has had little or no application in amateur work at the present writing.

Matching Sections

If the three antenna systems in Fig. 3-55 are redrawn in somewhat different fashion, as shown in Fig. 3-59, there results a system that differs in no consequential way from the matching stubs previously described, but in which the stub formed by A and B together is called a "quarter-wave matching section." The justification for this is that a quarter-wave section of line — especially one with a short-circuit at one end — is similar to a resonant circuit, as described earlier in this chapter. It is therefore possible to use it to transform impedances as described in connection with Fig. 3-45, by tapping at the appropriate point along the line.

The usual procedure recommended for adjusting such a system is first to resonate the antenna independently to the operating frequency, then resonate the matching section before it is connected to either line or antenna. (In the case of the open-ended section at A in Fig. 3-59, one end may be shorted temporarily while the section is adjusted to resonance.) Then the section and antenna are connected together, and finally the line is tapped along the matching section until a position is found where standing waves on the transmission line disappear.

In many cases this procedure does not result in the minimum possible s.w.r. on the line. The

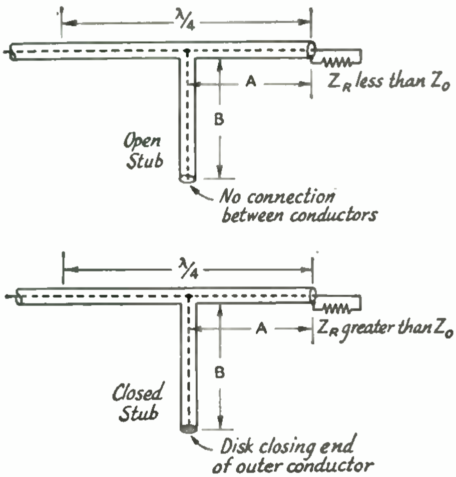


Fig. 3-58 — Open and closed stubs on coaxial lines.

reason can be found on examination of Figs. 3-56 and 3-57. The sum of A and B is not even approximately a quarter wavelength when the original s.w.r. is fairly low. At an s.w.r. of 2, for example, it is approximately 0.3 wavelength for the closed stub and 0.2 wavelength for the open stub. At an s.w.r. of 5 it is slightly over 0.26 wavelength for the closed stub and slightly under 0.24 wavelength for the open stub. Above an s.w.r. of 10 the length is substantially ¼ wavelength in either case. In other words, the proper length for the matching section is ¼ wavelength *only* when the s.w.r. is quite high to start with. A small correction in matching-section length is necessary, for minimum s.w.r. on the line, even when the original s.w.r. is of the order of 10 or higher.

This peculiarity of tuning is not unique with "quarter-wave" transmission lines, but applies also in the case of ordinary LC circuits. It arises because there are three ways in which "resonance" in a parallel circuit can be specified. When the Q of the circuit is at all high there is no significant difference between the three, but if the Q is low the difference may be considerable. When the s.w.r. is low the Q of the matching section also is low.

Figs. 3-56 and 3-57 give design data for matching sections, A being the distance from the antenna to the point at which the line is connected, and $A + B$ being the total length of the matching section. The curves apply only in the case where the characteristic impedances of the matching section and transmission line are the same. Equations are available for the case where the matching section has a different Z_0 than the line, but are somewhat complicated and will not be given here, in view of the fact that it is generally possible to make the line and matching section similar in construction.

When experimental adjustments are made in the length of a matching section, a closed section is more convenient than an open section, particularly if the matching section is constructed of wire. This is because it is easy to construct a "shorting bar" — which actually may be a wire the same length as the line spacing, and provided with clips for making contact to the line — and slide it along the matching section. About the only means available for adjusting the length of an open section is to clip off short lengths from the open end. This obviously does not allow much latitude in finding the optimum length. An open section can always be made into a closed section by extending its length ¼ wavelength, so that the over-all length becomes approximately ½ wavelength. (A ¼-wave shorted transmission line has very high input impedance and so can be connected across the open end of the matching section without affecting its operation.) Adjustments can then be made by moving the shorting bar, which in effect changes the length of the formerly open section because the length of the added quarter-wave stub remains the same so long as the operating frequency is not changed.

Matching with Lumped Constants

It was pointed out earlier that the purpose of a matching stub is to cancel the reactive component of line impedance at the point of connection. In other words, the stub is simply a reactance of the proper kind and value shunted across the line. It does not matter what physical shape this reactance takes. It can be a section of transmission line or a "lumped" inductance or capacitance as desired. Fig. 3-60A shows the case of a line terminated in a load less than its characteristic impedance, calling for an open (capacitive) stub for impedance matching. A condenser having the same value of reactance can be used just as well, as shown in Fig. 3-60B. There are cases where, from an installation standpoint, it may be considerably more convenient to use a condenser in place of a stub. Furthermore, if a variable condenser is used for X_C it becomes possible to adjust the capacitance to the exact value required.

The proper value of capacitance may be determined with the aid of Figs. 3-56 and 3-12. As an example, suppose that the antenna is a close-spaced array fed by a 300-ohm line, and that

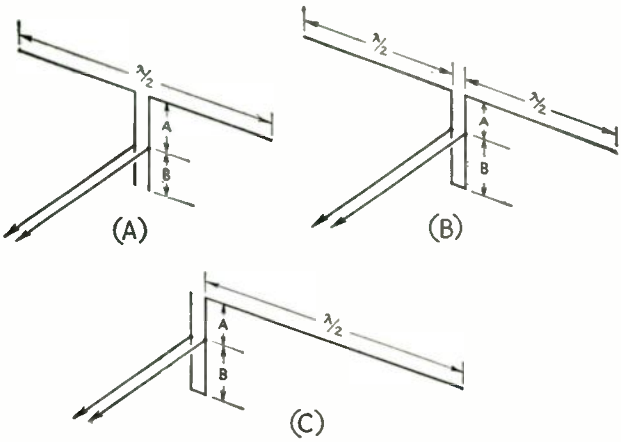


Fig. 3-59—Application of matching sections to common antenna types.

the standing-wave ratio at the load has been determined to be 15 to 1. From Fig. 3-56 dimension A is 0.04 wavelength and B is 0.206 wavelength. From Fig. 3-12, 0.206 wavelength corresponds to a value of X_C/Z_0 of 0.28, so $X_C = 0.28 \times 300 = 84$ ohms. If the frequency is 14.2 Mc., for instance, 84 ohms corresponds to a capacitance of 134 μμfd. A 150-μμfd. variable condenser connected across the line 0.04 wavelength from the antenna terminals would provide ample adjustment range. The r.m.s. voltage across the condenser is $E = \sqrt{PZ_0}$, and for 500 watts, for example, would be $E = \sqrt{500 \times 300} = 386$ volts. The peak voltage is 1.41 times the r.m.s. value, or 545 volts. With 100 per cent amplitude modulation the peak would be twice as large.

The impedance-transforming properties of a coil-and-condenser resonant circuit also can be used for matching the load to the transmission line. If the line is disconnected from terminals *1-2* in Fig. 3-60C, we have an ordinary LC circuit with resistance in series with the inductance. By proper choice of values for X_L and X_C, the impedance between terminals *1-2* can be made purely resistive and of a value equal to the characteristic impedance of the transmission line. The reactances required to meet this condition are given by

$$X_L = Z_R\sqrt{\frac{Z_0}{Z_R} - 1} \text{ ohms}$$

and

$$X_C = \frac{Z_0}{\sqrt{\frac{Z_0}{Z_R} - 1}} \text{ ohms}$$

X_L and X_C do not have the same value unless the Q of the matching circuit is high. This is not generally the case because an extremely large ratio of Z_0 to Z_R (that is, the s.w.r. prior to matching) is required for high Q. The s.w.r. has to be practically 100 to 1 for a Q of 10, for example.

As shown in Fig. 3-60C, the inductive reactance should be divided into two equal parts, one in each side of the circuit, when a balanced line is used. Each coil, in other words, should have half the total inductance, provided they are mounted so there is no mutual inductance. The reactance values can be converted to inductance and capacitance by the usual formulas or charts.

Both systems described above can be used for matching the line to a load impedance higher than the characteristic impedance. Thus an inductance can be shunted across the line instead of the condenser in Fig. 3-60B, at the point where a closed stub would be required. The same procedure is used for finding the required reactance, but Fig. 3-57 is used instead of Fig. 3-56, and the X_L/Z_0 curve is used in Fig. 3-12. In the circuit of Fig. 3-60C the inductive and capacitive reactances are interchanged, with one-half the total capacitive reactance (twice the required capacitance) in series with each side of the line

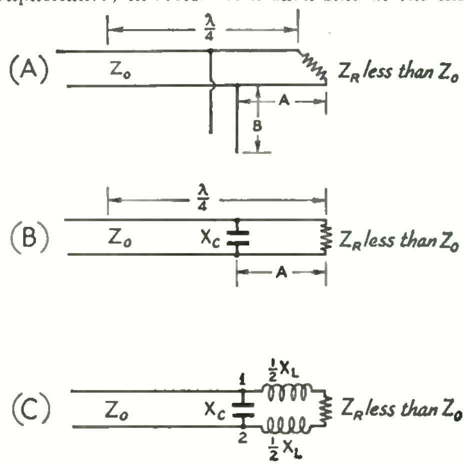


Fig. 3-60—The open stub (A) may be replaced by a condenser of the same reactance (B) connected at the same point on the line. A matching circuit using lumped constants is shown at C.

and with the inductive reactance across the line. The formulas given above are likewise interchanged. In practice, however, this system is most likely to be used when the load impedance is less than the characteristic impedance of the line.

Inductive Coupling

It is possible to match impedances by using inductively-coupled resonant circuits connected to the line and antenna. Although such a method has little application in most antenna systems, it is quite useful when a line is to be matched to a rotatable beam antenna. In such a case it avoids the necessity for any direct metallic connection between the antenna and line, and so allows continuous rotation without slip rings or other forms of contactor.

In its usual form the coupling is between two metallic loops or rings of rather large diameter constructed of copper tubing. These constitute the inductances in the primary (line) and secondary (antenna) circuits. The reactance of the secondary ring is tuned out by inserting condensers C_2 and C_3 in Fig. 3-61, having a total reactance equal to that of the secondary loop. This permits the antenna or driven element to be the proper length for resonance at the operating frequency.

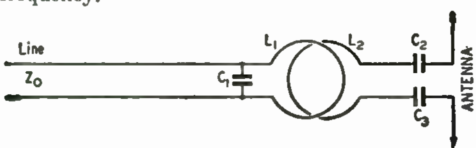


Fig. 3-61 — **Matching with inductively-coupled resonant circuits. This method is useful with low-impedance rotatable directive arrays.**

To obtain sufficient coupling between the two circuits it is necessary that their Qs be fairly high. The impedance of a center-fed antenna element is always so low that the secondary circuit must be series resonant to operate at all, and the lower the antenna impedance the better from the standpoint of attainable Q. On the other hand, the design of the primary circuit depends on the characteristic impedance of the transmission line and the reactance that it is possible to obtain in L_1. A single-turn loop approximately one foot in diameter will have a reactance in the neighborhood of 75 ohms at 14 Mc. In a parallel-resonant circuit terminating a 600-ohm line this will give a Q of roughly $600/75 = 8$, when resonated by a condenser of the same reactance. This is a satisfactory value for adequate coupling, particularly if the same size of loop is used in the secondary circuit with an antenna having an impedance of 20 ohms or less. To use a series-resonant circuit in the primary side it would be necessary that the transmission line be 50- or 75-ohm coaxial cable or Twin-Lead, and that the reactance of the primary loop be in the neighborhood of 500 ohms. This would require a coil of several turns, which is somewhat undesirable from a constructional standpoint. Consequently, the system is chiefly used with an open-wire transmission line.

C_1, C_2 and C_3 should be adjustable. As a preliminary step, L_1C_1 should be tuned to resonance at the operating frequency with the line disconnected and L_2 removed. (Resonance may be determined with a calibrated grid-dip meter or other suitable equipment.) $L_2C_2C_3$ also should be tuned to resonance, with the antenna disconnected and replaced by a short-circuit, and with L_1 removed. C_2 and C_3 should be adjusted to have the same capacitance at resonance. The antenna should likewise be resonated with $L_2C_2C_3$ removed. The complete system may then be connected up as shown in Fig. 3-61, leaving the condenser settings untouched. Provision should be made for varying the coupling between L_1 and L_2. Starting with loose coupling, the transmitter is then connected to the transmission line and the coupling between L_1 and L_2 increased until the desired power output is obtained. The s.w.r. should be measured, and if not close to unity small adjustments may be made to the variable condensers to make it as low as possible.

In a rotating system using this method, every effort should be made to keep the spacing between L_1 and L_2 constant as the antenna is turned. If there is any variation in spacing the power input to the antenna will depend on its position.

It is possible to dispense with condensers C_2 and C_3 and simply connect L_2 directly to the antenna terminals. This detunes the driven element, but the secondary circuit as a whole can be brought back to resonance by shortening the antenna until it shows a capacitive reactance equal to the inductive reactance of L_2. Alternatively, the antenna may be left at the normal resonant length, in which case reactance will be coupled into the primary circuit. This may be tuned out, within limits, by appropriate adjustment of C_1. However, this method will, in general, require closer coupling between L_1 and L_2; also, it is likely to be more sensitive to changes in coupling as the antenna is rotated, and the proper circuit constants are difficult to determine.

With loops as described above, approximately 150 $\mu\mu$fd. is required at C_1 and 300 $\mu\mu$fd. each at C_2 and C_3, for 14-Mc. operation. Loops of the same size at 28 Mc. will require half these values of capacitance.

● ADJUSTMENT OF MATCHING SYSTEMS

When a nonresonant line is to be used the choice of type of line and the method of matching is largely an individual matter; that is, there is no one system that is "best" for everybody. A study of the foregoing material will soon show that some systems are usable with certain antenna types but not with others; direct matching by using a line having a characteristic impedance equal to the antenna impedance, for example, is useful only with one type of antenna and a few types of line. In some systems there is little opportunity for adjustment; in others a wide range of adjustments is possible. In general, it is better to choose a system that permits adjustment if the antenna impedance is not known fairly accurately beforehand.

In the adjustment of any matching system there are two separate operations that must be performed. The first is that of resonating the antenna so that the load presented to the transmission line is purely resistive. The second is transforming the antenna impedance to the value that matches the line impedance. Both operations require the use of measuring equipment of one type or another.

Resonating the Antenna

The resonant frequency of an antenna can be determined by methods similar to those used with ordinary tuned circuits. One method, for example, is to use a grid-dip meter coupled to one end through a small capacitance, as already suggested in Fig. 3-39, for checking feeder resonance. This scheme is often not practicable, since it may not be possible to reach the end of the antenna after it is erected. It is always best to check the frequency with the antenna in its actual operating position.

A better method makes use of the fact that nodes and loops in the standing waves on a mismatched transmission line will occur at points that are exact multiples of ¼ wavelength from the end of the line only when the line is terminated in a pure resistance. A standing-wave indicator will be required, but it will be necessary anyway if the termination is to be adjusted for minimum s.w.r. The indicator must be one that will respond to current (or voltage) — not, like the "Micro-Match," to reflected power alone. It is difficult to check standing waves on coaxial and 75-ohm parallel-conductor lines, but for the purpose of measuring the resonant frequency of the antenna a temporary line (such as 300-ohm Twin-Lead) can be substituted.

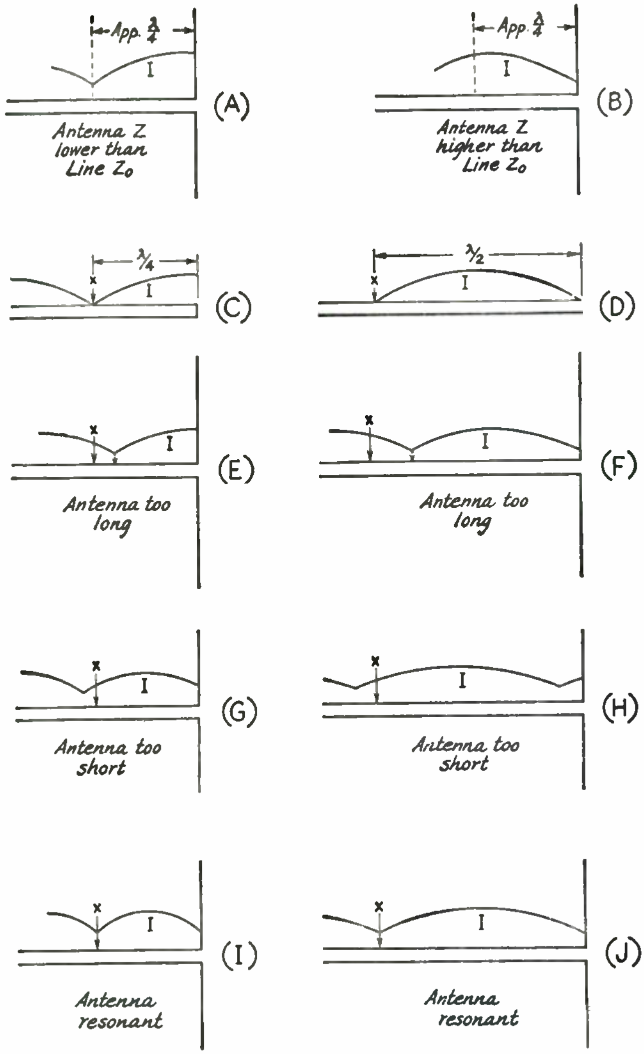


Fig. 3-62 — Adjusting an antenna to resonance by location of current nulls on the transmission line.

As a preliminary, it is desirable to know whether the antenna impedance is higher or lower than the line impedance. If this is not known it can be determined very easily by connecting the line to the antenna, supplying enough power to give a suitable reading on the standing-wave indicator, and checking the current along the line starting from the antenna. As the indicator is moved away from the antenna the current will either increase or decrease (except if the antenna happens to match the line), as shown in Fig. 3-62 at A and B, until a point approximately ¼ wavelength from the antenna is reached. If the current *decreases* moving away from the antenna, the antenna impedance is lower than the characteristic impedance of the line; if the current *increases*, the antenna impedance is higher.

The next step is to disconnect the line from the antenna and if the antenna impedance (Z_R) was lower than Z_0, short circuit the line as in Fig. 3-62C. In the opposite case, the end of the line should be left open, as at D. Then (with reduced power) measure the current along the line and locate the first current node, denoted by X in both drawings. With the shorted line, this locates a point ¼ wavelength from the end with high accuracy; with the open line, the distance is ½ wavelength. The node is used in preference to a loop in the latter case because its position can be determined much more accurately. This measurement takes into account the velocity factor of the line and so requires no correction. Mark the position of the node for future reference.

Next, reconnect the line to the antenna (removing the short, if used) and again measure the current starting from the antenna. The current at the node will not be as small as it was with the transmission line alone, but the node will be easily identifiable. If it has shifted toward the antenna, as in E and F, the antenna is too long; if the shift is away from the antenna the antenna is too short, as in G and H. After appropriate adjustments in antenna length have been made it should be possible to make the node coincide with point X, as shown at I and J. This indicates that the antenna is resonant at the operating frequency.

Although the description above and the drawings are in terms of the first current node going away from the antenna, it should be obvious that any current node along the line may be used. The standing-wave measurements can, in fact, be made at any convenient point along the line, just so long as a node can be located. It is not even necessary to know the relationship between Z_R and Z_0, since a measurement can first be made with the antenna connected, a node located, and the transmission line tried both open and short-circuited to see which brings a node closest to the one found with the antenna connected. The remainder of the procedure is the same. Also, a standing-wave indicator that responds to voltage instead of current may be used; the only difference is that a voltage node occurs ¼ wavelength from the antenna when Z_R is *higher* than Z_0, and ½ wavelength away when Z_R is *lower* than Z_0.

Adjusting the Termination

When a transmission line is to be matched to a simple half-wave dipole it will suffice, in most cases, to follow the design information previously given for the type of line and matching system to be employed. In this case, in fact, it is often satisfactory to cut the antenna to length by formula and not go to the trouble of establishing exact resonance.

Aside from resonating the antenna, there is little that can be done about adjusting systems such as a dipole fed at the center with 75-ohm line, or a folded dipole fed with 300-ohm line. If the antenna height can be altered, it may be possible to bring about a better match by using the height that makes the radiation resistance the optimum value for good matching. From the standpoint of power lost in the line, an s.w.r. of 2 to 1 is perfectly satisfactory, as pointed out earlier. The chief advantage of a lower s.w.r. is that little or no retuning of the coupling circuit at the transmitter is required when shifting frequency within a band. This point is covered later in this chapter.

When a ¼-wave line balancer "bazooka" or impedance transformer is used it is advisable to resonate it before installation. This can be done without much difficulty if a grid-dip meter is available. In the system shown in Fig. 3-42A, the stub should be cut slightly long and shorted at the bottom with a small loop — just large enough to permit coupling to a grid-dip meter. A half-inch loop should be sufficient. The open end is then clipped a little at a time until the stub is resonant at the desired frequency. The resonant length should include the length of the connections to the antenna at the open end. After resonating, the conductors at the bottom end may be soldered directly together. The small error introduced by the coupling loop will not be significant except at frequencies approaching the u.h.f. region. The system at B in Fig. 3-42 may be adjusted by a similar process, except that in this case the connection at the bottom will serve as the coupling loop. *Both* lines should be clipped equally to bring the system to resonance, leaving enough conductor for making connection to the antenna at the top, but not connecting the two lines together until after resonance has been established. The system shown in Fig. 3-42C may be resonated by cutting a small longitudinal slot near the bottom of the sleeve, just large enough to take a single-turn loop that is in turn link-coupled to the grid-dip meter. If the sleeve is a little long to start with, a bit at a time can be cut off the top until the stub is resonant.

The ¼-wave transformer of Fig. 3-43 may be adjusted to resonance before being connected to the antenna by short-circuiting one end and coupling it inductively at that end to a grid-dip meter. The length of the short-circuiting conductor increases the length of the line slightly, but this can be compensated for by cutting half the length of the shorting bar from each line conductor after resonating, measuring the shorting-bar length between the centers of the line conductors. The only means available for adjustment of the impedance match with this system is to vary the spacing between the conductors of the linear transformer. Check of the s.w.r. with two or three different spacings will show which direction to take to bring about the best match.

In the delta and "T" systems there are two variables, the positions of the taps on the antenna being one. The other is the length B, Fig. 3-46, in the delta system, and the spacing, B, Fig. 3-47, in the "T". No specific directions can be given other than to try all possible combinations to find the one that gives the lowest s.w.r. It is usually better to try the antenna taps first, keeping them spaced equally distant from the center of the antenna. Where a specific antenna design using either of these systems is being followed, the dimensions given will serve as a starting point. If initial measurement of the s.w.r. shows it to be no higher than 2 to 1 there is almost nothing to be gained, in the way of efficiency, by spending a large amount of time to secure a "perfect" match.

Matching Stubs and Sections

There are also two variables when matching stubs or sections are used, dimensions A and B, Figs. 3-55 and 3-59. With this system considerable time can be saved if the antenna impedance or the s.w.r. (with no matching stub or section) can be measured. While accurate measurement is not easy — some of the pitfalls are discussed later in this chapter — even an approximate measurement will give a set of starting dimensions that is likely to save some cut-and-try.

Although there is no difference between a stub and a "¼-wave" matching section, the explanation of the adjustment procedure is somewhat more straightforward if the matching-section viewpoint is taken. Also, if the s.w.r. cannot be determined accurately before installing a stub, it is advisable to make the stub or matching section of the appropriate length to be closed at the far end. The length depends on whether Z_R, the antenna impedance, is larger or smaller than the Z_0 of the line. The two cases are shown in Fig. 3-63. Dimensions A and B correspond to the same designations in Figs. 3-55 and 3-59. Where an open stub would be used, the stub length is extended ¼ wavelength so the end can be closed. This is not done for electrical reasons but purely for convenience in adjustment, as explained earlier.

The adjustment procedure that follows is based on the same principle used in adjusting an antenna to resonance. In fact, the initial steps in the procedure are to determine whether Z_R is higher or lower than Z_0, then mark the position of a current node on the disconnected transmission line (leaving the end open if Z_R is higher than Z_0, and shorting it if Z_R is lower), and then adjust the antenna length with the matching section disconnected. Measure or estimate the s.w.r. with the antenna alone terminating the line, and set the shorting bar accordingly. If the s.w.r. is 4 or more, $A + B$ will be near enough to ¼ wavelength to use that length as a beginning. Next, attach the line on the matching section near the antenna and note whether the position of the current node has shifted. If it has, adjust the position of the shorting bar to bring it back to point X. Note the s.w.r., as indicated by comparative current readings at a loop and node (the meter does not have to read linearly with current for this purpose). Move the line taps down to a new position on the matching section, readjust the position of the shorting bar to bring the current node at X, and note the new s.w.r. Continue this process, adjusting the shorting bar each time, until the s.w.r. is minimum. If the line taps are moved down too far, the positions of the nodes and loops along the line will be interchanged, indicating that the proper point for matching has been passed.

This procedure is based on the fact that for proper matching the load must be resistive; this is accomplished by setting the shorting bar to tune out the reactance, as indicated by the fact that the current nodes occur at the same point as with the line open or shorted. The match between the load and line resistances is a separate adjustment, the position of the line taps. However, the two adjustments interlock and the final settings must be approached step by step.

In case a variable condenser is to be substituted for an open stub (Fig. 3-60B) an equivalent adjustment procedure may be followed. At each

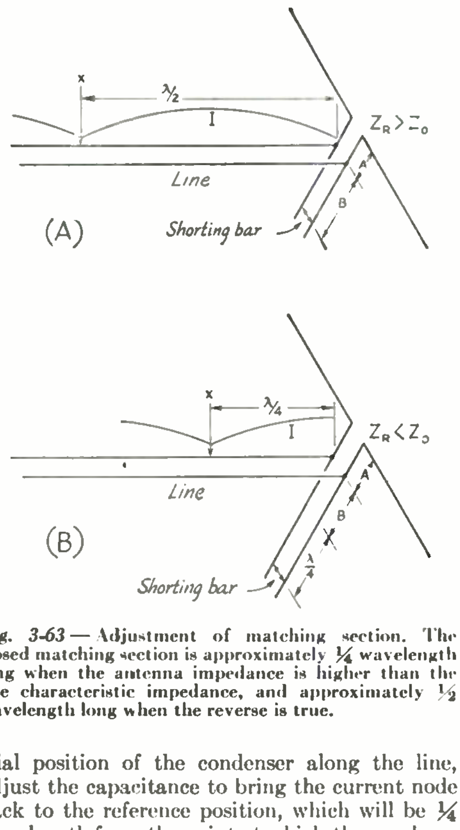


Fig. 3-63 — Adjustment of matching section. The closed matching section is approximately ¼ wavelength long when the antenna impedance is higher than the line characteristic impedance, and approximately ½ wavelength long when the reverse is true.

trial position of the condenser along the line, adjust the capacitance to bring the current node back to the reference position, which will be ¼ wavelength from the point at which the condenser is attached. Then check the s.w.r. and try a new position on the line for the condenser, exploring in this way until the s.w.r. is minimized.

With the LC circuit of Fig. 3-60C, changing the inductance in each coil corresponds to changing the position of the condenser on the line in the preceding paragraph. The corresponding procedure should be followed until a match is obtained.

Coupling the Line to the Transmitter

In many respects the problem of coupling the transmission line to the transmitter is similar to that of coupling the antenna to the transmission line. That is, the question of impedance matching is involved. The input impedance of the transmission line is the load for the transmitter. As described earlier in this chapter, the input impedance can fall anywhere in a wide range of values, depending on the length of the line, the standing-wave ratio, and the characteristic impedance of the line. Also, it can be either a pure resistance or a combination of both reactance and resistance.

On the other hand, the final amplifier of the transmitter requires a definite value of load impedance for optimum efficiency and/or linearity, and this impedance must be a pure resistance. In most cases the required impedance will be in the vicinity of a few thousand ohms; the exact value depends on the tube operating conditions such as plate voltage, grid bias, amount of grid excitation available, and so on. Fortunately, it is seldom necessary to know the actual load impedance required. When the tube manufacturer's recommendations are followed with respect to tube operating conditions the value of full-load plate current is always specified, and plate current becomes the criterion for loading. If the load resistance is too high ("light" loading) the plate current will be below the recommended value; if the resistance is too low ("heavy" loading) the plate current will be too great. The main object of the coupling system is to transform the input impedance of the transmission line into the required value of tube load resistance.

IMPEDANCE TRANSFORMATION

A resonant circuit exhibits properties similar to those of a transformer, and may be used to change a given value of impedance into a new value that may be either higher or lower. In Fig. 3-64A this transformer action is secured by tapping the load resistance, *R*, across part of the coil in a circuit that is tuned to resonance at the operating frequency. The impedance **(reflected impedance)** appearing at terminals *AB* is larger than *R* and will be a pure resistance if *R* is a pure resistance. The smaller the number of turns across which *R* is connected, in proportion to the total number of turns in the coil, the greater the impedance step-up to terminals *AB*. There is no theoretical limit to the step-up that can be obtained if the *LC* circuit has no inherent losses. However, the coil always has some internal resistance, so there is a definite practical limit. The maximum impedance that can exist across terminals *AB* is proportional to the *Q* of the circuit with *R* disconnected. With a coil and condenser having reasonably low losses, the self-impedance of the circuit will be in the tens of thousands of ohms — much higher than the range of impedances that will have to be handled in the line-to-transmitter coupling system. Practically speaking, then, such a circuit can be used for matching any of the impedances likely to be encountered.

Fig. 3-64B shows a circuit that also provides a method for impedance transformation, using a capacitive "tap" instead of tapping on the inductance. In this case, decreasing the capacitance of C_b (while increasing the capacitance of C_a correspondingly to maintain resonance) has the same effect as moving the tap toward the top of the coil in Fig. 3-64A. That is, it reduces the impedance step-up ratio. Among other advantages, this type of circuit gives very smooth control. However, the range of impedance transformation, with variable condensers of practicable size, is not as great as with the circuit at A.

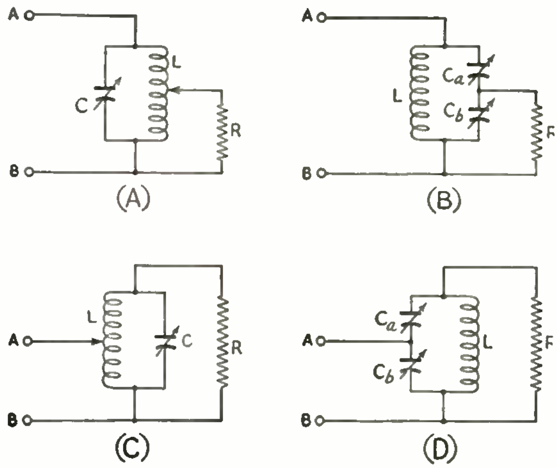


Fig. 3-64 — Impedance transformation with a resonant circuit.

Should it happen that the load impedance is higher than the impedance required by the source of power, the same circuits can be used by reversing the terminals. This is shown at *C* and *D* in Fig. 3-64. With *R* connected across the whole circuit, its resistance can be transformed to a lower value when the input terminals are tapped across part of the coil, as at C, or across C_b in Fig. 3-64D. The nearer the tap is to the bottom end of the coil, or the larger the capacitance of C_b compared with C_a, the smaller the resistance reflected between terminals *AB*.

Series Matching

If the resistance of R, the load, is very low — below 50 ohms, say — it may be difficult to get a suitable adjustment when R is tapped on the coil as in Fig. 3-64A. In the circuit of Fig. 3-64B it may be impossible to get a variable condenser of large-enough capacitance when R is very small. In such cases the load may be connected in *series* in the tuned circuit, as shown in Fig. 3-65. In series-connected circuits the impedance step-up to terminals AB depends on the LC ratio; the higher the LC ratio the greater the step-up.

These circuits are not suited to large values of R because there is a practical limit to the LC ratio that can be obtained. In many applications the Q of the circuit may be so low, if R is more than 100 ohms or so, that the series circuit is not usable even with the largest practicable LC ratio.

Inductive Coupling

It is possible to use the circuits of Fig. 3-64 and 3-65 exactly as shown, with terminals AB connected to the final amplifier tube and the input terminals of the transmission line replacing the load resistance R. However, for several reasons it is more desirable to use inductively-coupled circuits. For one thing, the plate voltage applied to the final tubes appears on the tank coil in series-fed circuits, and it is dangerous to allow this voltage to appear on the transmission line. For another, a single tuned circuit is not selective enough to prevent other frequencies, such as harmonics of the operating frequency generated in the amplifier, from being coupled into the line and thus into the antenna.

Three general types of inductively-coupled circuits are shown in simplified form in Fig. 3-66. The circuit at A corresponds to Fig. 3-64A, but the load resistance is connected to a second coil, L, instead of being tapped across a portion of the tank coil. If the two coils are very tightly coupled (interwound) the impedance transformation will be about the same as in Fig. 3-64A, when L has the same number of turns as the number across which R is tapped in Fig. 3-64A. With looser coupling the loading is "lighter" and the impedance across the L_pC_p circuit increases.

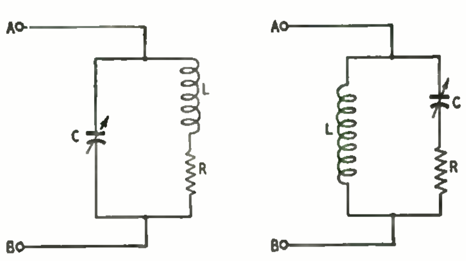


Fig. 3-65 — **Matching with a series-resonant circuit. If the circuit Q is 10 or more, the impedance reflected between terminals AB is equal to X^2R, where X is the reactance of either the coil or condenser.**

The circuit of Fig. 3-66B also is similar to Fig. 3-64A, except that the entire circuit is inductively coupled to the final tank circuit. Fig. 3-66C is the series-tuned circuit of Fig. 3-65 inductively coupled to the final tank circuit. In these two cases the behavior of the secondary circuit, LCR, is as already described. However, there is another

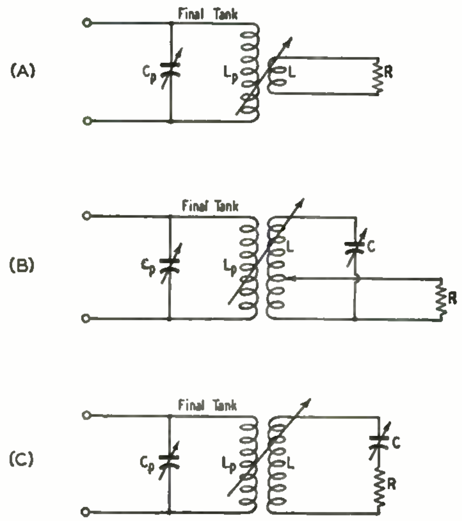


Fig. 3-66 — **Inductively-coupled impedance-transforming circuits.**

variable, the degree of coupling between the secondary and primary (final tank) circuits. When the coupling between the two coils is very loose, the loading on the final tank circuit is "light"; its impedance is high, therefore, and the amplifier draws less than normal plate current. As the coupling between the coils is made tighter the final tank becomes more heavily loaded; the impedance across its terminals decreases and the amplifier plate current rises. When the circuit constants are properly chosen it is readily possible to load the amplifier tank circuit to the desired degree even with quite loose coupling between the coils. On the other hand, it is also possible, when the circuits are not properly designed, that even the tightest coupling that is physically attainable will not load the amplifier heavily enough to make it draw normal plate current.

The degree of coupling between two resonant circuits, for a given physical separation of the coils L and L_p, depends on the Qs of the circuits. Good transmitter design calls for a tank-circuit Q (under loaded conditions) of at least 12, and this is high enough for good coupling in all cases. The effective Q of the secondary (antenna coupling) circuit should be at least 5 if the amplifier is to be loaded properly with reasonable separation between the coils. Aside from loading considerations, it is desirable to use a secondary Q in the vicinity of 10 because this increases the selectivity of the coupling circuit and thus helps discriminate against other frequencies, such as harmonics, that may be present in the output of the transmitter.

The Q of the secondary circuit depends on how heavily this circuit itself is loaded. Thus in Fig. 3-66B increasing the number of turns across which R is tapped increases the loading and reduces the Q. As the Q is decreased, tighter coupling between L and L_p is required for transferring the same amount of power. The proper tap position is the

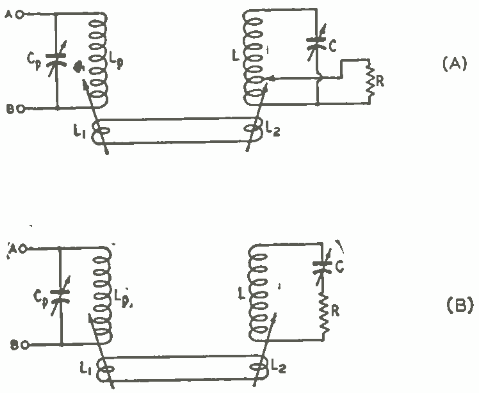


Fig. 3-67 — Link coupling to replace inductive coupling.

one that brings the Q down to about 10. When this is the case the voltage across the ends of the LC tank will be approximately the same as the voltage across the final-amplifier tank, and the circulating current in the tank also will be approximately the same. If L heats more than the final tank coil, the current is too high; that is, the Q is too high and R should be tapped across more of the coil. If the coupling between the two coils has to be quite tight to make the amplifier load properly, the Q is too low and R should be tapped across a smaller number of turns.

The secondary Q in the simple transformer circuit of Fig. 3-66A is zero. The impedance transformation and power transfer depend entirely on the number of turns that can be used in L and the tightness of coupling between L_p and L.

Link Coupling

Varying the coupling between the two coils of Fig. 3-66B and 3-66C by changing the spacing between them usually is rather inconvenient, because the coupling-circuit coil, L, usually will be about the same size as the final tank or primary coil, L_p. Common practice is to use link coupling between the two coils. Illustrative circuits using link coupling are shown in Fig. 3-67. Provision should be made for varying the coupling between L_1 and L_p as well as between L_2 and L, as indicated in the drawings.

There are two ways of looking on link coupling for design purposes. From one viewpoint the link circuit is merely a means for obtaining mutual inductance between L_p and L so they will be magnetically coupled even though they may be several feet apart. What is wanted, in this case, is a rather large current in the link circuit. This will be obtained when both L_1 and L_2 have a small number of turns compared with the number of turns in the coils to which they are coupled. From the mutual-inductance standpoint, the operation of a link-coupled circuit is exactly the same as the inductively-coupled circuits described in the preceding section.

The other viewpoint is that the link circuit is really a transmission line with coupling coils at both ends. In this case the coupling between L_1 and L_p is adjusted so that the amplifier would draw normal plate current if the link circuit were terminated in a resistance equal to its characteristic impedance, instead of being terminated in L_2. Then the loading on the antenna-coupling circuit, LC, is adjusted by tapping R across the proper number of turns on L, together with adjustment of the coupling between L and L_2, until the line is actually terminated in a pure resistance equal to its characteristic impedance. Thus we have impedance matching at both ends of the link.

The second method, although somewhat more complicated in adjustment, is desirable when the link circuit has appreciable length, or when the material used in the link line has considerable capacitance per unit length. If the link line is more than a few feet long, or is constructed of solid-dielectric coaxial cable, every attempt should be made to match impedances at both ends of the link circuit. The line may heat excessively, resulting in poor efficiency of power trans-er, if this is not done.

Complex Loads

In the foregoing discussion it has been assumed that the load, R, is a pure resistance. This is so only when the transmission line is matched or nonresonant, or, if a resonant line, when the line length is an exact multiple of a quarter wavelength long, measured from a resonant antenna. In all other cases the input impedance of the line will have both reactive and resistive components.

The four general cases with reactive lines are shown in Fig. 3-68. In A, the line has inductive reactance along with resistance. The parallel equivalent of the input impedance is used in this case because the line is in parallel with the LC matching circuit. When the line is tapped on L, both R' and X'_L are reflected in the same ratio. Thus if R' is 500 ohms and the tap is adjusted so that it appears as 3000 ohms across the entire circuit, the inductive reactance, X'_L, also will be stepped up in the ratio of 6 to 1. Tapping a line showing inductive reactance across part of L is therefore equivalent to connecting a larger inductive reactance across the LC circuit. Since L and the reflected inductance are in parallel, the

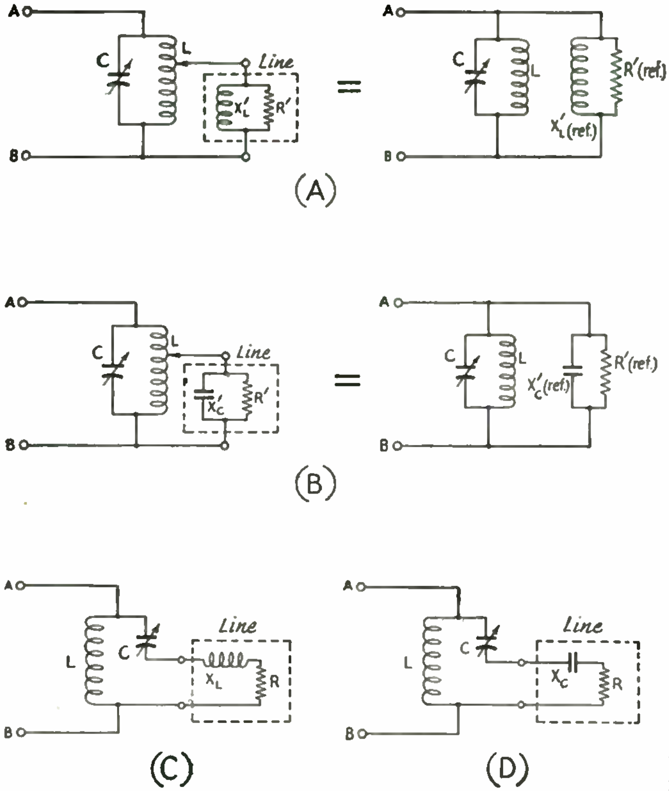


Fig. 3-68 — **Effect of complex line impedance on parallel and series matching circuits. In the parallel circuit, an equivalent inductance or capacitance is reflected across the circuit. In the series circuit, the line reactance adds equivalent inductance or capacitance in series.**

resonant frequency of the entire circuit becomes higher than the resonant frequency of LC alone. Consequently, the capacitance of C has to be increased to bring the circuit back to resonance.

In Fig. 3-68B the line is equivalent to a resistance and capacitance in parallel. Just as in Fig. 3-68A, the capacitive reactance is stepped up in the same ratio that the resistance is stepped up. In this case the LC circuit is shunted by an equivalent capacitance, which tunes it to a lower frequency. Consequently the capacitance of C has to be decreased to bring the entire circuit back to resonance.

The corresponding series circuits are shown in Fig. 3-68C and D. (These would be used when R is very small, as already explained.) When the line shows inductive reactance along with resistance, the effect is the same as though additional inductance were used in the tuned circuit. Therefore the circuit becomes tuned to a lower frequency and the capacitance of C has to be decreased to bring it back to resonance. When the line shows capacitive reactance the opposite is true; the equivalent capacitance of the line is in series with the tuning capacitance, C, and thus lowers the total capacitance in the circuit. Consequently the circuit becomes tuned to a higher frequency and the capacitance of C has to be increased to bring it back to resonance.

In general, the necessary retuning can be accomplished by adjustment of C as described. However, there may be cases where the line reactance is so high, in the series case, or the reflected reactance is so low, in the parallel case, that the tuning condenser does not have enough range to compensate for it. Within limits, this can be compensated for in the series case by inserting additional reactance, of the same value as the line reactance but of the opposite kind, in series with the circuit. If a very large reactance is required (large coil or very small condenser) it may be difficult to arrive at satisfactory values and the circuit should be changed to the parallel equivalent.

An alternative method of compensating for the line reactance in the parallel case is shown in Fig. 3-69. Here the compensating reactance is shunted directly across the line terminals. Its value should be adjusted so that it is equal to the line reactance, in which case the impedance is a pure resistance. Under these conditions, tapping the compensated line on L will not change the tuning of the LC circuit, if the Q of the circuit (loaded) is in the neighborhood of 10 or more.

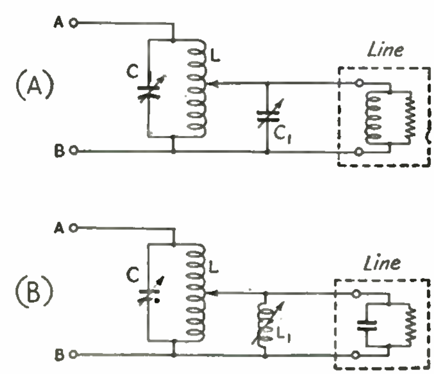


Fig. 3-69 — **Direct compensation of line input reactance before impedance transformation.**

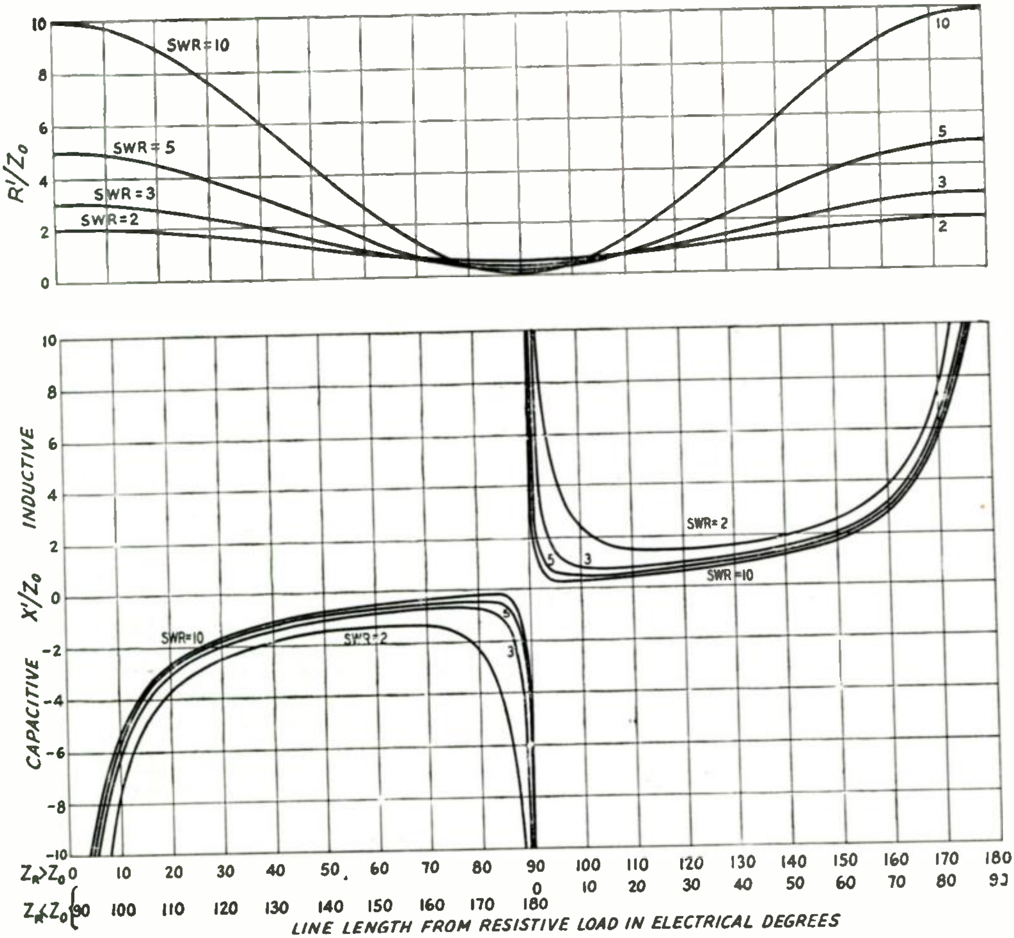


Fig. 3-70 — **Normalized resistance and reactance components of line input impedance, equivalent parallel circuit. To find actual values of reactance multiply values given by the characteristic impedance of the transmission line.**

COUPLING-CIRCUIT DESIGN

An understanding of the principles just outlined will enable anyone to recognize the causes of difficulties that may be encountered in coupling the transmission line to the transmitter. The proper solution can then be arrived at by cut-and-try methods. However, it is possible to determine in advance the approximate circuit conditions that will exist, and in many cases this will save a good deal of time in experimental work.

The input impedance of the transmission line can be determined if the characteristic impedance of the line, its electrical length, and the standing-wave ratio are known. The first two are a matter of choice, while the last can either be measured or estimated with good-enough accuracy. The only requirement is that the antenna should be operated at or quite close to resonance so that it presents an essentially resistive load to the line. In most cases it will suffice to take the value of resistance, at the feed point, from the published data on the type of antenna to be used. The s.w.r. is then Z_R/Z_0 or Z_0/Z_R, whichever is larger, Z_R being the antenna impedance and Z_0 the characteristic impedance of the line.

Knowing these three quantities, the resistive and reactive components of the input impedance may be found from the charts of Figs. 3-70 to 3-73, inclusive. Figs. 3-70 and 3-71 give the resistance and reactance in the equivalent parallel circuit, Fig. 3-71 being on an enlarged scale so the lower values can be read more easily. Figs. 3-72 and 3-73 are similar charts of the equivalent series circuit. In both cases the resistance and reactance are "normalized" to the line Z_0, so that they apply to all lines. Actual values are obtained by multiplying the figure taken from the chart by the Z_0 of the line.

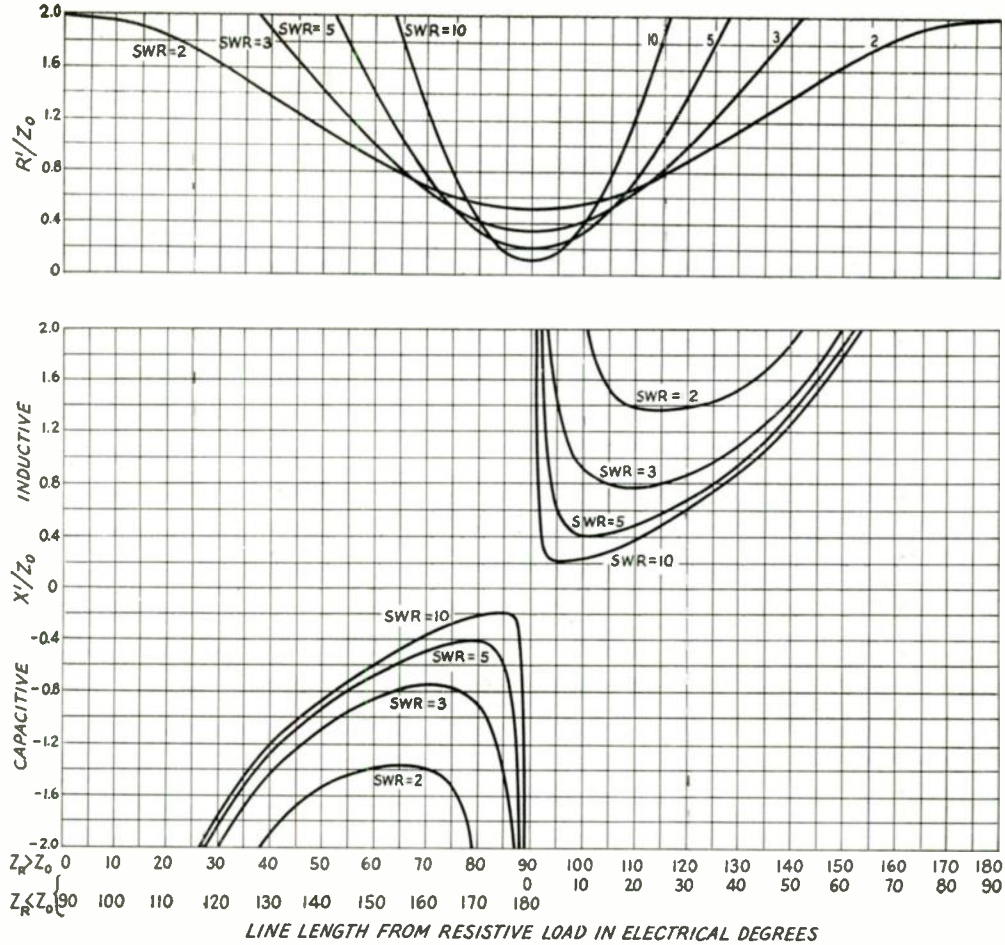


Fig. 3-71 — Enlarged scale (from Fig. 3-70) for low values of resistance and reactance, equivalent parallel circuit.

Reactance Compensation

The line length can be converted into electrical degrees by the following formula:

$$\text{Length in degrees} = \frac{0.37}{V} lf$$

where l = Length of line in feet
f = Frequency in Mc.
V = Line velocity factor

For example, if the frequency is 7150 kc. (7.15 Mc.) and the line is a 100-foot length of 300-ohm Twin-Lead, the electrical length is $0.37/0.82 \times 100 \times 7.15 = 322$ degrees. We can bring this within the range of the charts by subtracting the largest possible multiple of 180 degrees, since the impedance repeats at each half-wave interval along the line. In this case the largest multiple is 1, so we have $322 - 180 = 142$ degrees.

Now suppose that the line is terminated in the center of a half-wave antenna, without any special matching arrangements. The impedance of the antenna will be approximately 70 ohms, at average heights, so the s.w.r. is 300/70, or about 4 to 1. From Figs. 3-71 and 3-73 we arrive at the following (Z_R less than Z_0):

Parallel: $R' = 1.5 \times 300 = 450$ ohms
$X' = 0.9 \times 300 = 270$ ohms
Series: $R = 0.4 \times 300 = 120$ ohms
$X = 0.7 \times 300 = 210$ ohms

It was necessary to interpolate between the curves for s.w.r. = 3 and s.w.r. = 5 in this case, and although the figures are approximate they are well within the necessary limits of accuracy. As shown by the charts, the reactance is capacitive.

The series resistance and reactance are just about low enough in this example to permit the use of a series-tuned coupling circuit. On the other hand, the parallel resistance and reactance are

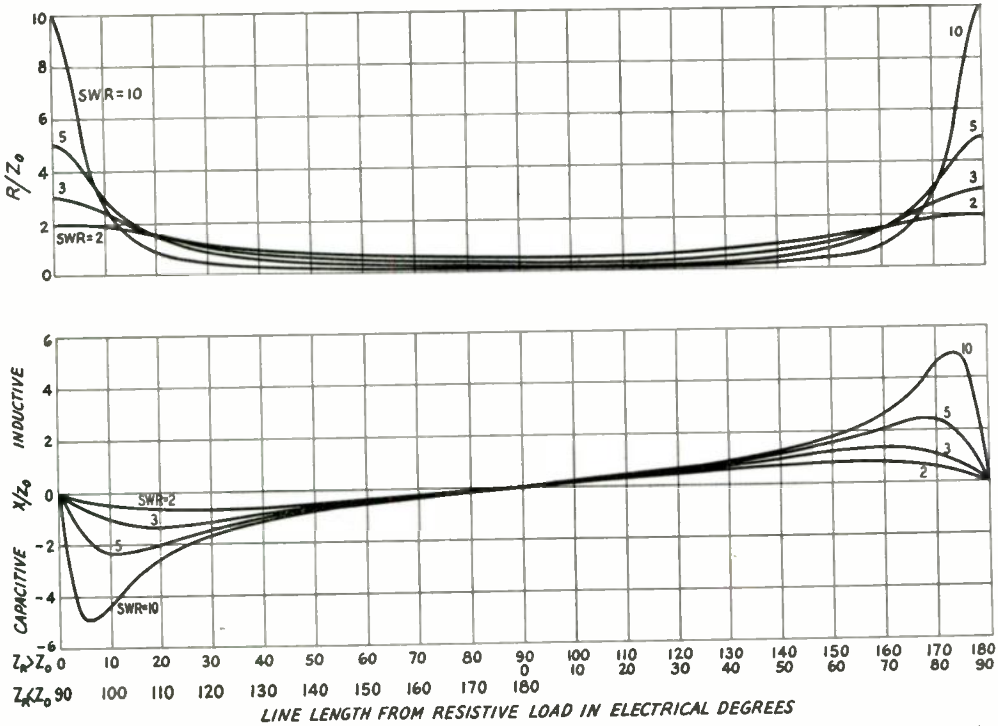


Fig. 3-72 — Normalized resistance and reactance components of line input impedance, equivalent series circuit. To find actual values of reactance multiply values given by the characteristic impedance of the transmission line.

high enough to be used with a parallel-tuned circuit. In view of these conditions either method could be used.

The Coupling Tank

At this point it is necessary to consider the LC constants in the coupling circuit. It has already been mentioned that the Q should be approximately 10. Practical considerations such as the capacitance range available in variable condensers limit the choice of L and C. A reasonable choice is a capacitance of 100 μμfd. for 3.5 Mc. If the same LC ratio is maintained on all bands, the coil and condenser reactances will be the same on all harmonic frequencies. We then have the following:

Frequency	*Capacitance*	*Inductance*
3.5 Mc.	100 μμfd.	21.0 μh.
7 Mc.	50 μμfd.	10.5 μh.
14 Mc.	25 μμfd.	5.25 μh.
21 Mc.	17 μμfd.	3.5 μh.
28 Mc.	13 μμfd.	2.6 μh.

As shown by the charts of Figs. 3-74 and 3-75, these inductances and capacitances correspond to a reactance of about 450 ohms. With the load resistance in series, the circuit Q is equal to the reactance divided by R, and so will be 10 when R is 45 ohms. With the load resistance in parallel, the circuit Q is R' divided by the condenser or coil reactance, and so will be 10 when R' as reflected across the whole circuit is 4500 ohms.

The fundamental circuits at B and D in Fig. 3-68 are the ones that apply in the example above. In the parallel circuit we must step up R' to 4500 ohms, so the impedance ratio is 4500/450 = 10. The reflected reactance therefore will be 10 × 270 = 2700 ohms. From Fig. 3-75 the corresponding capacitance is approximately 8 μμfd. at 7 Mc., so connecting the line across the proper number of turns on L will increase the tuning capacitance across the whole circuit by 8 μμfd. This is easily compensated by a corresponding reduction in the capacitance of C.

As a rough estimate, the number of turns across which the line should be connected is equal to the total number of turns in the coil divided by the square root of the impedance ratio. If the coil has 12 turns, for example, the turns required for the taps are given by $12/\sqrt{10} = 12/3.16 = 3.8$, or approximately 4 turns. This relationship is not exact, and so should be used only to indicate where to start in adjusting the circuit.

In the series circuit we have 120 ohms of resistance and 210 ohms of capacitive reactance connected in series with L and C. Since the reactance of C is approximately 450 ohms, the total

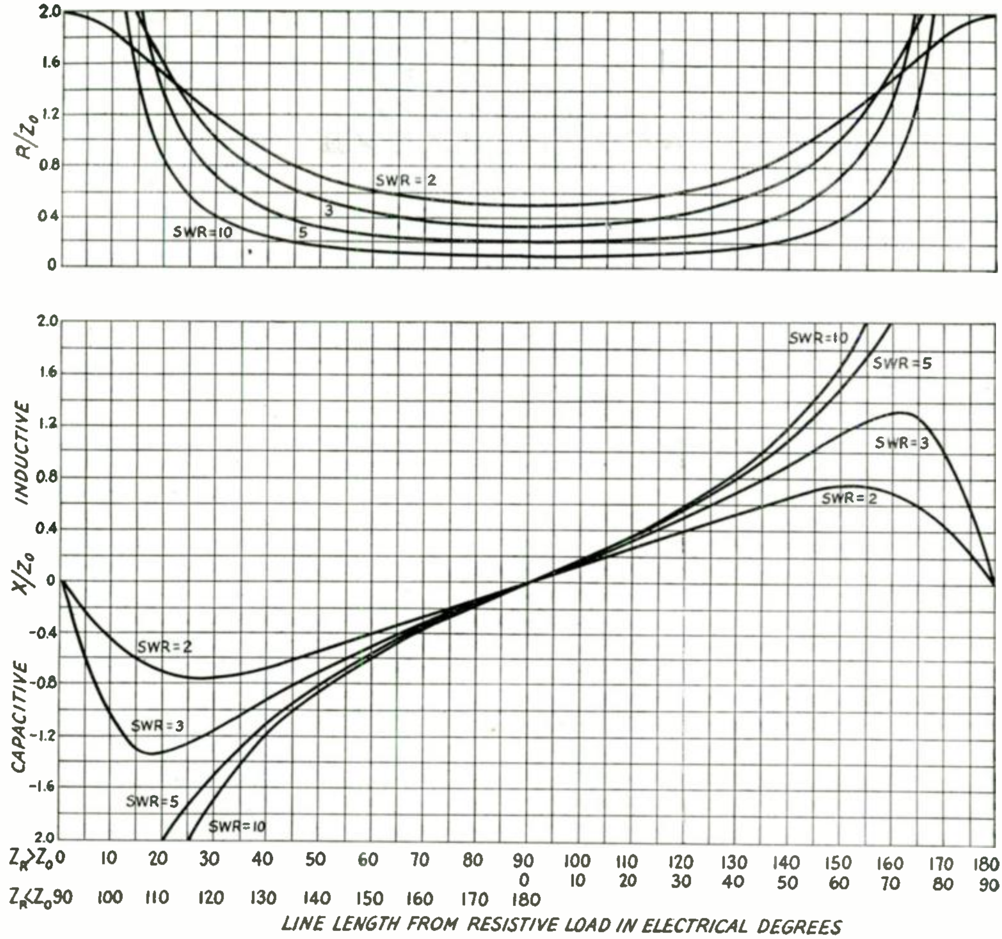


Fig. 3-73 — Enlarged scale (from Fig. 3-72) for low values of resistance and reactance, equivalent series circuit.

reactance is 450 + 210 = 660 ohms. This could be compensated for by adding 210 ohms of inductive reactance in series to make a total of 660 ohms. From Fig. 3-74 this would require an additional inductance of 4.8 μh., or a total of about 15.3 μh. The circuit Q then would equal 660/120, or 5.5. Alternatively (and this is the usual practice) the reactance of C can be decreased, by increasing its capacitance, to 240 ohms, so that its reactance plus the line reactance equals 450 ohms, the reactance of the coil. From Fig. 3-75 the capacitance required for a reactance of 240 ohms is 95 μμfd., which would be within the range of a 100-μμfd. condenser. Under these conditions the Q of the circuit would be X/R = 450/120 = 3.75. With either method of compensation the circuit Q is somewhat low for good selectivity, but probably not too low to permit adequate coupling to the final tank when either inductive or link coupling is used.

Other cases can readily be worked out by the same method. With series tuning, the Q should always be checked to make sure it is adequate for good coupling, and it is generally advisable to shift to parallel tuning if the Q cannot be brought near 10.

Additional Reactance Compensation

If calculations such as those in the example worked out above show that the tank condenser will have to be detuned very far from the capacitances recommended in the previous section, or if the capacitance has to be increased beyond the maximum range of the condenser, it is better to compensate directly at the line input terminals as shown in Fig. 3-69. In such a case the value of reactance as given by the charts of Figs. 3-70 and 3-71 is used without transformation. In the example, the line reactance in the parallel equivalent circuit was 270 ohms, capacitive. We therefore would need an inductive reactance of 270 ohms to shunt across the line terminals. From

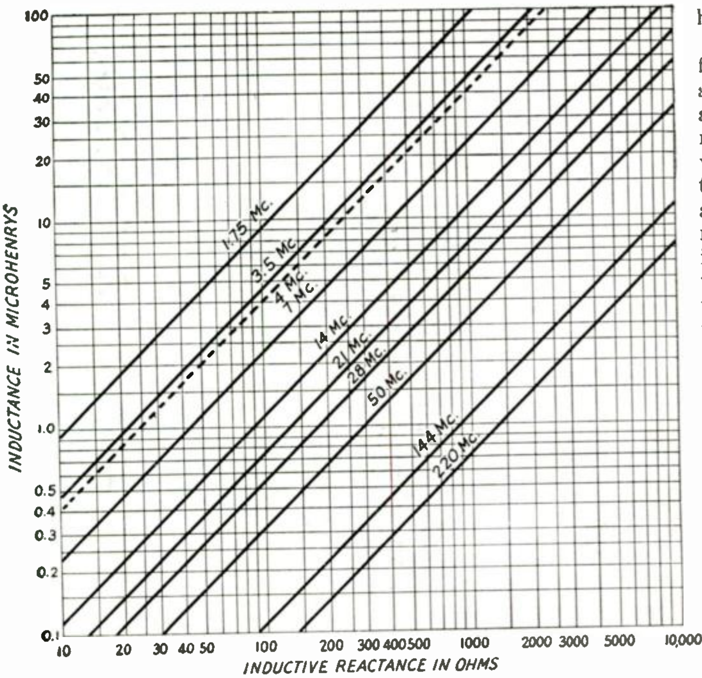


Fig. 3-74 — **Inductive-reactance chart. Values given for the frequency shown are close enough, for most calculations, throughout the band. The maximum difference occurs in the 3.5–4.0 Mc. band.**

Fig. 3-74 the required inductance is approximately 6.5 μh. for direct compensation of the line reactance.

Besides the physical difficulty of retuning the coupling tank to resonance when the reactive effects are pronounced, an attempt to compensate wholly by retuning the tank circuit may lead to excessive power loss in the tank. This is particularly likely to be the case when a relatively low value of inductive reactance is reflected across the tank circuit, because this effectively reduces the L/C ratio and increases the current circulating in the tank.

The action of the parallel-tuned circuit is somewhat similar to that of an autotransformer. The circulating tank current and the line current are more or less in phase opposition in the section of the coil included between the line taps, so the resultant current is smaller than the larger of the two. On the other hand, there is no such opposition in the part of the coil that is not included between the taps. The result is that the losses in this part of the coil are greater than in the portion between the taps. If the circulating current becomes large as a result of reflected reactance the power loss in the end sections of the coil may be considerable. The additional power loss not only means decreased efficiency but also may make part of the coil too hot for safety, especially when the turns are supported by insulating material having a low softening point.

Whether or not the reflected reactance will be such as to require compensation at the line terminals is not so much a matter of the actual value of line reactance but of the relationship between reactance and resistance. This relationship can be expressed in terms of the ratio of the total current into the line to the "active" current (that is, the current in the resistive component of the equivalent parallel circuit). This ratio varies with the line length and s.w.r. as shown in Fig. 3-76. For any value of s.w.r. the ratio is most unfavorable (highest) when the line length is an odd multiple of 45 degrees or ⅛ wavelength.

It might seem desirable to select a value of the current ratio above which the reactance compensation should be applied at the line input terminals and below which it would be satisfactory to retune the coupling tank. However, there is no clear-cut method for doing it. If the coupling coil is built of conductor having ample size to handle the power, the losses will not be significant even when the current ratio is quite large. But if the coil is wound with small wire and is operating near its maximum power capacity when the circuit is delivering power to a purely resistive load, a small increase in the current ratio may cause objectionable heating. Fig. 3-76 is useful in showing the regions in which difficulties may be anticipated. As shown by the curves, the range of line lengths over which it may be necessary to use direct compensation increases rapidly with an increase in the s.w.r.

CIRCUIT DETAILS AND ADJUSTMENT

The coupling circuits have been covered in basic form in the preceding sections. Actual details of circuit arrangement vary with the requirements. For example, unbalanced (coaxial) lines require a different type of treatment than balanced (parallel-conductor) lines. With the latter, it is desirable that the entire circuit be balanced to ground. When the s.w.r. is high, the coupling circuit must be able to handle widely-different line input impedances, particularly if the same antenna is used on several bands. On the other hand, simpler coupling circuits can be used when the s.w.r. is very low, especially with low-impedance lines. Finally, it is always desirable to take precautions against harmonic radiation.

Coupling to Nonresonant Coaxial Lines

Several circuits suitable for coupling a transmitter to a coaxial line are shown in Fig. 3-77. It is assumed that the antenna is matched to the line so that the s.w.r. is very low. In the untuned circuits at A and B in Fig. 3-77 the line is matched to the load required by the transmitter through simple transformer action. This calls for rather close coupling between the link and the final tank coil. Fig. 3-77A shows how a single-ended tank can be coupled to the line. The outer conductor of the line should be connected to the side of the link nearest to the grounded end of the tank coil, when the two coils are end to end. If the link is wound over one end of the tank coil the side farthest from the end of the tank coil should be grounded. In either case, this method of grounding tends to reduce capacitive coupling between the two coils.

With a balanced tank, the link should be at the center of the amplifier tank coil, as indicated in Fig. 3-77B. In this case either side of the link may be grounded.

With untuned coupling the adjustment is quite simple. The only control over the loading on the amplifier is the coupling between the two coils, and this is varied until the amplifier takes the desired plate current. With each change in the coupling, the amplifier tank condenser should be adjusted to resonance, as indicated by minimum plate current. If the line is properly terminated the coupling will have very little effect on the setting of the condenser. However, if there is an appreciable s.w.r. and the line impedance contains reactance, tightening the coupling between the two coils will couple an increasing amount of reactance into the amplifier tank. This makes it necessary to retune the tank with each change in coupling.

Use of a separate antenna coupling circuit is shown in Fig. 3-77C. This arrangement provides additional selectivity — the untuned coupling scheme does not — but necessitates additional adjustments. The *LC* circuit should have a *Q* of about 10, which requires a reactance in both coil and condenser of about 500 ohms for 50-ohm line and 750 ohms for 75-ohm line. Corresponding values of inductance and capacitance can be taken from Figs. 3-74 and 3-75.

With this method fairly loose coupling should be used, at the start, between the

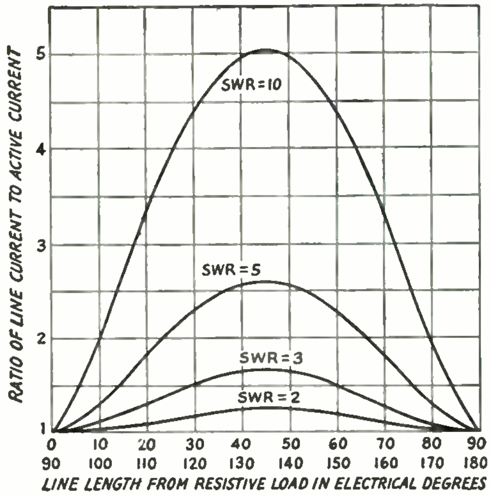


Fig. 3-76 — Effect of standing-wave ratio on ratio of total line input current to "active" current (the current that represents power delivered by the line to the antenna). These curves repeat every 90 degrees (¼ wavelength) of line.

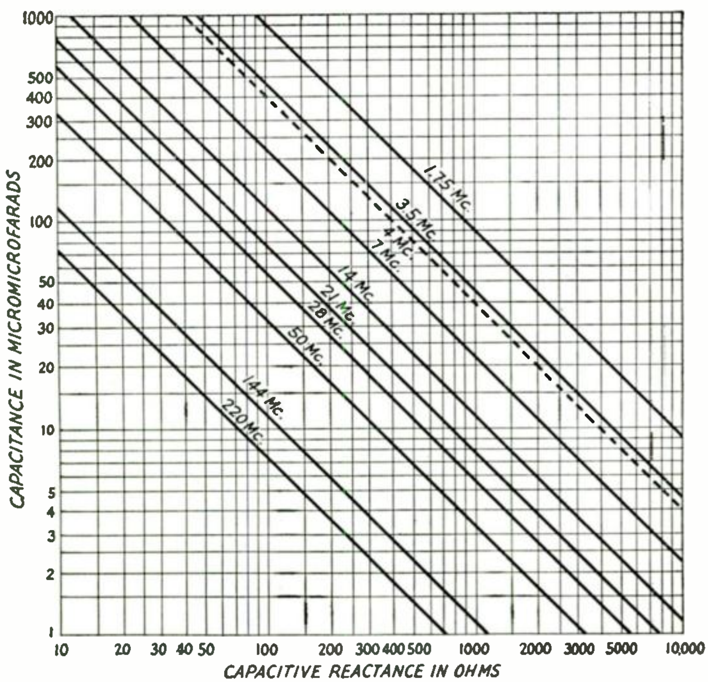


Fig. 3-75 — Capacitive-reactance chart. Values given for the frequency shown are close enough, for most calculations, throughout the band. The maximum difference occurs in the 3.5–4.0 Mc. band.

links and the tank coils. The final tank condenser is first adjusted to resonance, shown by minimum plate current, and then the coupling tank is tuned by means of C. As C is varied the plate current will rise to a peak, at resonance, and then fall off again. The coupling at both ends of the link circuit can then be increased a bit and the tank circuits retuned as before. This process should be continued until the amplifier is drawing the desired plate current. When the circuits are properly coupled the plate current should fall off when the coupling tank circuit is detuned on either side of resonance. With too-tight coupling the plate current will rise with such detuning.

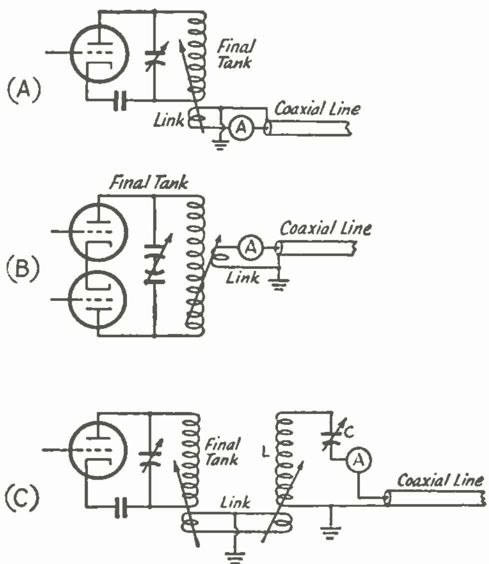


Fig. 3-77 — Untuned coupling to nonresonant coaxial lines (A and B); link coupling with series-tuned matching circuit (C). The number of turns required in the link coil varies from 1 or 2 at 28 Mc. and higher to 4 or 5 at 3.5 Mc. In the event that adequate coupling cannot be secured with the circuits at A and B, the circuit at C should be used. Alternatively, the line input circuit can be tuned as shown in Fig. 3-81.

The method of adjustment when the link circuit is treated as a low-impedance transmission line is considered later. In the case of a balanced amplifier, the link at the amplifier tank circuit should be coupled at the center of the coil in the same way as in Fig. 3-77B.

Coupling to Nonresonant Parallel-Conductor Lines

Since the currents in parallel-conductor lines should be balanced, it is good practice to use a balanced circuit for feeding power into such lines. If there is any circuit unbalance to ground the voltage applied to one line wire may differ from that applied to the other, with the result that the two wires will not carry the same current. This point requires particular attention when the coupling circuit is mounted on a metal chassis, and when the line is one having a characteristic impedance of 300 ohms or more.

Untuned coupling may be used with 75-ohm parallel-conductor lines. The circuits are the same as those given in Fig. 3-77 at A and B, except that the link either should be grounded at its center or not grounded at all. The method of adjustment is also the same.

Untuned coupling is not very successful with lines of 300 ohms and over when the line actually is operating at a low standing-wave ratio. With these high-impedance lines the number of coupling turns may need to be as great as half the number of turns in the amplifier tank coil. Moreover, the maximum coefficient of coupling between the tank coil and a "swinging link" seldom exceeds about 70 per cent, even at the maximum-coupling position. The voltage induced in the coupling coil is proportionately small, and it becomes difficult to obtain adequate power transfer. Also, a coefficient of coupling less than 100 per cent means that part of the magnetic flux set up by current flowing in the coupling coil does not link with the amplifier tank coil. This in turn means that there is reactance (leakage reactance) in the secondary circuit, and this reactance acts in series with the input resistance of the line to increase the total impedance of the circuit. Increasing the number of turns in the coupling coil is seldom beneficial, when the coupling is initially inadequate, because the reactance increases with the number of turns.

The only remedy is to tune the coupling circuit. The separate coupling tank usually is link-coupled to the amplifier tank circuit, as suggested in Fig. 3-78. If the link is grounded, as is desirable, it should be grounded at the amplifier-tank end. The circuit of Fig. 3-78 uses a split-stator or balanced condenser, so that the capacitance to chassis or ground will be the same on each side of the circuit. Care should be used to mount the inductance, L, in such a way that each end is exposed in the same way, physically, to near-by metal masses such as chassis, panels and shields. At the amplifier-tank end, the link should be at the "cold" end of the coil when the amplifier is single-ended, and at the center when the amplifier circuit is balanced.

The method of adjustment is as follows: First loosen the coupling between the link coil and L, and also between the amplifier tank coil and its link. Tune the amplifier plate circuit to resonance, as shown by minimum plate current. Then, with the taps on L fairly close together and equidistant from the center of the coil, adjust C to resonance. The resonance point will be indicated by a peak in the amplifier plate current. Recheck the amplifier tuning by again adjusting for minimum plate current; the condenser setting should be essentially the same. Next, increase the coupling between the links and the tank coils a little, recheck the tuning, and note the plate current. Continue this process just so long as tuning C through resonance, without touching the amplifier tank condenser, causes the plate current to pass through a peak.

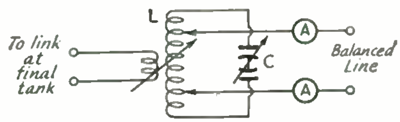


Fig. 3-78 — Link-coupled circuit for coupling to a nonresonant open-wire line. L and C should, in general, have the same physical construction and electrical ratings as the final-amplifier tank circuit. However, the LC ratio can be varied over reasonably wide limits without affecting the efficiency to any marked degree. The primary requirement is that the LC circuit should be resonant at the operating frequency.

If tuning C off resonance causes the plate current to rise or makes it necessary to reset the amplifier tank condenser to resonance (after obtaining an initial setting for the value of coupling in use) and the plate current at resonance is above the desired value, increase the spread between the taps on L and continue. With the proper coupling and tap settings the tuning of the coupling tank should not materially affect the amplifier tuning, and the plate current should reach the desired load value as C is tuned through resonance, dropping off on either side. Under these conditions a moderate frequency range can be covered without retuning the coupling tank.

After a suitable set of adjustments has been reached it is a good plan to let the transmitter run for a short period, then shut off power and see if L is showing excessive heat. If L is warmer than the amplifier tank coil and both are of the same conductor size and approximate inductance, the Q in the coupling tank is higher than necessary and the taps on L may be moved farther apart. Flashover in the condenser also indicates that the Q is higher than necessary, assuming that both C and the final tank condenser have the same plate spacing.

Before attaching the line to the coil, check the setting of C at which the coupling circuit is resonant. This may be done in the same way as the resonance check with the line attached, except that very loose coupling will have to be used. If attaching the line makes any material difference in the condenser setting at which resonance occurs the line is not "flat."

A series-resonant coupling circuit should not be used with high-impedance lines because it is practically impossible to use a large-enough inductance at L and a small enough capacitance at C to give the circuit a reasonable Q.

Coupling to Resonant Lines

With resonant transmission lines the use of a tuned coupling circuit is essential, except in a few cases that are so specialized that they need not be considered. The transmission line is usually of the open-wire type and will have a characteristic impedance of 500 to 600 ohms. (It was pointed out earlier that 300-ohm Twin-Lead also can be used, but the power that can be handled is limited by the standing-wave ratio.) If the s.w.r. is high enough the input impedance will be quite low at certain line lengths, and in these cases series tuning can be used. The common circuit is shown in Fig. 3-79A. Link coupling to the final tank circuit is shown, since this is the most convenient way to secure variable inductive coupling. However, the amplifier-tank and coupling-circuit coils may be coupled inductively as suggested in Fig. 3-66.

The series tuning capacitance is preferably divided between two condensers, C_1 and C_2 in Fig. 3-79A. These should be connected with their stator plates to the ends of the coil, L, as shown in the diagram. If the circuit is otherwise symmetri-

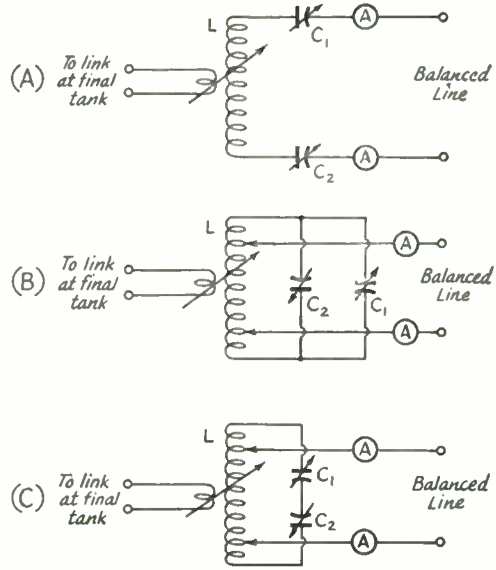


Fig. 3-79 — Matching resonant lines by series- and parallel-tuned circuits. B and C show how the same two condensers used for series tuning can be combined for parallel tuning to provide either a high maximum or low minimum capacitance. The capacitance of C_1 and C_2 should be 75 to 100 $\mu\mu$fd, with a voltage rating equal to that of the final tank condenser. Suggested values for L are tabulated in the text in the section on coupling-tank design. The same values may be used in the series-tuned circuit, except that at some line lengths (input impedance capacitive) it will be necessary to make L larger, at low frequencies, in order to resonate the circuit. This may be done by connecting a "loading" inductance in series. The amount of inductance required may be found by experiment or determined from the design data given in the text.

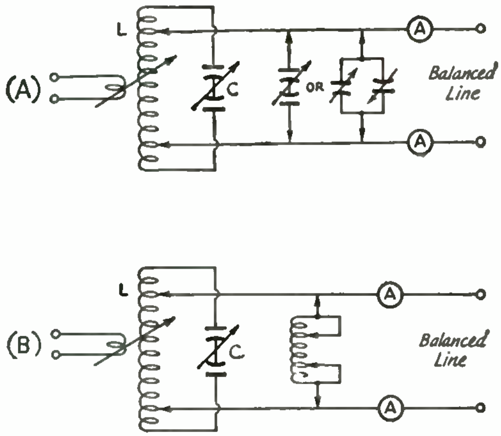


Fig. 3-80 — **Compensating for line input reactance, to make the input impedance purely resistive before tapping on coupling circuit.**

cally laid out (see discussion under coupling to nonresonant lines) the use of two condensers will keep the capacitance to ground the same on both sides of the line. The circuit formed by L, C_1 and C_2 should be capable of being tuned to resonance when the line terminals are short-circuited.

In adjusting a series-tuned circuit, start with loose coupling at both ends of the link. Resonate the amplifier tank circuit, as indicated by minimum plate current, and then tune C_1 and C_2 to resonance, keeping both at the same capacitance. This will be indicated by a rise in plate current. Then tighten the coupling by means of the links, a little at a time, until the amplifier takes normal plate current. Recheck the settings of the amplifier tank condenser, as well as C_1 and C_2, each time the coupling is changed. Use the loosest coupling at the links that will permit normal loading on the amplifier. Detuning C_1 and C_2 from resonance should cause the plate current to fall when the proper value of coupling is used.

The parallel-tuned circuit of Fig. 3-78 is suitable for use with resonant lines when the input impedance is in the range calling for parallel tuning. Alternative circuits are shown at B and C in Fig. 3-79. These make use of the same two condensers employed in the series-tuned circuit at A, and thus form a useful and economical combination for a transmitter that works on several bands. On the lower frequencies the two condensers are connected in parallel (B), but with the rotor of one and the stator of the other on the same end of the coil. This preserves the capacitance balance to ground. At high frequencies, where a low minimum capacitance is necessary, the two condensers can be connected in series (C). In this case the two rotors should be connected together. This is equivalent to the circuit of Fig. 3-78, which uses a split-stator condenser. Note that the stator of C_1 can be connected permanently to the top end of L_1 and the stator of C_2 to the bottom end. It is necessary only to switch rotor connections to use either series tuning, parallel tuning with the condensers in series, or parallel tuning with the condensers in parallel.

The adjustment procedure for parallel tuning is the same as in the case of a nonresonant line (Fig. 3-78) except that the capacitance setting for resonance is not necessarily the same with the line connected as with it disconnected. When two condensers are used, as in B and C, Fig. 3-79, both should be maintained at the same capacitance. The proper adjustment is the one that loads the amplifier to full plate current with the loosest coupling at the links.

In general, parallel tuning can be used under practically all conditions. Series tuning is seldom actually *required*. However, if the s.w.r. is quite high and the line length is such as to make the input impedance minimum (current loop at the input terminals) it may be easier to find the optimum operating adjustments with series than with parallel tuning.

At certain line lengths, as already discussed, it may be necessary to use shunting reactances across the line in order to allow the parallel-tuned circuit to operate at good efficiency. It is desirable to maintain physical symmetry with respect to the line when using such reactances, as indicated in Fig. 3-80. If the design procedure outlined earlier cannot be followed through, the proper values can be determined experimentally as follows:

First, resonate the LC circuit alone with very loose coupling to the amplifier tank circuit. Note the condenser setting. Then tap the line across a small portion of L and reresonate the circuit. If the capacitance has to be increased, a shunt condenser is required across the line. If the capacitance is lower, a shunt coil is needed. When the line is tapped on L it may be necessary to increase the coupling to the amplifier tank because of the loading on the coupling circuit, but the taps should be kept close enough together so that loose coupling can be used.

Use a trial value of shunt inductance or capacitance as required, and check the setting of the tuning condenser at resonance. Vary the shunt inductance or capacitance until the setting of C is the same as with the line disconnected. Then increase the spread between the taps, tighten the coupling, and check again, adjusting the shunt capacitance or inductance, if necessary, so that resonance is obtained at the original setting of C. Continue until the coupling is such that the amplifier is fully loaded.

It is a simple matter to determine whether such shunt condensers and coils are necessary. If it is difficult to find resonance in the coupling circuit when the line is tapped on the coil, or if part of the coil heats badly when the transmitter is allowed to run for a few minutes, the shunting reactances should be used. If the coil does not heat under continuous operation, and the tuning procedure follows the normal lines described previously, the shunt elements are not necessary even though C may have to be retuned considerably when the line is attached.

The Link Circuit as a Transmission Line

In describing the adjustment of link-coupled circuits the link was considered simply as providing mutual inductance between the amplifier-tank and coupling-circuit coils. When the link is treated as a matched transmission line the procedure should be modified somewhat.

Any type of transmission line can be used for the link, theoretically. However, it should be noted that with high-impedance lines it is just as difficult to couple power into the line, if an attempt is made to use untuned coupling, whether the line is short or long. In using a matched link circuit, therefore, it is advisable to choose 50- or 75-ohm line for the link. The coaxial type is best from a constructional standpoint, because the outside of the shield is "cold" for r.f. The line can therefore be run wherever desired without regard for its proximity to other conductors.

The proper procedure to follow with a matched link is first to disconnect the link coil at the far end from the transmitter and substitute a resistive load equal to the characteristic impedance. This load or dummy antenna must be capable of dissipating the power output of the transmitter. The coupling between the amplifier tank coil and the link then should be adjusted to make the amplifier draw normal plate current, with its tank circuit tuned to resonance. After this the coupling between these two coils should not be touched.

If it is not possible to obtain sufficient coupling in this test, the input circuit of the link must be tuned. This can be done by inserting a condenser in series with the "input" link, L_1, as shown in Fig. 3-81. The circuit formed by C_1L_1 should resonate at the operating frequency, and to make this possible it may be necessary to increase the number of turns in L_1 over the number usually supplied in manufactured link coils. The Q of the L_1C_1 circuit needs to be only large enough to make it possible to get adequate power transfer. A Q of 3 to 4 is usually sufficient. The reactance of both L_1 and C_1 therefore should be 3 to 4 times the characteristic impedance of the line used in the link circuit. A high-Q circuit is not desirable at this point because it increases the voltage across C_1, for a given amount of power, and because its greater selectivity is an operating inconvenience in that C_1 has to be readjusted when changing frequency. With low Q it is possible to work over most of a band with a compromise setting of C_1.

Once the link line is properly coupled at its input end the tuning procedure is the same as described previously, except that only the coupling between L_2 and L is varied to change the loading on the amplifier. The link is properly matched when LC is tuned to resonance and the amplifier draws normal plate current.

Another method of adjustment is to use a standing-wave indicator (see later in this chapter) at the input end of the link circuit, adjusting the tuning and coupling at the output end to obtain minimum s.w.r. This may be done at any convenient power level, and with loose coupling between the amplifier tank coil and L_1. Once this adjustment is made the coupling between L and L_2 should not be changed. The power input to the amplifier then can be adjusted to the desired value by varying the coupling between L_1 and the amplifier tank coil. The two methods are shown in Fig. 3-81.

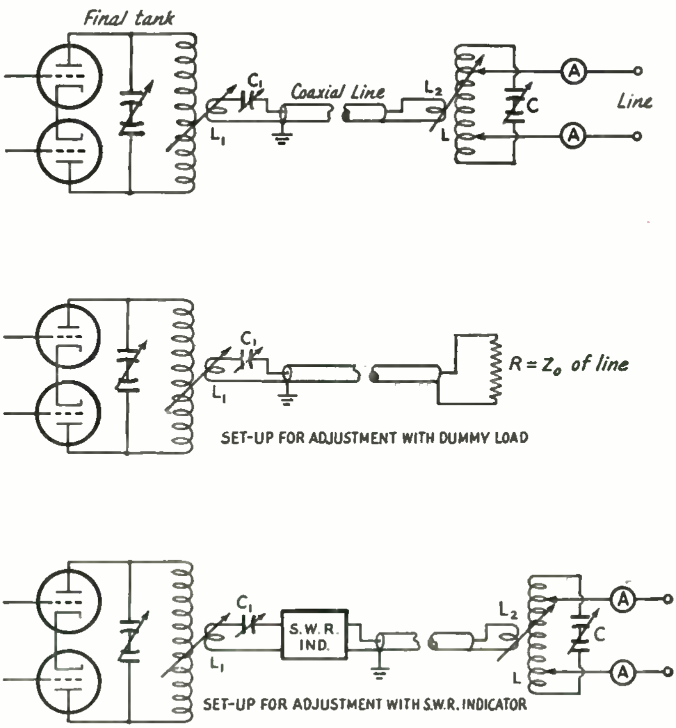


Fig. 3-81 — Using the link circuit as a matched transmission line, with suggested set-ups for adjustment.

The link arrangement shown can of course be used with a single-ended amplifier, or with any of the coupling circuits previously described.

Measurement of Line Input Current

Although the circuits in Figs. 3-77 to 3-81, inclusive, show r.f. ammeters connected in the proper places for measuring the line current, no mention was made of them in the adjustment procedure. This is because they are not essential to the operation of the system. If available, however, they may be used to advantage in tuning up.

The input impedance of the line is unaffected by any adjustments made in the coupling circuit. Consequently, the larger the current flowing into the line the greater the power fed to the antenna, under a given set of conditions.

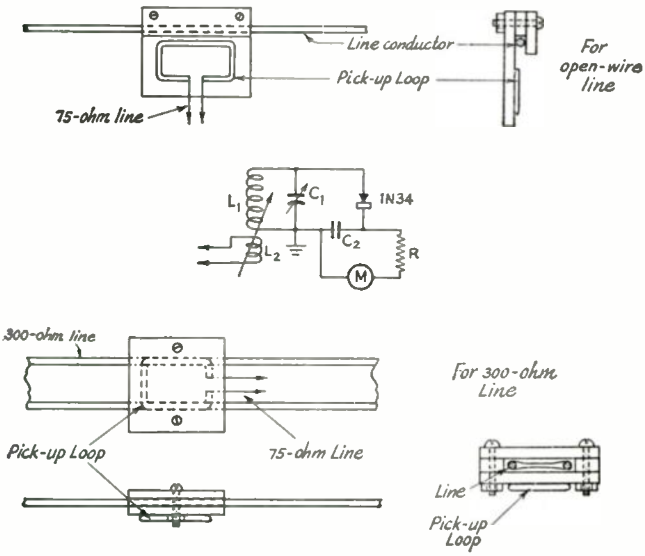


Fig. 3-82 — Circuit and suggested construction for standing-wave indicator using crystal-detector wavemeter. L_1C_1 can be any LC combination capable of being tuned to the operating frequency. L_2 usually will consist of a turn or two of the same diameter as L_1, preferably movable with respect to L_1 so the coupling can be adjusted. C_2 should be 0.001 μfd. or so. The series resistor, R, may be omitted if maximum sensitivity is desired, but will improve the linearity of the readings; see text for discussion. M may be a 0–1 milliammeter or any low-range d.c. instrument.

When the line is perfectly matched the current at the input terminals is equal to $\sqrt{P/Z_0}$, where P is the power and Z_0 is the characteristic impedance of the line. If there are standing waves on the line this relationship does not hold. In such a case the current that will flow into the line is determined by the line length, s.w.r., and whether the antenna impedance is higher or lower than the line impedance. Fig. 3-22 shows how the maximum current to be expected will vary with the standing-wave ratio. This information can be used in selecting the proper range for the ammeter.

For example, assume that a 600-ohm open-wire line is to be used and that the power is 500 watts. The antenna is to be an all-band affair and the highest s.w.r. to be expected is about 10 to 1. With perfect matching, the current would be $\sqrt{500/600} = 0.91$ amp. From Fig. 3-22, the current at a loop will be 3.16 times the matched current when the s.w.r. is 10, so the maximum current to be expected under any conditions is 3.16 $\times$ 0.91 = 2.9 amp. An instrument having a range of 0–3 or 0–4 amperes would be suitable. The minimum current to be expected would be the maximum divided by the s.w.r., or 2.9/10 = 0.29 amp.

Two ammeters, one in each line conductor, have been shown in the balanced-line circuits. The use of two instruments gives a check on the line balance, since the currents should be the same. However, one meter can be switched from one conductor to the other. If only one instrument is used it is preferably left out of the circuit except when adjustments are being made, since it will add capacitance to the side in which it is inserted and thus cause some unbalance. This is particularly important when the instrument is mounted on a metal panel. When coaxial line is used the meter can be left in at all times, since it merely adds a little shunt capacitance that is easily compensated for in the tuning.

Since the resistive component of the input impedance of a resonant line is seldom known accurately, the r.f. current is of little value as a check on power input to the line. However, it shows in a relative way the efficiency of the system as a whole. The set of coupling adjustments that results in the largest line current with the least plate current is the one that delivers the greatest power to the antenna with the lowest plate power input. Also, in case an amplifier is not operated properly — for example, the grid excitation may be low — the r.f. meter will show when maximum power output has been obtained. In such cases maximum output, as shown by the largest obtainable r.f. meter reading, may occur at a value of plate current less than the rated plate current of the amplifier tubes.

For adjustment purposes, it is possible to substitute small flashlight lamps for the r.f. ammeters. Their relative brightness shows when the current increases or decreases. They have the advantage of cheapness and such small physical size that they do not unbalance the input circuit.

Standing-Wave Measurements

Two general types of instruments are used in making standing-wave measurements. One type measures the current or voltage along the transmission line, either directly or indirectly. With this type of indicator it is possible to locate the points of maximum and minimum current or voltage, as well as to determine the s.w.r. when the instrument is properly calibrated.

The second type is the "directional bridge," or balancing circuit, which responds to power traveling in one direction along the line but not in the other. By measuring the amount of power traveling in each direction the standing-wave ratio can be determined. When properly calibrated, instruments of this type are direct reading. However, they cannot be used for locating loops and nodes along the transmission line.

Both types have their applications. The current or voltage indicator is especially useful for checking the resonant frequency of an antenna by the method described earlier in this chapter, as well as in matching procedures based on the positions of loops and nodes along the transmission line. The bridge-type indicator saves time and effort during adjustment of matching systems in procedures where it is not necessary to determine the positions of loops and nodes, since it does not have to be moved along the line to check the s.w.r. Consequently, it will show instantly whether an adjustment has resulted in an improvement.

VOLTAGE AND CURRENT INDICATORS

Any sort of indicator that will respond more or less linearly to r.f. voltage or current can be used to detect standing waves, if only a rough check is needed. Many amateurs use a neon bulb for this purpose. If the transmitter power is high enough to make the bulb glow, the glow will increase as the bulb is moved along the line toward a voltage loop (current node) and decrease as it is moved toward a voltage node (current loop). It is not possible to make real measurements with such a simple indicator, but the bulb readily will show the effect of adjustments to the matching system. When the s.w.r. is high the bulb will glow brightly when held near the line at a voltage loop, and dimly or not at all near a current loop. As the s.w.r. is brought down by a better match between the antenna and the line there will be less variation in brightness as the bulb is moved along the line. If there is no observable variation, or if the brightness simply drops off without showing peaks and valleys as the bulb is moved along the line from the transmitter to the antenna, the impedance match can be considered to be quite satisfactory.

A less sensitive indicator, but one that will work if the transmitter power is high enough, is a flashlight bulb or dial light connected to a one- or two-turn loop a few inches in diameter. The lamp will be brightest when coupled to the line at a current loop, and least bright when coupled at a voltage loop. Uniform brightness as the loop is moved along the line, keeping the same coupling to the line at all points, indicates that the s.w.r. is small. This type of indicator will work with open-wire lines, but cannot be coupled closely enough to Twin-Lead to give an indication.

As an alternative to the flashlight lamp, a thermomilliammeter can be connected to a similar loop to indicate current along the line. Such meters are expensive, but have the advantage that the relative value of line current can be measured providing care is taken to keep the coupling between the loop and line exactly the same at all points.

Wavemeter Indicator

A sensitive indicator, and one that can be calibrated for actual measurement of s.w.r., is shown in Fig. 3-82. It uses a small pick-up loop with one-half to one inch of conductor parallel to the line conductor or conductors. The r.f. output of the loop is fed through a flexible r.f. line of low impedance, such as receiving-type 75-ohm Twin-Lead, to the coupling link L_2. L_1C_1 is a circuit capable of being tuned to the frequency in use; the L/C ratio is not critical and any convenient values can be used. The r.f. voltage developed across the tuned circuit is rectified by the crystal detector and the resulting current read by the meter M, which can have a range of 0–1 ma. or less. The 75-ohm line can have any convenient length, enough so that the "trolley" shown in the drawing can be moved along the line independently of the meter itself. The trolley construction is used so that the coupling between the loop and the line will be constant.

The coupling between L_1 and L_2 should be adjusted so that the meter does not go off scale when L_1C_1 is tuned to resonance. An indicator of this type responds to current, and so will give maximum readings at a current loop and minimum readings at a current node. It has the advantage that because of the selective circuit it will respond only to the frequency to which it is tuned. The simple voltage and current indicators discussed above will respond to whatever harmonic voltage and current may be present, in addition to the fundamental, and thus may give inaccurate indications. This is particularly true of voltage indicators that are nonselective, since for a given physical separation between the indicator and the line the coupling becomes greater the higher the frequency.

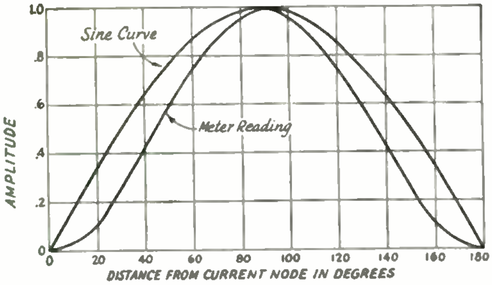


Fig. 3-83 — Typical curve obtained by the standing-wave method described in the text. Comparison with the sine curve shows the correction that must be applied to the meter reading to calibrate it for standing-wave measurements.

Calibration

A crystal-detector indicator is generally not linear — that is, the current flowing in the meter circuit is not directly proportional to the voltage developed in L_1C_1 — unless the resistance, *R*, is 10,000 ohms or more. This reduces the sensitivity, but the use of series resistance is recommended if the lower sensitivity can be tolerated. Alternatively, *R* may be reduced or omitted entirely, in which case the instrument may be calibrated by the following method: Make up a temporary transmission line about a wavelength long, short it at the far end, and feed in a little power from the transmitter at the input end. With the coupling between L_1 and L_2 adjusted so that the meter reads exactly full scale at a current loop near the transmitter, move the indicator along the line and locate accurately the positions of two consecutive nodes. Measure the distance between them, and then convert the length into inches per degree by dividing the length in inches by 180. Starting at a node, measure the current every 10 or 15 degrees along the line until the next node is reached, then plot the results on graph paper. Plot a half sine wave on the same paper, as shown in the typical curve given in Fig. 3-83. As a convenience, plot another curve showing the relative r.f. current, as given by the sine curve, against the corresponding meter readings. Such a curve is shown in Fig. 3-84. When making s.w.r. calculations the values given by the correction curve should be used.

The meter readings in Fig. 3-83 were obtained experimentally, using a 0–1 milliammeter with no resistance in series. Without calibration, the s.w.r. as computed from the meter readings always is higher than the actual s.w.r. This discrepancy increases with the magnitude of the s.w.r. As shown by Fig. 3-84, the maximum actual s.w.r. that can be read with any pretense of accuracy is about 5 to 1, but from the meter readings alone it would be calculated to be 1.0/0.025 or 40 to 1.

This method of calibration is based on the fact that the current distribution in a low-loss short-circuited transmission line is sinusoidal, for all practical purposes. If the line loss is low and there are no "antenna" currents on the line, the nulls will be quite sharp and the readings on one side of the current loop will duplicate those on the other side, at equal distances from the current loop. If there is any considerable discrepancy (more than a few per cent) between the two sets of readings there is probably a parallel component of current on the line. To avoid it, choose a line length that, considering the two wires to be in parallel and forming a single conductor, is not resonant. This means that the physical length of the line should not be close to a multiple of a quarter wavelength in space.

In using the instrument for checking standing waves it is necessary to avoid direct pick-up of r.f. energy by the tuned circuit. It is therefore advisable to build the circuit in a metal box. Also, loose coupling between L_1 and L_2 is desirable because it helps eliminate any "antenna" effect of the 75-ohm coupling line. The meter should give no indication until the pick-up loop is brought quite close to the transmission line being checked; a residual reading from strong fields about the antenna will obscure the readings from the transmission line itself.

BRIDGE-TYPE INDICATORS

Bridge-type indicators are the only kind that it is practicable to use with coaxial line, since there is no way in which to move a current or voltage indicator inside such a line over the distances necessary, at low frequencies, for locating loops

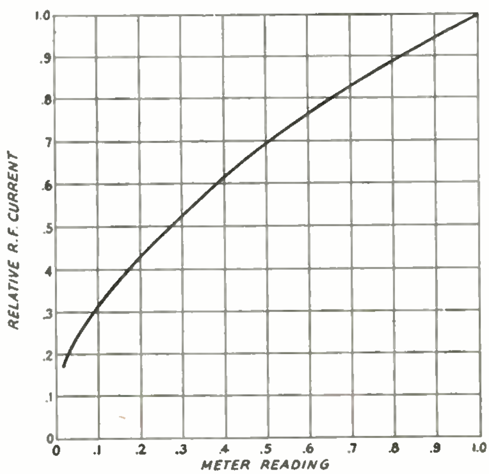


Fig. 3-84 — Calibration curve for the experimentally-determined readings shown in Fig. 3-83. Without calibration the instrument is useful for checking the effect of matching adjustments, since the better the impedance match the smaller the current variation along the line, but cannot be relied upon for measurement of the actual value of s.w.r.

and nodes. The principle on which such bridges operate is that of balancing a circuit so that a voltmeter will not respond to the voltage of a wave traveling in one direction along the line, but will respond to the voltage of a wave traveling in the other direction. As explained earlier in this chapter, there will be no reflection from a load equal to the Z_0 of the line, and so when the line is perfectly matched a bridge connected to

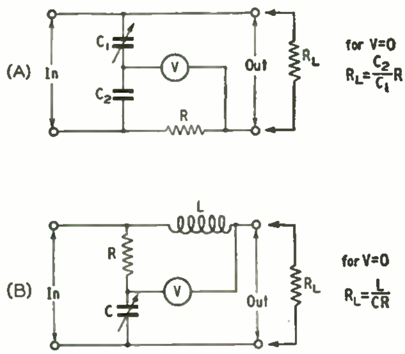


Fig. 3-85 — **Bridge circuits suitable for measurement of standing-wave ratio. (A) resistance-capacitance bridge; (B) Maxwell bridge. Equations for balance (null indication on voltmeter) are given to the right of each circuit.**

read voltage returning from the load toward the transmitter will show no indication. If a returning voltage can be measured at all, there are standing waves on the line. The relative amplitudes of the outgoing and returning voltages give a measure of the standing-wave ratio.

Two fundamental types of bridge circuit capable of giving readings proportional to the outgoing and reflected voltages independently are shown in Fig. 3-85. If stray capacitance, inductance, and coupling between parts of the circuit can be kept to negligible proportions, both circuits will function at any frequency. In practice, they can be made to work satisfactorily at frequencies up to 30 Mc. if sufficient care is used in the selection of components and in the construction of the bridge. This requires that the circuit elements be as nearly "pure" as possible. That is, condensers must have low inductance; inductances, if used, must have low distributed capacitance; resistors must be noninductive and have low shunt capacitance. Half-watt and one-watt carbon resistors are generally satisfactory in the low-resistance values.

In the construction of a bridge particular care must be taken to keep the leads short, since a relatively small length of wire will introduce enough unwanted reactance at the higher frequencies to have an appreciable effect on the accuracy. In laying out the parts every effort should be made to avoid forming loops through which r.f. can flow and induce a voltage in any other part of the circuit. The crystal rectifier and the voltmeter circuit, especially, should be so placed that the r.f. circuits cannot induce voltage in a loop inadvertently formed by the rectifier leads.

In measuring an s.w.r. that is higher than 1, the bridge is in an unbalanced condition and current flows through the voltmeter. The higher the reading on the voltmeter the more serious the error introduced by such current flow. Therefore the voltmeter impedance must be very much higher than the impedance of any arm of the bridge. The voltmeter impedance will be increased by using resistance in series with the milliammeter; it is advisable to use the highest value of series resistance that will permit full-scale readings at the r.f. power level used. In addition, the voltmeter should be linear. As discussed in connection with the wavemeter indicator described earlier, a resistance of 10,000 ohms or more in series with the meter is desirable, to overcome the nonlinearity of the crystal rectifier.

Balancing

The equations to the right of the diagrams in Fig. 3-85 show the relationships between L, C and R in the bridge, for proper balance when a purely-resistive load of value R_L is connected to the terminals marked "Out" while power is being applied to the terminals marked "In." When these equations are satisfied the voltmeter will read zero, indicating a standing-wave ratio of 1.

Practical circuits corresponding to those in Fig. 3-85 are given in Figs. 3-86 and 3-87. To check either bridge for a desired value of line characteristic impedance it is necessary to have a noninductive resistor of the same resistance as the Z_0 of the line. The resistor is first connected to the "In" terminals and enough r.f. power applied at the "Out" terminals to cause a full-scale reading on the voltmeter. Then the resistor is connected to the "Out" terminals and

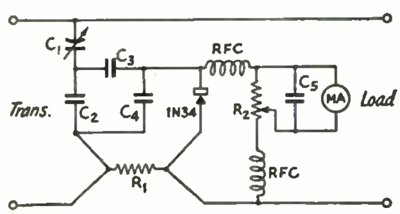


Fig. 3-86 — **Practical circuit diagram of the *RC* bridge ("Micro-Match") standing-wave indicator.**
C_1 — 3–15-μμfd. midget variable.
C_2, C_4 — 220-μμfd. mica.
C_3 — 82-μμfd. mica.
C_5 — 0.0047-μfd. mica.
R_1 — 1.1-ohm resistor (nine 10-ohm 1-watt carbon resistors in parallel).
R_2 — 5000-ohm potentiometer.
MA — 0–1 d.c. milliammeter.
RFC — 2.5-mh. r.f. choke.

the same power applied to the "In" terminals. C_1 in either circuit is then adjusted to make the voltmeter reading as low as possible. If the bridge is carefully constructed as described above, the reading will be zero, providing the test resistor has negligible inductance and capacitance. If the reading is only one or two per cent of the full-scale reading the instrument can be considered satisfactory at that frequency, but if the reading is not fairly close to zero a different arrangement of parts should be tried, with the object of reducing the stray couplings mentioned above.

Once operating satisfactorily with a nonreactive resistor as a load, the bridge is ready for use on the line. Measurements are made in the same way as in calibrating; that is, the line is first connected to the "In" terminals with power applied to the "Out" terminals and the power level adjusted to give a full-scale reading; then the connections are reversed without changing the power. The s.w.r. will be approximately as given by the curve of Fig. 3-88 for either type of bridge.

Calibration

Fig. 3-88 assumes that the voltmeter resistance is high compared with the r.f. impedances, and that the voltmeter is linear. It is advisable to make an actual calibration in terms of s.w.r., where actual measurement of s.w.r. is necessary. This can be done as follows: First, balance the bridge for the line Z_0 by using a nonreactive resistor of the same value as described above. Then, without touching C_1, use a new value of test resistance (nonreactive resistors must be used), adjust the power input for full-scale reading with power applied to the "Out" terminals and the resistor connected across the "In" terminals; then reverse the bridge and note the meter reading. The s.w.r. at that reading is equal to R_2/R_1, where R_1 is the value of the first re-

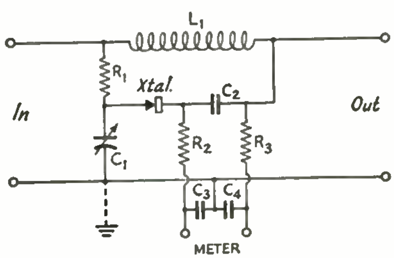


Fig. 3-87 — **Practical circuit diagram of the Maxwell-bridge standing-wave indicator. The meter should have a full-scale range of 1 milliampere or less.**
C_1 — 10–100-μμfd. Ceramicon variable.
C_2 — 470-μμfd. mica.
C_3, C_4 — (Optional) 100-μμfd. mica.
R_1 — 500 ohms, nonreactive.
R_2, R_3 — 10,000 ohms, ½-watt carbon.
L_1 — Approx. 29 turns No. 18, diameter 0.6 inch, 2.5 inches long.
XTAL — 1N34 or equivalent.

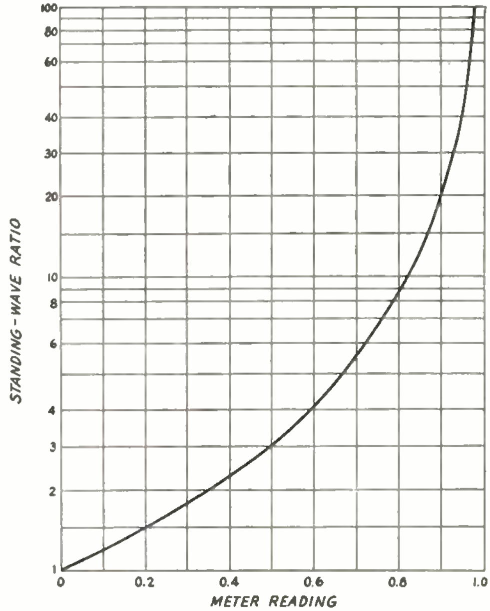


Fig. 3-88 — **Standing-wave ratio as a function of voltmeter reading for the bridge circuits of Figs. 3-86 and 3-87. This curve assumes that the voltmeter impedance does not affect the operation of the bridge and that the voltmeter scale is linear.**

sistor (equal to the line Z_0) and R_2 the value of the second. A calibration curve can be made by using several different resistance values in this fashion.

It is sometimes found that substituting different resistors that should give the same s.w.r., but one higher and the other lower in value than the line Z_0, gives different meter readings. For example, if the bridge is balanced to give zero reading with 70 ohms, either a 14-ohm or 350-ohm resistor should result in the same voltmeter reading. If they do not, the fault may be in the voltmeter (too low resistance and r.f. impedance) or in the bridge itself. It is more difficult to get accurate readings when the bridge "ratio" is high; that is, when the impedances of the arms in one side of the bridge are considerably different from the impedances in the other side. In the circuit of Fig. 3-86 the bridge ratio is high, since the resistance in the arm in series with the load is only about an ohm. In the circuit of Fig. 3-87 the bridge ratio varies with the applied frequency and the line impedance for which it is balanced, but should not exceed about 10 to 1 under balanced conditions.

An accurate calibration is not at all essential if the bridge is to be used simply as an aid to adjusting a matching system. In such case the object is to get the s.w.r. as low as possible and not particularly to find its exact value. The s.w.r. will be lowest when the meter reading is lowest.

Errors in measurement increase with the value of s.w.r., and as shown by Fig. 3-88 the readings are quite "broad" at low values of s.w.r. For all practical purposes, a reading that does not exceed 20 per cent of full-scale value indicates that the s.w.r. is low enough to cause no appreciable increase in line loss as compared with exact matching.

The bridge shown in Fig. 3-86 can be adjusted for balance in the range of line impedances from 70 ohms to 300 ohms, and the resistor is capable of carrying a current of a little less than 3 amperes. The Maxwell bridge, with the constants given, will work over the range 50-500 ohms, and the amount of power that can be handled is limited by the power capacity of the resistor. The power dissipated in the resistor will be largest under conditions that make the voltage across the bridge reach high values. This will occur when the s.w.r. is high and there happens to be a voltage loop at the line input terminals. The voltage also will increase with the value of Z_0, for a given input power. If R_1 is a 1-watt resistor the power should not exceed a watt or two when Z_0 is near 500 ohms, but much more power can be safely handled when Z_0 is below 100 ohms.

Resistance Bridge

Fig. 3-89 is the circuit of a resistance-arm bridge that is particularly useful for coaxial lines. Since all arms are fixed in value, the bridge must be built for a particular line impedance (50 or 75 ohms in the usual case) and cannot be adjusted like the bridges previously described. The same precautions with respect to construction should be followed, although layout is somewhat less critical in this type of circuit so long as the shielding shown in the diagram is employed.

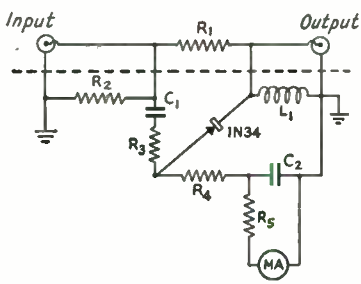


Fig. 3-89 — Circuit diagram of resistance bridge for measuring s.w.r., as adapted for coaxial lines. This circuit operates at very low power level and provision must be made for reducing the transmitter power to a low value when using it. The part of the circuit shown above the dotted line should be shielded from the remainder of the circuit.

C_1, C_2 — 470-$\mu\mu$fd. mica.
R_1 — 1-watt composition resistor, value equal to impedance of line being measured.
R_2 — 10 ohms, 1 watt.
R_3, R_4 — 56-ohm 1-watt composition. Exact value not important but the two resistors must have the same value.
R_5 — 470 ohms, ½ watt.
L_1 — Good r.f. choke at operating frequency. Not required if antenna system is closed type that offers d.c. return. At 28 Mc., 40 turns of No. 36 d.c.c. wound on a 1-watt 0.1-megohm resistor is satisfactory, or a regular 2.5-mh. choke may be used.
MA — 0-1 milliammeter.

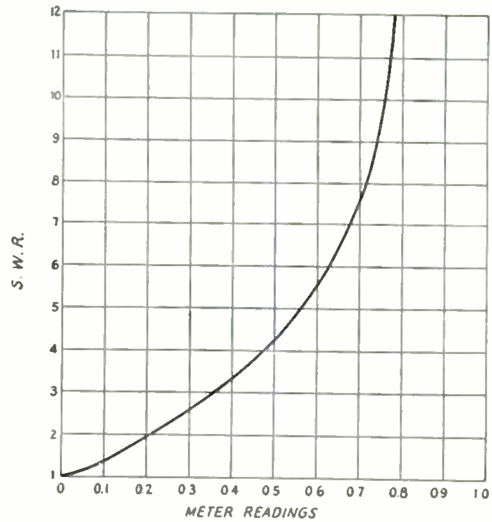


Fig. 3-90 — Calibration curve of the bridge-type s.w.r. indicator of Fig. 3-89. This curve will not apply if other circuit constants are used.

The bridge will be in balance when a pure resistance equal to R_1 is connected across the output terminals, and under these conditions there will be zero current through the milliammeter. Hence, to fit the bridge to the characteristic impedance of the transmission line to be used the resistance of R_1 must equal the line Z_0. Just as in the case of the other bridges, reflected waves on the line will cause its input impedance to differ from Z_0 and the meter will show an indication. Fig. 3-90 is a calibration curve for a bridge having the constants shown in Fig. 3-89.

This type of bridge does not require reversing the line and load connections to establish a reference voltage. It is only necessary to open-circuit the output terminals, adjust the voltage applied to the input terminals to give a full-scale reading on the milliammeter, and then connect the line to the output terminals. The s.w.r. is then indicated by the meter reading. Alternatively, the output terminals may be short-circuited instead of open-circuited when adjusting the input voltage to give the full-scale reading.

In either case the voltage regulation of the source of r.f. voltage must be good enough so that the output voltage does not change when the load is connected; the purpose of R_2 is to place a low-resistance load on the source so that changing the load connections will not change the voltage. The voltage regulation can be checked by setting the meter to full scale with the output terminals open, and then shorting them. If there is a change in meter reading the regulation is not good enough, or the construction is such that stray capacitance or inductance is affecting the readings. Inspection of Fig. 3-89 will show that in the one case the meter is reading the voltage across R_3 and in the other case is reading the voltage across R_4. Since both resistors have the same value the readings should be the same in both cases.

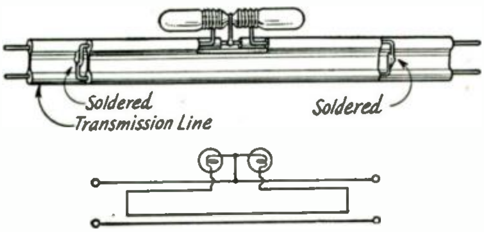


Fig. 3-91 — The "twin-lamp" standing-wave indicator. It requires only a few inches of 300-ohm line and two low-current dial lights. The electrical circuit is given in the lower drawing.

Because suitable nonreactive resistors are available only in small power ratings, the power capacity of a bridge of this type is limited to a watt or so. The sensitivity is high because the bridge ratio is near 1 to 1. Measurements should be made with a very low power source of r.f.; frequently, the leakage from a buffer stage in the transmitter will supply enough power for a full-scale reading.

This type of bridge may be calibrated by using several values of nonreactive resistors for the load, as described above in connection with the other bridges. Reversal of connections is not necessary in this case, however.

The "Twin-Lamp"

Fig. 3-91 shows a very popular type of standing-wave indicator for 300-ohm line. It is exceptionally simple to build and costs almost nothing. It works on the principle of separating the outgoing power from the reflected power on the line, and accomplishes this by a combination of inductive and capacitive coupling. When fastened to the transmission line the lamp nearest the transmitter will light when power is put into the line, but if the s.w.r. is close to 1 the lamp nearest the antenna will be dark. If the s.w.r. is high, both lamps will light to about the same brilliance.

The construction of the "twin-lamp" is shown in the drawing. It is made from a short length of Twin-Lead of the same type used in the transmission line, with the ends of the wires soldered together to form a loop. The length may be anything from a few inches to a foot or two, more length being required when the transmitter power is low, or when the device is used at the lower frequencies. At the exact center one of the conductors should be cut and the ends separated enough to permit soldering to the lamp-base shells. The tips of the lamp bases are then soldered together and to one conductor of the transmission line as shown. The small drawing gives the circuit.

The dial lights should be of the low-current type (60 ma.). To check the s.w.r., adjust the transmitter power until the lamp nearest the transmitter is burning at normal brightness. If the other lamp does not light, or lights only dimly, the s.w.r. will be below 2 to 1 in the average case. The "twin-lamp" should be fastened to the line with Scotch Tape to make sure that it is held in the proper position. As a convenience, a "twin-lamp" can be assembled on a short length of 300-ohm line which then can be inserted in the regular line whenever the s.w.r. is to be checked.

A coaxial version of the "twin-lamp" is shown in Fig. 3-92. In this case it is necessary to remove a section of the outer covering and shield braid, as shown at B in the drawing. The loop is a 12-inch length of 75-ohm Twin-Lead with the ends soldered together and cut at the center in the same way as the 300-ohm lead described above. This pick-up loop is then fitted into a groove cut in the polyethylene insulation in the cable so that it is parallel to and close to the inner conductor. A slot is cut in the braid as shown at C, and the braid is then slipped over the assembly, taking care that the leads from the loop do not touch it. The lamps are then soldered in place, with the common connection going to the shield braid, after which the vinyl covering can be put back and the whole assembly taped together. The section of cable should be fitted with connectors so it can be inserted in the regular line for making measurements.

Although the "twin-lamp" is not adapted to actual measurement of s.w.r., it is a thoroughly practical device for showing the effect of changes in matching adjustments, and for obtaining a rough idea of whether the s.w.r. is high or low.

ERRORS IN STANDING-WAVE MEASUREMENTS

It has been stated several times in this section that it is seldom necessary to measure the standing-wave ratio to any degree of accuracy. For the practical purpose of determining the optimum adjustment of a matching system a purely qualitative check on the s.w.r. is entirely sufficient. Exact measurement is necessary only when the s.w.r. enters into the calculation of antenna or matching-system constants.

For example, a little thought will show that if the s.w.r., line length and line Z_0 are known accurately the impedance of the antenna system to which the line is connected can be calculated to an equal degree of accuracy. This is of interest

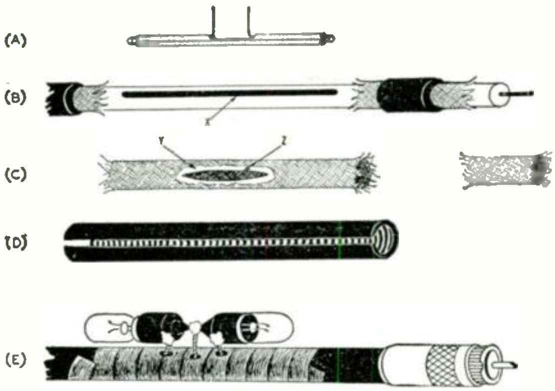


Fig. 3-92 — Steps in the construction of the "twin-lamp" for coaxial cable. The pick-up loop (A) is a length (approximately 12 inches) of 75-ohm Twin-Lead with the ends soldered together and covered with insulating material. There is no connection between the lamp circuit and the line except at the common connection between the bulbs; this point may be joined to the outer braid.

when the properties of an antenna are being investigated, but is not essential when the sole object is to get the system "tuned up" for best performance. Again, reasonably-accurate measurement of the s.w.r. is necessary if matching stubs are to be designed according to the charts given earlier in this chapter. But in most cases the antenna impedance will be known to a sufficiently close degree to make it possible to choose a matching system intelligently; then, after installation, a qualitative check will be enough to show whether any adjustment is required and, if so, what the effect of trial adjustments is

Antenna Currents

It is possible for considerable error to occur in making s.w.r. measurements whether the standing-wave indicator is accurately calibrated or not. The chief source of trouble is unbalance and the presence of antenna currents on the line. With coaxial lines these effects are not so important. The line is normally unbalanced or "single-ended" and thus one side can be grounded, thereby eliminating the effect of stray capacitances. Also, the antenna currents on coaxial lines flow on the *outside* of the shield and do not affect measurement of current or voltage inside the cable, so long as the measuring equipment itself also is shielded.

On parallel-conductor lines conditions are not so favorable. Antenna currents flow in parallel on both wires, leading to the representative current distribution shown in Fig. 3-36. It can be appreciated that even if the line is perfectly matched and there are no standing waves in the transmission-line current, a standing-wave indicator of the voltage or current type will respond to the standing waves in the antenna-current component on the line.

When such standing waves are present the current loops (and nodes) do not occur exactly opposite each other on the two line conductors. If measurement along each conductor separately shows this to be so, the s.w.r. as given by the instrument is meaningless. When the line is of open-wire construction it is possible to check each conductor separately, but with Twin-Lead it is difficult because the wires are so closely spaced. With this type of line the way to check for the presence of antenna currents is to take readings at equal small intervals along the line for a distance of at least a wavelength and plot the data graphically. If the resulting curve is irregular and does not show alternate loops and nodes at intervals corresponding to a quarter wavelength, taking the line velocity factor into account, there is an appreciable component of antenna current on the line. Before the actual s.w.r. can be checked this antenna current must be reduced to the point where it does not affect the measurements.

Bridge Indicators

In all bridge-type s.w.r. indicators there is inherently some unbalance with respect to ground, since the line is not symmetrical with respect to the input terminals of the bridge. Considering the unbalances (including those in the coupling circuit) to be lumped into equivalent capacitances, Fig. 3-93 shows that there are three such capacitances to be considered. C_1 is the stray capacitance to ground from the upper side of the

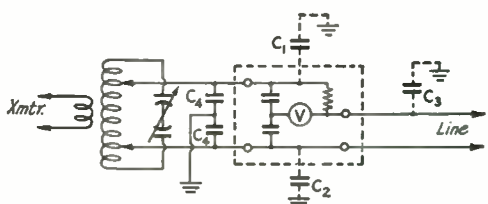


Fig. 3-93 — Stray capacitances that affect the accuracy of bridge measurements of s.w.r. when there are antenna currents on the transmission line. When the bridge operates at low power, condensers (C_4) connected across the input terminals as shown will help reduce the effects of unbalance. These may be mica condensers of about 500 $\mu\mu$fd. capacitance; the value is not critical but both condensers should be the same within ordinary tolerances.

bridge, C_2 is the sum of the capacitances of the lower side of the bridge and the lower line wire, and C_3 is the stray capacitance to ground from the upper line wire. An antenna current on the line will cause equal voltages to appear at the line terminals, if the line itself is perfectly balanced. But these voltages do not appear equally at the voltmeter terminals unless (1) C_3 and C_2 are substantially equal, so that the voltage is the same at both terminals; (2) the voltmeter impedance is high compared with the impedances of the bridge arms; (3) C_1 and C_2 are identical in value. If these conditions are not met some of the antenna current will flow through the voltmeter.

Unbalance can be checked simply by interchanging the line wires at the output terminals. If the system is well balanced the readings both ways will not differ by more than a few per cent of the full-scale reading of the voltmeter. However, it is not unusual to find that the s.w.r. is indicated to be nearly 1 to 1 with the line connected one way and several times as high when the connections are interchanged. When such a condition exists neither reading is dependable. The line must be detuned for antenna currents before attempting to make any measurements. It is helpful to make the over-all length (see Fig. 3-37) equal to an odd multiple of a quarter wavelength (using $234/f$ for a quarter wave) so that the voltage from antenna currents at the output terminals of the bridge will be minimum. Another expedient that is useful when the bridge operates at low power is to connect two condensers, C_4 in Fig. 3-93, in series across the input terminals of the bridge and ground the common connection between them. This tends to equalize the paths to ground for parallel currents. It may be necessary to try several different grounds to find the one that gives the best balance.

The preceding discussion has been in terms of parallel currents induced on the line by the field from the antenna. However, it applies equally well to parallel currents flowing into the line because of capacitive coupling between the transmitter tank and the coupling tank. Also, the presence of harmonics of appreciable amplitude will affect the accuracy of measurement. If the harmonic currents are of the parallel type, which is frequently the case when there is capacitive coupling between the transmitter and coupling tanks, they will cause the meter readings to differ when the line conductors are interchanged. Harmonic currents of the transmission-line type will simply lead to an inaccurate s.w.r. measurement, since the system probably will not be as well matched for harmonics as it is for the fundamental.

Harmonic Reduction

The increasing congestion in the radio-frequency spectrum makes it more necessary, as time goes on, to confine all the radiation from a transmitter to the frequency to which it is assigned. Radiation can take place at any frequency that is generated in the course of multiplication from the primary frequency control to the output amplifier. The strongest of these radiations usually are harmonics of the final frequency, and special attention must be given to reducing the intensity of radiated harmonics. This problem is particularly acute when, as in the case of television broadcasting, a near-by receiver is tuned to a channel in which the transmitter harmonic falls.

Transmitter harmonics that cause interference within a radius of several hundred yards are not always radiated by the antenna system. When the transmitter and receiver are in close proximity stray radiation from the transmitter wiring itself may be the principal source of interference. It must be emphasized that in such cases attention must be paid to harmonic reduction in the transmitter itself. Otherwise, means taken to prevent harmonic currents from flowing in the antenna system cannot be expected to effect any material improvement.

Antenna Systems

Antenna systems differ in their ability to discriminate against transmitter harmonics. Multiband antenna systems do not discriminate against harmonics at all, except insofar as the coupling system may be detuned for a particular harmonic. Systems that are designed primarily for operation on one band, and incorporate a matching system for nonresonant transmission-line operation, usually will discriminate against harmonics to a considerable degree. In general, such systems will have rather high discrimination against even harmonics, but sometimes are fairly well matched to the line at odd harmonics.

The folded dipole is essentially nothing but a transmission line at all even harmonics of the frequency for which it is designed, and so will not radiate such harmonics if they appear at the antenna as true transmission-line currents. At odd harmonics the folded dipole represents a fairly good impedance match for the line and so will be a good radiator.

Coupling to the Line at Harmonic Frequencies

Harmonics can be coupled into a transmission line in the same way that the fundamental power is coupled in, so long as the harmonics are present in the transmitter tank circuit. The coupling circuit will suppress them to an extent depending on its over-all selectivity. For this reason alone it is

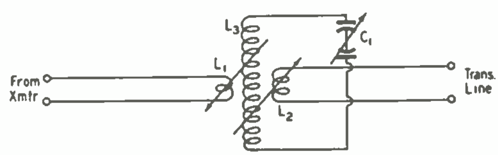


Fig. 3-94 — Double link coupling in the antenna tuner. With good separation between the coils the capacitive coupling between the link coil and transmission line is minimized.

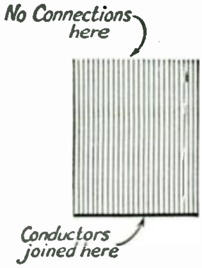


Fig. 3-95 — The Faraday shield, for providing electric but not magnetic shielding. The conductors should be small — about No. 18 wire — when the shield is used between coils of conventional diameter, and the spacing should be about equal to the wire diameter. The shield dimensions should be at least twice the diameter of the coils between which it is placed. It must be grounded.

highly desirable to use a line coupling circuit that is resonant in itself and is inductively coupled to the transmitter tank circuit. The reason is simply that two tuned circuits of equal Q have much higher selectivity than one circuit alone.

However, harmonics also can be coupled to the line through stray capacitances between coils and other parts of the final tank and antenna circuits. This capacitive coupling is smaller the greater the physical separation of the tuned-circuit coils. Greater separation can be used when the circuits operate at reasonably high Q — 10 to 20 — and so high-Q tuned circuits are useful both in reducing stray coupling and in adding selectivity of the ordinary type. It is advisable to use link coupling between the final tank and the antenna coupling circuit; the link coils, having only a few turns, will have relatively small capacitance to the coils in the tuned circuits and can be placed at points where the tuned-circuit coils are practically at ground potential. This, plus the fact that the stray capacitances at each end of the link circuit are in series, results in a worthwhile reduction in capacitive coupling.

An extension of the principle of using inductively-coupled circuits to reduce capacitive coupling is shown in Fig. 3-94. L_1 and L_3 are the link and tank coils of an antenna coupler of the type described earlier in this chapter. Instead of tapping the transmission line on L_3, it is coupled to L_3 by means of a separate coil, L_2. The number of turns required at L_2 varies with the frequency and the input impedance of the line, and is best determined experimentally. The three coils should be arranged so that L_1 and L_2, while both coupled to L_3, are not physically close to each other.

Capacitive coupling may be further reduced by using shielding between coils. The ordinary type of shielding will not serve, since it is a shield for both electric and magnetic fields. It is necessary to use a form of shield that will prevent electric lines of force from extending from one coil to the other, but will at the same time have little or no effect on the magnetic field. Such a shield can be constructed as shown in Fig. 3-95. It consists of a group of parallel conductors connected together at one end but not at the other. For the electric field it is practically equivalent to a solid sheet of metal. However, effective shielding of the magnetic field at radio frequencies depends on current flow in the shield, and with a shield of this type the only possible way for current to flow, when the shield is placed between two coupled coils, is across the conductor. The result is that the current is very small and the shield has little or no effect on the magnetic field. To shield a link coil that couples to the center of a tank coil two such shields must be used, one on each side of the link coil. They may be mounted on a swinging link assembly so that they move with the link coil.

A scheme that is similar in principle, and that can be applied readily to single-turn coils such as are used for link coupling at the higher frequencies, is shown in Fig. 3-96. The link is made of small-diameter solid-dielectric coaxial line, with the inner conductor making a one-turn loop back to the outer conductor. The latter is cut back so that it does not form a complete loop, hence the magnetic field will cause relatively little current to flow in it. However, it serves as an effective shield for the electric field when the outside of the cable is grounded, and thereby reduces capacitive coupling between the tank coil and link.

Parallel Currents

When energy is transferred to the transmission line by stray capacitive coupling the currents in both line wires are in the same phase; that is, the line simply acts like two conductors in parallel. In such a case the earlier remarks about the ability of different antenna systems to discriminate

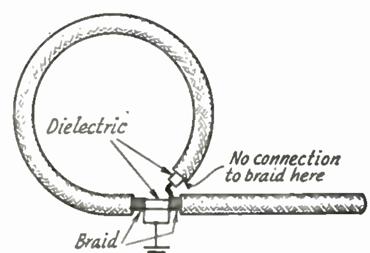


Fig. 3-96 — Shielded link coil made from coaxial cable. RG-59/U is satisfactory and can be formed into a turn of fairly small diameter. The outer conductor is opened at the point where the inner conductor returns so that current can flow inside the cable without having to flow back over the turn.

against harmonics simply do not apply. The system consisting of the line and antenna then acts as though the antenna terminals were joined together and fed by a single wire having the same length as the transmission line.

Whether such a system is resonant near a harmonic frequency depends on the length L in Fig. 3-37. It also depends on the arrangement of the coupling circuit and whether or not that circuit is grounded. If the whole system can be detuned at a particular harmonic the current at that harmonic will be small. However, when the harmonic is in the v.h.f. range and the over-all length L is large (many wavelengths at the harmonic frequency) a small change in length will carry the system from one resonance point to the next. Fig. 2-9, Chapter Two, explains how this comes about. If a particular harmonic is causing interference its intensity can be reduced, when the cause is capacitive coupling between the transmitter and line, by changing the line length a few inches at a time to determine the length at which the harmonic radiation is minimum. However, a length that reduces one harmonic may increase the radiation at another.

Linear Traps

The properties of short-circuited transmission lines can be used to suppress harmonics on a regular transmission line, providing the harmonic currents are true transmission-line currents. When the harmonics travel as parallel currents line traps have no effect.

It was explained earlier that the input impedance of a shorted quarter-wave line is resistive and is extremely high. Because of its high impedance, such a line can be connected across a

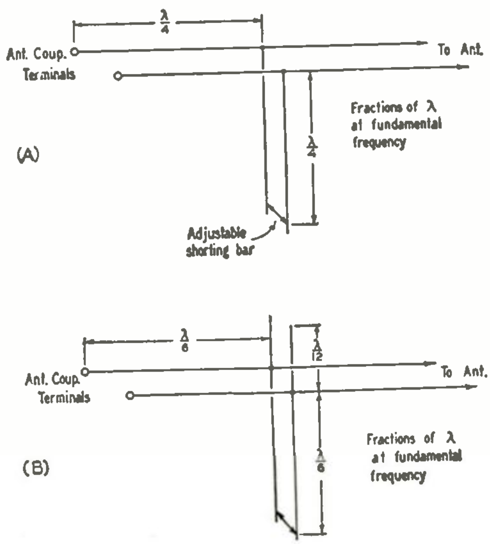


Fig. 3-97 — Shorted quarter-wave line sections as linear traps for harmonic suppression. A, trap dimensions for suppressing the second and all even harmonics. B, dimensions for a third-harmonic trap.

transmission line at any point along its length without affecting the line operation. However, at twice the frequency for which the quarter-wave line was designed it is a half wavelength long and, being shorted at its far end, "looks like" a short-circuit at its input terminals. Consequently it will short circuit a second harmonic traveling along the line to which it is connected. It will similarly act as a short-circuit for any even harmonic, since at such frequencies it is a multiple of a half wave in length.

A shorted quarter-wave line used as a trap for even harmonics is shown in Fig. 3-97A. While it can be placed at any point along the line, it will be most effective when connected at a point where the harmonic standing waves have a voltage loop or, as shown in the drawing, at a point ¼ wavelength from the line coupling apparatus. At even harmonics this distance is a multiple of a half wavelength and so a short-circuit is reflected at the line input terminals for the harmonic frequency.

For trapping out odd harmonics the trap length must be equal to a multiple of a half wavelength at the harmonic frequency. At the third harmonic this length is equal to ⅙ wavelength at the fundamental frequency. A shorted line section of this length has appreciable inductive reactance at the fundamental and so would affect the operation of the main transmission line. To overcome this the length of the trap is made ¼ wave at the fundamental, as shown in Fig. 3-97B, with the line connected ⅙ wavelength from the shorted end of the trap. The quarter-wave trap has a purely-resistive input impedance no matter where the line is tapped on it, and so puts only a resistive load on the line. If the trap has high Q (open-wire construction) the resistance will be sufficiently high to have no effect on the operation of the main transmission line. The open section of the trap is ¼ wave long at the third harmonic and so acts as an additional short across the line at the harmonic frequency.

For other odd harmonics it is necessary to change the position of the line along the trap to make the distance to the shorted end equal to a multiple of ½ wavelength at the particular harmonic considered. In general, it is best to keep the line as near the open end of the trap as possible.

The actual length of a linear trap should be adjusted while a check is made on the strength of the harmonic radiation. A movable shorting bar can be used for adjustment, the proper position being the one that reduces the harmonic strength as much as possible.

Coaxial Cable

When a coaxial transmission line is used there are two separate points to consider, harmonic currents flowing inside the line, and those on the outside of the outer conductor. There will be little or no harmonic current coupled to the outside of the outer conductor from the transmitter if the transmitter and coupling apparatus are shielded and the outer conductor of the line terminates in a grounded coupling where it enters the shield containing the antenna coupler. However, if harmonic currents flow inside the line they can flow back down over the outer conductor from the point where the latter is attached to the antenna.

The line-balancing arrangements shown in Fig. 3-42 will act as a short-circuit for even harmonics and so will tend to prevent such harmonic currents inside the line from reaching either the antenna or the outside of the outer conductor. However, the line balancer is effectively an open circuit for odd harmonics and so has no effect. A separate trap for a particular odd harmonic can be installed, using the same principles discussed above in connection with parallel-conductor lines. In computing the length of such a trap it is necessary to include the velocity factor of the cable, if coax is used for the trap. The junction between the trap and line is most conveniently made by using a small metal box fitted with four male couplings, two for the line going in and out, and two for the two sections of the trap.

Low-Pass Filters

A low-pass filter of the type having a so-called "infinite rejection" frequency may be used to attenuate all harmonics higher than the filter cut-off frequency, with especially great attenuation of one particular harmonic. Simple filters of this type consist merely of a parallel-tuned trap, adjusted to the rejection frequency, connected in series with the transmission line, followed by a shunt condenser to make the characteristic impedance of the filter equal the line impedance at all frequencies below the cut-off frequency.

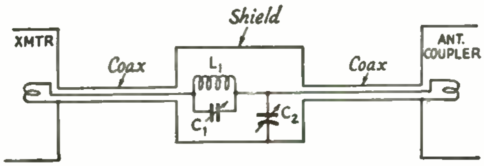


Fig. 3-98 — **Low-pass filter for suppressing harmonics, with maximum rejection at a selected harmonic. Design and adjustment data are given in the text.**

It must be emphasized that filters, like linear traps, will not work if the harmonic radiation is caused by parallel currents on the line. They can only operate when the harmonics to be suppressed are true transmission-line currents. Also, filters must be designed for a specific characteristic impedance and so can be used satisfactorily only with properly-matched nonresonant lines. They are best adapted to use with coaxial lines, partly because of the range of *LC* constants required but chiefly because with the coaxial line only transmission-line currents flow *inside* the line.

Probably the best method of using a low-pass filter is to use it in conjunction with an antenna coupler that is connected to the transmitter through a length of coaxial cable. This is indicated in Fig. 3-98. If the shielding is reasonably complete from the transmitter to the antenna coupler (including the shielding about both these units) there will be relatively little opportunity for harmonics generated in the transmitter to get on the outside of the shielding and thus bypass the filter. Design formulas for filters of this type are as follows:

For 50-ohm line:

$$L_1 = \frac{12}{f}$$

$$C_1 = \frac{2120}{f}$$

$$C_2 = \frac{4770}{f}$$

For 75-ohm line:

$$L_1 = \frac{18}{f}$$

$$C_1 = \frac{1420}{f}$$

$$C_2 = \frac{3180}{f}$$

where f is the frequency (in megacycles) at which maximum rejection is desired, L_1 is in microhenrys, and C_1 and C_2 are in micromicrofarads. The cut-off frequency is equal to $0.8f$; the filter will pass all frequencies below the cut-off frequency with substantially no loss.

In adjusting a filter used as shown in Fig. 3-98, it is advisable first to adjust the coupling apparatus so that the interconnecting or link line is terminated in a resistive load equal to its characteristic impedance. (See discussion earlier in this chapter.) This is best done by using a bridge-type s.w.r. indicator, leaving the filter out of the circuit. When the link line is properly matched, the filter may be inserted and, with the s.w.r. indicator connected between it and the transmitter, adjusted for minimum s.w.r. This is done by adjusting C_2, leaving C_1 at a setting estimated to be close to the design value. After matching, the s.w.r. indicator may be removed and C_1 adjusted to attenuate the harmonic as much as possible, using a near-by receiver to check the setting. If the link line is properly matched at the antenna coupler on each band, the filter will require no readjustment when working on any band below the cut-off frequency of the filter.

Multielement Directive Arrays

The gain and directivity that can be secured by intentionally combining antenna elements into an array represent a worth-while improvement both in transmitting and receiving. Power gain in an antenna is the same as an equivalent increase in the transmitter power. But, unlike increasing the power of one's own transmitter, it works equally well on signals received from the favored direction. In addition, the directivity reduces the strength of signals coming from the directions not favored, and so helps discriminate against a great deal of interference.

One common method of securing gain and directivity is to combine the radiation from a group of half-wave dipoles in such a way as to concentrate it in a desired direction. The way in which such combinations affect the directivity has been explained in Chapter Two. A few words of additional explanation may help make it clear how power gain is achieved.

In Fig. 4-1, imagine that the four circles, A, B, C, D, represent four dipoles so far separated from each other that the coupling between them is negligible. The point P is supposed to be so far away from the dipoles that the distance from P to each one is exactly the same (obviously P would have to be much farther away than it is in this drawing). Under these conditions the fields from all the dipoles will add up at P, if all four are fed r.f. currents in the same phase.

Let us say that a certain current, I, in dipole A will produce a certain value of field strength, E, at the distant point P. The same current in any of the other dipoles will produce the same field at P. Thus if only dipoles A and B are operating, each with a current I, the field at P will be $2E$.

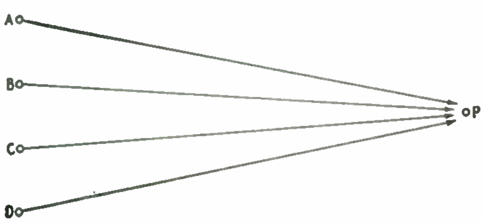


Fig. 4-1 — Fields from separate antennas combine at a distant point, P, to produce a field strength that exceeds the field produced by the same power in a single antenna.

With A, B, and C operating the field will be $3E$, and with all four operating with the same I, the field will be $4E$. Since the power received at P is proportional to the square of the field strength, the relative power received at P is 1, 4, 9 and 16, depending on whether one, two, three or four dipoles are operating.

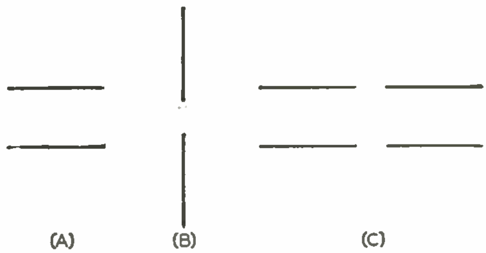


Fig. 4-2 — Parallel (A) and collinear (B) antenna elements. The array shown at C combines both parallel and collinear elements.

Now since all four dipoles are alike and there is no coupling between them, the same power must be put into each in order to cause the current I to flow. For two dipoles the relative power input is 2, for three dipoles it is 3, for four dipoles 4, and so on. The gain in each case is the relative received (or output) power divided by the relative input power. Thus we have:

Dipoles	*Relative Output Power*	*Relative Input Power*	*Power Gain*	*Gain in Db.*
A only	1	1	1	0
A and B	4	2	2	3
A, B and C	9	3	3	4.8
A, B, C and D	16	4	4	6

The power gain is directly proportional to the number of elements used.

It is well to have clearly in mind the conditions under which this relationship is true:

1) The fields from the separate antenna elements must be in phase at the receiving point.

2) The currents in all elements must be identical.

3) The elements must be separated in such a way that the current induced in one by another is negligible; i.e., the radiation resistance of each element must be the same as it would have been had the other elements not been there.

Very few antenna arrays meet all these conditions exactly. However, as a rough approximation it may be said that the power gain of a directive array consisting of dipole elements in which optimum values of element spacing are used is proportional to the number of elements. It is not impossible, though, for an estimate based on this rule to be in error by a factor of 2 or more.

Definitions

The "element" in a multielement directive array is the half-wave dipole. The length is not always an exact electrical half wavelength, because in some types of arrays it is desirable that the element show either inductive or capacitive reactance. However, the departure in length from a true half wave is ordinarily small (not more than 5%, in the usual case) and so has no appreciable effect on the radiating properties of the element.

Antenna elements in multielement arrays of the type considered in this chapter are always

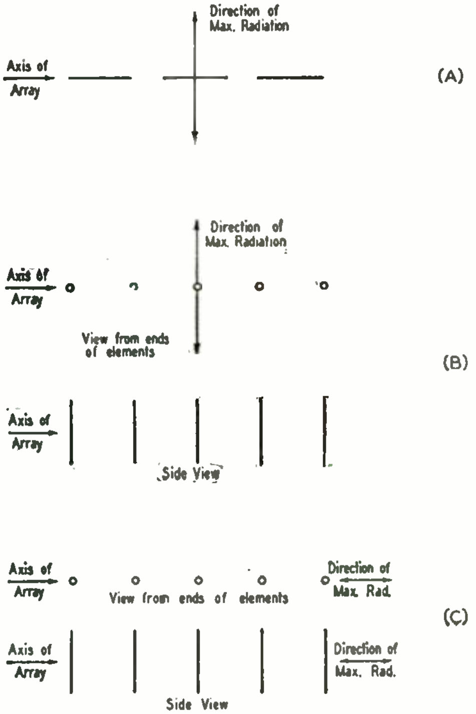


Fig. 4-3 — Representative broadside arrays are shown at A and B, the first with collinear elements, the second with parallel elements. An end-fire array is shown at C. Practical arrays may combine both broadside and end-fire directivity, including both parallel and collinear elements.

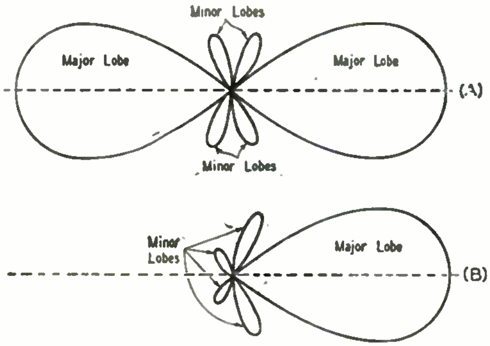


Fig. 4-4 — Typical bidirectional (A) and unidirectional (B) directive patterns. These drawings also illustrate the application of the terms "major" and "minor" to the pattern lobes.

either parallel, as at A in Fig. 4-2, or collinear (end-to-end), Fig. 4-2B. Fig. 4-2C shows an array combining both parallel and collinear elements. The elements can be either horizontal or vertical, depending on whether horizontal or vertical polarization is desired. There is seldom any reason for mixing polarization, so arrays are customarily constructed with all elements similarly polarized.

A **driven element** is one supplied power from the transmitter, usually through a transmission line. A **parasitic element** is one that obtains power solely through coupling to another element in the array because of its proximity to such an element.

A **driven array** is one in which all the elements are driven elements. A **parasitic array** is one in which one or more of the elements are parasitic elements. At least one element in a parasitic array has to be a driven element, since it is necessary to introduce power into the array.

A **broadside array** is one in which the principal direction of radiation is perpendicular to the axis of the array and to the plane containing the elements. An end-fire array is one in which the principal direction of radiation coincides with the direction of the array axis. These definitions are illustrated by Fig. 4-3.

A **bidirectional array** is one that radiates equally well in either direction along the line of maximum radiation. A bidirectional pattern is shown in Fig. 4-4 at A. A **unidirectional array** is one that has only one principal direction of radiation, as illustrated by the pattern at B in Fig. 4-4.

The **major lobes** of the directive pattern are those in which the radiation is maximum. Lobes of lesser radiation intensity are called **minor lobes.** The **beam width** of a directive antenna is the width, in degrees, of the major lobe between the two directions at which the relative radiated power is equal to one-half its value at the peak of the lobe. At these "half-power points" the field intensity is equal to 0.707 times its maximum value, or down 3 db. from maximum. Fig. 4-5 is an example of a lobe having a beam width of 30 degrees.

The term "gain" as used in this chapter is the power gain over a half-wave dipole of the same orientation and height as the array under discussion, and having the same power input.

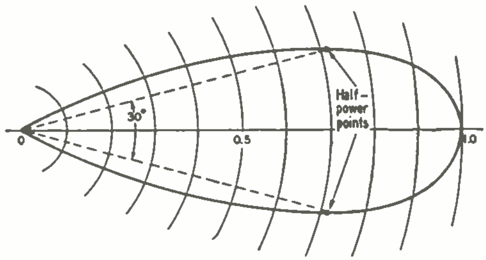


Fig. 4-5 — The width of a beam is the angular distance between the directions at which the received or transmitted power is one-half the maximum power.

Front-to-back ratio means the ratio of the power radiated in the favored direction to the power radiated in the opposite direction.

Phase

The term "phase" has the same meaning when used in connection with the currents flowing in antenna elements as it does in ordinary circuit work. For example, two currents are in phase when they reach their maximum values, flowing in the same direction, at the same instant. The direction of current flow depends on the way in which power is applied to the element.

This is illustrated in Fig. 4-6. Assume that by some means an identical voltage is applied to each of the dipoles at the end marked *A*. Assume also that the instantaneous polarity of the voltage is such that the current is flowing away from the point at which the voltage is applied. The arrows show the assumed current directions. Then the currents in elements *1* and *2* are completely in phase, since they are flowing in the same direction in space and are caused by the same voltage. However, the current in element *3* is flowing in the *opposite* direction in space because the voltage is applied to the opposite end of the element. The current in element *3* is therefore 180 degrees out of phase with the currents in elements *1* and *2*.

The phasing of driven elements depends on the direction of the element, the phase of the applied voltage, and the point at which the voltage is applied. In the systems used by amateurs the voltages applied to the elements are practically always exactly in or exactly out of phase with each other. Also, the axes of the elements are always in the same direction, since parallel or collinear elements are invariably used. The currents in driven elements in such systems are therefore always either exactly in or out of phase with the currents in other elements.

It is possible to use phase differences of less than 180 degrees in driven arrays — one important case is where the voltage applied to one set of elements differs by 90 degrees from the voltage applied to another set — but such systems have not met with much application in amateur work. The reason probably is that making provision for proper phasing is considerably more of a problem than in the case of simple 0- or 180-degree phasing.

In parasitic arrays the phase of the currents in the parasitic elements depends on the spacing and tuning, as described later.

Ground Effects

The effect of the ground is the same with a directive antenna as it is with a simple dipole antenna. The reflection factors discussed in Chapter Two may therefore be applied to the vertical pattern of an array, subject to the same modifications mentioned in that chapter. In cases where the array elements are not all at the same height, the reflection factor for the *mean* height of the array must be used. The mean height is the average of the heights measured from the ground to the centers of the lowest and highest elements.

MUTUAL IMPEDANCE

Consider two half-wave dipoles that are fairly close to each other. When power is applied to one and current flows, a voltage will be induced in the second by the electromagnetic field and current will flow in it as well. The current in antenna No. 2 will in turn induce a voltage in antenna No. 1, causing a current to flow in the latter. The total current in No. 1 is then the sum (taking phase into account) of the original current and the induced current.

If the voltage applied to antenna No. 1 has not changed, the fact that the amplitude of the current flowing is different with antenna No. 2 present than it would have been had No. 2 not been there indicates that the presence of the second antenna has changed the impedance of the first. This coupling effect is called **mutual impedance.** The actual impedance of an antenna element is the sum of its self-impedance (the impedance with no other antennas present) and its mutual impedance with all other antennas in the vicinity.

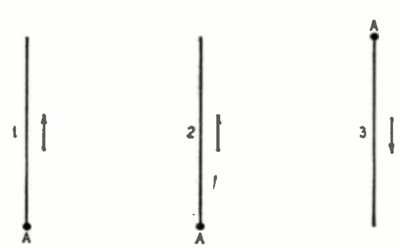


Fig. 4-6 — Illustrating phasing of currents in antenna elements.

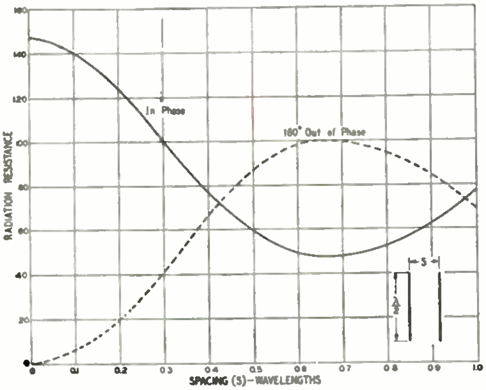


Fig. 4-7 — Radiation resistance measured at the center of one element as a function of the spacing between two parallel half-wave self-resonant antenna elements.

The magnitude and nature of the mutual impedance depends on the amplitude of the current induced in the first antenna by the second, and on the phase relationship between the original and induced currents. The amplitude and phase of the induced current depend on the spacing between the antennas and whether or not the second antenna is tuned to resonance.

Amplitude of Induced Current

The induced current will be largest when the two antennas are close together and are parallel. Under these conditions the voltage induced in the second antenna by the first, and in the first by the second, has its greatest value and causes the largest current flow. The coupling decreases as the parallel antennas are moved farther apart.

The coupling between collinear antennas is comparatively small, and so the mutual impedance between such antennas is likewise small. It is not negligible, however.

Phase Relationships

When the separation between the two antennas is an appreciable fraction of a wavelength a measurable period of time elapses before the field from antenna No. 1 reaches antenna No. 2, and there is a similar time lapse before the field set up by the current in No. 2 gets back to induce a current in No. 1. Hence the current induced in No. 1 by No. 2 will have a phase relationship with the original current in No. 1 that depends on the spacing between the two antennas.

The induced current can range all the way from being completely in phase with the original current to being completely out of phase with it. In the first case the total current is larger than the original current and the antenna impedance is reduced. In the second, the total current is smaller and the impedance is increased. At intermediate phase relationships the impedance will be lowered or raised depending on whether the induced current is mostly in or mostly out of phase with the original current.

Except in the special cases when the induced current is exactly in or out of phase with the original current, the induced current causes the phase of the total current to shift with respect to the applied voltage. The mutual impedance, in other words, has both resistive and reactive components. Consequently, the presence of a second antenna nearby may cause the impedance of an antenna to be reactive — that is, the antenna will be detuned from resonance — even though its self-impedance is entirely resistive. The amount of detuning depends on the magnitude and phase of the induced current.

Tuning Conditions

A third factor that affects the impedance of antenna No. 1 when No. 2 is present is the tuning of the latter. If No. 2 is not exactly resonant the current that flows in it as a result of the induced voltage will either lead or lag the phase it would have had if the antenna were resonant. This causes an additional phase advance or delay that affects the phase of the current induced back in No. 1. Such a phase lag has an effect similar to a change in the spacing between self-resonant antennas. However, a change in tuning is not exactly equivalent to a change in spacing because the two methods do not have the same effect on the amplitude of the induced current.

Mutual Impedance and Gain

The mutual impedance between antennas is important because it determines the amount of current that will flow for a given amount of power supplied. It must be remembered that it is the *current* that determines the field strength from the antenna. Other things being equal, if the mutual impedance between two antennas is such that the currents are greater, for the same total power, than would be the case if the two antennas were not coupled, the power gain will be greater than in the case discussed at the beginning of this chapter. On the other hand, if the mutual impedance is such as to reduce the current the gain will be less than if the antennas were not coupled.

The calculation of mutual impedance between antennas is a difficult problem, and data are available only for a few special cases. Two simple, but important, ones are shown in Figs. 4-7 and 4-8. These graphs do not show the mutual impedance but instead show a more useful quantity, the radiation resistance measured at the center of an antenna as it is affected by the spacing between two antennas.

As shown by the solid curve in Fig. 4-7, the radiation resistance at the center of either dipole, when the two are self-resonant, parallel, and operated in phase, decreases rapidly as the spacing between them is increased until the spacing is about 0.7 wavelength. The maximum gain is

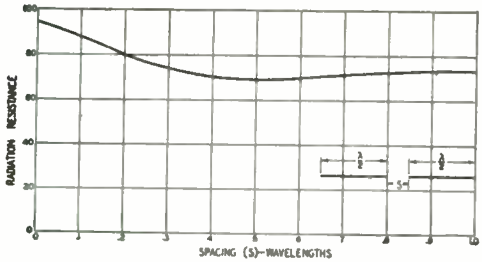


Fig. 4-8 — **Radiation resistance measured at the center of one element as a function of the spacing between the ends of two collinear self-resonant half-wave antenna elements operated in phase.**

secured from a pair of such elements when the spacing is in this region, because the current is larger for the same power and the fields from the two arrive in phase at a distant point placed on a line perpendicular to the line joining the two antennas (see Fig. 2-19, Chapter Two).

The broken curve in this figure, representing two antennas operated 180 degrees out of phase (end-fire), cannot be interpreted quite so simply. The radiation resistance decreases with decreasing spacing in this case. However, the fields from the two antennas add up in phase at a distant point in the favored direction only when the spacing is one-half wavelength (in the range of spacings considered). At smaller spacings the fields become increasingly out of phase, so the total field is less than the simple sum of the two. The latter factor decreases the gain at the same time that the reduction in radiation resistance is increasing it. As shown later in this chapter, the gain goes through a maximum when the spacing is in the region of ⅛ wavelength.

The curve for two collinear elements in phase, Fig. 4-8, shows that the radiation resistance decreases and goes through a broad minimum in the region of 0.3- to 0.5-wavelength spacing between the adjacent ends of the antennas. Since the minimum is not significantly less than the radiation resistance of an isolated antenna, the gain will not exceed the gain calculated on the basis of uncoupled antennas. That is, the best that two collinear elements will give, even with the optimum spacing, is a power gain of about 2 (3 db.). When the separation between the ends is very small — the usual method of operation — the gain is reduced.

Driven Arrays

Driven arrays may be either broadside or end-fire, and may consist of collinear elements, parallel elements, or a combination of both. The number of elements that it is practicable to use depends on the frequency and the space available for the antenna. Fairly elaborate arrays, using as many as 16 or even 32 elements, can be installed in a rather small space when the operating frequency is in the v.h.f. range. At lower frequencies the construction of antennas with a large number of elements would be impracticable for most amateurs.

It is characteristic of broadside arrays that the power gain is proportional to the length of the array but is substantially independent of the number of elements used, provided the optimum element spacing is not exceeded. This means, for example, that a 5-element array and a 6-element array will have the same gain, provided the elements in both are spaced so that the over-all array length is the same. Although this principle is seldom used for the purpose of reducing the number of elements, because of complications introduced in feeding power to each element in the proper phase, it does illustrate the fact that there is nothing to be gained by increasing the number of elements if the space occupied by the antenna is not increased proportionally.

Generally speaking, the maximum gain in the smallest linear dimensions will result when the antenna combines both broadside and end-fire directivity and uses both parallel and collinear elements. In this way the antenna is spread over a greater volume of space, which has the same effect as extending its length to a much greater extent in one linear direction.

Feeding Driven Arrays

Not the least of the problems encountered in constructing multielement driven arrays is that of supplying the required amount of power to each element and making sure that the currents in the elements are in the proper phase. The directive patterns given in this chapter are based on the assumption that each element carries the same current and that the phasing is exact. If the element currents differ, or if the phasing is not proper, the actual directive patterns will not be quite like those shown. Small departures will not greatly affect the gain, but may increase the beam width and introduce minor lobes — or emphasize those that exist already.

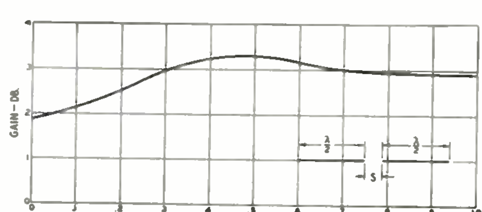


Fig. 4-9 — **Gain of two collinear half-wave elements as a function of spacing between the adjacent ends.**

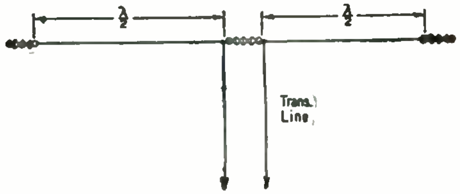


Fig. 4-10 — A two-element collinear array ("two half waves in phase"). The transmission line shown would operate as a tuned line. A matching section can be substituted and a nonresonant line used if desired.

If the directive properties of beam antennas are to be fully realized, care must be used to prevent antenna currents from flowing on transmission lines (see Chapter Three) used as interconnections between elements, as well as on the main transmission line. If radiation takes place from these lines, or if signals can be picked up on them, the directive effects may be masked by such stray radiation or pick-up. Although this may not greatly affect the gain either in transmission or reception, received signals coming from undesired directions will not be suppressed to the extent that is possible with a well-designed system.

COLLINEAR ARRAYS

Collinear arrays are always operated with the elements in phase. (If alternate elements in such an array are out of phase the system simply becomes a harmonic-type antenna.) A collinear array is a broadside radiator, the direction of maximum radiation being at right angles to the line of the antenna.

Power Gain

Because of the nature of the mutual impedance between collinear elements the radiation resistance is increased as shown in Fig. 4-8. For this reason the power gain does not increase in direct proportion to the number of elements. The gain with two elements as the spacing between them is varied is shown by Fig. 4-9. Although the gain is greatest when the end-to-end spacing is in the region of 0.3 to 0.5 wavelength, the use of spacings of this order is inconvenient constructionally and introduces problems in feeding the two elements. As a result, collinear elements are almost always operated with their ends quite close together — in wire antennas, usually with just a strain insulator between.

With very small spacing between the ends of adjacent elements the theoretical power gain of collinear arrays is approximately as follows:

2 collinear elements — 1.9 db.
3 collinear elements — 3.2 db.
4 collinear elements — 4.3 db.

More than four elements are rarely used.

Directivity

The directivity of a collinear array, in a plane containing the axis of the array, increases with its length. Small secondary lobes appear in the pattern when more than two elements are used, but the amplitudes of these lobes are low enough so that they are not important. In a plane at right angles to the array the directive diagram is a circle, no matter what the number of elements. Collinear operation, therefore, affects only the directivity in the plane containing the antenna. At right angles to the wire the pattern is the same as that of the half-wave elements of which it is composed.

When a collinear array is mounted with the elements vertical the antenna radiates equally well in all geographical directions. An array of such "stacked" collinear elements tends to confine the radiation to low vertical angles. For purposes of estimating the effect of ground reflection the height is taken as the height of the center of the array. Applying the ground-reflection factor for this height (using the reflection factors given in Chapter Two for half-wave antennas) to the directive pattern of the array will give the resultant vertical pattern, taking into account ground reflection.

If a collinear array is mounted horizontally the directive pattern in the vertical plane at right angles to the array is the same as the vertical pattern of a simple half-wave antenna at the same height (Chapter Two).

Two-Element Array

The simplest and most popular collinear array is one using two elements, as shown in Fig. 4-10.

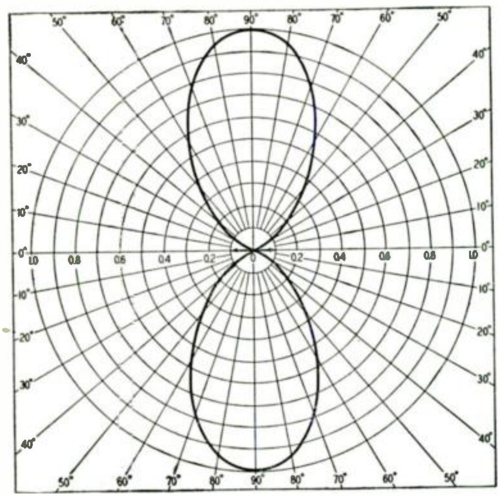


Fig. 4-11 — Free-space directive diagram for a two-element collinear array. Field strength is shown on a relative basis. This is the horizontal pattern at low wave angles when the array is horizontal.

This system is commonly known as "two half waves in phase," and the manner in which the desired current distribution is secured has been described in Chapter Three. The directive pattern in a plane containing the wire axis is shown in Fig. 4-11.

The impedance at the feed point in the array shown in Fig. 4-10 usually lies in the range 1200–1800 ohms. An open-wire transmission line can be used without matching, if desired; the standing-wave ratio should not exceed 2 or 3 to 1 with a 600-ohm line and the losses will not be consequential. If the antenna is to be matched to the line a matching stub or section such as is described in Chapter Three is an appropriate matching device.

Three- and Four-Element Arrays

When more than two collinear elements are used it is necessary to connect "phasing" stubs between adjacent elements in order to bring the currents in all elements in phase. It will be recalled from Chapter Two that in a long wire the direction of current flow reverses in each half-wave section. Consequently, collinear elements cannot simply be connected end to end; there must be some means for making the current flow in the same direction in all elements. In Fig. 4-12A the direction of current flow is correct in the two left-hand elements because the transmission line is connected between them. The phasing stub

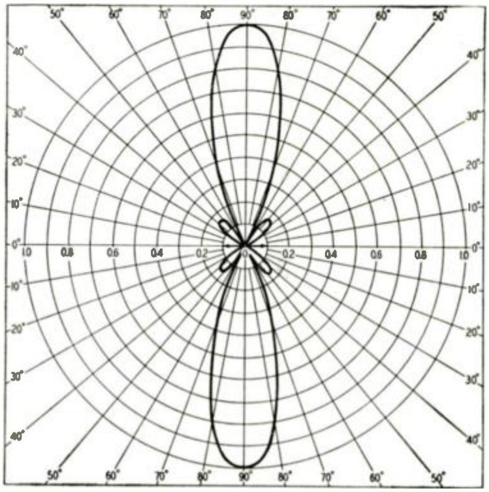


Fig. 4-13 — Free-space directive diagram for a four-element collinear array. Field strength is shown on a relative basis.

between the second and third elements makes the instantaneous current direction correct in the third element. This stub may be looked upon simply as the alternate half-wave section of a long-wire antenna folded back on itself to cancel its radiation. In Fig. 4-12A the part to the right of the transmission line has a total length of three half wavelengths, the center half wave being folded back to form a quarter-wave phase-reversing stub. No data are available on the impedance at the feed point in this arrangement, but various considerations indicate that it should not be over 1000 ohms.

An alternative method of feeding three collinear elements is shown in Fig. 4-12B. In this case power is applied at the center of the middle element and phase-reversing stubs are used between this element and both of the outer elements. The impedance at the feed point in this case is somewhat over 300 ohms and provides a close match to 300-ohm line. The s.w.r. will be less than 2 to 1 when 600-ohm line is used. Center feed of this type is somewhat preferable to the arrangement in Fig. 4-12A because the system as a whole is balanced. This assures more uniform power distribution among the elements. In A, the right-hand

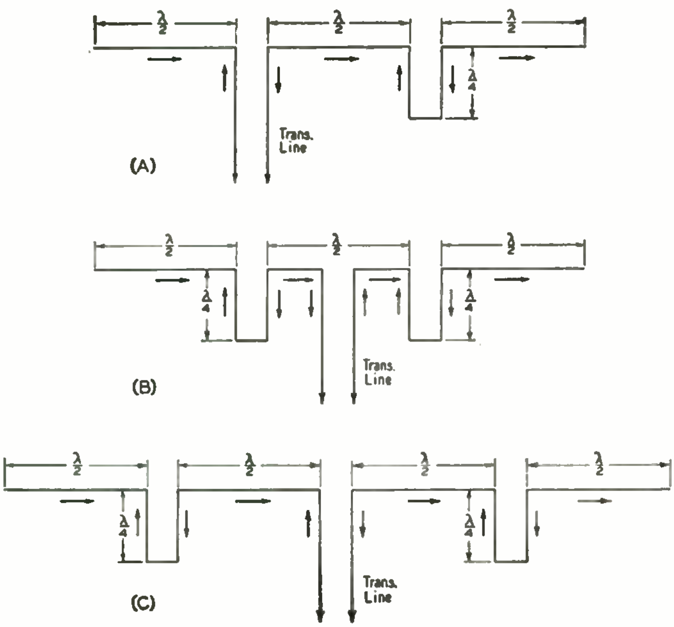


Fig. 4-12 — Three- and four-element collinear arrays. Alternative methods of feeding a three-element array are shown at A and B. These drawings also show the current distribution on the antenna elements and phasing stubs. A matched transmission line can be substituted for the tuned line by using a suitable matching section.

element is likely to receive somewhat less power than the other two because a portion of the fed power is radiated by the middle element before it can reach the one located at the extreme right.

A four-element array is shown in Fig. 4-12C. The system is symmetrical when fed between the two center elements as shown. As in the three-element case, no data are available on the impedance at the feed point. However, the s.w.r. with a 600-ohm line should be less than 2 to 1. Fig. 4-13 shows the directive pattern of a four-element array. The sharpness of the three-element pattern is intermediate between Figs. 4-11 and 4-13, with a small minor lobe at right angles to the array axis.

Collinear arrays can be extended to more than four elements. However, the simple two-element collinear array is the type most used, for the reason that it lends itself well to multiband operation. More than two collinear elements are seldom used because more gain can be obtained from other types of arrays.

Adjustment

In any of the collinear systems described the lengths of the radiating elements can be found from the formula 468/*f* (Mc.). The lengths of the

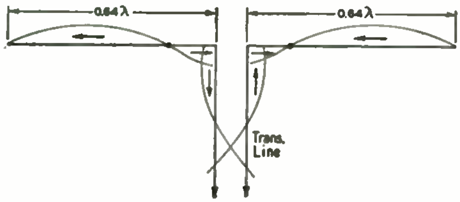


Fig. 4-14 — **The extended double Zepp. This system gives somewhat more gain than two half-wave collinear elements.**

phasing stubs can be found from the formulas given in Chapter Three for the type of line used. If the stub is open-wire line (500 to 600 ohms impedance) it is satisfactory to use a velocity factor of 0.975 in the formula for a quarter-wave line. On-the-ground adjustment is, in general, an unnecessary refinement. If desired, however, the following procedure may be used when the system has more than two elements:

Disconnect all stubs and all elements except those directly connected to the transmission line (in the case of feed such as is shown in Fig. 4-12B leave only the center element connected to the line). Adjust the remaining elements to resonance, preferably using the standing-wave method described in Chapter Three. When the proper length is determined, cut all other elements to the same length. Make the phasing stubs slightly long and use a shorting bar to adjust their length. Connect the remaining elements to the stubs and adjust the stubs to resonance, as indicated by maximum current in the shorting

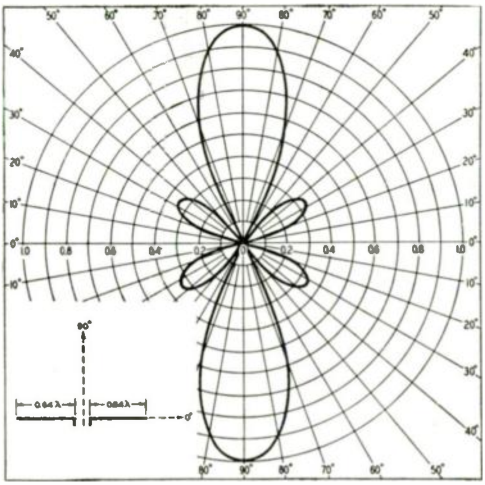


Fig. 4-15 — **Free-space directive diagram for the extended double Zepp. This is also the horizontal directional pattern when the elements are horizontal.**

bars or by the positions of the standing waves along the transmission line. When the whole system is resonant the position of the first current or voltage maximum along the transmission line should be the same as when the line is shorted or open, as described in Chapter Three. If more than three or four elements are used it is best to add elements two at a time (one at each end of the array), resonating the system each time before a new pair is added.

The Extended Double Zepp

An expedient that may be adopted to obtain the higher gain that goes with wider spacing in a simple system of two collinear elements is to make the elements somewhat longer than ½ wavelength. As shown in Fig. 4-14, this increases the spacing between the two in-phase half-wave sections at the ends of the wires. The section in the center carries a current of opposite phase, but if this section is short the current will be small because it represents only the outer ends of a half-wave section. Because of the small current and short length the radiation from the center is small. The optimum length for each element is 0.64 wavelength. At greater lengths the system tends to act as a long-wire antenna and the gain decreases.

This system is known as the "extended double Zepp." The gain over a half-wave dipole is approximately 3 db., as compared with slightly less than 2 db. for two collinear dipoles. The directional pattern in the plane containing the axis of the antenna is shown in Fig. 4-15. As in the case of all other collinear arrays, the free-space pattern in the plane at right angles to the antenna elements is the same as that of a half-wave antenna; i.e., is circular.

BROADSIDE ARRAYS WITH PARALLEL ELEMENTS

To obtain broadside directivity with parallel elements the currents in the elements must all be in phase. At a distant point lying on a line perpendicular to the axis of the array and also perpendicular to the plane containing the elements, the fields from all elements add up in phase. The situation is similar to that pictured in Fig. 4-1 in this chapter and in Fig. 2-19, Chapter Two.

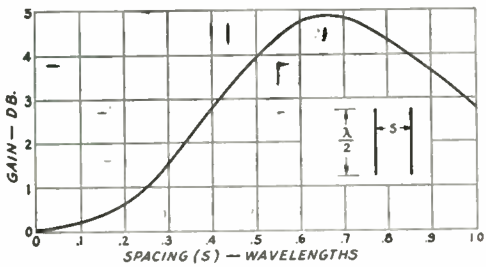


Fig. 4-16 — Power gain as a function of the spacing between two parallel elements operated in phase (broadside).

Broadside arrays of this type theoretically can have any number of elements. However, practical limitations of construction and available space usually limit the number of broadside parallel elements to two, in the amateur bands below 30 Mc., when horizontal polarization is used. More than four such elements seldom are used even at v.h.f.

Power Gain

The power gain of a parallel-element broadside array depends on the spacing between elements as well as on the number of elements. The way in which the gain of a two-element array varies with spacing is shown in Fig. 4-16. The greatest gain is obtained when the spacing is in the vicinity of 0.7 wavelength.

The theoretical gains of broadside arrays having more than two elements are approximately as follows:

No. of Parallel Elements	*Db. Gain with ½-Wave Spacing*	*Db. Gain with ¾-Wave Spacing*
3	5	7
4	6	8.5
5	7	10
6	8	11

The elements must, of course, all lie in the same plane.

Directivity

The sharpness of the directive pattern depends on the spacing between elements and on the number of elements. Larger element spacing will sharpen the main lobe, for a given number of elements. The two-element array has no minor lobes when the spacing is ½ wavelength, but small minor lobes appear at greater spacings. When three or more elements are used the pattern always has minor lobes.

The vertical directive pattern of such an array when the elements are vertical is the same as that for a simple half-wave dipole at the same height. The patterns are given in Chapter Two. When the array elements are horizontal the vertical pattern is the product of the broadside pattern for the particular array used multiplied by the ground-reflection factors given in Chapter Two. For the purpose of applying the ground-reflection factor the height of the array is taken as the mean height above ground. The horizontal directive pattern of a horizontally-polarized parallel-element broadside array is the same as that of a simple dipole.

Two-Element Arrays

The elements of a broadside array must be connected by transmission lines that supply power in the proper phase to each element. Three methods of interconnection for a two-element array are given in Fig. 4-17. In A, the main transmission line is connected to the "phasing line" at its center. The two halves of the phasing line, *AB* and *AC*, are simply in parallel, insofar as the main

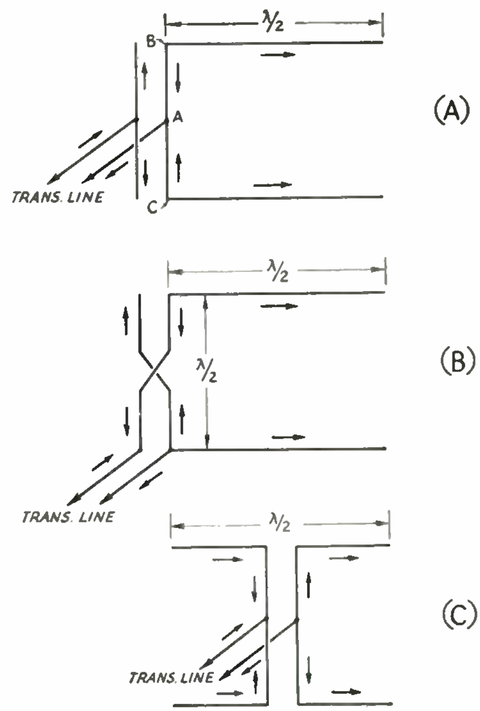


Fig. 4-17 — Two-element broadside arrays, showing different methods of supplying power.

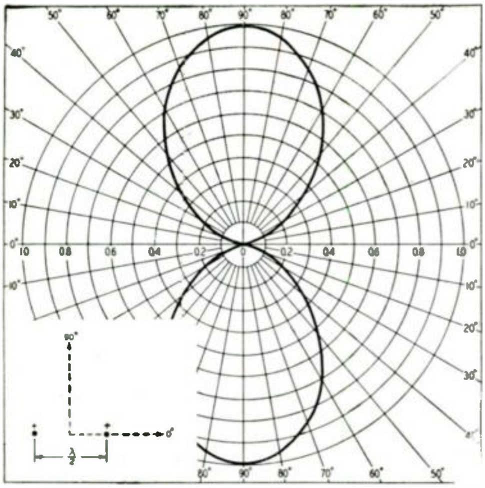


Fig. 4-18 — Free-space directive diagram of a two-element broadside array for an element spacing of ½ wavelength. This drawing gives the low-angle horizontal pattern of a vertically-polarized array.

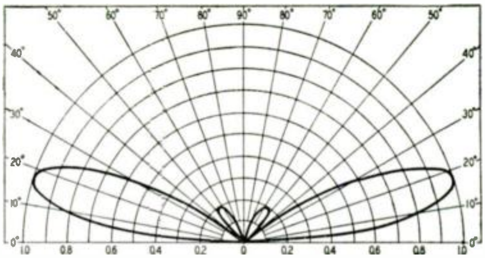


Fig. 4-19 — Vertical pattern broadside to a two-element in-phase array with horizontal elements. This pattern is for a mean height of ¾ wavelength; i.e., lower element ½ wavelength high and upper element one wavelength high. At low wave angles, the horizontal pattern of such an array is the same as for a half-wave dipole.

transmission line is concerned, so the currents in the phasing line flow in opposite directions, with respect to the junction *A*. This brings the currents in the array elements in phase. The phasing line can be any convenient length in this case, so the spacing between the two elements can be any value desired. Although no data are available on impedances, a rough estimate indicates that in most practical cases the impedance will be well below 100 ohms at the point where the main transmission line joins the phasing line, assuming a half-wave phasing line having a Z_0 of about 600 ohms. If the phasing line is not exactly ½ wavelength long the impedance will be reactive as well as resistive.

In B, the main transmission line is connected at the junction of the phasing line and one element. In this case it is necessary to transpose the phasing line somewhere along its length so that the element currents will be in the proper phase. This is shown by the arrows indicating relative direction of current flow. The impedance at the feed point will be resistive and of the order of a few thousand ohms when the elements and phasing line all have electrical lengths of ½ wavelength. With this type of feed the spacing between the elements is determined by the electrical length of the phasing line; it must be an *electrical* half wave long to bring the element currents in proper phase. Open-wire lines are always used as phasing lines in this type of system because their electrical length is nearest to the length of an actual half wavelength in space. If the velocity factor of the phasing line is much less than 1 the antenna elements will perforce be considerably less than a half wavelength apart, and this will reduce the gain.

A third method of feeding is shown at C. This is the best of the three, insofar as symmetry is concerned. The spacing between the two elements can be any desired value. However, when the spacing is one-half wavelength the impedance at the point where the main transmission line is connected is resistive and can be calculated with the aid of Fig. 4-7. For example, Fig. 4-7 shows that the radiation resistance of each element is approximately 60 ohms at half-wave spacing. If the Z_0 of the phasing line is 600 ohms, the impedance reflected at the transmission-line terminals will be $(600)^2/60 = 6000$ ohms, since with half-wave spacing the phasing line is ¼ wave long from the element to the junction. As the reflected resistances from both elements are in parallel, the resistive impedance seen by the transmission line is $6000/2 = 3000$ ohms.

The arrays shown in Fig. 4-17 may be installed either vertically or horizontally, depending on the type of polarization desired. The free-space directive diagram given in Fig. 4-18 is also the horizontal pattern of the array at low wave angles when the elements are vertical. The vertical pattern for a horizontally-polarized two-element array for a mean height of ¾ wavelength is given in Fig. 4-19. The pattern for other heights may be found by multiplying the pattern of Fig. 4-18 by the ground-reflection factor for the actual mean height.

Multielement Arrays

Three- and 4-element arrays are shown in Fig. 4-20. In the 3-element array with half-wave spacing (A) the array is fed at the center. This is the most desirable point in that it tends to keep the power distribution between elements uniform. However, the transmission line could be connected at either *B* or *C*.

When the spacing is greater than ½ wavelength the phasing lines must be one wavelength long and are not transposed between elements. This is shown at B in Fig. 4-20. With this arrangement any element spacing up to one wavelength can be used, if the phasing lines can be folded as suggested in the drawing.

The 4-element array at C is fed at the center of the system to make the power distribution between elements as uniform as possible. However, the transmission line could be connected at either *B*, *C*, *D* or *E*. In such case the section of phasing line between *B* and *D* must be transposed in order to make the currents flow in the same direction in all elements. The 4-element array at C and the 3-element array at B have approximately the same gain when the element spacing in the latter is ¾ wavelength.

An alternative feeding method is shown at D. This system can also be applied to the 3-element arrays, and will result in better symmetry in any case. It is only necessary to move the phasing line to the center of each element, making connection to both sides of the line instead of one only.

The free-space pattern for a 4-element array with half-wave spacing is shown in Fig. 4-21. This is also approximately the pattern for a 3-element array with ¾-wave spacing. The major lobe of a 3-element array with half-wave spacing is intermediate in sharpness between a 2-element and a 4-element array.

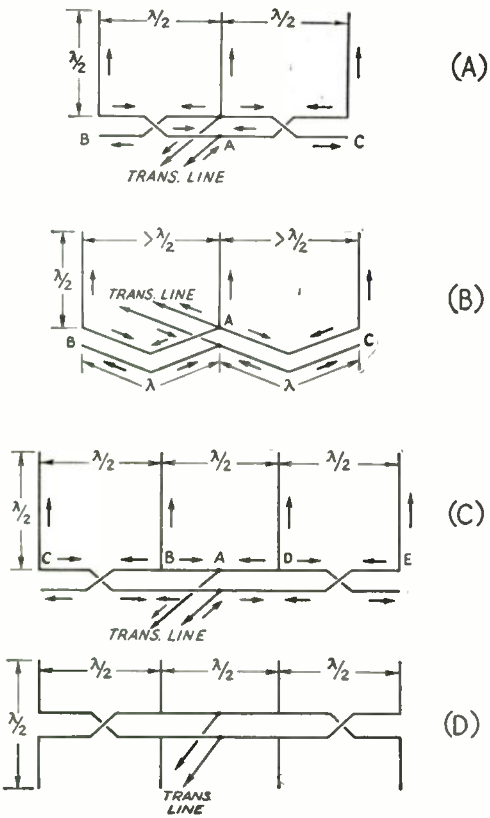


Fig. 4-20 — Methods of feeding three- and four-element broadside arrays with parallel elements.

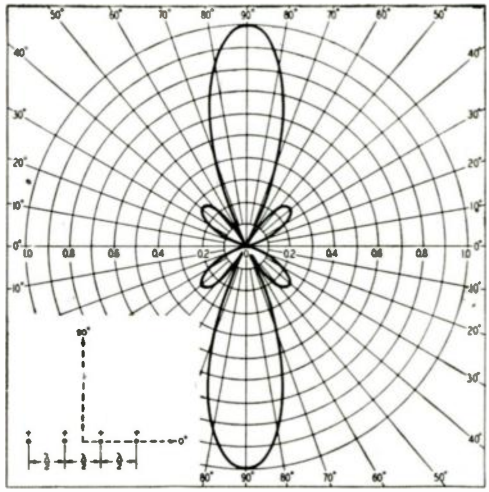


Fig. 4-21 — Free-space directive diagram of a four-element broadside array using parallel elements. This is also the horizontal directive pattern at low wave angles for a vertically-polarized array.

Larger arrays can be designed and constructed by following the phasing principles shown in the drawings.

No accurate figures are available for the impedances at the various feed points indicated in the drawings. It can be estimated to be in the vicinity of 1000 ohms when the feed point is at a junction between the phasing line and a half-wave element, becoming smaller as the number of elements in the array is increased. When the feed point is midway between end-fed elements as in Fig. 4-20C, the impedance of a 4-element array as seen by the transmission line is in the vicinity of 200–300 ohms, with open-wire phasing lines. The impedance at the feed point with the antenna shown at D should be about 1500 ohms.

● END-FIRE ARRAYS

The term "end-fire" covers a number of different methods of operation, all having in common the fact that the maximum radiation takes place along the array axis, and that the array consists of a number of parallel elements in one plane. End-fire arrays can be either bidirectional or unidirectional. In the bidirectional type commonly used by amateurs there are only two elements, and these are operated with currents 180 degrees out of phase. Unidirectional end-fire driven arrays have not had much amateur use because the element phasing is neither 0 nor 180 degrees and tends to be complicated from an adjustment standpoint. (Instead, unidirectional antennas as used by amateurs are practically all based on the use of parasitic elements as described later in this chapter.)

Two-Element Arrays

In the two-element array with equal currents out of phase the gain varies with the spacing between elements as shown in Fig. 4-22. The maximum gain is in the neighborhood of ⅛-wave spacing. Below 0.05-wave spacing the gain decreases rapidly, since the system is approaching the spacings used for nonradiating transmission lines.

The radiation resistance at the center of either element is very low at the spacings giving the

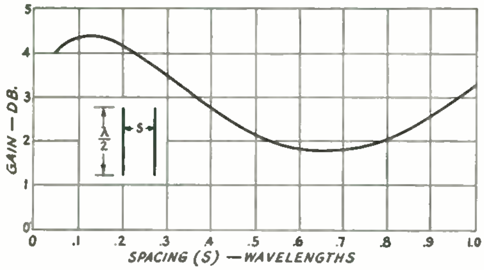


Fig. 4-22 — **Gain of an end-fire array consisting of two elements fed 180 degrees out of phase, as a function of the spacing between elements. Maximum radiation is in the plane of the elements and at right angles to them at spacings up to 0.5 wavelength, but the direction changes at greater spacings.**

greatest gain, as shown by Fig. 4-7. The spacings most frequently used are ⅛ and ¼ wavelength, at which the resistances are about 8 and 32 ohms, respectively. When the spacing is ⅛ wavelength it is advisable to use good-sized conductors — preferably tubing — for the elements because with the radiation resistance so low the heat losses in the conductor can represent an appreciable portion of the power supplied to the antenna. Excessive conductor loss will mean that the theoretical gain cannot be realized.

Three methods of feeding bidirectional end-fire elements are shown in Fig. 4-23. In A, one section of the phasing line is transposed to bring the element currents in proper phase. The method at B is suitable for close-spaced (i.e., ⅛-wave) arrays because each half of the connecting wire is only 1/16 wave long and carries very little current. Hence there is very little radiation from the wires joining the ends of the elements to the transmission line even though the currents are in phase. The center-fed arrangement shown at C is especially useful when the antenna is to be operated on two bands — for example, 14 and 28 Mc. — the higher of which is the second harmonic of the lower.

Because of the very low radiation resistance when the spacing, S in Fig. 4-23, is ⅛ wavelength, the s.w.r. on the transmission line is very high. No figures are available for the end-fed cases, but it can be estimated to be 20 to 1 or higher. With center feed using a 600-ohm line the s.w.r. is over 30 to 1. If the transmission line has any considerable distance to run, it is advisable to match it to the antenna by using a matching section of the type described in Chapter Three. Such a matching section should be of open-wire construction in view of the high s.w.r. The line itself, of course, can be any type capable of carrying the transmitter power. If the transmission line does not have to run more than a wavelength or two it may be of open-wire construction and operated as a tuned line.

With ¼-wave spacing the increased radiation resistance will lower the s.w.r. considerably. With center feed it will be about 10 to 1, and should not exceed that figure with end feed.

In a close-spaced array fed through a tuned transmission line the element lengths are not critical; the only point to watch is to preserve the symmetry of the system as a whole. When a matching section is used, however, it is necessary to adjust the system accurately to the particular frequency to be used most. The low radiation resistance makes the antenna a sharply-tuned affair, and so relatively small departures from the design frequency will throw off the impedance match.

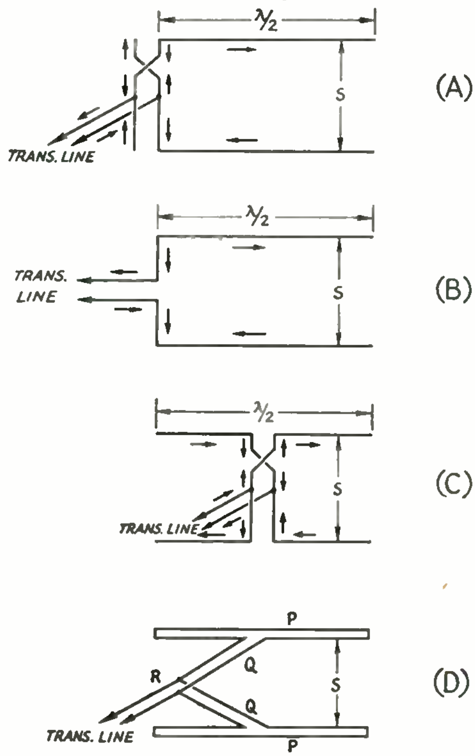


Fig. 4-23 — **Parallel-element end-fire array with various methods of feed, showing current distribution. Matching sections for making the transmission line nonresonant, as described in Chapter Three, may be used in the first three cases. The distance S may be selected from Fig. 4-22.**

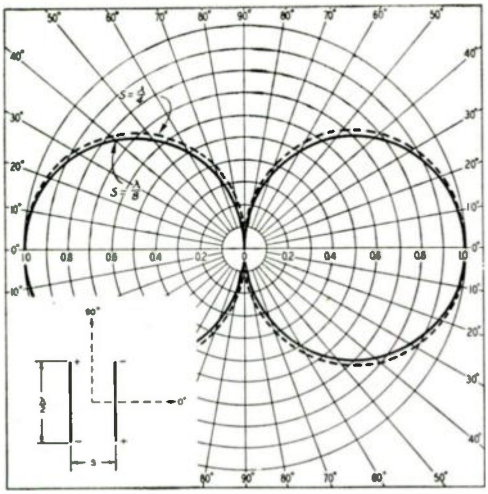


Fig. 4-24 — Free-space directive diagram of a two-element end-fire array with 180-degree phasing, in the plane containing the two parallel elements.

Another way of overcoming the high s.w.r. on the transmission line, and at the same time reducing the resistance loss in the antenna elements, is to use a folded-dipole (see Chapter Three) arrangement as indicated at D in Fig. 4-23. In this way the impedance at the element terminals is stepped up, and then there is a further impedance step-up in the section *Q*, which is a quarter-wave *Q*-type matching transformer. A number of combinations are listed below:

S, wave-length	*No. of conductors in dipole*	*Z_0 of ¼-wave matching section*	*Z_0 of main transmission line*
⅛	1	75	300
⅛	2	75	75
⅛	3	300	600
⅛	4	300	300
¼	1	75	75
¼	2	300	300
¼	3	600	600

In each case the s.w.r. on the transmission line will be well below 2 to 1. Twin-Lead will be satisfactory for the 75- and 300-ohm line. The velocity factor of the line must be taken into account in determining the physical length of the ¼-wave matching transformer. In all the arrangements listed above except those using plain dipoles as elements the frequency characteristic of the antenna will be broadened by the folded-dipole action.

The free-space directive pattern in the plane containing the array is given in Fig. 4-24 and the corresponding pattern in the plane at right angles to the array plane is given in Fig. 4-25. Fig. 4-24 is also the horizontal directive pattern of the array at low wave angles when the ele-

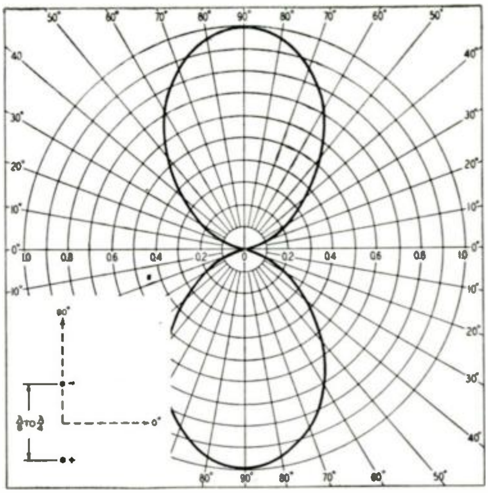


Fig. 4-25 — Free-space directive diagram of a two-element end-fire array with 180-degree phasing, in the plane at right angles to the plane containing the elements.

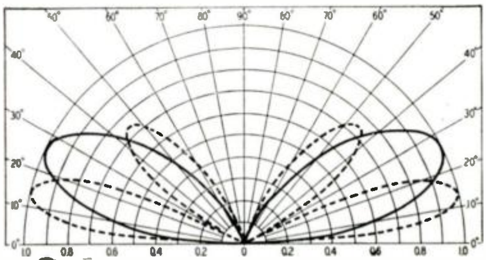


Fig. 4-26 — Vertical pattern of a horizontally-polarized two-element end-fire array. Solid curve, height ½ wavelength; broken curve, height 1 wavelength.

ments are horizontal, while Fig. 4-25 is the horizontal pattern at low wave angles when the elements are vertical. The vertical pattern of a horizontally-polarized array is shown in Fig. 4-26.

Unidirectional End-Fire Arrays

Two parallel elements spaced ¼ wavelength apart and fed equal currents 90 degrees out of phase will have a directional pattern, in the plane at right angles to the plane of the array, as represented in Fig. 4-27. The maximum radiation is in the direction from the element in which the current leads to the element in which the current lags. In the opposite direction the fields from the two elements cancel.

One way in which the 90-degree phase difference can be obtained is shown in Fig. 4-28. Each element must be matched to its transmission line, the two lines being of the same type except that one is an electrical quarter wavelength longer than the other. The length *L* can be any convenient value. Open quarter-wave matching sections are shown, but half-wave shorted sections

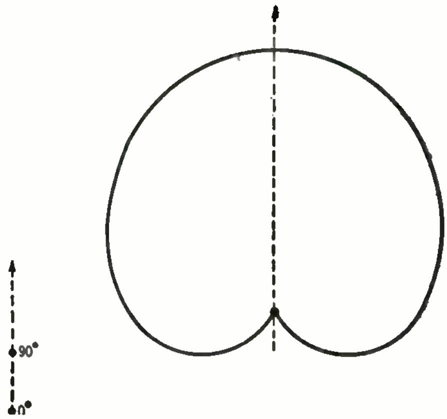


Fig. 4-27 — Representative pattern for a two-element end-fire array with 90-degree phasing, in the plane perpendicular to the plane containing the elements.

could be used instead. The two transmission lines are connected in parallel at the transmitter coupling circuit.

When the currents in the elements are neither in phase nor 180 degrees out of phase the radiation resistances of the elements are not equal. This complicates the problem of feeding equal currents to the elements. If the currents are not equal one or more minor lobes will appear in the pattern and decrease the front-to-back ratio. The adjustment process is likely to be tedious and requires field-strength measurements in order to get the best performance.

More than two elements can be used in a unidirectional end-fire array. The requirement for unidirectivity is that there must be a progressive phase shift in the element currents equal to the spacing, in electrical degrees, between the elements. Because of the difficulty of feeding the elements properly this type of multielement array has had no application in amateur work.

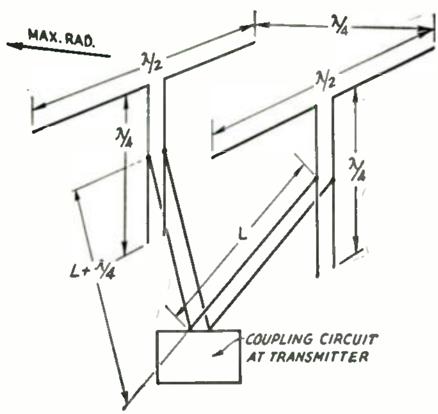


Fig. 4-28 — Unidirectional two-element end-fire array and method of obtaining 90-degree phasing.

COMBINATION DRIVEN ARRAYS

Broadside, end-fire and collinear elements can readily be combined to increase gain and directivity, and this is in fact usually done when more than two elements are used in an array. Combinations of this type give more gain, in a given amount of space, than plain arrays of the types just described.

The combinations that can be worked out are almost endless, but in this section we shall describe only a few of the simpler types that are in common use. The drawings that follow all show the elements arranged for horizontal polarization, which is customary on the frequencies below 30 Mc. where these arrays find their greatest application. For vertical polarization the arrays should be rotated 90 degrees so that the elements are vertical — that is, "stood on end."

Other methods of interconnecting elements than those shown in the drawings may be used. However, the methods shown are recommended over others for two reasons: The antenna system is symmetrical with respect to the feed point, thus making the current distribution among elements as uniform as possible; the lengths of phasing lines (and antenna elements as well) are not critical so long as the lengths of lines radiating from a junction are all the same. With other feed methods this may not be true, and it becomes necessary to use the methods described in Chapter Three to ensure that elements and phasing lines are exactly resonant at the design frequency, if maximum performance is to be secured from the antenna. This adjustment process can be rather difficult as well as tedious. If the feed arrangements shown in the drawings are followed the lengths of wire elements can be found from 468/*f* (Mc.), the element spacings from 984/*f* (Mc.) multiplied by the fraction of wavelength desired, and the phasing lines can simply be cut to fit, keeping all lines the same length.

Gain of Combination Arrays

The accurate calculation of the power gain of a multielement array requires a knowledge of the mutual impedances between all elements. For approximate purposes it is sufficient to assume that each *set* (collinear, broadside, end-fire) will have the gains as given earlier, and then simply add up the gains for the combination. This neglects the effects of cross-coupling between sets of elements. However, the array configurations are such that the mutual impedances from cross-coupling should be relatively small, particularly when the spacings are ¼ wavelength or more, so that the estimated gain should be reasonably close to the actual gain.

Four-Element End-Fire and Collinear Array

The array shown in Fig. 4-29 combines collinear in-phase elements with parallel out-of-phase elements to give both broadside and end-fire

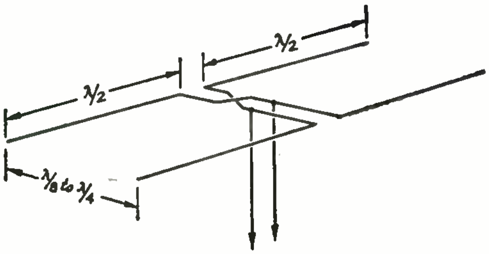


Fig. 4-29 — A four-element array combining collinear broadside elements and parallel end-fire elements.

directivity. It is popularly known as a "two-section W8JK" or "two-section flat-top beam." The approximate gain calculated as described above is 6.2 db. with ⅛-wave spacing and 5.7 db. with ¼-wave spacing. Directive patterns are given in Figs. 4-30 and 4-31.

The impedance between elements at the point where the phasing line is connected is of the order of several thousand ohms. The s.w.r. with an unmatched line consequently is quite high, and this system should be constructed with open-wire line (500 or 600 ohms) if the line is to be resonant. To use a matched line a closed stub 3/16 wavelength long can be connected at the transmission-line junction shown in Fig. 4-29, and the transmission line itself can then be tapped on this matching section at the point resulting in the lowest line s.w.r. This point can be determined by trial as described in Chapter Three.

With ¼-wave spacing the s.w.r. on a 600-ohm line is estimated to be in the vicinity of 3 or 4 to 1.

This type of antenna can be operated on two bands having a frequency ratio of 2 to 1, if a resonant transmission line is used. For example, if designed for 28 Mc. with ¼-wave spacing between elements it can be operated on 14 Mc. as a simple end-fire array (Fig. 4-23) having ⅛-wave spacing.

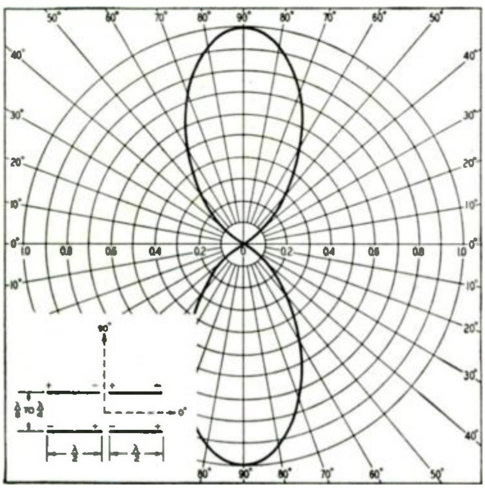


Fig. 4-30 — Free-space directive diagram of the antenna shown in Fig. 4-29, in the plane of the antenna elements. The pattern in the plane perpendicular to the element plane is the same as Fig. 4-25.

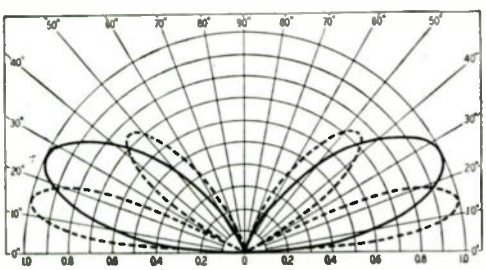


Fig. 4-31 — Vertical pattern of the four-element antenna of Fig. 4-29 when mounted horizontally. Solid curve, height ½ wavelength; broken curve, height 1 wavelength. Fig. 4-30 gives the horizontal pattern.

Four-Element Broadside Array

The four-element array shown in Fig. 4-32 is commonly known as the "lazy-H." It consists of a set of two collinear elements and a set of two parallel elements, all operated in phase to give broadside directivity. The gain and directivity will depend on the spacing, as in the case of a simple parallel-element broadside array. The spacing may be chosen between the limits shown on the drawing, but spacings below ⅜ wavelength are not worth while because the gain is small. Approximate gains are as follows:

⅜-wave spacing — 4.4 db.
½-wave spacing — 5.9 db.
⅝-wave spacing — 6.7 db.
¾-wave spacing — 6.6 db.

Half-wave spacing is generally used. Directive patterns for this spacing are given in Figs. 4-33 and 4-34.

With half-wave spacing between parallel elements the impedance at the junction of the phas-

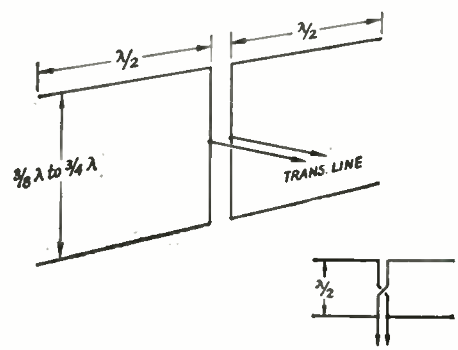


Fig. 4-32 — Four-element broadside array ("lazy-H") using collinear and parallel elements.

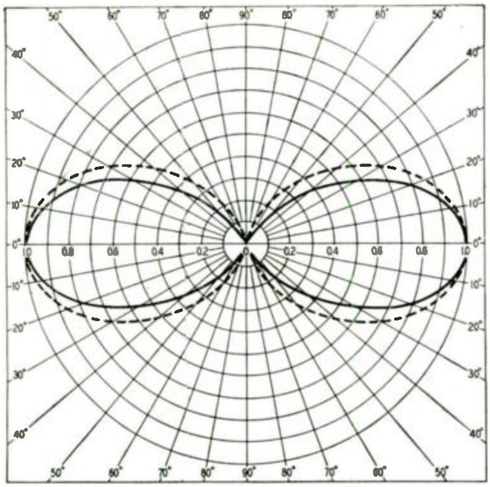


Fig. 4-33 — Free-space directive diagrams of the four-element antenna shown in Fig. 4-32. The solid curve is the horizontal directive pattern at low wave angles when the antenna is mounted with the elements horizontal. The broken curve is the free-space vertical pattern of a horizontally-polarized array, broadside to the array. Actual pattern in the presence of ground may be found by multiplying this pattern by the ground-reflection factors given in Chapter Two.

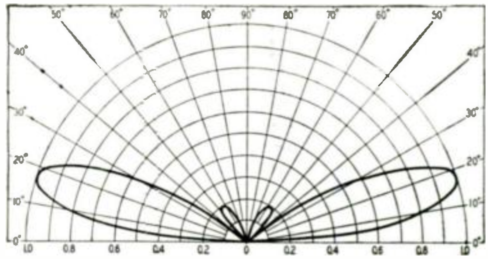


Fig. 4-34 — Vertical pattern of the four-element broadside antenna of Fig. 4-32, when mounted with the elements horizontal and the lower set ½ wavelength above ground. "Stacked" arrays of this type give best results when the lowest elements are at least ½ wave high. The gain is reduced and the wave angle raised if the lowest elements are close to ground.

ing line and transmission line is resistive and is in the vicinity of 100 ohms. With larger or smaller spacings the impedance at this junction will be reactive as well as resistive. Matching stubs are recommended in cases where a nonresonant line is to be used. They may be calculated and adjusted as described in Chapter Three, after first determining the position of the current loop or node (on the transmission line) nearest the junction and after measuring the standing-wave ratio.

The system shown in Fig. 4-32 may be used on two bands having a 2-to-1 frequency relationship. It should be designed for the higher of the two frequencies, using ¾-wave spacing between parallel elements. It will then operate on the lower frequency as a simple broadside array with ⅜-wave spacing.

An alternative method of feeding is shown in the small diagram in Fig. 4-32. In this case the elements and the phasing line must be adjusted exactly to an electrical half wavelength. The impedance at the feed point will be resistive and of the order of 2000 ohms.

Four-Element Broadside and End-Fire Array

The array shown in Fig. 4-35 combines parallel elements in broadside and end-fire directivity. Approximate gains can be calculated by adding the figures from Figs. 4-16 and 4-22 for the element spacings used. The smallest (physically) array — ⅜-wave spacing between broadside and ⅛-wave spacing between end-fire elements — has a gain of about 6.8 db. and the largest — ¾- and ¼-wave spacing, respectively — about 8.5 db. The optimum element spacings are ⅝ wave broadside and ⅛ wave end-fire, giving an over-all gain of about 9.3 db. Directive patterns are given in Figs. 4-36 and 4-37.

The impedance at the feed point will not be purely resistive unless the element lengths are correct and the phasing lines are exactly a half wavelength long. (This requires somewhat less than half-wave spacing between broadside elements.) In this case the impedance at the junction is estimated to be over 10,000 ohms. With other element spacings the impedance at the junction will be reactive as well as resistive, but in any event the standing-wave ratio will be quite large. An open-wire line can be used as a resonant line, or a matching section may be used for nonresonant operation.

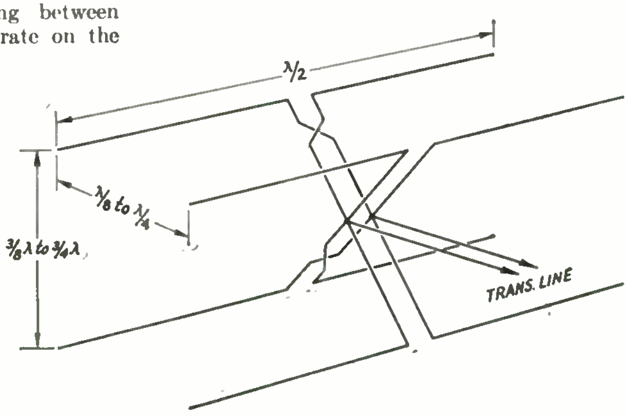


Fig. 4-35 — Four-element array combining both broadside and end-fire elements.

Eight-Element Driven Array

The array shown in Fig. 4-38 is a combination of collinear and parallel elements in broadside and end-fire directivity. The gain can be calculated as described previously, using Figs. 4-9, 4-16 and 4-22. Common practice is to use half-wave spacing for the parallel broadside elements and ⅛-wave spacing for the end-fire elements. This gives a gain of approximately 10 db. Directive patterns for an array using these spacings are given in Figs. 4-39 and 4-40.

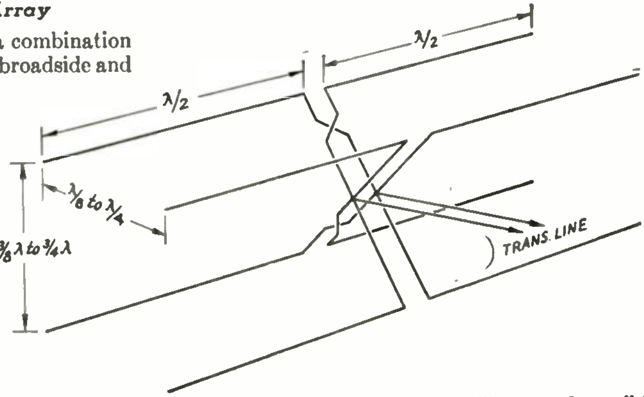


Fig. 4-38 — Eight-element driven array combining collinear and parallel elements for broadside and end-fire directivity.

Although even approximate figures are not available, the s.w.r. with this arrangement will be high. Matching stubs are recommended for making the line nonresonant. Their position and length can be determined by measuring the s.w.r. and locating the current loop or null nearest the junction of the transmission line and phasing lines. The procedure is described in Chapter Three.

This system can be used on two bands related in frequency by a 2-to-1 ratio, providing it is designed for the higher of the two with ¾-wave spacing between the parallel broadside elements and ¼-wave spacing between the end-fire elements. On the lower frequency it will then operate as a four-element antenna of the type shown in Fig. 4-35, with ⅜-wave broadside spacing and ⅛-wave end-fire spacing. For two-band operation a resonant transmission line will have to be used.

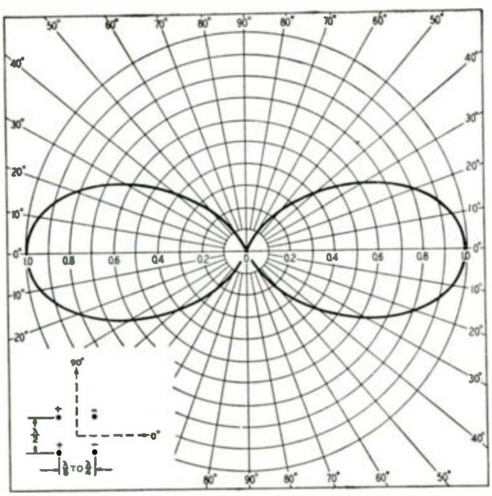


Fig. 4-36 — Free-space pattern of the four-element antenna shown in Fig. 4-35, in the plane perpendicular to the array axis. The pattern in the plane containing a set of end-fire elements is the same as Fig. 4-24.

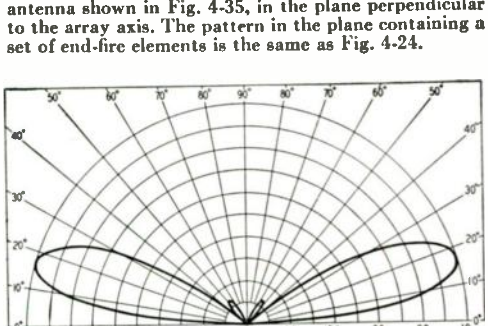


Fig. 4-37 — Vertical pattern of the antenna shown in Fig. 4-35 at a mean height of ¾ wavelength (lowest elements ½ wave above ground) when the antenna is horizontally polarized. For optimum gain and low wave angle the mean height should be at least ¾ wavelength.

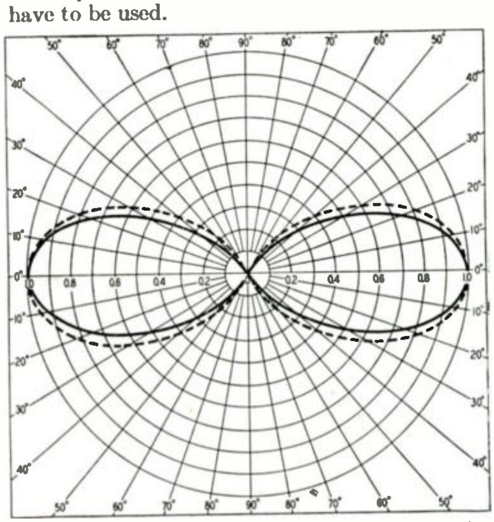


Fig. 4-39 — Free-space directive diagrams of the eight-element array shown in Fig. 4-38. The solid curve is the horizontal pattern of the antenna at low wave angles when the antenna is horizontally polarized. Broken curve, free-space vertical pattern of a horizontally-polarized array, broadside to the array. The vertical pattern in the presence of ground may be found by applying the ground-reflection factors in Chapter Two, for the actual mean height.

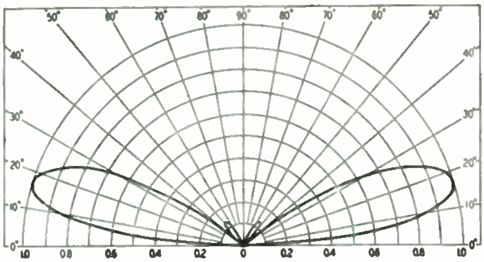


Fig. 4-40 — Vertical pattern of the antenna of Fig. 4-38 when mounted horizontally at a mean height of ¾ wavelength. This is the minimum height that should be used for realizing good gain and a low wave angle.

Checking Phasing

In the antenna diagrams earlier in this chapter the relative direction of current flow in the various antenna elements and connecting lines was shown by arrows. In laying out any antenna system it is necessary to know that the phasing lines are properly connected; otherwise the antenna may have entirely different characteristics than anticipated. The phasing may be checked either on the basis of current direction or polarity of voltages. There are two rules to remember:

1) In every half-wave section of wire, starting from an open end, the current directions reverse. In terms of voltage, the polarity reverses at each half-wave point, starting from an open end.

2) The currents in transmission lines always must flow in opposite directions in adjacent wires. In terms of voltage, the polarities always must be opposite.

Examples of the use of current direction and voltage polarity are given at A and B, respectively, in Fig. 4-41. The half-wave points in the system are marked by the small circles. When current in one section flows toward a circle, the current in the next section must also flow toward it, and vice versa. In the four-element antenna shown at A, the current in the upper right-hand element cannot flow toward the transmission line, because then the current in the right-hand section of the phasing line would have to flow upward and thus would be flowing in the same direction as the current in the left-hand wire. The phasing line would simply act like two wires in parallel in such a case.

The drawing at C shows the effect of transposing the phasing line. This transposition reverses the direction of current flow in the lower pair of elements, as compared with A, and thus changes the array from a combination collinear and end-fire arrangement into a collinear-broadside array.

The drawing at D shows what happens when the transmission line is connected at the center of a section of phasing line. Viewed from the main transmission line the two parts of the phasing line are simply in parallel, so the half wavelength is measured from the antenna element along the upper section of phasing line and thence along the transmission line. The distance from the lower elements is measured in the same way. Obviously the two sections of phasing line should be the same length. If they are not, the current distribution becomes quite complicated; the element currents are neither in phase nor 180 degrees out of phase, and the elements at opposite ends of the lines do not receive the same power. To change the element current phasing at D into the phasing at A, simply transpose the wires in one section of the phasing line; this reverses the direction of current flow in the antenna elements connected to that section of phasing line.

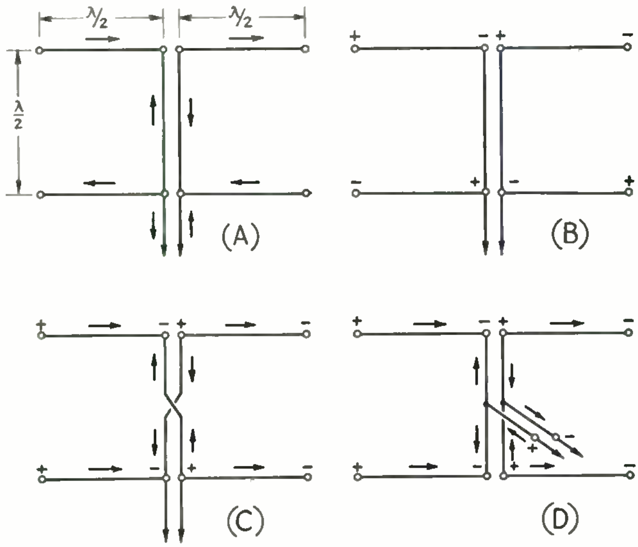


Fig. 4-41 — Methods of checking the phase of currents in elements and phasing lines.

OTHER DRIVEN SYSTEMS

Two other types of driven antennas are worthy of mention, although their use by amateurs has been rather limited. The Sterba array, shown at A in Fig. 4-42, is a broadside radiator consisting of both collinear and parallel elements, with ½-wave spacing between the latter. Its distinctive feature is the method of closing the ends of the system. For direct current and low-frequency a.c. the system forms a closed loop, which is advantageous in that heating currents can be sent through the wires to melt the ice that forms in cold climates. There is comparatively little

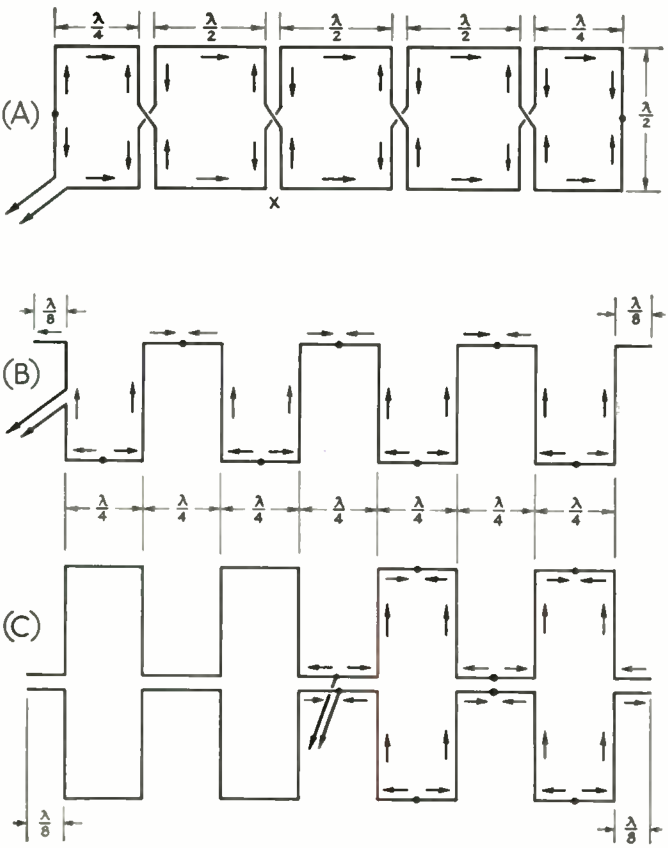


Fig. 4-42 — The Sterba array (A) and two forms of the Bruce array (B and C).

radiation from the vertical connecting wires at the ends because the currents are relatively small and are flowing in opposite directions with respect to the center (the voltage loop is marked with a dot in this drawing).

The system obviously can be extended as far as desired. The approximate gain is the sum of the gains of one set of collinear elements and one set of broadside elements, counting the two ¼-wave sections at the ends as one element. The antenna shown, for example, is about equivalent to one set of four collinear elements and one set of two parallel broadside elements, so the total gain is approximately 4.3 + 4.0 = 8.3 db. Horizontal polarization is the only practicable type at the lower frequencies, and the lower set of elements should be at least ½ wavelength above ground for best results.

When feeding at the point shown the impedance is of the order of 600 ohms. Alternatively, this point can be closed and the system fed between any two elements, as at *X*. In this case a point near the center should be chosen so that the power distribution between elements will be as uniform as possible. The impedance at any such point will be 1000 ohms or less in systems with six or more elements.

The Bruce array is shown at B in Fig. 4-42. It consists simply of a single wire folded so that the vertical sections carry large currents in phase while the horizontal sections carry small currents flowing in opposite directions with respect to the center (indicated by the dot). The radiation consequently is vertically polarized. The gain is proportional to the length of the array but is somewhat smaller, because of the short radiating elements, than is obtainable from a broadside array of half-wave parallel elements of the same over-all length. The array should be 2 or more wavelengths long to secure a worth-while gain. The system can be fed at any current loop; these occur at the centers of the vertical wires.

Another form of the Bruce array is shown at C. Because the radiating elements have twice the height, the gain is increased. The system can be fed at the center of any of the connecting lines.

Parasitic Arrays

Multielement arrays containing parasitic elements are called "parasitic" arrays even though at least one and sometimes more than one of the elements is driven. A parasitic element obtains its power through electromagnetic coupling with a driven element, as contrasted with receiving it by direct connection to the power source.

As explained earlier in this chapter in the section on mutual impedance, the amplitude and phase of the current induced in an antenna element depend on the spacing between it and the driven element to which it is coupled, and on its tuning. The fact that the relative phases of the currents in driven and parasitic elements can be adjusted is very advantageous. For example, the spacing and tuning can be adjusted to approximate the conditions that exist when two driven elements ¼ wavelength apart are operated with a phase difference of 90 degrees (which gives the unidirectional pattern as shown in Fig. 4-27). However, complete cancellation of radiation in the rear direction is not possible when a parasitic element is used. This is because it is usually not possible to make *both* amplitude and phase reach desired values simultaneously. Nevertheless, a properly-designed parasitic array can be adjusted to have a very large front-to-back ratio.

The unidirectional characteristic and relatively-simple electrical configuration of an array using parasitic elements make it especially useful for antenna systems that are to be rotated to aim the beam in any desired direction. Rotatable antennas of this type find wide application at 14 Mc. and above.

Reflectors and Directors

Although there are one or two special cases where a parasitic array will have a bidirectional (but usually not symmetrical) pattern, in most applications the pattern tends to be unidirectional. A parasitic element is called a **director** when it makes the radiation maximum in the direction perpendicular to the driven element and along the line from the driven to the parasitic element, as shown at A in Fig. 4-43. When the maximum radiation is in the opposite direction, as at B, the parasitic element is called a **reflector.**

Whether the parasitic element operates as a director or reflector is determined by the relative phases of the currents in the driven and parasitic elements. At the element spacings commonly used (¼ wavelength or less) the current in the parasitic element will be in the right phase to make the element act as a reflector when the tuning is adjusted to the low-frequency side of resonance. The parasitic element will act as a director when the tuning is adjusted to the high-frequency side of resonance. The proper tuning is ordinarily accomplished by adjusting the lengths of the parasitic elements. However, the elements can be "loaded" at the center with lumped inductance or capacitance to achieve the same purpose. If the parasitic element is self-resonant the element spacing determines whether it will act as a reflector or director.

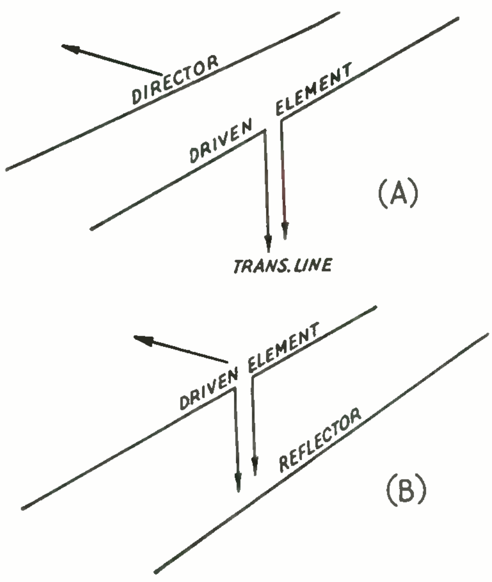


Fig. 4-43 — Antenna systems using a single parasitic element. In A the parasitic element acts as a director, in B as a reflector. The arrows show the direction in which maximum radiation takes place.

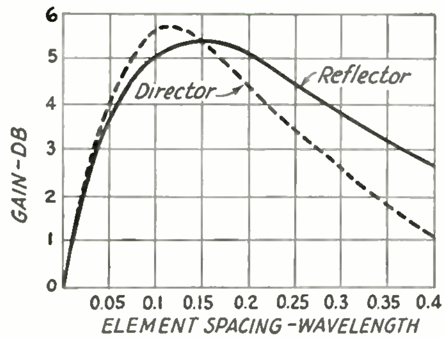


Fig. 4-44 — The maximum possible gain obtainable with a parasitic element over a half-wave antenna alone, assuming that the parasitic element tuning is adjusted for greatest gain at each spacing.

● ONE PARASITIC ELEMENT

The maximum gain obtainable with a single parasitic element, as a function of the spacing, is shown in Fig. 4-44. The two curves show the greatest gain to be expected when the element is tuned for optimum performance either as a director or reflector. The shift from director to reflector, with the corresponding shift in direction as shown in Fig. 4-43, is accomplished simply by tuning the parasitic element — usually, in practice, by changing its length.

With the parasitic element tuned to act as a director, maximum gain is secured when the spacing is approximately 0.1 wavelength. The peak is rather sharp, and the gain drops off rapidly at greater or smaller spacings. When the parasitic element is tuned to work as a reflector, the spacing which gives maximum gain is about 0.15 wavelength, with a fairly broad peak. The director will give slightly more gain than the reflector, but the difference is less than ½ db.

In only two cases are the gains shown in Fig. 4-44 secured when the parasitic element is self-resonant. These occur at 0.1- and 0.25-wavelength spacing, with the parasitic element acting as director and reflector, respectively. For reflector operation, it is necessary to tune the parasitic element to a lower frequency to secure maximum gain at all spacings less than 0.25 wave-

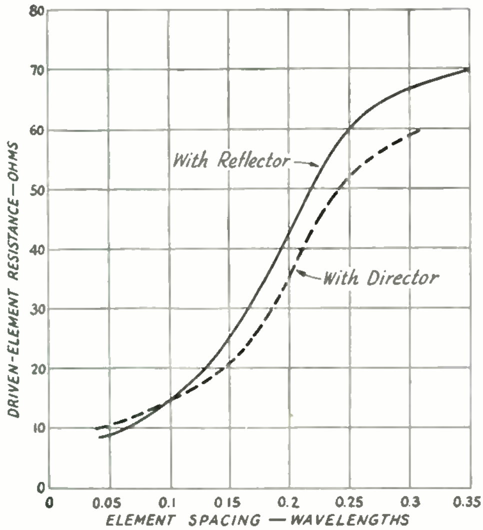


Fig. 4-45 — Radiation resistance at the center of the driven element as a function of element spacing, when the parasitic element is adjusted for the gains given in Fig. 4-44.

length, while at greater spacings the reverse is true. The closer the spacing the greater the detuning required. On the other hand, the director must be detuned toward a higher frequency (that is, its length must be made less than the self-resonant length) at spacings greater than 0.1 wavelength in order to secure maximum gain. The amount of detuning necessary becomes greater as the spacing is increased. At less than 0.1-wavelength spacing the director must be tuned to a lower frequency to secure the maximum gains indicated by the curve.

Radiation Resistance

The radiation resistance measured at the center of the driven element varies as shown in Fig. 4-45 for the spacings and tuning conditions that give the gains indicated by the curves of Fig. 4-44. These values, especially in the vicinity of 0.1-wavelength spacing, are quite low. The curves coincide at 0.1 wavelength, both showing a value of 14 ohms.

The low radiation resistance at the spacings giving highest gain is important in three ways. First, the radiation efficiency goes down because, with a fixed loss resistance, more of the power supplied to the antenna is lost in heat and less is radiated, as the radiation resistance approaches the loss resistance in magnitude. Second, the selectivity of the antenna system becomes higher as the radiation resistance decreases. This means that optimum performance can be secured over only a narrow band of frequencies as compared with the frequency-performance characteristic of a higher-resistance antenna. Third, the number of suitable feeder systems becomes limited, and adjustment becomes more critical.

The loss resistance can be decreased by using low-resistance conductors for the antenna elements. This means, principally, large-diameter conductors, usually tubing of aluminum, copper, or copper-plated steel. Such conductors have mechanical advantages as well, in that it is relatively easy to provide adjustable sliding sections for changing length, while the fact that they can be largely self-supporting makes them well adapted for rotatable antenna construction. With half-inch or larger tubing the loss resistance in any two-element antenna should be negligible.

With low radiation resistance the standing waves of both current and voltage on the antenna reach considerably higher maximum values than is the case with the antenna alone. For this reason losses in insulators at the ends of the elements become more serious. The use of tubing rather than wire helps reduce the end voltage, and furthermore, the tubing does not require support at the ends, thus eliminating the insulators and one source of power loss.

Self-Resonant Parasitic Elements

The special case of the self-resonant parasitic element is of interest, since it gives a good idea of the performance as a whole of two-element systems, even though the results can be modified by detuning the parasitic element. Fig. 4-46 shows gain and radiation resistance as a function of the element spacing for this case. Relative field strength in the direction *A* of the small drawing is indicated by Curve A; similarly for Curve B. The front-to-back ratio at any spacing is the difference in the values at that spacing for Curves A and B. Whether the parasitic element is functioning principally as a director or reflector is determined by whether Curve A or Curve B is on

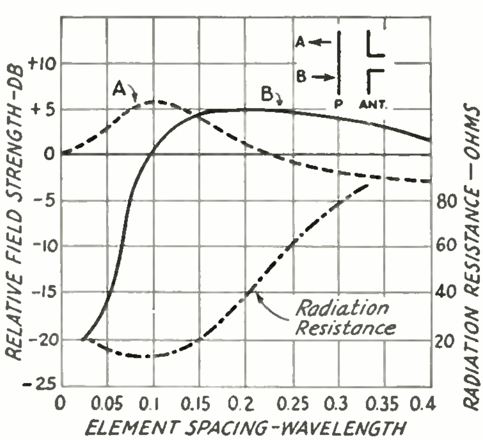


Fig. 4-46 — Gain in a two-element parasitic array as a function of element spacing when the parasitic element is self-resonant.

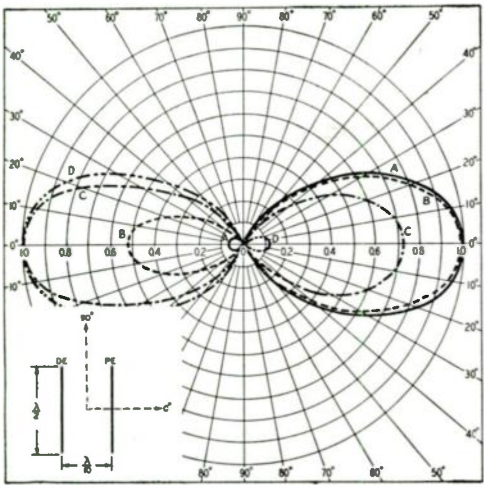


Fig. 4-47 — "Horizontal" directive pattern of a horizontally-polarized two-element parasitic array at a height of 1¼ wavelengths. This pattern is for a wave angle of 12 degrees. The curves represent the following conditions, approximately:

A — Parasitic element tuned for maximum gain as a director.
B — Parasitic element self-resonant.
C — Parasitic element tuned for maximum gain as a reflector.
D — Parasitic element tuned for maximum front-to-back ratio as a reflector.

The spacing between elements is 0.1 wavelength.

The patterns should not be compared for gain, since they are plotted on a relative basis to an arbitrarily-chosen maximum of 1.0.

top; it can be seen that the principal function shifts at about 0.14-wavelength spacing. That is, at closer spacings the parasitic element is principally a director, while at greater spacings it is chiefly a reflector. At 0.14 wavelength the radiation is the same in both directions; in other words, the antenna is bidirectional with a gain of about 4 db.

The front-to-back ratios that can be secured with the parasitic element self-resonant are not very great except in the case of extremely close spacings. Spacings of the order of 0.05 wavelength are hardly practicable with outdoor construction since it would be difficult, if not impossible, to make the elements sufficiently stable mechanically. Better practice is to use spacings of at least 0.1 wavelength and detune the parasitic element for greatest attenuation in the backward direction.

The radiation resistance increases rapidly for spacings greater than 0.15 wavelength, while the gain, with the parasitic element acting as a reflector, decreases quite slowly. If front-to-back ratio is not an important consideration, a spacing as great as 0.25 wavelength can be used without much reduction in gain, while the radiation resistance approaches that of a half-wave antenna alone. Spacings of this order are particularly suited to antennas using wire elements, such as multielement arrays consisting of combinations of collinear and broadside elements.

Front-to-Back Ratio

The tuning conditions that give maximum gain forward do not give maximum signal reduction, or attenuation, to the rear. It is necessary to sacrifice some gain to get the highest front-to-back ratio. The reduction in backward response is brought about by adjustment of the tuning or length of the parasitic element. With a reflector, the length must be made slightly greater than that which gives maximum gain, at spacings up to 0.25 wavelength. The director must be shortened somewhat to achieve the same end, with spacings of 0.1 wavelength and more. The tuning condition, or element length, which gives maximum attenuation to the rear is considerably more critical than that for maximum gain, so that a good front-to-back ratio can be secured without sacrificing more than a small part of the gain.

For the sake of good reception, general practice is to adjust for maximum front-to-back ratio rather than for maximum gain. Larger front-to-back ratios can be secured with the parasitic element operated as a director rather than as a reflector. With the optimum director spacing of 0.1 wavelength, the front-to-back ratio with the director tuning adjusted for maximum gain is only 5.5 db. (the back radiation is equal to that from the antenna alone). By proper director tuning, however, the ratio can be increased to 17 db.; the gain in the desired direction is in this case 4.5 db., or 1 db. less than the maximum obtainable.

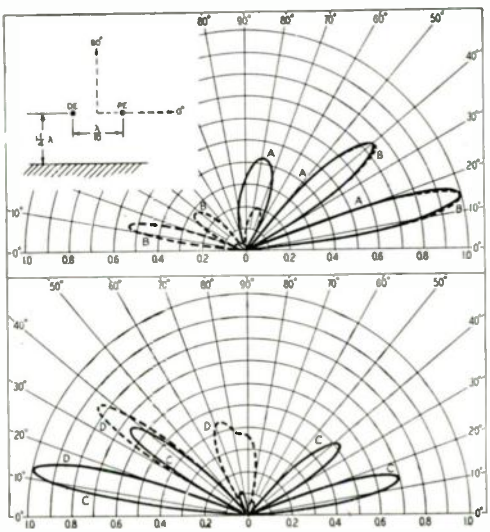


Fig. 4-48 — Vertical pattern of a horizontally-polarized two-element array under the conditions given in Fig. 4-47. This pattern is in the vertical plane at right angles to the antenna elements.

Directional Patterns

The directional patterns obtained with two-element arrays will vary considerably with the tuning and spacing of the parasitic element. Typical patterns are shown in Figs. 4-47 and 4-48, for four cases where the parasitic element tuning or length is approximately adjusted for optimum gain as a director, for self-resonance, for optimum gain as a reflector, and for optimum front-to-back ratio as a reflector. Over this range of adjustment the width of the main beam does not change significantly. These patterns are based on experimental measurements by J. L. Gillson, W3GAU.

TABLE 4-I
Measured Gains of 3- and 4-Element Arrays

Parasitic-Element Spacing			*Gain in Db. over Dipole*	
Directors		*Reflector*	*I*	*II*
2nd	*1st*			
		0.15	4.8	
	0.1	0.15	8.0	7.8
	0.15	0.1	8.1	7.8
	0.2	0.1	8.4	8.3
	0.2	0.15	8.8	8.7
0.1	0.1	0.15	9.3	8.7
0.2	0.2	0.15	10.2	
0.2	0.2	0.2	10.4	9.6

TWO AND THREE PARASITIC ELEMENTS

It is readily possible to use more than one parasitic element in conjunction with a single driven element. With more than one parasitic element the optimum gain and directivity result when one is used as a reflector and the second and third as directors. Even more elaborate arrangements are occasionally used, but principally at frequencies of the order of 144 Mc. or higher. In the range 14 to 54 Mc., where the simple parasitic array finds its greatest use, the popular types are the "three-element beam" — a driven element with one reflector and one director — and the "four-element beam" — a driven element with one reflector and two directors. These are shown in Figs. 4-49 and 4-50.

The power gains of three- and four-element parasitic arrays cannot be estimated — even to an approximation — as satisfactorily as in the case of driven elements. This is because the amplitudes and phases of the currents in the parasitic elements depend on the spacing and tuning, and because at the close spacings used there is considerable coupling between the parasitic elements themselves as well as between the parasitic and driven elements. The number of variables is so large that calculation becomes impracticable. Information on such arrays is consequently based on experimental measurements in which one or more of the variables (the spacing, for example) is arbitrarily held constant. Such measurements are subject to several possible inaccuracies. To mention a few, gain measurements depend on the accuracy of calibration of field-strength measuring equipment, accurate determination of the power input to the antenna, elimination of stray radiation from transmission lines, and so on.

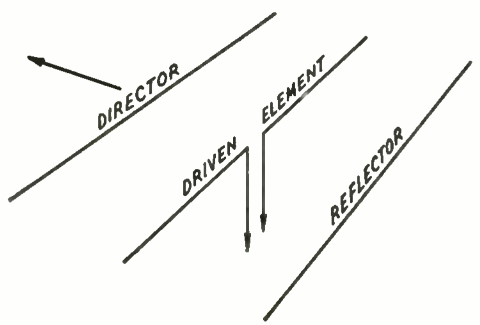


Fig. 4-49 — **Antenna system using a driven element and two parasitic elements, one as a reflector and one as a director.**

The results of a series of experimental measurements made by R. G. Rowe, W2FMF, with model antennas (antennas scaled down to operate at about 140 Mc., to eliminate some of the variables such as ground effects that are encountered at lower frequencies) are shown in Table 4-I. Measurements were made with two different types of field-intensity measuring equipment to check the results. As can be seen from the two gain columns, the results were in good agreement. In all cases shown in the table the element lengths were adjusted for maximum gain. In round figures, it can be said that a three-element array so adjusted has a gain of approximately 8 db. and a four-element array a gain of approximately 10 db. The gain depends somewhat upon the element spacing.

The front-to-back ratio depends critically on the element tuning in a multielement beam. As in the case of the two-element array, the maximum front-to-back ratio does not occur at the adjustment that gives greatest gain, but relatively little gain is sacrificed in adjusting for maximum front-to-back.

When the antenna is to be operated over a range of frequencies the gain, directive pattern, and front-to-back ratio can be expected to change with frequency. If the frequency applied to an antenna designed for, say, 29.0 Mc., is changed to 29.5 Mc., the physical lengths of the parasitic elements being fixed, the electrical lengths of the elements automatically become greater. If the frequency is increased sufficiently it is possible for an element to shift from acting as a director and become a reflector. If the frequency is *lowered* too much below the design value, a reflector can change into a director. In the average case a director is made roughly 5% shorter than the resonant length while a director is about 5% longer than the resonant length. The antenna can therefore be operated over a frequency range of about plus or minus 5% before a reversal of parasitic element operation takes place. This range is great enough to include the entire width of any band on which parasitic arrays are likely to be used, provided the antenna is designed for the center of the band. The changes in front-to-back ratio when working over such a range will be much more marked than the change in gain, since the adjustment for the former is considerably more critical.

Radiation Resistance

The radiation resistance at the center of the driven element in a parasitic array is always lower than the radiation resistance of a half-wave element alone. In multielement systems its value may be quite low with close-spaced elements. For example, the three-element array with the director 0.1 wavelength and the reflector 0.15 wavelength from the driven element will have a resistance in the vicinity of 8 to 10 ohms. With a four-element array at the same spacings and having a second director 0.1 wavelength from the first (all elements in the same plane) the resistance can drop as low as 4 to 6 ohms. The actual value in either case will depend on the parasitic element tuning.

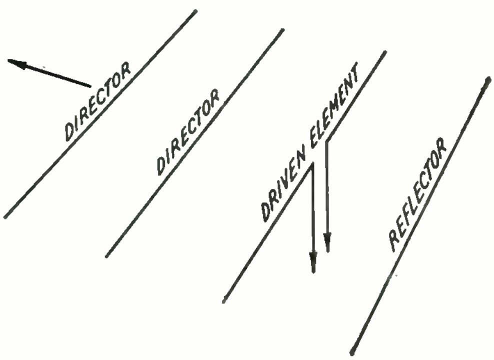


Fig. 4-50 — A "four-element" antenna system, using two directors and one reflector in conjunction with a driven element.

The low values of radiation resistance are accompanied by a high degree of selectivity in the antenna; that is, its impedance is constant over only a small frequency range. These changes in impedance make it troublesome to couple power from the transmitter to the line. Such difficulties can be reduced by using wider spacing — in particular, using spacings of the order of 0.2 wavelength or more — and by adopting expedients such as the folded dipole (for the driven element) that flatten out the impedance-frequency characteristic as seen by the transmission line.

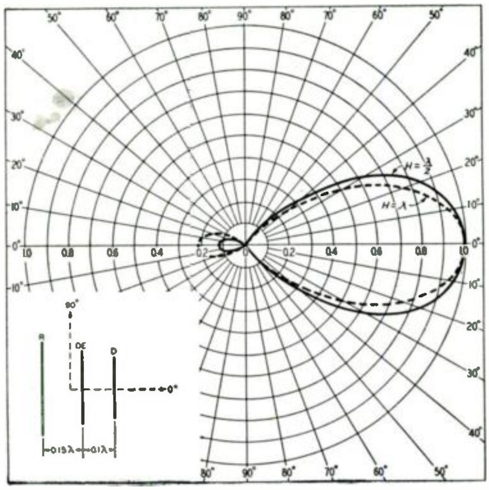


Fig. 4-51 — Measured radiation patterns of a horizontally-polarized 3-element array having the director 0.1 wavelength and the reflector 0.15 wavelength from the driven element. Element tuning adjusted for maximum gain. These "horizontal" patterns are for the wave angles at which the lowest lobe (Fig. 4-52) has its maximum. The wave angles are 28 degrees at a height of ½ wavelength and 12 degrees at a height of 1 wavelength.

Directive Patterns

There is, unfortunately, no existing analysis of three- and four-element parasitic arrays on which theoretical directive patterns can be based. The patterns of Figs. 4-51 to 4-54 inclusive are, like those of Figs. 4-47 and 4-48, based on experimental measurements made by W3GAU with model antennas. They show that the beam is somewhat sharper, as is to be expected, when the parasitic-element tuning is adjusted for maximum gain. Increasing the height of the antenna will of course lower the wave angle since the shape and amplitude of the vertical lobes are determined by the ground-reflection factors given in Chapter Two as well as by the free-space pattern of the antenna itself.

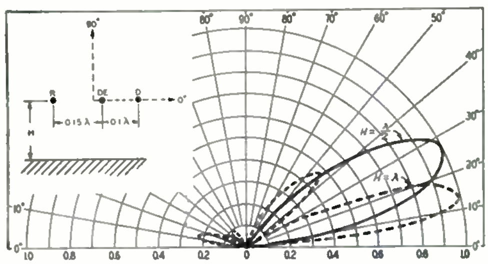


Fig. 4-52 — Vertical patterns of the antenna of Fig. 4-51 in the vertical plane at right angles to the direction of the antenna elements.

FEEDING AND ADJUSTMENT

The problems of matching and adjusting parasitic arrays for maximum performance are the same in principle as with other antenna systems. However, the adjustment process cannot be carried out quite as readily because parasitic elements are not, in general, tuned to resonance. Relative field-strength measurements usually are required if the adjustment procedure is to lead to the optimum operating conditions.

Methods of Feed

The driven element in a parasitic array is a load for the transmission line in the same way that a driven element in any antenna system is such a load. It differs from the load presented by a simple dipole only in that the resistance is quite low, the impedance may be both reactive and resistive under the optimum operating conditions, and the reactance usually changes more rapidly with frequency. These things mean that a fairly large impedance step-up is required for matching practicable lines, that it may be necessary to tune out some reactance, and that the amount of mismatch will increase more rapidly than with a simple dipole when the applied frequency is varied from that at which the line is matched.

Practically any of the matching systems detailed in Chapter Three are applicable. For example, an open quarter-wave matching section (Fig. 3-59A) can be used if the transmission line is to be 300- to 600-ohm parallel-conductor line. The quarter-wave transformer (Q) method also can be used, as shown in Fig. 3-43, with 75-ohm Twin-Lead for the transformer. This particular value of Z_0 will match a driven element having a resistive impedance of 9 to 10 ohms to a 600-ohm open-wire line, and will result in only a 2-to-1 mismatch if the driven-element impedance is as low as 4 or 5 ohms or as high as 18 to 20 ohms. The loss in a quarter-wave section of 75-ohm transmitting-type Twin-Lead, even though the s.w.r. in the matching transformer is rather high, will not be of any real consequence. Alternatively, the matching transformer may be made of 50- or 75-ohm coaxial cable. However, in such case some method of line balancing (such as those shown in Fig. 3-42) should be used. The delta match (Fig. 3-46) also may be used.

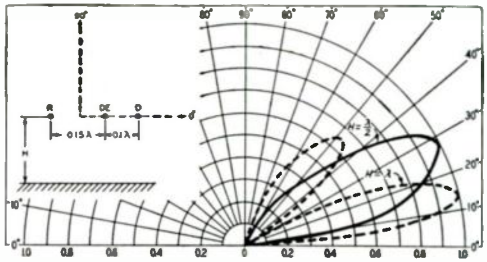


Fig. 4-54 — Vertical patterns of the antenna of Fig. 4-53 in the vertical plane at right angles to the direction of the antenna elements.

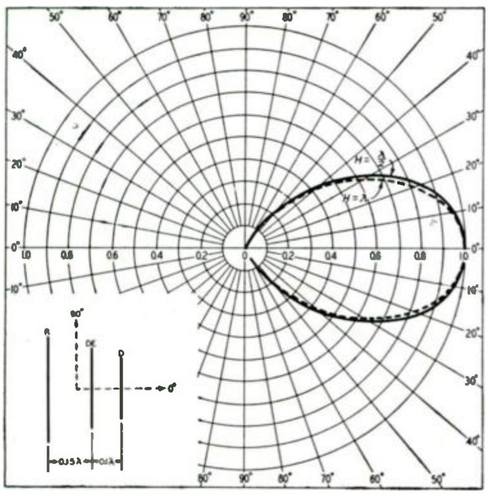


Fig. 4-53 — Measured radiation patterns of a horizontally-polarized 3-element parasitic array having the director 0.1 wavelength from the driven element and the reflector 0.15 wavelength from the driven element. Element tuning adjusted for maximum front-to-back ratio. Heights and wave angles are same as in Fig. 4-51.

The choice of a matching system is affected by constructional considerations, since parasitic arrays are frequently built to be rotated in operation. The "T"-match (Fig. 3-47) is a favorite with many amateurs because it fits in well, constructionally, when the driven element is made of tubing, and is relatively easy to adjust from a mechanical standpoint. Another matching system that is useful when the antenna is to be continuously rotatable is the inductive coupling shown in Fig. 3-61.

The adjustment procedure with all these matching systems is described in detail in Chapter Three.

It is possible to design a coupling system that will permit operating a parasitic array on two bands having a frequency ratio of 2 to 1. In this case the array operates as a combination parasitic-collinear array on the higher of the two frequencies. Design methods are rather involved and will not be treated here. However, a specific antenna system for 14- and 28-Mc. operation is described in Chapter Twelve.

Broadening the Response

It has already been pointed out that the tuning conditions giving maximum gain with parasitic elements are not highly critical. However, the varying amounts of reactance coupled into the driven element, as well as the fact that the radiation resistance at the center of the driven element is very low, cause the impedance to change rapidly when the applied frequency is varied above or below the design frequency.

This impedance change can be made less rapid by using fairly wide spacing between elements, as already mentioned. It is also beneficial to use elements having a fairly large ratio of diameter to length because, as explained in Chapter Two, the impedance change with frequency is noticeably reduced when the antenna conductor has a large diameter.

Another method of broadening the frequency-response curve is to use the folded-dipole principle in connecting the transmission line to the driven element. As described in Chapter Three, the folded dipole has the effect of reducing reactance variations, as viewed by the transmission line, with changes in frequency. A simple two-conductor folded dipole does not have a great-enough impedance step-up ratio for matching most types of transmission line, when the dipole is used as the driven element in a close-spaced array. However, it is readily possible to obtain a desired impedance ratio by using a multiconductor dipole or by using a two-conductor folded dipole constructed of conductors of unequal diameters.

In multiconductor folded dipoles using conductors of the same diameter the impedance ratio is equal to the square of the number of conductors, as explained in Chapter Three. Suppose the radiation resistance of the driven element is 5 ohms and that it is to be fed with 300-ohm line. Then the required impedance ratio is 300/5 = 60. The whole number whose square is nearest to 60 is 8, so the dipole should have 8 conductors to make the input impedance of the driven element as close as possible to 300 ohms. Theoretically, all the conductors should have the same diameter, but experimental work by G. N. Carmichael, W4GCA, indicates that the ratios are realized to a good approximation when the conductor to which the transmission line is connected is made of tubing while the other conductors are of wire, at least in the physical arrangement used by W4GCA and shown in Fig. 4-55. Table 4-II shows the number of conductors to use with various line impedances in three-and four-element arrays. The figures are based on using close-spaced elements — 0.1 wavelength for directors and 0.15 wavelength for reflectors. The length of the driven element can be calculated from $475/f$ (Mc.) decreased by twice the diameter of the cage. The cage diameter is not critical, a diameter of 8 inches being satisfactory both electrically and mechanically when a number of wires are used.

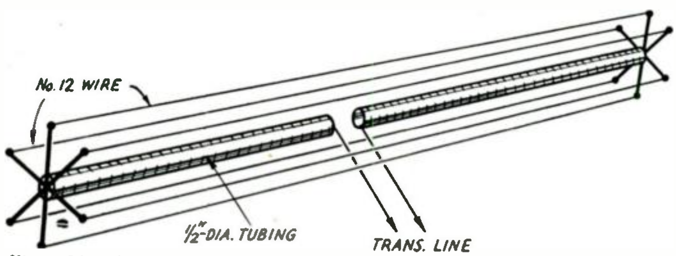


Fig. 4-55 — Multiconductor folded dipole for use as the driven element in a parasitic array. The number of conductors required depends on the number of antenna elements and the line impedance, as shown in Table 4-II.

The design of a folded-dipole driven element using two conductors of unequal diameter can be based on the nomogram of Fig. 3-51. For example, an 8-ohm driven element to be fed by 75-ohm cable requires an impedance step-up of 75/8 = 9.4. From Fig. 3-51, and choosing a ratio of 3 for R_2/R_1, it is found that D/R_1 is approximately 8. If R_1 is ¼ inch and R_2 is ¾ inch, the center-to-center spacing is 8 × ¼ = 2 inches. By using end clamps that permit adjusting the spacing between the two conductors it is possible to adjust the system for a close match after the antenna is built.

In using either of the folded-dipole methods it has been found that designing the antenna for the center of the band permits working over the whole band without significant loss of gain at either end.

Adjusting Parasitic Arrays

There are two separate processes in adjusting an array with parasitic elements. One is the determination of the optimum element lengths, depending on whether maximum gain or maximum front-to-back ratio is desired. The other is matching the antenna to the transmission line. The second is usually dependent on the first, and the results observed on adjusting the element tuning may be meaningless unless the line is equally well matched under all tuning conditions.

The approximate element lengths for maximum

TABLE 4-II
Number of Conductors Required in a Folded-Dipole Driven Element in Parasitic Arrays

	Line Impedance in Ohms	*Conductors in Driven Element*
Two Parasitic Elements	72	3
	150	4
	300	6
	500	7
	600	8
Three Parasitic Elements	72	4
	150	7
	300	8
	500	9
	600	10

gain are as follows:

$$\textbf{Driven element: } \textit{Length} \text{ (ft.)} = \frac{475}{f \text{ (Mc.)}}$$

$$\textbf{Director: } \textit{Length} \text{ (ft.)} = \frac{455}{f \text{ (Mc.)}}$$

$$\textbf{Reflector: } \textit{Length} \text{ (ft.)} = \frac{500}{f \text{ (Mc.)}}$$

These are average lengths determined experimentally for elements having a length/diameter ratio of 200 to 400, and with element spacings from 0.1 to 0.2 wavelength.

Many amateurs have found that very satisfactory results are secured simply by cutting the elements to the lengths given by formula. It has been a rather common experience that, after a considerable amount of time has been spent in trying all possible adjustments, the dimensions finally determined to be optimum are very close to those given by the formulas and the actual difference in gain is negligible. It appears safe to say, therefore, that in the average case there is probably little to be realized, in the way of increased gain, by spending much time in adjusting element lengths. The front-to-back ratio can often be improved, however, since it is much more critical with respect to element tuning. In general, the reflector tuning is the more critical.

If the array is put up by formula, the only adjustment that need be made is to match the driven element to the transmission line. The adjustment procedure for each type of matching arrangement is described in Chapter Three. If a folded-dipole driven element of the multiconductor type is used no adjustment is necessary, but it is advisable to check the s.w.r. on the transmission line. If it is more than 2 to 1 or so better results may be obtained by using one more or one less conductor in the driven element. If the impedance at the feed point is lower than the line Z_0 an additional conductor should be used, and vice versa. The method of determining whether the load resistance is higher or lower than the line Z_0 is outlined in Chapter Three (Fig. 3-62). With an unequal-conductor folded dipole the match may be brought about by changing the conductor spacing.

Test Set-up

The only practicable method of adjusting parasitic element lengths for best performance is to measure the field strength from the antenna as adjustments are made. Measurements on a relative basis are entirely satisfactory for the purpose of determining the operating conditions that result in the maximum output or greatest front-to-back ratio. For this purpose the measuring equipment does not need to be calibrated; the only requirement is that it indicate whether the signal is stronger or weaker.

If the help of a near-by amateur owning a receiver with an S-meter can be enlisted, the S-meter indications can be used to indicate the relative field strength. A few precautions must be taken if this method is to be reliable. The receiving antenna must have the same polarization as the transmitting antenna under test (this is usually horizontal), and should be reasonably high above its surroundings. The receiving system should be checked for pick-up on the transmission line to make sure that the indications given by the receiver are caused entirely by energy picked up by the receiving antenna itself. This can be checked by temporarily disconnecting the line from the antenna (but leaving it in place) and observing the signal strength on the S-meter. If the reading is not several S points below the reading with the antenna connected the readings cannot be relied upon when adjusting the transmitting antenna for maximum gain. In checking the front-to-back ratio, the stray pick-up at the receiving installation must be well below the smallest signal received via the antenna, if the adjustments are to mean anything at all.

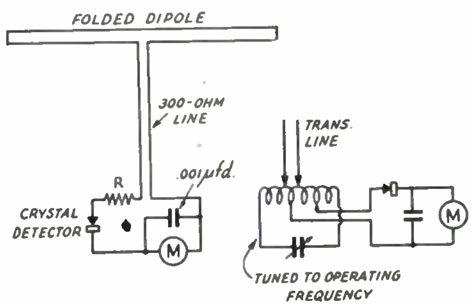


Fig. 4-56 — Field-strength measurement set-up. The folded dipole should be at least as high as the antenna under test and should be three or more wavelengths away. *R* should be a 300-ohm carbon resistor to provide a proper load for the line, so that a line of any desired length can be used. If the sensitivity is not high enough with this arrangement the alternative connections at the right will result in increased meter readings. The taps should be adjusted for maximum reading, keeping the transmission-line taps spaced equally on either side of the coil center-tap. The indicating meter, *M*, may be either a microammeter or 0–1 milliammeter, depending on the sensitivity required.

Another method of checking field strength is to use a field-strength indicator of the crystal-detector type. The preferable method of using such an indicator is to connect it to a dipole antenna mounted some distance away and at a height at least equal to that of the transmitting antenna. There should be no obstructions between the two antennas, and both should have the same polarization. The receiving dipole need not be a half-wave long, although that length is desirable because it will increase the ratio of energy picked up on the antenna to energy picked up by stray means. To prevent coupling effects the distance between the two antennas should be at least three wavelengths. At shorter distances the mutual impedance may be large enough to cause the receiving antenna to tend to become part of the transmitting system, which can lead to false results. A recommended type of indicating system is shown in Fig. 4-56. The transmission line should drop vertically down to the indicator, to avoid stray pick-up. This pick-up can be checked as described in the preceding paragraph. If the distance between the two antennas is such that greater sensitivity is needed a reflector may be placed ¼ wavelength behind the receiving dipole.

Adjustment Procedure

It is advisable first to set the element lengths to those given by the formulas and then match the driven element to the transmission line, obtaining as low an s.w.r. as possible. In subsequent adjustments a close watch should be kept on the s.w.r. and the transmitter power input should be maintained at exactly the same figure throughout. If the s.w.r. changes enough to affect the coupling at the transmitter when an adjustment is made, but not enough to raise the line loss significantly (see Fig. 3-21), readjust the coupling to bring the input back to the same value. If the line loss increases more than a half db. or so, rematch at the driven element.

The experience of most amateurs in adjusting parasitic arrays indicates that there is not a great deal of preference in the order in which elements are tuned, but that there is slightly less interlocking if the director is first adjusted to give maximum gain and the reflector is then adjusted to give either maximum gain or maximum front-to-back ratio, whichever is desired. After the second parasitic element has been adjusted, go back and check the tuning of the first to make sure that it has not been thrown out of adjustment by the mutual coupling. If there are three parasitic elements, the other two should be checked each time an appreciable change is made in one. The actual lengths should not be very far from those given by the formulas when the optimum settings are finally determined. As already pointed out, the reflector length may be somewhat greater when adjusted to give maximum front-to-back ratio.

Radiation from the transmission line must be eliminated, or at least reduced to a very low value compared with the radiation from the antenna itself, if errors are to be avoided. Conditions are usually favorable to low line radiation in horizontally-polarized rotatable parasitic arrays because the line is usually symmetrical with respect to the antenna and is brought away perpendicular to it, at least for a half wavelength or so. Nevertheless the line radiation can be appreciable unless the line is detuned as described in Chapter Three. With coaxial line some method of line balancing at the antenna always should be incorporated.

After arriving at the optimum adjustments at the frequency for which the antenna was designed, the performance should be checked over a frequency range either side of the design frequency to observe the sharpness of response. If the field strength falls off rapidly with frequency, it may be desirable to shorten the director a bit to increase the gain at frequencies above resonance and lengthen the reflector slightly to increase it at frequencies below resonance. Do not confuse the change in s.w.r. with the change in antenna gain. The antenna itself may give good gain over a considerable frequency range, but the s.w.r. may vary between wide limits in this range. To check the *antenna* behavior, keep the power input to the transmission line constant and rematch the driven element to the line, as suggested above, whenever the line losses increase appreciably. If such rematching is found necessary over the band of frequencies to be used, it is advisable to change to a matching system such as the folded dipole that is inherently capable of operating over a wide frequency band.

Adjustment by Reception

As an alternative to applying power to the array and checking the field strength, it is possible to adjust the array by measuring received signal strength. It is impracticable to do this on distant signals because of fading. The most reliable method is to erect a temporary antenna of the same type recommended for field-strength measurements (Fig. 4-56) and excite it from a low-power oscillator. The same precautions with respect to distance between the two antennas apply.

In this method, as in the one where the transmitting antenna is excited, it is necessary to minimize line radiation and pick-up if the results are to be reliable. The same tests may be applied. However, it is less easy to keep the s.w.r. under control. In the receiving case the s.w.r. on the transmission line depends on the load presented by the *receiver* to the line. Under most conditions the s.w.r. will be reasonably constant over an amateur band, although its value may not be known. However, the energy transfer from the antenna to the line depends on the mismatch between the driven element and the line. There is no convenient way to check this in the receiving case. About all that can be done is to apply power to the array after a set of tuning conditions has been reached, and then rematch at the driven element if necessary. After rematching the measurement will have to be repeated. Thus double checking is necessary if the results are to be comparable with those obtained by the field-strength method.

COMBINATION ARRAYS WITH PARASITIC ELEMENTS

Parasitic elements may be combined with collinear or broadside driven elements to form multielement arrays giving higher gain and directivity than the simple parasitic arrays described. In such cases, one set of parasitic elements is associated with each driven element.

Arrays of this type are not used by amateurs to any extent at frequencies below 30 Mc., probably because of the constructional difficulties involved. Another factor that operates against too-high gain and directivity at the lower frequencies is the fact that amateur work is seldom concentrated in one direction. A sharp beam is most useful when the antenna can be rotated, and there are, of course, limitations on the size of rotating structure that can be built. The use of high-gain multielement arrays is consequently confined to the v.h.f. region where the shorter wavelengths make them entirely practicable, constructionally.

There is no theoretical reason why elements cannot be combined in any desired way, so long as the principles discussed in this chapter are observed. However, there are always two considerations that influence the selection of an element arrangement: the distribution of power to the elements, and construction that is satisfactory from a mechanical standpoint. Arrays for the v.h.f. bands that have been found to meet these problems satisfactorily are described in Chapter Nine.

CHAPTER 5

Long-Wire Antennas

The power gain and directive characteristics of the harmonic wires (which are "long" in terms of wavelength) described in Chapter Two make them useful for long-distance transmission and reception on the higher frequencies. In addition, long wires can be combined to form antennas of various shapes that will increase the gain and directivity over a single wire. The term "long wire" as used in this chapter means any such configuration, not just a straight-wire antenna.

Long Wires vs. Multielement Arrays

In general, the gain achieved with long-wire antennas is not as great, when the space available for the antenna is limited, as can be secured from the multielement arrays in Chapter Four. To offset this the long-wire antenna has advantages of its own.

The construction of long-wire antennas is simple both electrically and mechanically, and there are no especially critical dimensions or adjustments. The long-wire antenna will work well and give satisfactory gain and directivity over a 2-to-1 frequency range; in addition, it will accept power and radiate it well on any frequency for which its over-all length is not less than about a half wavelength. Since a wire is not long at high frequencies unless its length is at least equal to a half wavelength on 3.5 Mc., any long-wire antenna can be used on all amateur bands that are useful for long-distance communication.

As between two directive antennas having the same theoretical gain, one a multielement array and the other a long-wire antenna, many amateurs have found that the long-wire antenna is actually more effective in communication. This seems to be particularly true in reception, and possibly is the result of the fact that, because of its greater effective area, the long-wire antenna is exposed to more of the energy in an incoming wave. That is, its pick-up efficiency (Chapter Two) possibly is better than that of the more concentrated multielement array. Another factor is that long-wire antennas have directive patterns that are sharp in both the horizontal and vertical planes, and tend to concentrate the radiation at the low vertical angles that are most useful at the higher frequencies.

General Characteristics of Long-Wire Antennas

Whether the long-wire antenna is a single wire running in one direction or is formed into a V, rhombic, or some other configuration, there are certain general principles that apply and some performance features that are common to all types.

The first of these is that the power gain of a long-wire antenna as compared with a half-wave dipole is not considerable until the antenna is really "long" — long, that is, when the lengths are measured in wavelengths rather than in a specific number of feet. The reason for this is that the fields radiated by elementary lengths of wire along the antenna do not combine, at a distance, in as simple a fashion as the fields from half-wave dipoles used as described in Chapter Four. There is no point in space, for example, where the distant fields from all points along the wire are exactly in phase (as they are, in the optimum direction, in the case of two or more collinear or broadside dipoles when fed with in-phase currents). Conse-

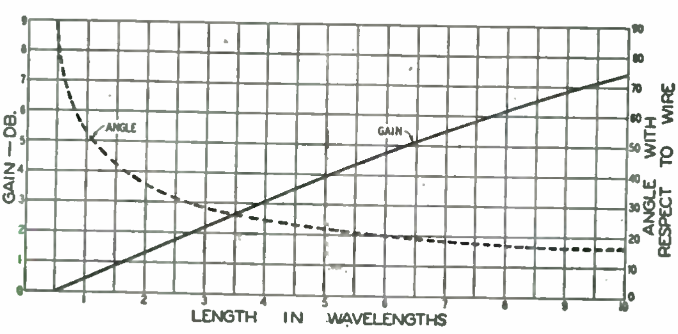


Fig. 5-1 — Theoretical gain of a long-wire antenna over a dipole as a function of wire length. The angle, with respect to the wire, at which the radiation intensity is maximum also is shown.

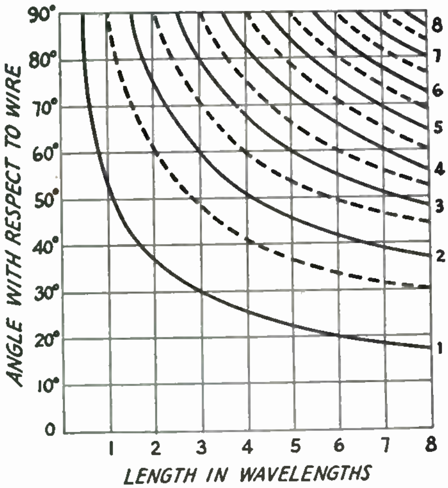


Fig. 5-2 — Angles at which radiation from long wires is maximum (solid curves) and zero (broken curves). The major lobe, No. 1, has the power gains given by Fig. 5-1. Secondary lobes have smaller amplitude, but the maxima may exceed the radiation intensity from a half-wave dipole.

quently, the field strength at a distance always is less than would be obtained if the same length of wire were cut up into properly-phased separately-driven dipoles. As the wire is made longer the fields combine to form an increasingly intense main lobe, but this lobe does not develop rapidly until the wire is several wavelengths long. The longer the antenna, the sharper the lobe becomes; and since it is really a cone of radiation about the wire in free space, it becomes sharper in all planes. Also, the greater the length the smaller the angle with the wire at which the maximum radiation occurs.

Because many points along a long wire are carrying currents in different phase (usually with different current amplitude as well) the field pattern at a distance becomes more complex as the wire is made longer. This complexity is manifested in a series of minor lobes, the number of which increases with the wire length. The intensity of radiation from the minor lobes is frequently as great as, and sometimes greater than, the radiation from a half-wave dipole. The energy radiated in the minor lobes is not available to improve the gain in the major lobe, which is another reason why a long-wire antenna has to be long to give appreciable gain.

Driven and parasitic arrays of the simple types described in Chapter Four do not have minor lobes of any great consequence. For that reason they frequently seem to have better directivity than long-wire antennas, because their response in directions other than that at which the antenna is aimed is well down. This will be so even if a multielement array and a long-wire antenna have the same actual gain in the favored direction. For amateur work, particularly with directive antennas that cannot be rotated, the minor lobes of a long-wire antenna have some advantages. In most directions the antenna will be as good as a half-wave dipole, and in addition will give high gain in the most favored direction; thus a long-wire antenna (depending on the design) frequently is a good all-around radiator in addition to being a good directive antenna.

In the discussion of directive patterns of long-wire antennas in this chapter, it should be kept in mind that the radiation patterns of resonant long wires are based on the assumption that each half-wave section of wire carries a current of the same amplitude. As pointed out in Chapter Two, this is not exactly true, since energy is radiated as it travels along the wire. For this reason it is to be anticipated that, although the theoretical pattern is bidirectional and identical in both directions, actually the radiation (and reception) will be best in one direction. This effect becomes more marked as the antenna is made longer.

Wave Angles

The wave angle at which maximum radiation takes place from a long-wire antenna depends on the same factors that operate in the case of simple dipoles and multielement antennas. That is, the directive pattern in the presence of ground is found by multiplying the free-space vertical-plane pattern of the antenna by the ground-reflection factors for the particular antenna height used. These factors are discussed in Chapter Two.

As mentioned a few paragraphs ago, the free-space radiation pattern of a long-wire antenna has a major lobe that forms a cone around the wire. The angle at which maximum radiation takes place becomes smaller, with respect to the wire, as the wire length is increased. For this reason a long-wire antenna is primarily a low-angle radiator when installed horizontally above the ground. Its performance in this respect is improved by selecting a height that also tends to concentrate the radiation at low wave angles.

Antenna systems formed from ordinary horizontal dipoles that are not stacked have, in most cases, a rather broad vertical pattern; the wave angle at which the radiation is maximum therefore depends chiefly on the antenna height. However, with a long-wire antenna the wave angle at which the major lobe is maximum can never be as great as the angle at which the first null occurs (see Fig. 5-2) even if the height is very low. (The efficiency may be less at very small heights, partly because of increased losses in the ground and partly because the pattern is affected in such a way as to put a greater proportion of the total power into the minor lobes.) The result is that when radiation at wave angles below 15 or 20 degrees is under consideration a long-wire antenna is less sensitive to height than are the multielement arrays or a simple dipole. To assure good results, however, the antenna should have a height equivalent to at least a half wavelength at 14 Mc. — that is, a minimum height of about 30 feet. Greater heights will give a worth-while improvement at wave angles below 10 degrees.

With an antenna of fixed physical length and height, both length and height increase, in terms of wavelength, as the frequency is increased. The over-all effect is that both the antenna and the ground reflections tend to keep the system operating at high effectiveness throughout the frequency range. At low frequencies the wave angle is raised, and high wave angles are useful at 7 and 3.5 Mc. At high frequencies the converse is true. Good all-around performance usually results on all bands when the antenna is designed to be optimum in the 14-Mc. band.

Calculating Length

In this chapter lengths are always discussed in terms of wavelength. There can obviously be nothing very critical about wire lengths in an antenna system that will work over a frequency range including several amateur bands. The antenna characteristics change very slowly with length, except when the wires are short (i.e., around one wavelength) and there is no need to try to establish exact resonance at a particular frequency.

The formula for harmonic wires given in Chapter Two is quite satisfactory for determining the lengths of any of the antenna systems to be described. For convenience, the formula is given here in the following form:

$$\textit{Length}\ \text{(feet)} = \frac{984\ (N - 0.025)}{\textit{Freq.}\ \text{(Mc.)}}$$

where N is the number of full waves on the antenna. In cases where exact resonance is desired for some reason (for obtaining a resistive load for a transmission line at a particular frequency, for example) it is best established by trimming the wire length until measurement of the resonant frequency shows it to be correct.

Long Single Wires

The directional characteristics of wires that are multiples of a half wave in length have been discussed in Chapter Two. For convenience, the power gain over a dipole in free space and the angle with the wire at which the radiation is maximum are repeated in Fig. 5-1. The solid curve shows that the gain in decibels increases almost linearly with the length of the antenna. The gain does not become appreciable until the antenna is about four wavelengths long, where it is equivalent to doubling the transmitter power (3 db.). The actual gain over a half-wave dipole when the antenna is at a practical height above ground will depend on the way in which the radiation resistances of the long-wire antenna and the comparison dipole are affected by the height. This point was discussed in connection with the half-wave dipole in Chapter Two. The exact way in which the radiation resistance of a long wire varies with height depends on its length. In general, the resistance does not fluctuate as much, in terms of percentage, as does the resistance of a half-wave antenna. This is particularly true at heights from one-half wavelength up. Consequently, at a particular height the gain over a half-wave antenna at the same height will depend principally on how much the radiation resistance of the latter departs from the free-space value of 73 ohms on which the gain curve of Fig. 5-1 is based. The gain will be greatest at heights that make the dipole radiation resistance reach its highest values, approximately.

The minor lobes in the directive pattern have a maximum intensity approximately equivalent to that from a half-wave dipole. The nulls bounding the lobes are fairly sharp and are frequently somewhat obscured, in practice, by irregularities in the pattern. The locations of nulls and maxima for antennas up to eight wavelengths long are shown in Fig. 5-2.

Orientation

The broken curve of Fig. 5-1 shows the angle with the wire at which the radiation intensity is maximum. As shown in Chapter Two, there are two main lobes to the directive patterns of long-wire antennas; each makes the same angle with respect to the wire and each, considered in free space, is the solid formed by rotating the wire on its axis.

When the antenna is mounted horizontally above the ground, the situation depicted in Fig. 5-3 exists. Only one of the two lobes is considered in this drawing, and its lower half is cut off by the ground. The maximum intensity of radiation in the remaining half occurs through the broken-line semicircle; that is, the angle (B) between the wire direction and the line marked "wave direction" is the angle given by Fig. 5-1 for the particular antenna length used.

In the practical case, there will be some wave angle (A) that is optimum for the frequency and the distance between the transmitter and receiver. Then for that wave angle the wire direction and the optimum geographical direction of transmission are related by the angle C. If the wave angle is very low, B and C will be practically equal. But as the wave angle becomes higher the angle C becomes smaller; in other words, the best direction of transmission and the direction of the wire more nearly coincide. They coincide exactly when C is zero; that is, when the wave angle is the same as the angle given by Fig. 5-1.

The maximum radiation from the antenna can be aligned with a particular geographical direction at a given wave angle by means of the following formula:

$$\cos C = \frac{\cos B}{\cos A}$$

In most amateur work the chief requirement is that the wave angle should be as low as possible, particularly at 14 Mc. and above. In such case it is usually satisfactory to make angle C the same as is given by Fig. 5-1.

It should be borne in mind that only the *maximum* point of the lobe is represented in Fig. 5-3. Radiation at higher or lower wave angles in any given direction will be proportional to the way in which the actual pattern shows the field strength to vary as compared with the maximum point of the lobe.

Tilted Wires

Fig. 5-3 shows that when the wave angle is equal to the angle which the maximum intensity of the lobe makes with the wire the best transmitting or receiving direction is that of the wire itself. If the wave angle is less than the lobe angle the best direction can be made to coincide with the direction of the wire by tilting the wire enough to make the lobe and wave angle coincide. This is shown in Fig. 5-4, for the case of a one-wavelength antenna tilted so that the maximum radiation from one lobe is horizontal to the left, and from the other is horizontal to the right (zero wave angle). The solid pattern can be visualized by imagining the plane diagram rotating about the antenna as an axis.

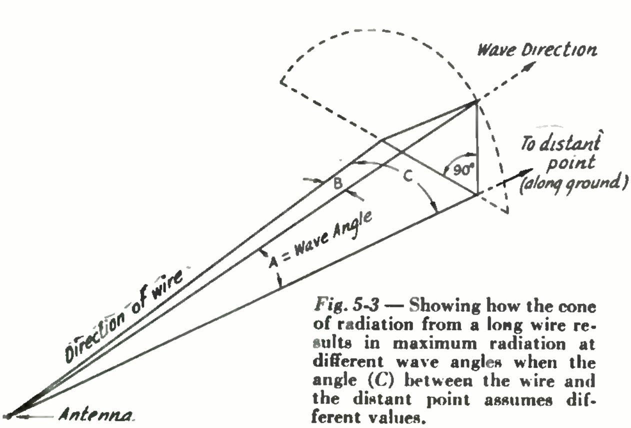


Fig. 5-3 — Showing how the cone of radiation from a long wire results in maximum radiation at different wave angles when the angle (C) between the wire and the distant point assumes different values.

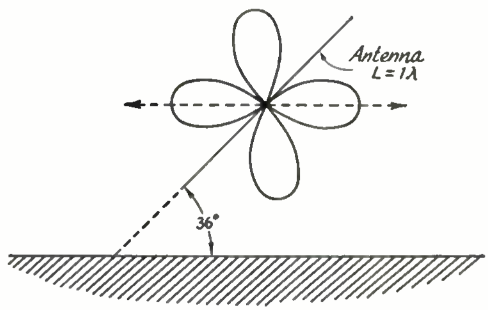


Fig. 5-4 — Alignment of lobes for horizontal transmission by tilting a long wire in the vertical plane.

Since the antenna is neither vertical nor horizontal in this case, the radiation is part horizontally polarized and part vertically polarized. Computing the effect of the ground becomes complicated, because the horizontal and vertical components must be handled separately. In general, the directive pattern at any given wave angle becomes unsymmetrical when the antenna is tilted. For small amounts of tilt (less than the amount that directs the lobe angle horizontally) and for low wave angles the effect is to shift the optimum direction closer to the line of the antenna. This is true in the direction in which the antenna slopes downward. In the opposite direction the low-angle radiation is reduced.

Feeding Long Wires

It has been pointed out in Chapter Three (Fig. 3-31) that a harmonic antenna can be fed only at the end or at a current loop. Since a current loop changes to a node when the antenna is operated at any even multiple of the frequency for which it is designed, a long-wire antenna will operate as a true long wire on all bands only when it is fed at the end.

A common method of feeding is to use a resonant open-wire line as shown at A in Fig. 5-5. This system will work on all bands down to the one, if any, at which the antenna is only a half wave long. Any convenient line length can be used if the transmitter is matched to the line input impedance by the methods described in Chapter Three.

Two arrangements for using nonresonant lines are given at B and C. The one at B is useful for one band only, since the matching section must be a quarter wave long, approximately, unless a different matching section is used for each band. In C, the Q-section impedance should be adjusted to match the antenna to the line as described in Chapter Three, using the values of radiation resistance given in Fig. 2-23. This method is best suited to working with a 600-ohm transmission line. Although it will work as designed only on one band, the antenna can be used on other bands by treating the line and matching transformer as a resonant line. In such case, as mentioned earlier, the antenna will not radiate as a true long wire on even multiples of the frequency for which the matching system is designed.

The end-fed arrangement, although the most convenient when tuned feeders are used, suffers the disadvantage that there is likely to be a considerable antenna current on the line, as described in Chapter Three. In addition, the antenna reactance changes rapidly with frequency for the reasons outlined in Chapter Two (Figs. 2-9 and 2-10). Consequently, when the wire is several wavelengths long a relatively small change in frequency — a fraction of the width of a band — may require major changes in the adjustment of the transmitter-to-line coupling apparatus. Also, the line becomes unbalanced at all frequencies between those at which the antenna is exactly resonant. This leads to a considerable amount of radiation from the line. The unbalance can be overcome by using two wires in one of the arrangements described in succeeding sections.

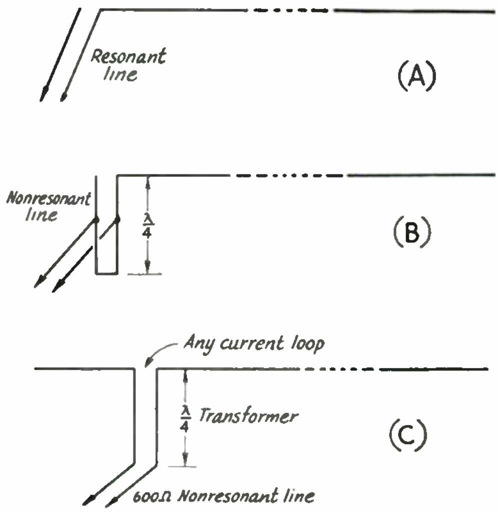


Fig. 5-5 — Methods of feeding long single-wire antennas.

Combinations of Resonant Long Wires

The directivity and gain of long wires may be increased by using two wires so placed in relation to each other as to make the fields from both combine to produce the greatest-possible field strength at a distant point. The principle is similar to that used in forming the multielement arrays described in Chapter Four. However, the maximum radiation from a long wire occurs at an angle of less than 90 degrees with respect to the wire, so different physical relationships have to be used.

Parallel Wires

One possible method of using two (or more) long wires is to place them in parallel, with a spacing of ½ wavelength or so, and feed the two in phase. In the direction of the wires the fields will add up in phase. However, since the wave angle is greatest in the direction of the wire, as shown by Fig. 5-3, this method will result in rather high-angle radiation unless the wires are several wavelengths long. The wave angle can be lowered, for a given antenna length, by tilting the wires as described earlier. With a parallel arrangement of this sort the gain should be about 3 db. over a single wire of the same length, at spacings in the vicinity of ½ wavelength.

● ECHELON ANTENNA

A better arrangement, and one that illustrates the principle of combining parallel long wires to obtain desired directive effects, is shown in Fig. 5-6. In this case the wires are still parallel but are also staggered; that is, the wires form two sides of a parallelogram but not of a rectangle. *A* and *B* are corresponding points on each of the wires. All the lines shown in Fig. 5-6 are in the same plane, that containing the antenna wires. The angle α between *YY* and the upper wire is made equal to the angle at which the radiation from a

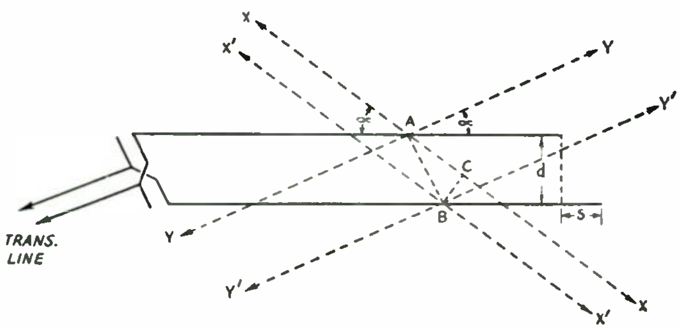


Fig. 5-6 — The echelon antenna, using two parallel long wires. The effect of the stagger distance, *s*, and spacing, *d*, on the directive pattern are discussed in the text.

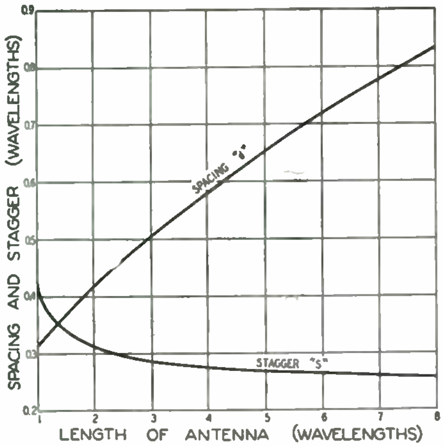


Fig. 5-7 — **Chart for determining stagger distance and spacing of an echelon antenna as a function of wire length.**

wire of that length is maximum, as given by Fig. 5-1. The angle between the wire and XX is the same, when XX lies in the direction of maximum radiation from the other pair of lobes in the plane directive diagram. $Y'Y'$ is parallel to YY, and $X'X'$ is parallel to XX, since it is assumed that any point considered along YY will be so far distant that the waves from A and B follow parallel paths; similarly for any point along XX.

If the stagger between the wires is such that the line joining A and B is perpendicular to YY, waves radiated from A and B will travel exactly the same distance to reach any distant point along YY. This will be true of any pair of corresponding points along the antennas. If the two wires are fed in phase and carry the same currents, the fields at the distant point will add in phase. However, if the two wires are fed equal currents out of phase, the fields at a distant point along YY will cancel.

Now if the distance between the wires is such that the distance AC (BC being perpendicular to XX) is equal to one-half wavelength, there will be a difference of exactly one-half wavelength in the distance traveled by waves from A and B to reach a distant point along XX. Again this will be true of any pair of corresponding points along the wires. If the two wires are fed equal currents out of phase, the additional 180-degree phase delay because of the spacing makes the waves arrive in phase at the distant point. However, if the wires are fed equal currents in phase the delay caused by the spacing makes the waves cancel along the line XX.

Thus by using long wires in this "echelon" arrangement the radiation can be increased in one direction, say XX, and simultaneously canceled in the other, YY. By changing the phasing, YY will be the favored direction and cancellation will take place along XX. The preferred method is to feed the two wires out of phase, since this permits the transmission line to be balanced as shown in Fig. 5-6, and the mutual impedance between the wires is such as to increase the gain. In such case the maximum direction is XX and the lobes that would lie along YY, in the case of a single wire, will be eliminated. However, in a practical installation these directions can be interchanged, if desired, simply by changing the phasing of the elements. The principles discussed in Chapter Four in connection with element phasing apply here equally well.

The amount of stagger can be calculated from

$$s\text{ (feet)} = \frac{492 \sin \alpha}{f\text{ (Mc.)} \sin 2\alpha}$$

and the distance between wires is given by

$$d\text{ (feet)} = \frac{492 \cos \alpha}{f\text{ (Mc.)} \sin 2\alpha}$$

where α = angle of maximum radiation from a single wire. Fig. 5-7 gives the information in graphical form, with the lengths expressed in fractions of a wavelength.

Using wires in echelon has one disadvantage: the fact that the spacing and stagger distances are critical with respect to both wavelength and the length of the wires confines the operation to the frequency band for which the antenna is designed. This is probably one reason why it has not had much amateur use, despite the fact that it offers a choice of two bidirectional beams. The broad frequency characteristic that constitutes one of the principal advantages of the single long wire is retained in the other configurations described below.

THE V ANTENNA

Instead of using two long wires parallel to each other they may be placed in the form of a horizontal V, with the angle at the apex of the V twice the angle given by Fig. 5-1 for the particular length of wire used. The currents in the two wires should be out of phase. Under these conditions the plane directive patterns of the individual wires combine as is indicated in Fig. 5-8. Along a line in the plane of the antenna and bisecting the V the fields from the individual wires reinforce each other at a distant point. The other pair of lobes in the plane pattern is more or less eliminated, so that the pattern becomes essentially bidirectional.

The directional pattern of an antenna of this type is sharper in both the horizontal and vertical planes than the patterns of the individual wires composing it. Maximum radiation in both planes is along the line bisecting the V. There are minor lobes in both the horizontal and vertical patterns but if the legs are long in terms of wavelength the amplitude of the minor lobes is small. When the antenna is mounted horizontally above the ground (the usual method) the wave angle at which the radiation from the major lobe is maximum is determined by the height, but cannot exceed the angle values shown in Fig. 5-1 for the leg length used. Only the minor lobes give high-angle radiation.

The gain and directivity of a V depend on the length of the "legs." Fig. 5-9 shows the gain and gives the maximum apex angle to use with legs up to 10 wavelengths long. The gains shown in this figure are approximate only; the solid curve is simply the gain of a single wire plus 3 db. for the effect of the second wire. The actual gain will be modified by the mutual impedance between the sides of the V, and data are available for only two cases, for leg lengths of 1 and 8 wavelengths. Although the broken gain curve in Fig. 5-9 merely joins the theoretical gains for these two cases, it is probably closer to the actual gain than the solid curve.

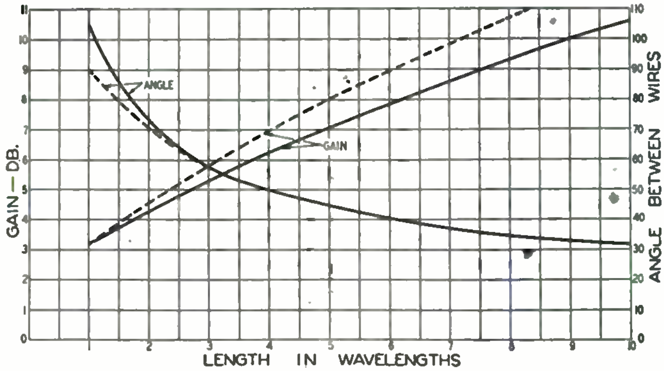


Fig. 5-9 — **Apex angle of V antenna as a function of the leg length in wavelengths. Curves of estimated gain also are shown. See text for discussion of these curves.**

When the leg length is small there is some advantage in reducing the apex angle of the V because this changes the mutual impedance in such a way as to increase the gain of the antenna. The approximate change realized is given by the broken section in the curve marked "angle." For example, the optimum apex angle in the case of a one-wavelength (each leg) V antenna is 90 degrees.

Lobe Alignment

It is possible to align the lobes from the individual wires with a particular wave angle by the method described in connection with Fig. 5-3. At very low wave angles the change in the apex angle is extremely small; for example, if the desired wave angle is 5 degrees the apex angles given in Fig. 5-9 will not be reduced more than a degree or so, even at the longest leg lengths which might be used.

When the legs are long, alignment does not necessarily mean that the greatest signal strength will be secured at the wave angle for which the apex angle is chosen. It must be remembered that the polarization of the radiated field is the same as that of a plane containing the wire. As illustrated by the diagram of Fig. 5-3, at any wave angle other than zero the plane containing the wire and passing through the desired wave angle is not horizontal. In the limiting case where the wave angle and the angle of maximum radiation from the wire are the same the plane is vertical, and the radiation at that wave angle is vertically polarized. At in-between angles the polarization consists of both horizontal and vertical components.

When two wires are used in a V these planes have opposite slopes. In the plane bisecting the V, this makes the horizontally-polarized components of the two fields add together numerically, but the vertically-polarized components are out of phase and cancel completely. As the wave angle is increased the horizontally-polarized components become smaller, so the intensity of horizontally-polarized radiation decreases. On the other hand, the vertically-polarized components become more intense but always cancel each other. The over-all result is that although alignment for a given wave angle will increase the useful radiation at that an-

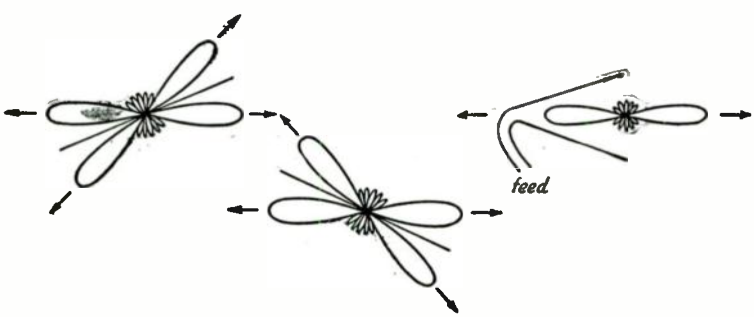


Fig. 5-8 — **Two long wires and their respective patterns are shown at the left. If these two wires are combined to form a V whose apex angle is twice that of the major lobes of the wires, and the wires are excited out of phase, the radiation along the bisector of the V adds and the radiation in the other directions tends to cancel.**

gle, the wave angle at which maximum radiation occurs (in the direction of the line bisecting the V) is always below the wave angle for which the wires are aligned. As shown by Fig. 5-10, the difference between the apex angles required for optimum alignment of the lobes at wave angles of zero and 15 degrees is rather small, even when the legs are many wavelengths long.

For long-distance transmission and reception the lowest possible wave angle usually is the best. Consequently, it is good practice to choose an apex angle between the limits represented by the two curves in Fig. 5-10. The actual wave angle at which the radiation is maximum will depend on the shape of the vertical pattern and the height of the antenna above ground.

Multiband Design

When a V antenna is used over a range of frequencies — such as 14 to 28 Mc. — the characteristics will not be greatly altered if the legs are sufficiently long at the lowest frequency. The apex angle, at zero wave angle, for a 5-wavelength V (each leg approximately 350 feet long at 14 Mc.) is 44 degrees. At 21 Mc., where the legs are 7.5 wavelengths long, the optimum angle is 36 degrees, and at 28 Mc. where the leg length is 10 wavelengths it is 32 degrees. Such an antenna will operate well on all three frequencies if the apex angle is about 35 degrees. From Fig. 5-10, a 35-degree apex angle with a 5-wavelength V will align the lobes at a wave angle of something over 15 degrees, but this is not too high when it is kept in mind that the maximum radiation actually will be at a lower angle. At 28 Mc. the apex angle is a little large, but the chief effect will be a small reduction in gain and a slight broadening of the horizontal pattern, together with a tendency to reduce the wave angle at which the radiation is maximum. The same antenna can be used at 7 Mc. and 3.5 Mc., and on these bands the fact that the wave angle is raised is of less consequence, since high wave angles are useful. The gain will be small, however, because the legs are not long at these frequencies.

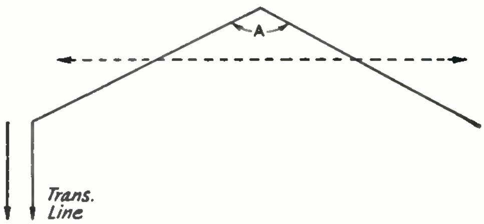


Fig. 5-11 — The obtuse-angle V antenna. Angle *A* is equal to 180 degrees minus the angle given in Fig. 5-9.

Other V Combinations

The gain can be increased about 3 db. by stacking two Vs one above the other, a half wavelength apart, and feeding them so that the legs on one side are in phase with each other and out of phase with the legs on the other side. This will result in a lowered angle of radiation. The bottom V should be at least a quarter wavelength above the ground and preferably a half wavelength.

Two V antennas can be broadsided to form a "W," giving an additional 3-db. gain. However, two transmission lines are required and this, plus the fact that five poles are needed to support the system, renders it normally impractical for the amateur. It is used by many commercial short-wave stations.

The V antenna can be made unidirectional by using another V placed an odd multiple of a quarter wavelength in back of the first and exciting the two with a phase difference of 90 degrees. The system will be unidirectional in the direction of the antenna with the lagging current. However, the V reflector is not normally employed by amateurs at low frequencies because it restricts the use to one band and requires a fairly elaborate supporting structure. Stacked Vs with driven reflectors could, however, be built for the 200- to 500-Mc. region without much difficulty. The over-all gain for such an antenna (two stacked Vs, each with a V reflector) is about 6 db. greater than the gains given by the curves.

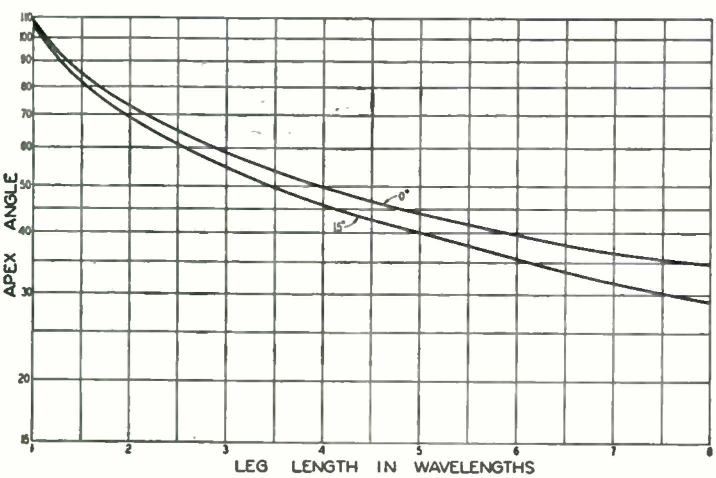


Fig. 5-10 — Apex angle of V antenna for alignment of main lobe at different wave angles, as a function of leg length in wavelengths.

Feeding the V

The V antenna is most conveniently fed by tuned feeders, since they permit multiband operation. Although the length of the wires in a V beam is not at all critical, it is important that both wires be of the same electrical length. Balanced feeder currents (in a tuned line) give sufficient indication of balanced lengths in the antenna proper.

If it is desired to use a nonresonant line, probably the most appropriate matching system is that using a stub or quarter-wave matching section. The adjustment is as described in Chapter Three.

● OBTUSE-ANGLE Vs

In the type of V just described the angle between the wires always is 90 degrees or less when the wires are more than 1 wavelength long. Another type of V can be formed by folding one wire as shown in Fig. 5-11 to align the lobes of maximum radiation. The angle *A* in this case is equal to 180 degrees less twice the angle given by Fig. 5-1 for a single wire having a length equal to that of one leg. For example, if each leg of the obtuse-angle V of Fig. 5-11 is 3 wavelengths (total length 6 wavelengths) the angle given by Fig. 5-1 is 29 degrees. The angle *A* in Fig. 5-11 will then be 180 − 58 = 122 degrees.

The obtuse-angle V is seldom used in the form shown in Fig. 5-11 because twice the over-all length is required and the gain is less than with an acute-angle V having the same leg length. This is because of the way the mutual impedance between legs compares in the two cases. However, the obtuse-angle V has the advantage that a change in frequency causes the major lobe from one leg to shift in one direction while the lobe from the other leg shifts the opposite way. This tends to make the optimum direction stay the same over a considerable frequency range. The pattern broadens and the gain is reduced when the antenna is operated at frequencies far removed from that for which it is designed.

● THE RESONANT RHOMBIC ANTENNA

The diamond-shaped or rhombic antenna shown in Fig. 5-12 can be looked upon as two acute-angle Vs placed end-to-end or as two obtuse-angle Vs placed side-by-side. This arrangement has two advantages over the simple V that have caused it to be favored by amateurs. For the same total wire length it gives somewhat greater gain than the V; a rhombic 4 wavelengths on a leg will have more gain than a V 8 wavelengths on a leg, for example. And, because of the compensating effect mentioned in connection with the obtuse-angle V, the directional pattern is less affected by frequency when the antenna is used over a wide frequency range. The disadvantage of the rhombic as compared with the V is that one additional support is required.

The same factors that govern the design of the V antenna apply in the case of the resonant rhombic. The angle *A* in the drawing is the same as that for a V (Fig. 5-9) having a leg length equal to *L*. If it is desired to align the lobes from individual wires with the wave angle, the curves of Fig. 5-10 may be used, again using the length of one leg in taking the data from the curves. The diamond-shaped antenna also can be operated as a nonresonant antenna, as described later in this chapter, and much of the discussion in that section applies to the resonant rhombic as well.

The direction of maximum radiation with a resonant rhombic is given by the arrows in Fig. 5-12; i.e., the antenna is bidirectional. There are minor lobes in other directions, their number and intensity depending on the leg length. When used at frequencies below the v.h.f. region the rhombic antenna is always mounted with the plane containing the wires horizontal. The polarization in this plane, and also in the perpendicular plane that bisects the rhombic, is horizontal. At 144 Mc. and above, the dimensions are such that the antenna can be mounted with the plane contain-

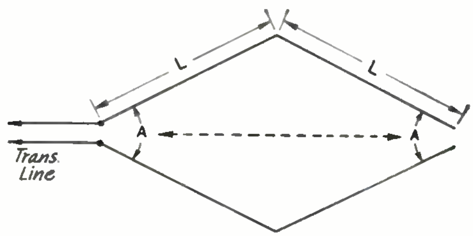


Fig. 5-12 — **The resonant rhombic or diamond-shaped antenna. All legs are the same length, and opposite angles of the diamond are equal.**

ing the wires vertical if vertical polarization is desired.

When the rhombic antenna is to be used on several amateur bands it is advisable to choose the apex angle, *A*, on the basis of the leg length in wavelengths at 14 Mc. This point is covered in more detail in connection both with the V and with the nonresonant rhombic. Although the gain on higher frequency bands will not be quite as favorable as if the antenna had been designed for the higher frequencies, the system will radiate well at the low angles that are necessary at such frequencies. At frequencies below the design frequency the greater apex angle of the rhombic (as compared with a V of the same total length) is more favorable to good radiation than in the case of the V.

The resonant rhombic antenna can be fed in the same way as the V. Resonant feeders are necessary if the antenna is to be used in several amateur bands.

Nonresonant Long-Wire Antennas

All the antenna systems previously considered have been based on resonant operation; that is, with standing waves of current and voltage along the wire. Although most antenna designs are based on using resonant wires, resonance is by no means a necessary condition for the wire to radiate and intercept electromagnetic waves.

In Fig. 5-13, let us suppose that the wire is parallel with the ground (horizontal) and is terminated by a load Z equal to its characteristic impedance, Z_0. The load Z can represent a receiver matched to the line. The resistor R is also equal to the Z_0 of the wire. A wave coming from the direction X will strike the wire first at its far end and sweep across the wire at an angle until it reaches the end at which Z is connected. In doing so it will induce voltages in the antenna and currents will flow as a result. The current flowing toward Z is the useful output of the antenna, while the current flowing toward R will be absorbed in R. The same thing is true of a wave coming from the direction X'.

The greatest possible power will be delivered to the load Z when the individual currents induced as the wave sweeps across the wire all combine properly on reaching the load. The currents will reach Z in optimum phase when the time required for a current to flow from the far end of the antenna to Z is exactly one-half cycle longer than the time taken by the wave to sweep over the antenna. Since a half cycle is equivalent to a half wavelength in space, the length of the antenna must be one-half wavelength greater than the distance traversed by the wave from the instant it strikes the far end of the antenna to the instant that it reaches the near end. This is shown by the small drawing, where AC represents the antenna, BC is a line perpendicular to the wave direction, and AB is the distance traveled by the wave in sweeping past AC. AB must be one-half wavelength shorter than AC. Similarly, AB' must be the same length as AB for a wave arriving from X'.

A wave arriving at the antenna from the opposite direction Y (or Y'), will similarly result in the largest possible current at the *far* end. However, the far end is terminated in R, which is equal to Z_0, so all the power delivered to R by a wave arriving from Y will be absorbed in R. The current traveling to Z will produce a signal in Z in proportion to its amplitude. If the antenna length is such that all the individual currents arrive at Z in such phase as to add up to zero, there will be no current through Z. At other lengths the resultant current may reach appreciable values. The lengths that give zero amplitude are those which are odd multiples of ¼ wavelength, beginning at ¾ wavelength. The response from the Y direction is greatest when the antenna is any even multiple of ½ wavelength long; the higher the multiple, the smaller the response.

Directional Characteristics

The explanation above considered the phase but not the relative amplitudes of the individual currents reaching the load. When the appropriate correction is made, the angle with the wire at which radiation or response is maximum is given by the curve of Fig. 5-14. The response drops off gradually on either side of the maximum point, resulting in lobes in the directive pattern much like those for harmonic antennas, except that the system is substantially unidirectional. Typical patterns are shown in Fig. 5-15. When the antenna length is 3/2 wavelength or greater there are also angles at which secondary maxima occur; these secondary maxima (minor lobes) have their peaks approximately at angles for which the length AB, Fig. 5-13, is less than AC by any odd multiple of one-half wavelength. When AB is shorter than AC by an *even* multiple of a half wavelength the induced currents cancel each other completely at Z, and in such cases there is a null for waves arriving in the direction perpendicular to BC.

The antenna of Fig. 5-13 responds to horizontally-polarized signals when mounted horizontally. If the wire lies in a plane that is vertical with respect to the earth it responds to vertically-polarized signals. By reciprocity, the characteristics for transmitting are the same as for receiving. For average conductor diameters and heights above ground, the Z_0 of the antenna is of the order of 500 to 600 ohms.

It is apparent that an antenna operating in this way has much the same characteristics as a transmission line. When it is properly terminated at both ends there are traveling waves, but no standing waves, on the wire. Consequently the current is substantially the same all along the wire. Actually, it decreases in the direction in

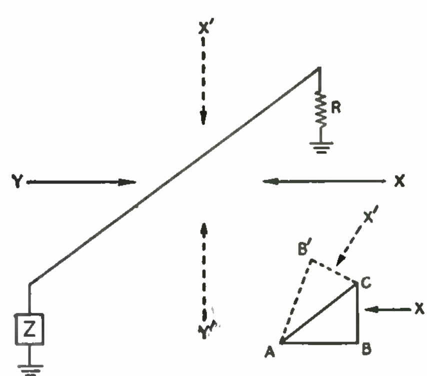


Fig. 5-13 — Nonresonant long-wire antenna.

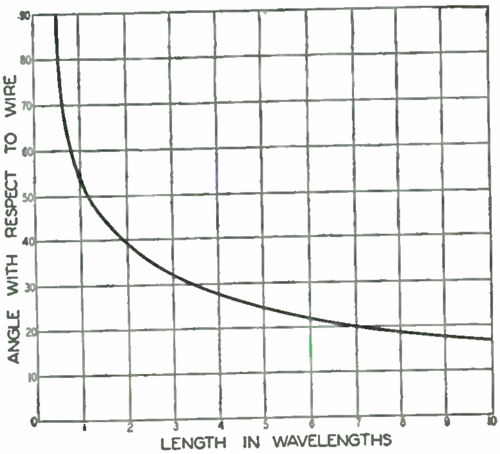


Fig. 5-14 — **Angle with respect to wire axis at which the radiation from a nonresonant long-wire antenna is maximum.**

which the current is flowing because of energy loss by radiation as well as by ohmic loss in the wire and the ground. The antenna can be looked upon as a transmission line terminated in its characteristic impedance, but having such wide spacing between conductors (the second conductor in this case is the image of the antenna in the ground) that radiation losses are by no means inconsequential.

A wire terminated in its characteristic impedance will work on any frequency, but its directional characteristics change with frequency as shown by Fig. 5-14. To give any appreciable gain over a dipole the wire must be at least a few wavelengths long. The angle at which maximum response is secured can be in any plane that contains the wire axis, so in free space the major lobe will be a cone. In the presence of ground, the discussion given in connection with Fig. 5-3 applies, with the modification that the angles of best radiation or response are those given in Fig. 5-14 rather than by Fig. 5-1. As comparison of the two curves will show, the difference in optimum angle between resonant and nonresonant wires is quite small.

THE INVERTED V ANTENNA

Two nonresonant long wires may be combined in much the same fashion as the resonant wires discussed earlier in this chapter. One example is the "inverted V" or "half rhombic" antenna shown in Fig. 5-16. For example, suppose that each of the sides, or legs, of the V has a length such that the major lobe of *each* wire makes an angle of 36 degrees with the wire. Then the "tilt angle," ϕ (half the total angle included between the wires), may be adjusted so that the major lobes of radiation from the two wires will add in the desired direction. The tilt angle then becomes the complement of the angle which the main lobe makes with each wire. This is shown in Fig. 5-16, for horizontal transmission to the right in the figure. In directions other than to the right in the plane containing the antenna the radiation from the individual wires will tend to cancel more or less completely.

For signals coming from the optimum direction the polarization is vertical when the plane containing the wires is vertical.

The advantage of combining wires as shown in Fig. 5-16 is the same as previously mentioned for the obtuse-angle V. That is, as the frequency is changed the optimum direction for one leg shifts in the opposite direction to that for the other leg. This tends to maintain the same direction of maximum response (or transmission) over a wide frequency range, although it is accompanied by a broadening of the pattern.

The form of tilted wire antenna shown in Fig. 5-16 is simple in construction, since only one pole is required. However, for all but quite high frequencies — 28 Mc. and higher — the pole height needed to provide the proper tilt angle for a leg length great enough to provide appreciable gain is beyond the facilities available to most amateurs.

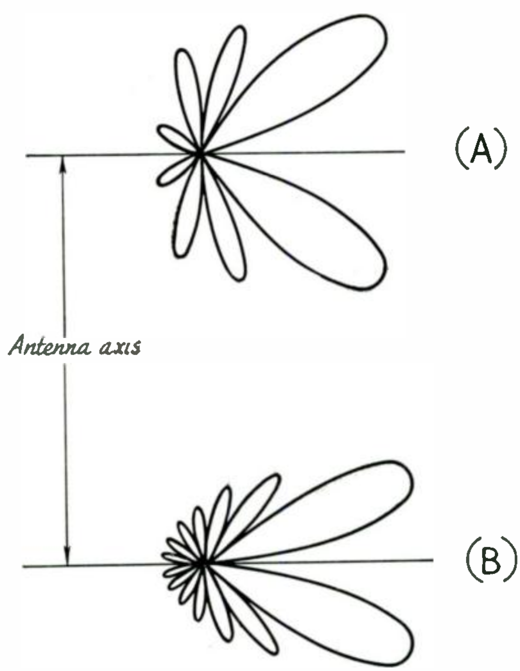


Fig. 5-15 — **Typical radiation patterns (cross section of solid figure) for terminated long wires. (A) length two wavelengths; (B) four wavelengths; both for an idealized case in which there is no decrease of current along the wire. In practice, the pattern is somewhat distorted by wire attenuation.**

Also, it is difficult to secure a satisfactory termination because of variation in ground resistance with weather conditions. This might be overcome by the use of a large ground screen under the antenna and extending a half wavelength or so beyond the wire in all directions, but the installation of such a screen probably would be impracticable in most locations. A form of transmission-line termination also may be used, with the far end of the antenna connected through the terminating resistor to the center of a half-wave wire running parallel to the ground and perpendicular to the line of the antenna. The impedance at the center of such a wire will be low, and currents induced by the incoming waves from the desired direction will balance out. However, such a termination is good only for the frequency for which the wire is cut, so that a wide frequency range is not possible. These difficulties are overcome by the use of the diamond-shaped, or rhombic, antenna.

THE NONRESONANT RHOMBIC ANTENNA

The nonresonant rhombic antenna, shown schematically in Fig. 5-17, consists of two tilted-wire antennas of the type shown in Fig. 5-16 placed side by side. The terminating resistor is connected between the far ends of the two sides, and is made approximately equal to the characteristic impedance of the antenna as a unit. The rhombic may be constructed either horizontally or vertically, but practically always is horizontal at frequencies below 54 Mc., since the pole height required is considerably less. Also, horizontal polarization is equally, if not more, satisfactory at these frequencies.

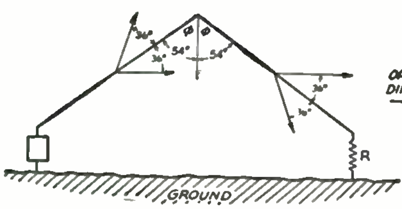


Fig. 5-16 — **The tilted-wire antenna, showing the directions of the main lobes for each wire for legs two wavelengths long, when the tilt angle ϕ is adjusted for alignment of the lobes.**

The basic principle of combining lobes of maximum radiation from the four individual wires constituting the rhombus or diamond is the same in either the nonresonant type shown in Fig. 5-17 or the resonant type described earlier in this chapter. The included angles should differ slightly because of the differences between resonant and nonresonant wires, but comparison of Figs. 5-1 and 5-14 show that the differences are almost negligible.

Tilt Angle

It is a matter of custom, in dealing with the nonresonant or "terminated" rhombic, to talk about the "tilt angle" (ϕ in Fig. 5-17) rather than the angle of maximum radiation with respect to an individual wire. The tilt angle is simply 90 degrees minus the angle of maximum radiation. In the case of a rhombic antenna designed for zero wave angle the tilt angle is 90 degrees minus the values given in Fig. 5-14.

Fig. 5-18 shows the tilt angle as a function of the antenna leg length. The curve marked "0°" is for a wave angle of zero degrees; that is, maximum radiation in the plane of the antenna. The other curves show the proper tilt angles to use when aligning the major lobe with a desired wave angle. For a wave angle of 5 degrees the difference in tilt angle is less than one degree for the range of lengths shown. Just as in the case of the resonant V and resonant rhombic, alignment of the wave angle and lobes always results in still higher radiation at a lower wave angle, and for the same reason, but also results in the highest possible radiation at the desired wave angle.

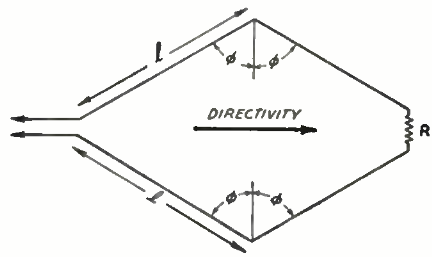


Fig. 5-17 — **The nonresonant rhombic antenna.**

The broken curve marked "Optimum Length" shows the leg length at which maximum gain is secured at a chosen wave angle. Increasing the leg length beyond the optimum will result in lessened gain, and for that reason the curves do not extend beyond the optimum length. Note that the optimum length becomes greater as the desired wave angle is smaller. Leg lengths over 6 wavelengths are not recommended because the directive pattern becomes so sharp that the antenna performance is highly variable with small changes in the angle, both horizontal and vertical, at which an incoming wave reaches the antenna. Since these angles vary to some extent in ionospheric propagation, it does not pay to attempt to use too great a degree of directivity.

Multiband Design

When a rhombic antenna is to be used over a considerable frequency range it is worth paying some attention to the effect of the tilt angle on the gain and directive pattern at various frequencies. For example, suppose the antenna is to

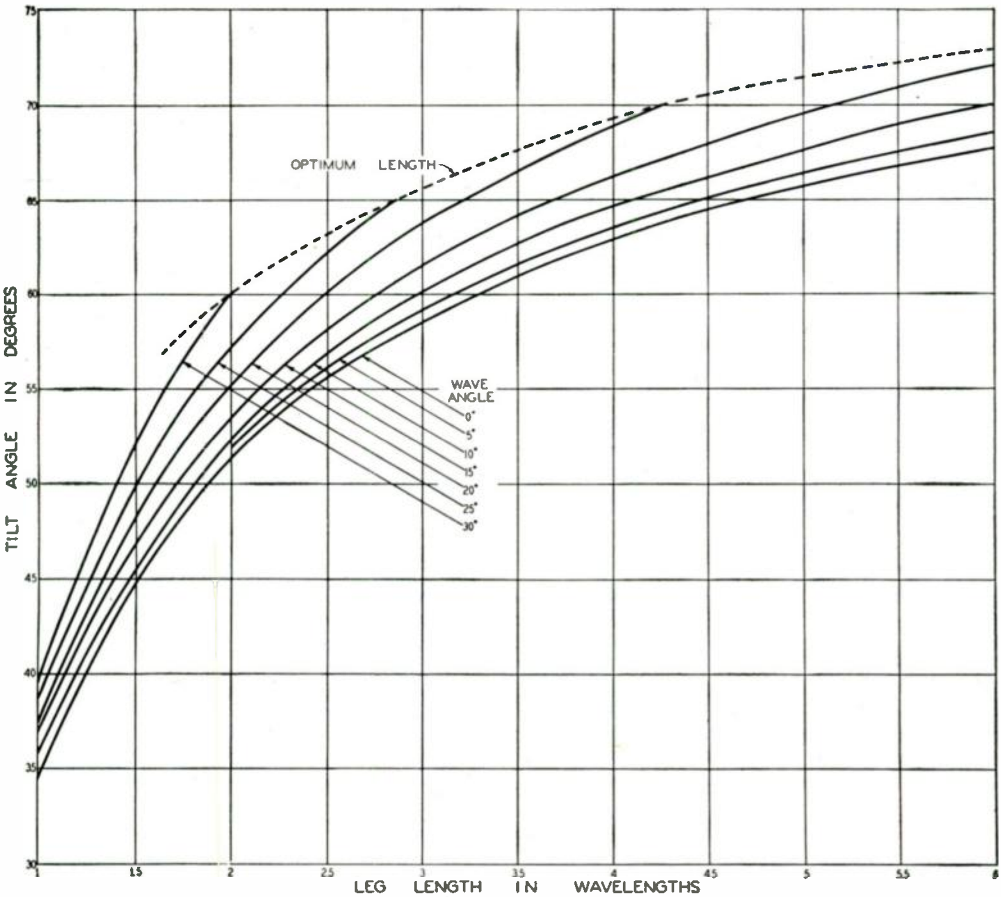


Fig. 5-18 — Rhombic-antenna design chart. For any given leg length, the curves show the proper tilt angle to give maximum radiation at the selected wave angle. The broken curve marked "Optimum Length" shows the leg length that gives the maximum possible output at the selected wave angle. The optimum length as given by the curves should be multiplied by 0.74 to obtain the leg length for which the wave angle and main lobe are aligned.

be used at frequencies up to and including the 28-Mc. band, and that the leg length is to be 6 wavelengths on the latter frequency. For zero wave angle the optimum tilt angle is 68 degrees, and the calculated free-space directive pattern in the vertical plane bisecting the antenna is shown in the right-hand drawing of Fig. 5-19. At 14 Mc. this same antenna has a leg length of three wavelengths, which calls for a tilt angle of 58.5 degrees for maximum radiation at zero wave angle. The calculated patterns for tilt angles of 58.5 and 68 degrees are shown in the left-hand drawing in Fig. 5-19, and it is seen that if the optimum tilt for 28-Mc. operation is used the gain will be reduced and the wave angle raised at 14 Mc. However, it will do no harm to use the tilt angle that gives maximum output at 15 degrees on 14 Mc., and as given by Fig. 5-18 this tilt angle is 61.5 degrees. The pattern with this tilt angle is shown in Fig. 5-19 for both the 14- and 28-Mc. cases. The effect at 28 Mc. is to decrease the gain by more than 6 db. and to split the radiation in the vertical plane into two lobes, one of which is at a wave angle too high to be useful at this frequency.

Now from Fig. 5-20 the free-space gain of a rhombic 6 wavelengths on a leg is about 13 db. and the gain of one 3 wavelengths on a leg is 10 db. If the tilt angle is made optimum for 28 Mc. the gain will be reduced roughly 3 db. on 14 Mc., consequently the gain is about 7 db. On the other hand, if the 15-degree tilt angle is chosen for 14-Mc. operation the gain at 28 Mc. will be about 9 db. The angle should be chosen on the basis of the frequency at which the best possible performance is desired. In most cases 14 Mc. should be the favored frequency, since the gain is inherently less the lower the frequency because the legs are shorter in terms of wavelength.

The patterns of Fig. 5-19 are in the vertical plane through the center of the antenna only. In vertical planes making an angle with the antenna axis the patterns may differ considerably. The effect of a tilt angle that is smaller than the optimum is to broaden the horizontal pattern, so at 28 Mc. the antenna in the example would be less directive in the horizontal plane than would be the case if it were designed for optimum performance at that frequency. It must also be pointed out that the patterns given in Fig. 5-19 are free-space patterns and must be multiplied by the ground-reflection factors for the actual antenna height used, if the actual vertical pattern is to be determined.

Power Gain

The theoretical power gain of a nonresonant rhombic antenna over a dipole, both in free space, is given by the curve of Fig. 5-20. This curve is for zero wave angle and includes an allowance of 3 db. for power dissipated in the terminating resistor. The actual gain of an antenna mounted horizontally above the ground, as compared with a dipole at the same height, can be expected to vary a bit either way from the figures given by the curve. The power lost in the terminating resistor is probably less than 3 db. in the average case, since more than half of the power input is radiated before the end of the antenna is reached. However, there is also more power loss in the wire and in the ground under the antenna than in the case of a simple dipole, so the 3 db. loss figure is probably a representative estimate.

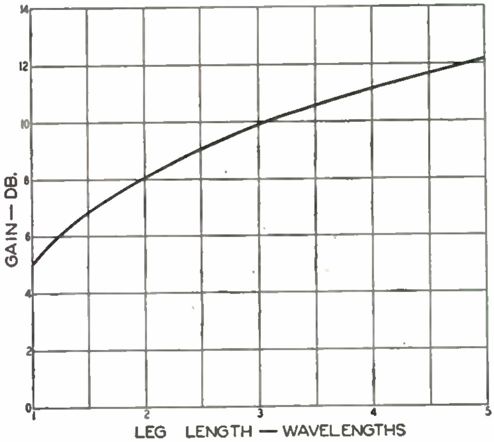


Fig. 5-20 — **Theoretical gain of a nonresonant rhombic antenna over a half-wave dipole in free space. This curve includes an allowance of 3 db. for loss in the terminating resistor.**

Termination

Although there is no marked difference in the gain obtainable with resonant and nonresonant rhombics of comparable design, the nonresonant antenna has the advantage that over a wide frequency range it presents an essentially resistive and constant load to the transmitter coupling apparatus. In addition, nonresonant operation makes the antenna substantially unidirectional, while the "unterminated" or resonant rhombic is always bidirectional, although not symmetrically so. In a sense, it can be considered that the power dissipated in the terminating resistor is simply power that would have been radiated in the other direction had the resistor not been there, so the fact that some of the power (about one-third) is used up in heating the resistor does not mean an actual loss in the desired direction.

The characteristic impedance of an ordinary rhombic antenna, looking into the input end, is of the order of 700 to 800 ohms when properly terminated in a resistance at the far end. The terminating resistance required to bring about the matching condition usually is slightly higher than the input impedance because of the loss of energy through radiation by the time the far end is reached. The correct value usually will be found to be of the order of 800 ohms, and should be determined experimentally if the flattest possible antenna is desired. However, for average work a noninductive resistance of 800 ohms can be used with the assurance that the operation will not be far from optimum.

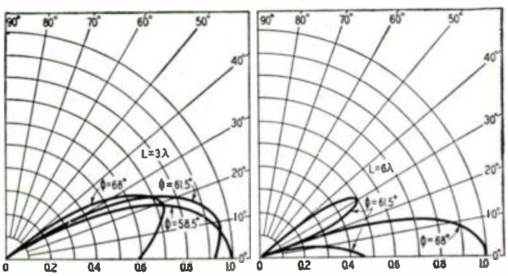


Fig. 5-19 — **Showing the effect of tilt angle on the free-space vertical pattern of a nonresonant rhombic antenna having a leg length of 3 wavelengths at one frequency and 6 wavelengths at twice the frequency. These patterns apply only in the direction of the antenna axis.**

The terminating resistor must be practically a pure resistance at the operating frequencies; that is, its inductance and capacitance should be negligible. Ordinary wire-wound resistors are not suitable because they have far too much inductance and distributed capacitance. Small carbon resistors have satisfactory electrical characteristics but will not dissipate more than a few watts and so cannot be used, except when the transmitter power does not exceed 10 or 20 watts or when the antenna is to be used for reception only. The special resistors designed either for use as "dummy" antennas or for terminating rhombic antennas should be used in other cases. To allow a factor of safety, the total rated power dissipation of the resistor or resistors should be equal to half the power output of the transmitter.

To lower capacity effects it is desirable to use several units, say three, in series even when one alone will safely dissipate the power. The two outer units should be identical and each should have one-fourth to one-third the total resistance, with the center unit making up the difference. The units should be installed in a weatherproof housing at the end of the antenna to protect them and to permit mounting without mechanical strain. The connecting leads should be short so that little extraneous inductance is introduced.

Alternatively, the terminating resistance may be placed at the end of an 800-ohm line connected to the end of the antenna. This will permit placing the resistors and their housing at a point convenient for adjustment rather than at the top of the pole. The line length is not critical, since it operates without standing waves and hence is nonresonant.

Multiwire Rhombics

The input impedance of a rhombic antenna constructed as in Fig. 5-17 is not quite constant as the frequency is varied. This is because the varying separation between the wires causes the characteristic impedance of the antenna to vary along its length. The variation in Z_0 can be minimized by a conductor arrangement that increases the capacitance per unit length in proportion to the separation between the wires.

The method of accomplishing this is shown in Fig. 5-21. Three conductors are used, joined together at the ends but with increasing separation

Fig. 5-21 — **Three-wire rhombic antenna. Use of multiple wires improves the impedance characteristic of a nonresonant rhombic.**

as the junction between legs is approached. As used in commercial installations having legs several wavelengths long, the spacing between wires at the center is 3 to 4 feet. Since all three wires should have the same length, the top and bottom wires will be slightly farther from the support than the middle wire. Using three wires in this way reduces the Z_0 of the antenna to approximately 600 ohms, thus providing a better match for a practicable open-wire line, in addition to smoothing out the impedance variations over the frequency range.

A similar effect, although not quite so good, is obtained by using two wires instead of three. It has been found that, with the 3-wire system, the gain of the antenna is slightly greater (of the order of 1 db.) than when only a single conductor is used.

Front-to-Back Ratio

It is theoretically possible to obtain an infinite front-to-back ratio with a terminated rhombic antenna, and in practice very large values can actually be secured. However, when the antenna is terminated in its characteristic impedance the infinite front-to-back ratio can be secured only at frequencies for which the leg length is an odd multiple of a quarter wavelength, as described in the section on nonresonant long wires. The front-to-back ratio is smallest at frequencies for which the leg length is a multiple of a half wavelength.

When the leg length is not an odd multiple of a quarter wave at the frequency under consideration, the front-to-back ratio can be made very high by slightly decreasing the value of terminating resistance. This permits a small reflection from the far end of the antenna which cancels out, at the input end, the residual response. With large antennas the front-to-back ratio may be made very large over the whole frequency range by experimental adjustment of the terminating resistance. Modification of the terminating resistance also permits "steering" the back null over a small horizontal range so that signals coming from a particular spot not exactly to the rear of the antenna may be minimized.

Ground Effects

Reflections from the ground play exactly the same part in determining the vertical directive pattern of a horizontal rhombic antenna that they play with other horizontal antennas. Consequently, if a low wave angle is desired it is necessary to make the height great enough to bring the reflection factor into the higher range of values given by the charts in Chapter Two. At 14 Mc. heights of ¾ to 1 wavelength are desirable. The antenna will work at heights as low as ½ wavelength, of course, but with reduced radiation at the lower angles, particularly below 10 degrees. In this respect it is under no greater or lesser handicap than any other antenna system that is close to the ground.

Alignment of Lobes, Wave Angle, and Ground Reflections

When maximum antenna response is desired at a particular wave angle (or maximum radiation is desired at that angle) the major lobe of the antenna cannot only be aligned with the wave angle as previously described but also with a maximum in the ground-reflection factor. When this is done it is no longer possible to consider the antenna height independently of other aspects of rhombic design. The wave angle, leg length, and height become mutually dependent.

This method of design is of particular value when the antenna is built to be used over fixed transmission distances for which the optimum wave angle is known. It has had wide application in commercial work with nonresonant rhombic antennas, but seems less desirable for amateur use where, for the long-distance work for which rhombic antennas are built, the lowest wave angle that can be achieved is the most desirable. Alignment of all three factors is limited in application because it leads to impracticable heights and leg lengths for small wave angles. Consequently, when a fairly broad range of low wave angles is the objective it is more satisfactory to design for a low wave angle and simply make the antenna as high as possible.

Fig. 5-22 shows the lowest height at which ground reflections make the radiation maximum at a desired wave angle. It can be used in conjunction with Fig. 5-18 for complete alignment of the antenna. For example, if the desired wave angle is 20 degrees, Fig. 5-22 shows that the height must be 0.75 wavelength. From Fig. 5-18, the optimum leg length is 4.2 wavelengths and the tilt angle is just under 70 degrees. A rhombic antenna so designed will have the maximum possible output that can be obtained at a wave angle of 20 degrees; no other set of dimensions will be as good. However, it will have still greater output at some angle lower than 20 degrees, for the reasons given earlier. When it is desired to make the maximum output of the antenna occur at the 20-degree wave angle, it may be accomplished by using the same height and tilt angle, but with the leg length reduced by 26 per cent. Thus for such alignment the leg length should be 4.2 × 0.74 = 3.1 wavelengths. However, the output at the 20-degree wave angle will be smaller than with 4.2-wavelength legs, despite the fact that the smaller antenna has its maximum radiation at 20 degrees. The reduction in gain is about 1.5 db.

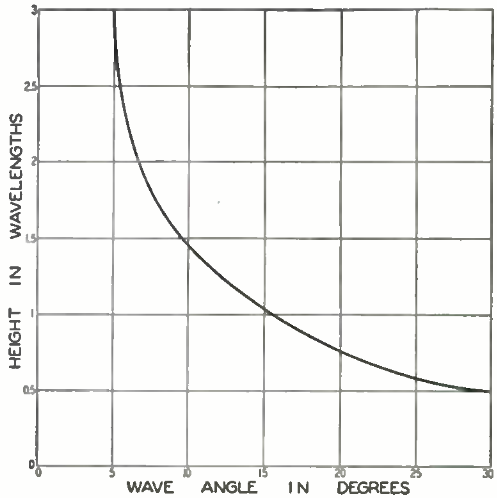


Fig. 5-22 — Antenna height to be used for securing maximum radiation at a desired wave angle. This curve applies to any type of horizontal antenna.

Methods of Feed

If the broad frequency characteristic of the rhombic antenna is to be fully utilized the feeder system used with it must be similarly broad. This practically dictates the use of a transmission line of the same characteristic impedance as that shown at the antenna input terminals, or approximately 700 to 800 ohms. Data for the construction of such lines will be found in Chapter Three. It will be found, however, that the spacing required is rather awkward, and also that rather small wire must be used. Both these considerations are disadvantageous mechanically, and the radiation from the line also tends to be high at 28 Mc. because of the wide spacing.

While the usual matching stub can be used to provide an impedance transformation to more satisfactory line impedances, this limits the operation of the antenna to a comparatively narrow range of frequencies centering about that for which the stub is adjusted. On the whole, the best plan is to connect a 600-ohm line directly to the antenna and accept the small mismatch which results. The operation of the antenna will not be adversely affected, and since the standing-wave ratio is quite low (1.33 to 1) the additional loss over the perfectly-matched condition will be unappreciable even for rather long lines. The chief disadvantage is that at some frequencies a slight readjustment of the coupling to the transmitter may be necessary to maintain constant input.

Multiband Antennas

For operation between 3.5 and 30 Mc., the majority of amateurs find it impractical to use a separate antenna for each band. This is no real hardship, because the only real advantages to having separate antennas for each band are that they allow "flat" feedlines to be used and also advantage can be taken of certain directional

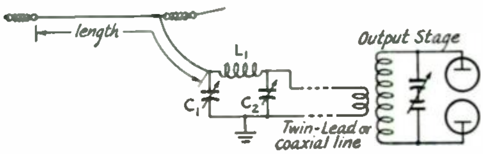


Fig. 6-1 — Any length of wire can be used for an antenna by feeding it with a π-section coupler. The values of C_1, C_2 and L_1 must be found by experiment but, in general, 150-μμfd. variables can be used for C_1 and C_2. The ground connection can be made back to the transmitter chassis or directly to a water pipe.

characteristics the antennas may have. This chapter will describe single antenna systems that can be used for all amateur bands between 3.5 and 30 Mc., while Chapters Seven and Eight will deal with more specialized antennas that are normally used for only one or two bands.

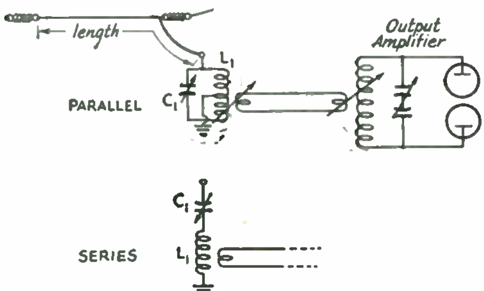


Fig. 6-2 — If the length of the wire is made 135 feet, the antenna can be fed with a parallel-tuned circuit on all amateur bands. L_1C_1 should resonate to the band in use, and can be a duplicate of the final-amplifier tank inductance and capacity. A ground may or may not be necessary.

If the wire is 67 feet long, a series-tuned circuit is required for 3.5-Mc. operation. L_1C_1 should resonate in the 80-meter band, and a good ground (or quarter wavelength of wire) should be used. Parallel tuning can be used on 7 Mc. and higher.

ANTENNAS WITHOUT FEEDLINES

The simplest (and probably least effective) multiband antenna is a random length of No. 12 or No. 14 wire. Power can be fed to the wire through a "π-section" coupler on practically any frequency, by connecting it as shown in Fig. 6-1. If the wire is made either 67 or 135 feet long, the wire can be fed through the coupler, or with a tuned circuit as in Fig. 6-2.

If a 28- or 50-Mc. rotary beam has been in-

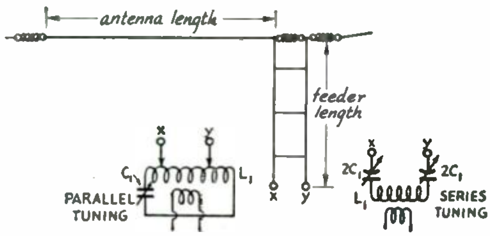


Fig. 6-3 — The end-fed antenna using tuned feeders will require a series- or parallel-tuned coupling circuit, depending upon the feeder length and the band (see Table 6-I). L_1C_1 should tune to the operating frequency. For the parallel-tuned circuit, the capacity C_1 should be at least 100 μμfd. for 3.5 Mc., 50 for 7 Mc. and 25 for 14 Mc. In the series-tuned circuit, the values of $2C_1$ will be considerably lower (roughly one-half to one-fourth), and consequently L_1 will be larger. A single series-tuning condenser can be used, of a value equal to half of $2C_1$.

stalled, in many cases it will be possible to use the beam's feedline as an antenna on the lower frequencies. Connecting the two wires of the feedline together at the station end will give a random-length wire that can be conveniently coupled to the transmitter as in Figs. 6-1 and 6-2. The rotary system at the far end will serve only to "end-load" the wire and will not have much other effect.

One disadvantage of all such directly-fed systems is that part of the antenna is practically within the station, and there is a good chance that trouble with r.f. feed-back will be encountered. The r.f. within the station can often be

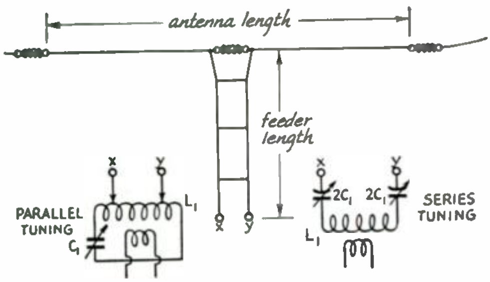


Fig. 6-4 — The center-fed all-band antenna requires parallel or series tuning of the coupling circuit, depending upon the system dimensions and the frequency band (see Table 6-II). L_1C_1 should resonate in the range of the operating frequency. For greatest ease of coupling, a low ratio of L_1 to C_1 should be used with parallel tuning, and a high ratio with series tuning.

satisfactory to return to the transmitter chassis for the ground, or directly to a convenient water pipe. If neither of these works well (or is not available), a length of No. 12 or No. 14 wire (approximately ¼ wavelength long) can often be used to good advantage. This wire is connected at the point in the circuit that is shown grounded, and it can be run out and down the side of the house, or supported a few feet above the ground if the station is on the first floor or in the basement. It should not be connected to actual ground at any point.

END-FED ANTENNAS

When a straight-wire antenna is fed at one end by a two-wire feedline, the length of the antenna portion becomes fairly critical if radiation from the feedline is to be held to a minimum. However, there is often no great harm in some radiation from the feedline, and one of the most popular antenna systems for multiband operation is the "end-fed" or "Zepp-fed" antenna shown in Fig. 6-3. (The "Zepp-fed" designation derives from an early use of the antenna on dirigible balloons.) The antenna length (length of "flat top") is made a half wavelength at the lowest operating frequency. The feeder length can be anything that is convenient, but feeder lengths that are multiples of a quarter wavelength generally give trouble with parallel currents and radiation from the feeder portion of the system. The feeder length given in Table 6-I is free of this trouble. The feedline is generally an open-wire line of No. 14 solid spaced 4 or 6 inches with ceramic or plastic spacers, but 300-ohm Twin-Lead can be used in low-power installations up to a few hundred watts without trouble.

If one has room for only a 67-foot flat top and

minimized by choosing a length of wire so that a *current loop* occurs at or near the transmitter. This means using a wire length of a quarter wavelength, or an odd multiple of a quarter wavelength. Obviously this can be done for only one band, in the case of even harmonically-related bands, since the wire length that gives a current loop at the transmitter will give a voltage loop at two (or four) times that frequency.

When operating with a straight-wire antenna, as in Figs. 6-1 and 6-2, it is wise to try different types of grounds on the various bands, to see which will give the best results. In many cases it will be

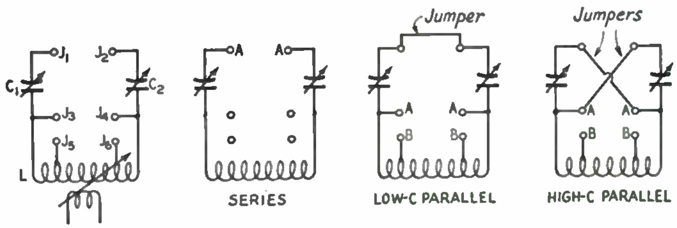


Fig. 6-5 — Wiring diagram of an antenna coupler for tuned lines. For 3.5- to 30-Mc. operation, C_1 and C_2 are 75-μμfd. condensers with plate spacing equal to that used in the final-amplifier tank condenser. J_1 through J_6 are banana-plug jacks mounted on the condensers or on stand-off insulators. The coil L is a transmitting-type coil similar in power rating to that used in the output amplifier. By using jumpers between the jacks, the three types of circuits shown can be obtained. The feedline connects to *A-A* or *B-B*.

Some experimenting with the number of turns in L may be required. In general, higher inductance will be used in the series circuit than in the parallel circuits, and sometimes the final-amplifier coil from the next lower-frequency band can be used. When the coil is tapped in parallel operation, the approximate position of the taps can be found from Table 6-II, if one of those particular antenna-and-feedline systems is being used.

To maintain balance to ground, the settings of C_1 and C_2 should be held the same. Insulated extension shafts on the condensers should be used to eliminate hand-capacity effects while tuning.

TABLE 6-I
End-Fed Antennas Using Tuned Lines

The antenna lengths are approximate and can be trimmed to the favorite operating frequency by the method given in the text.

Antenna Length (feet)	*Feeder Length (feet)*	*Band (Mc.)*	*Type of Coupling Circuit*	
			Open-Wire Line	*300-Ohm Twin-Lead*
135	45	3.5	Series	Series
		7	Series	Parallel
		14	Series	Parallel
		21	Series	Series
		28	Parallel	Series
67	45	7	Series	Parallel
		14	Series	Parallel
		21	Series	Series
		28	Parallel	Series

yet wants to operate in the 3.5-Mc. band, the feedline can be tied together at the transmitter end and the entire system treated as a random-length wire fed directly, as in Fig. 6-1.

While it is generally sufficient to cut the antenna to length by tape measurement, a closer check can be obtained by inserting an r.f. ammeter in the feedline at x and y in Fig. 6-3. If two similar meters are not available, one meter can be moved from one feeder to the other, closing the circuit in the unmetered feeder with a short jumper wire. When the antenna (flat-top) length is correct, the meter readings in either feeder will be the same, provided there are no parallel currents on the feedline. The best insurance against parallel currents is to use a feeder length that is not a multiple of a quarter wavelength, e.g., the 45-foot length recommended in Table 6-I. The flat top can be made long at first and trimmed at the far end until a balanced condition is obtained at the favorite operating frequency. It will be found that the currents will balance over only a narrow range of frequencies, because of the changes in current distribution with frequency. Also, the balance points will not be exactly harmonically related but will move to frequencies slightly higher than exact multiples when checked at higher-frequency bands.

TABLE 6-II
Center-Fed Antennas Using Tuned Lines

Antenna Length (feet)	*Feeder Length (feet)*	*Band (Mc.)*	*Type of Coupling Circuit* — *Open-Wire Line*	*Type of Coupling Circuit* — *300-Ohm Twin-Lead*
135	42	3.5	Parallel (large coil)	Parallel (tap across ½ coil)
		7	Parallel (tap across ⅓ coil)	Parallel (tap across ½ coil)
		14	Parallel (tap across ⅔ coil)	Series
		21	Parallel (tap across ⅔ coil)	Series
		28	Series (very low *C*)	Series at ends of band, parallel (tap across ½ coil) at center
135	77½	3.5	Parallel	Parallel (tap across ½ coil)
		7	Parallel (tap across ½ coil)	Parallel (tap across ¼ coil)
		14	Parallel (tap across ½ coil)	Parallel (tap across ½ coil)
		21	Parallel (tap across ¼ coil)	Parallel (tap across ½ coil)
		28	Parallel (tap across ¼ coil)	Parallel (tap across ¼ coil)
67	42½	3.5	Series	Series
		7	Parallel	Parallel (tap across ⅓ coil)
		14	Parallel (tap across ½ coil)	Series
		21	Parallel (tap across ¼ coil)	Parallel (tap across ½ coil)
		28	Parallel (tap across ¼ coil)	Parallel (tap across ½ coil)
67	65½	3.5	Parallel (coil larger than normal)	Parallel (coil larger than normal)
		7	Series	Series
		14	Parallel (tap across ⅔ coil)	Parallel (tap across ⅓ coil)
		21	Series	Parallel (tap across ½ coil)
		28	Parallel (tap across ½ coil)	Series

CENTER-FED ANTENNAS

Probably the most satisfactory all-band antennas are those using tuned feedlines to the center of the antenna, as in Fig. 6-4. Because each half of the flat top is the same length, the feedline currents will be balanced at all frequencies unless, of course, unbalance is introduced by one half of the antenna being closer to ground (or a grounded object) than the other. For best results and to maintain feed-current balance, the feedline should run away at right angles to the antenna. In cases where this is completely impractical, the angle should be made as close to 90° as possible. The length of the flat top is not critical (provided both halves are the same length), nor is the length of the feedline, but some combinations will give less trouble with parallel line currents than will others. Recommended combinations are given in Table 6-II.

The feedline can be the same as that used for end-fed antennas. The 300-ohm Twin-Lead can be used for powers up to a few hundred watts, and open-wire line is good for any amateur power. The coupling circuits in Table 6-II will, in general, be duplicates of the *LC* combination used in the final amplifier, for the parallel-tuned cases. The series-tuned coupling circuits generally require a larger inductance and a smaller capacity. By using two condensers in the series-tuned circuit, a lower minimum capacity can be obtained (useful at 21 and 28 Mc.) and the circuit balance is retained. However, it is often quite satisfactory to use only one series tuning condenser, provided it can tune the range and withstand the developed voltage.

When the flat-top length is made less than a half wavelength at the lowest operating frequency, there will be a slight reduction in effectiveness at this frequency, but it is not too important until the length is made a quarter wavelength. In other words, there is no need to relinquish 3.5-Mc. operation simply because one doesn't have enough room to string up an antenna 135 feet long.

● Vs AND RHOMBICS

Although "V" beams and rhombic antennas are generally designed for 14 or 28 Mc., they are made up of so much wire that they can be used down to 3.5 Mc. by proper coupling to the transmitter. The directional effects will not be very noticeable on 3.5 Mc., but it may be found that the antenna is not as effective over short distances (100 miles) as a simpler type of antenna. This is because the antenna system may tend to confine the radiation to lower angles. However, for greater distances the long-wire Vs and rhombics will prove quite satisfactory at 3.5 Mc.

● VERTICAL ANTENNAS

A vertical antenna is generally a poor all-band antenna because there is an increase in high-angle radiation at the higher frequencies, and this is the exact opposite of the desired characteristic. It will work, of course, and under some conditions it will result in good signals at a distance, but its performance will be inferior at 14, 21 and 28 Mc. to almost any other type of antenna. However, a 66-foot vertical antenna fed in the center with a tuned line is a good antenna for 3.5, 7 and 14 Mc., and a 33-foot vertical fed similarly in the center is reasonably satisfactory for 7-, 14-, 21- and 28-Mc. work.

Antennas for 3.5 and 7 Mc.

Obviously any of the antenna systems discussed under "Multiband Antennas" (Chapter Six) will work on 3.5 and 7 Mc. This chapter will discuss these and other antennas that are best suited for low-frequency work.

TUNED-LINE END- OR CENTER-FED ANTENNAS

The end-fed and center-fed antennas shown in Figs. 6-3 and 6-4 are quite widely used for 3.5- and 7-Mc. operation. The center-fed system has a slight edge, because it is inherently balanced on both bands and there is less chance for feeder radiation and r.f. feed-back troubles, but either system will give a good account of itself. On these frequencies the height of the antenna is not too important, and anything over 35 feet will work well for average operation. The advantage of height will show up when trying to work DX.

HALF-WAVELENGTH ANTENNAS

An untuned or "flat" feedline is a logical choice on any band, because the losses are low, but it generally limits the use of the antenna to one band. However, the half-wave antenna fed with untuned line is one of the most popular systems on the 3.5- and 7-Mc. bands. If the antenna is a single-wire affair, its impedance is in the vicinity of 70 ohms. The most logical way to feed the antenna is with 72-ohm Twin-Lead or coaxial line. The heavy-duty Twin-Lead and the coaxial line present support problems, but these can be overcome by using a small auxiliary pole to take the weight of the line. The line should come away from the antenna at right angles, and it can be of any length.

A "folded dipole" shows an impedance of 300 ohms, and so it can be fed directly with any length of 300-ohm Twin-Lead. The feedline should come away from the antenna at as close to a right angle as possible. The folded dipole can be made of ordinary wire spaced by lightweight wooden or plastic spacers, 4 or 6 inches long, or a piece of 300-ohm Twin-Lead can be used for the folded dipole. However, if Twin-Lead is used for the dipole, the ends should be shorted through condensers of 500 $\mu\mu$fd. (3.5 Mc.) or 250 $\mu\mu$fd. (7 Mc.)

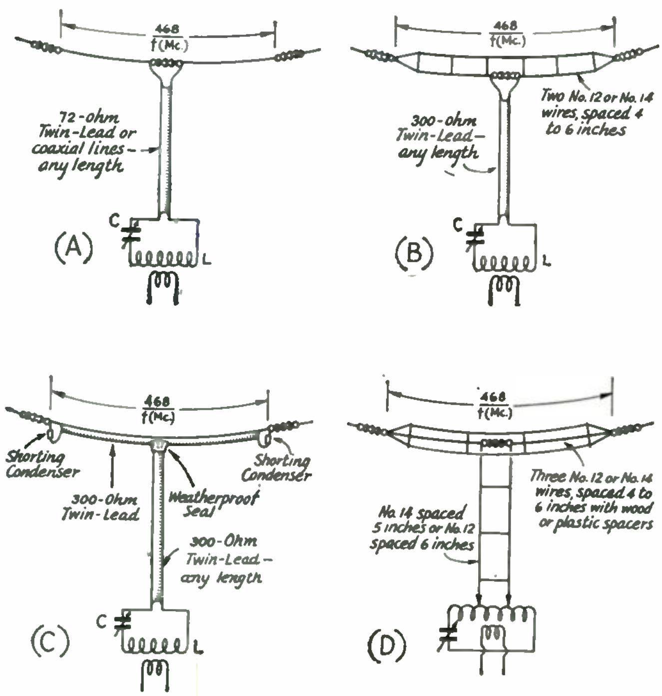


Fig. 7-1 — Half-wavelength antennas for single-band operation. The multiwire types shown in B, C and D have a slight advantage in that they will offer a better match over a wider range of frequencies, but otherwise the performances are identical. The feedlines can be any length in excess of a quarter wavelength, but they should run away from the antenna at a right angle for as great a distance as possible. In the coupling circuits shown, *LC* should resonate to the operating frequency. In the series-tuned circuit of A, B and C high *L* and low *C* are recommended, and in D the inductance and capacity should be similar to the output-amplifier tank, with the feeders tapped across about ⅓ the coil.

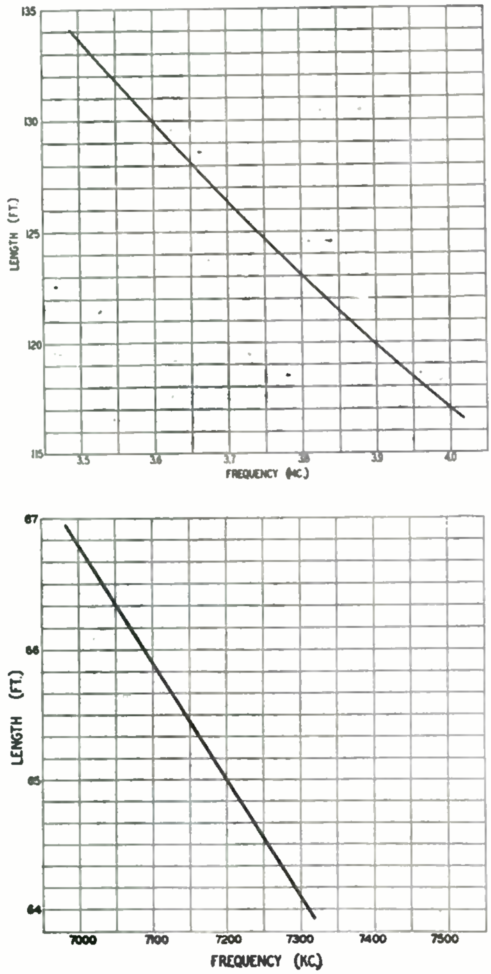


Fig. 7-2 — **The above charts can be used to determine the length of a half-wave antenna of wire.**

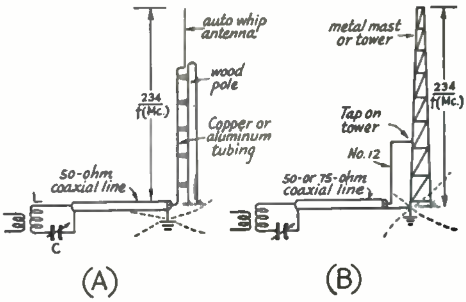


Fig. 7-3 — **Vertical antennas are effective for 3.5- or 7-Mc. work. The quarter-wavelength antenna shown at A is fed directly with 50-ohm coaxial line, and the resulting standing-wave ratio is about 1.5 to 1. If a grounded antenna is used as at B, the antenna can be shunt-fed with either 50- or 75-ohm coaxial line. The tap for best match will have to be found by experiment — the line running up the side of the antenna should be spaced from 6 to 12 inches from the antenna. The length (height) of the antenna can be computed from the formula, or it can be obtained from Fig. 7-2 by using just one-half the length indicated in the chart. For example, at 3.9 Mc., the length is 120′/2 = 60′.**

if best feedline matching is to be obtained. Small 500-volt mica receiver-type condensers will be satisfactory.

A folded dipole can be fed with a 600-ohm open-wire line with only a 2-to-1 standing-wave ratio, but a nearly perfect match can be obtained with 600-ohm open line and a three-wire dipole.

The three types of half-wavelength antennas just discussed are shown in Fig. 7-1. One advantage of the two- and three-wire antennas over the single wire is that they offer a better match over a band. This is particularly important if full coverage of the 3.5-Mc. band is contemplated.

While there are many other methods of matching lines to half-wavelength antennas, the three mentioned are the most practical ones. It is possible, for example, to use a quarter-wavelength transformer of 150-ohm Twin-Lead to match a single-wire half-wavelength antenna to 300-ohm feedline. But if 300-ohm feedline is to be used, a folded dipole offers an excellent match without the necessity for a matching section.

The formula shown above each antenna in Fig. 7-1 can be used to compute the length at any frequency, or the length can be obtained directly from the charts in Fig. 7-2.

As mentioned before, the height of an antenna is not too important at these frequencies, but it should be at least 35 feet for good results. The advantages of greater height will become increasingly apparent when working DX.

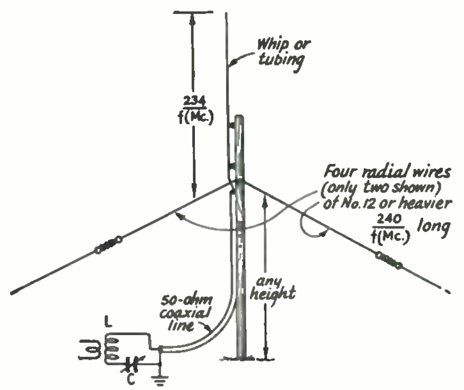


Fig. 7-4 — **A ground-plane antenna is effective for DX work on 7 Mc. Its base can be any height above ground. Feeding the antenna directly with 50-ohm coaxial cable will result in a low standing-wave ratio. The length of the vertical radiator can be computed from the formula, or it can be obtained from Fig. 7-2 by using just one-half the length indicated in the chart. The radial wires are 2.5% longer. For example, at 7.1 Mc., the radiator is 65′ 11″/2 = 33′; the radials are 1.025 × 33 = 33′ 10″.**

VERTICAL ANTENNAS

Vertical antennas find some favor on 3.5 and 7 Mc., and if one has a suitable location he can get good results with a vertical antenna on these frequencies. There is a good chance that BCI troubles will be a little more severe with a vertical antenna, but good results in DX work can be obtained, because of the low-angle radiation from the vertical.

For 3.5-Mc. work, the vertical can be a quarter wavelength long (if one can get the height), or it can be something less than this and "top-loaded." The bottom of the antenna has only to clear the ground by inches. Probably the best construction

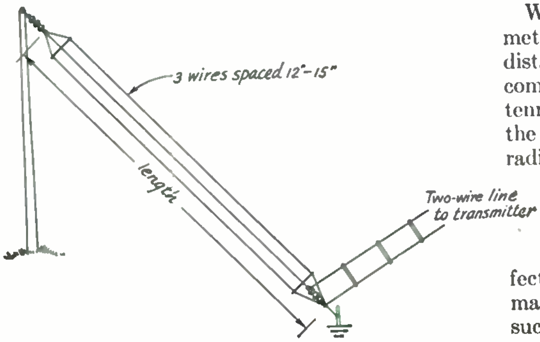


Fig. 7-5 — A three-wire quarter-wave dipole for 3.5 or 7 Mc. offers a good match for 300-ohm line. The antenna does not have to run exactly vertical, but a fairly good ground is important. The length is calculated from 234 ÷ *freq.* (Mc.), but, because the antenna is broad, this figure is not critical.

of a quarter-wavelength vertical involves running copper or aluminum tubing alongside a wooden mast, if a metal tower is not available. If a grounded tower is used, the antenna can be "shunt-fed," as shown in B of Fig. 7-3. A good ground system is helpful in feeding a quarter-wavelength vertical antenna, and the ground can be either a convenient water-pipe system, or several radial wires extending out from the base of the antenna for about a quarter wavelength.

The size of a "ground-plane" antenna makes it a little impractical for 3.5-Mc. work, but one can be used at 7 Mc. to good advantage, particularly on DX work. This type of antenna can be placed higher above ground than an ordinary vertical without decreasing the low-angle radiation. The vertical member can be a length of self-supporting tubing at the top of a short mast, and the radials can be lengths of wire used also to support the mast. The radials do not have to be exactly horizontal, as shown in Fig. 7-4.

Another vertical or semivertical antenna that will work well on one band (either 3.5 or 7 Mc.) is shown in Fig. 7-5. A quarter-wavelength three-wire dipole of this type offers a good match for 300-ohm line over a fairly wide frequency range. The wires should be spaced from about 12 to 15 inches either side of the driven center conductor, and wood spacers boiled in paraffin are quite adequate for the job. The ground should be a good one made directly to a water pipe near the base of the antenna, or a counterpoise or ground-radial system can be used. It is not important that the three-wire dipole be exactly vertical — part of the far end can run horizontally or nearly so, or the entire antenna can slope as much as 45°.

If the length of the three-wire dipole is made 55 feet, the antenna will work fairly well on both 3.9 and 14 Mc.

PHASED ARRAYS

When a directional antenna is required for 80-meter work — for special schedules over a long-distance path — the most practical system is a combination of vertical quarter-wavelength antennas fed in the correct phase, such as some of the broadcast stations use to obtain special radiation patterns. Since it is generally not very practical for an amateur to set up such an elaborate system, one usually has to forego the luxury of directive antennas at 80 meters.

Arrays using horizontal elements are not effective at these frequencies unless they can be made over 100 feet high, a condition that makes such systems impractical for amateur work. However, in the event that such heights can be obtained, phased arrays of horizontal elements will give added signal strength at a distance.

Phased arrays with horizontal elements can be used to advantage at 7 Mc., though, if they can be placed at least 60 feet above ground. Any of the usual combinations will be effective. If a

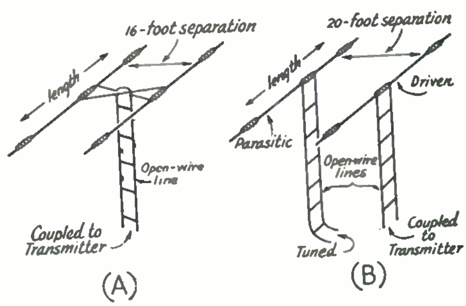


Fig. 7-6 — Directional antennas for 7 Mc. To realize any advantage from these antennas, they should be at least 40 feet high. The system at A is bidirectional, and that at B is unidirectional in a direction depending upon the tuning condition of the parasitic element. The length of the elements in either antenna should be exactly the same, but any length from 60 to 150 feet can be used. If the length of the antenna at A is made from 60 to 80 feet, it will be bidirectional along the same line on both 7 and 14 Mc. The system at B can be made to work on 7 and 14 Mc. in the same way, by keeping the length down to from 60 to 80 feet, but the tuning of the parasitic element will be different on the two bands.

bidirectional characteristic is desired, the W8JK type of array, shown at A in Fig. 7-6, is a good one. If a unidirectional characteristic is required, two elements can be mounted about 20 feet apart and provision included for tuning one of the elements as either a director or reflector, as shown in Fig. 7-6B. Such a system will not show much directivity on near-by signals, but a worth-while increase will be noted at distances of several thousand miles. The parasitic element is tuned at the end of its feedline with a series- or parallel-tuned circuit (whichever would normally be required to couple power into the line), and the proper tuning condition can be found by using the system for receiving and listening to distant stations along the line of maximum radiation of the antenna. The tuning of the feedline to the parasitic element will peak up the signal — that setting of the tuning condenser is noted, and then a signal in the opposite direction is peaked.

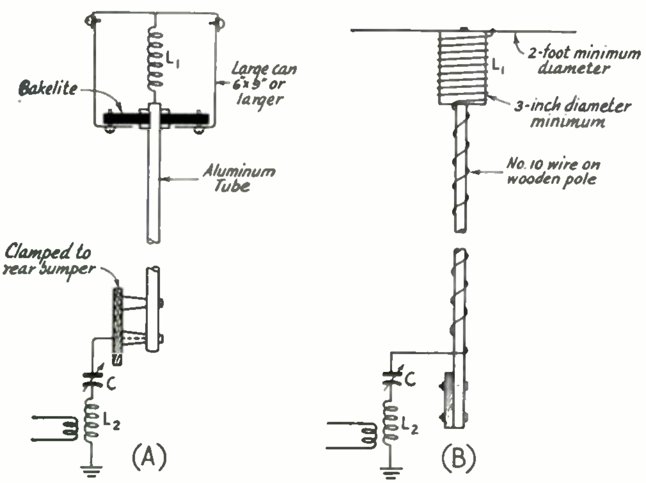


Fig. 7-7 — Two types of construction for "top-loaded" antennas for mobile work. At A a large cracker can is used as the loading capacity and a protective shield for the loading inductance, L_1; the antenna at B uses a large disk of aluminum for the capacity. Adjustment procedure will be found in the text.

● ANTENNAS FOR MOBILE WORK

The reason for the poor operation of "whip" antennas at low frequencies is the difficulty of putting any heavy current into the portion of the antenna "in the clear" behind and above the car. This can be overcome by "top loading" the antenna or by using a large loop antenna.

The antenna of Fig. 7-7A uses a large circular pretzel or cracker can for the top-loading capacity and to protect the loading coil, L_1. The larger the can the smaller will be the coil, but the coil will be large under any circumstances, generally ending up with more inductance than would normally be used in a transmitter tank at the same frequency. The antenna at B uses the same principle, but a disk of thin aluminum is used instead of the can. The coil L_1 is wound on a wooden form, and the active radiator of No. 10 wire is wound around the supporting pole a few times just to make L_1 a little smaller. The wire in L_1 and the wire on the pole can be covered with Glyptal or other protective lacquer after the adjustments have been completed. The height should be whatever is safe for the car and the driving habits of the operator.

The size of L_1 can be found by removing a few turns at a time, while maintaining constant input (loading) to the transmitter and observing the signal strength at a distance of a few miles, on the S-meter of a receiver. Make your adjustments at the low-frequency end of the band, to insure leaving enough coil for the rest of the band, and remove one turn at a time as the observed strength begins to level off. An alternative procedure is to disconnect the antenna from condenser C and connect C across L_2. With very loose coupling, resonate L_2C to the test frequency and note the setting of C. Reconnect the antenna to condenser C in the proper manner, and prune turns from L_1 until the system loads the transmitter best at the setting of C obtained in the first test.

With either the antenna at A or B, L_2 should be a large coil (high inductance) of good construction and high Q. This means using heavy wire or small copper tubing, and winding the coil with a diameter of at least 2½ or 3 inches. C should be a 100-$\mu\mu$fd. variable for 80-meter work, and a 50-$\mu\mu$fd. variable will suffice for 40 meters. The coil L_2 should tune to the band with condenser C across it and set at about 75 to 90 per cent of maximum.

If the coil L_1 in Fig. 7-7B is tapped, and coil L_2 is made plug-in, the antenna can be used on two bands.

Another type of antenna that can be used for mobile work is the "whip" antenna mounted at the top of the rear fender and bent forward and connected to the body of the car above one of the windows. It requires an excellent low-resistance connection to the car at this point, and the loop should extend as high above the car body as possible. It is fed in the same way as the previous antennas, through a series-tuned circuit.

It is important that the ground connection to the car be a good one. This requires scraping the paint from the car body and fastening a good lead with lock washers and well-tightened bolts. The connection can be made inside the trunk rack.

CHAPTER 8

Antennas for 14, 21 and 28 Mc.

The antenna systems discussed in Chapters Six and Seven can, of course, be used on 14, 21 and 28 Mc. with good results. The half-wavelength antenna of 3.5 Mc., fed with tuned feeders, becomes a multiwavelength antenna at these higher frequencies, and the directional characteristics become a little more apparent. The half-wavelength antennas fed with flat lines (Chapter Seven) can be scaled down for the higher frequencies, and they will give a good account of themselves. Suitable lengths are given in Fig. 8-1, and Fig. 3-38 should be referred to if feeder resonances are to be avoided. The vertical antennas (and the ground-plane in particular) can be used at these frequencies and will give good low-angle radiation, but when used for receiving the man-made noise pick-up is likely to be greater than with a horizontal antenna. The directional pattern of a half-wavelength antenna becomes apparent at these frequencies, and it is not at all unwise to provide two half-wavelength antennas for these bands, at right angles to each other, with a suitable switching arrangement to permit using one antenna or the other, depending upon the direction of the desired signal. Better still is to use a single half-wavelength antenna that can be rotated at least 135°. In either the case of the switchable antennas or the rotatable half-wave, it is highly desirable to use a carefully-balanced feedline, to insure against unwanted pick-up of signals on the feedline and consequent loss of directional effects.

On the other hand, if one wishes to avoid the directional effects at these frequencies, the antenna shown in Fig. 8-2 can be used. Both wires are horizontal and at right angles to each other. Since it is a balanced antenna, it will work on the three bands. If designed for 14 Mc. (each leg 33 feet long), it will be practically nondirectional on 14, 21 and 28 Mc., and it will show only slight directional effects on 7 Mc., along the bisector of the 90-degree angle.

When using a long-wire antenna at the higher frequencies (one of the antennas described in Chapter Six), it will pay to lay out its direction to take advantage of the major lobes that develop at the higher frequencies (see Chapter Two), if this is at all possible. Further, if the antenna is not to run horizontal, it should be kept in mind that the major useful radiation from a long sloping wire is in the general direction of the slope although, of course, there will be useful radiation at other angles as well.

DIRECTIONAL ANTENNAS

Because a wavelength is smaller physically at these frequencies than at the lower ones, relatively larger antennas (in terms of wavelengths) can be built, to take advantage of the gain and directive effects of the more complicated types of antennas. A 7-Mc. half-wavelength antenna, fed with tuned line, becomes "two half waves in phase" at 14 Mc., with a gain of 1.8 db. over a 14-Mc. half-wavelength antenna. However, by making the antenna only a little longer it becomes an "extended double Zepp" and has a gain of 3 db. The proper length of an extended double Zepp can be obtained from Fig. 8-3, and its physical construction is the same as for the center-fed antenna of Fig. 6-4. On the design frequency and at half this frequency, the maximum radiation is at right angles to the wire. At higher frequencies, the pattern tends toward the "X" shapes obtained from 1- and 2-wavelength antennas.

As one goes to the higher-frequency bands, it becomes increasingly important to confine the radiation to low angles above ground, for maximum signal strength on both transmitting and receiving. Since the vertical pattern of a simple antenna depends upon its height above ground, it would be wise to choose a height that confines the radiation to the lower angles. This is difficult to do, however, because the exact height above electrical ground is hard to determine. The solution is to use an antenna system that is less dependent upon ground height. While there is no antenna that is entirely free of this effect, con-

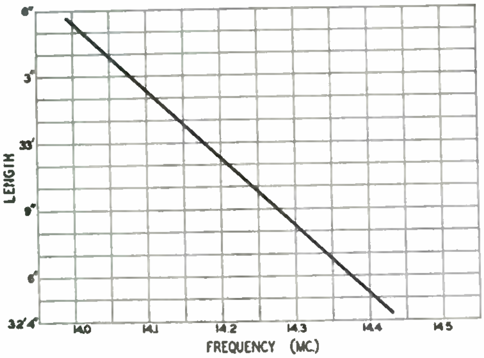
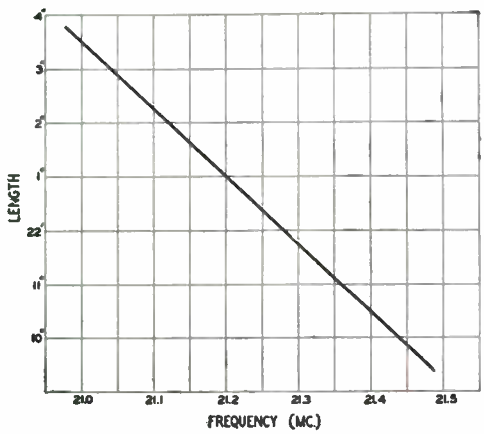
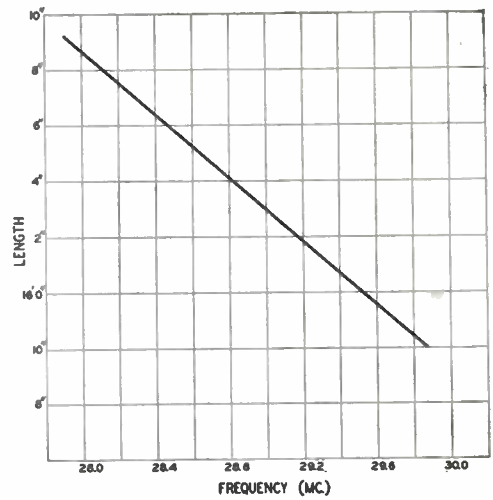


Fig. 8-1 — Charts for determining the length of a half-wavelength antenna of wire at 14, 21 and 28 Mc., based on 468/*f* (Mc.). Constructional details are shown in Fig. 7-1.





siderable improvement can be obtained. The ground-plane antenna will have the same pattern as a half-wavelength vertical with its center at the same height as the base of the ground plane. Simple end-fire or broadside arrays using horizontal elements will increase the low-angle radiation. They are also fairly easy types of antennas to build. The end-fire (W8JK beam) is generally easier to install because it doesn't require as much height to be effective, but its Q is higher and the system is somewhat harder to feed. The feed problem can be reduced by using multiwire elements. Practical types of simple end-fire and broadside arrays are shown in Figs. 8-4 and 8-5.

The antenna of Fig. 8-4A shows the general form and constructional details of the popular "W8JK" beam. Although it is generally strung between two supports in a horizontal position, it can be hung vertically from a single support. It is useful for the amateur with little room because it can be used on the three bands, if tuned feeders are connected to points x and y. The dimensions are not critical (see caption with the sketch), but the lengths running away from points x and y should all be equal. The folded-dipole version in Fig. 8-4B is a one-band affair, but has the advantage that the feed point (x and y) offers a more normal impedance value and the standing-wave ratio will be lower than in the first case. Open-wire line is recommended for use with either system. The longer-length elements will give several db. gain more than the shortest recommended lengths.

The antennas shown in Fig. 8-5 are "broadside" arrays. The system illustrated in A is the popular "lazy H," and it is an antenna that is good for DX work on the three bands. For best results on 14 through 28 Mc. the bottom wire should be 30 or 35 feet above ground, but the system is useful when the lower wire is only 15 or 20 feet high. Open-wire line should be used between the transmitter and points x and y. The larger dimensions will give several db. more gain

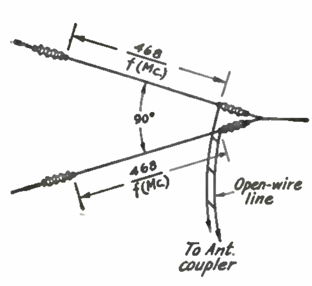
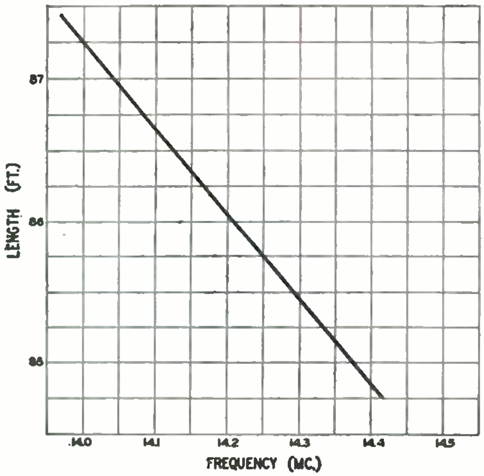
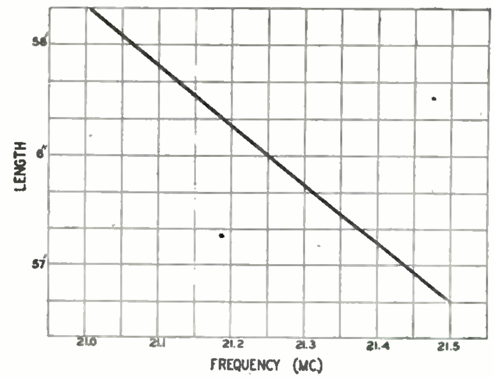


Fig. 8-2 — A simple nondirectional antenna for high-frequency work. It will work satisfactorily from half the design frequency to anything higher. Both wires should be horizontal. Designed for 14.2 Mc., the length of each leg is 33 feet.





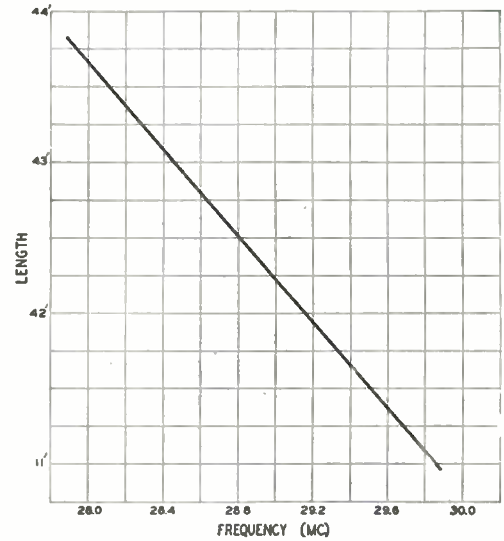


Fig. 8-3 — Charts for determining the over-all length of the flat-top portion of an extended double-Zepp antenna. The construction is the same as shown in Fig. 6-4.

than the minimum, but any intermediate lengths can be used, if care is taken to make the system symmetrical about the feed point. The "bi-square" array shown in Fig. 8-5B is a version of the "lazy H" that requires only one support instead of two, and its dimensions are such that it finds most application on 21 and 28 Mc. Open-wire line can be used between the transmitter and points x and y.

Directivity not only increases the strength of the transmitted and received signals, but the reduction of response in undesired directions is helpful under QRM conditions. The beams and antennas described so far are at best bidirectional.

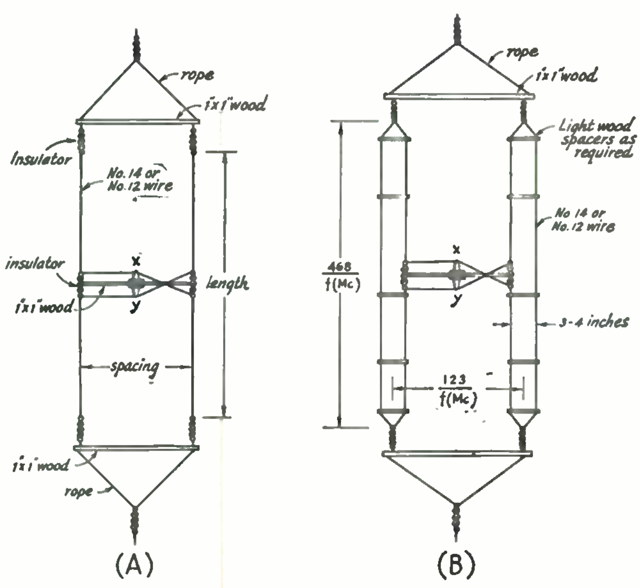


Fig. 8-4 — Two simple end-fire arrays for 14, 21 and 28 Mc. These arrays are called "W8JK" antennas, and are generally used horizontally, thus requiring two supports, but they can be hung vertically and rotated if desired. That shown at A will be useful on the three bands if the length is made 33 to 40 feet, with a spacing of 8½ feet.

The antenna shown at B is good for only one band, but it is easier to feed. Element lengths can be found from Fig. 8-1.

Either type of antenna is fed at points x and y with an open-wire line — the standing-wave ratio will be lower with B.

and an improvement in signal strength and operating convenience can be obtained by making them unidirectional. One of the simplest forms of unidirectional beams is shown in Fig. 8-6. Two folded dipoles spaced a quarter wavelength are fed by equal lengths of 300-ohm line. An extra length of line, electrically a quarter wavelength long, is switched in to one line or the other. This causes the phase difference between the two dipoles to be 90°, and the system becomes unidirectional along a line through the two antennas. Reversing the switch reverses the direction of maximum gain. A gain of about 5 db. and a front-to-back ratio of about 20 db. can be obtained with this simple system. It is desirable to place it at least a half wavelength above ground.

The bisquare antenna can be used in a somewhat similar fashion by using another section and tuning it either as a director or a reflector, as shown in Fig. 8-7. The driven element is fed at x

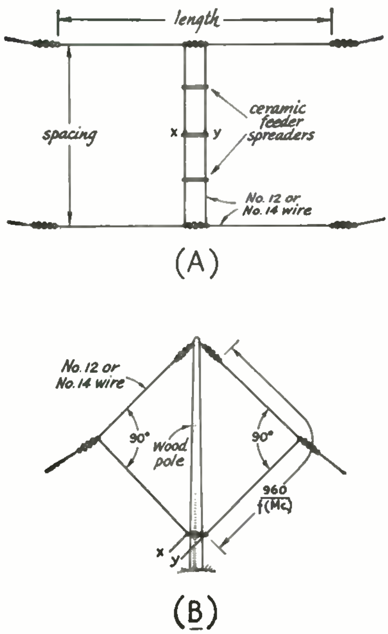


Fig. 8-5 — Two popular types of broadside arrays. The "lazy H" shown at A will work on 14, 21 and 28 Mc. if the length is made 33 to 40 feet, with a spacing of 16 to 24 feet. The larger dimensions give slightly more gain on all bands. The bottom wire should be at least 15 feet above ground for three-band operation, but better results will be obtained if the bottom wire is 30 or 35 feet above ground.

The "bisquare" antenna shown at B is a useful antenna for the 28-Mc. band. It has the advantage that only one support is required. The bottom of the antenna should be at least 3 feet above the ground, and preferably about 10 or 12 feet above electrical ground. The bisquare will show some end-fire directivity at half frequency, with vertical polarization. The total length of wire on one side is 45 feet 3 inches at 21.2 Mc., and 33 feet at 29 Mc.

Either antenna should be fed at points x and y with a tuned line.

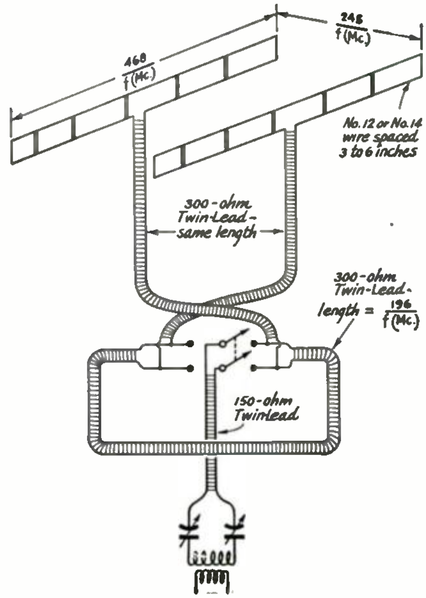


Fig. 8-6 — A simple reversible fixed beam for 14, 21 or 28 Mc. Two folded dipoles (see Fig. 7-1 for constructional details) are mounted a quarter wavelength apart and are fed with 300-ohm line. The directivity is made reversible by the d.p.d.t. switch (or relay) as shown. Element lengths can be obtained from Fig. 8-1. The spacing is 17 feet 3 inches at 14.2 Mc., 11 feet 7 inches at 21.2 Mc., and 8 feet 5 inches at 29 Mc. The phasing section is 13 feet 9 inches at 14.2 Mc., 9 feet 3 inches at 21.2 Mc., and 6 feet 9 inches at 29 Mc.

and y by a tuned line, and the switch S is opened when the parasitic element is used as a director. Under these conditions, C_1 is tuned for maximum signal coming from the left-hand side (Fig. 8-7). For signals coming in the opposite direction (parasitic element used as a reflector), the switch is closed and C_2 peaked for the best signal coming from the right-hand side. Once the two condensers

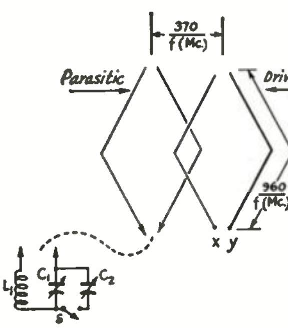


Fig. 8-7 — This unidirectional beam uses two bisquare elements, one driven and one tuned as either a reflector or director. The switch S is opened when the parasitic element is used as a director, and C_1 is tuned for maximum gain of the antenna in the left-hand direction. When the switch is closed, C_2 is tuned for maximum gain in the right-hand direction.

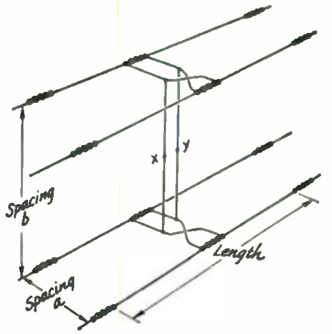


Fig. 8-8 — A bidirectional beam, made by combining the end-fire array of Fig. 8-4 with the broadside array of Fig. 8-5A. If the length is made from 33 to 40 feet, spacing *a* 8½ feet, and spacing *b* 16 to 24 feet, the beam will operate well on 14, 21 and 28 Mc. For any one band, the length can be from 0.5 to 1.2 wavelength, spacing *a* ⅛ wavelength, and spacing *b* 0.5 to 0.75 wavelength.

have been adjusted, the switch can be controlled remotely from the operating position if desired.

By combining the end-fire and broadside arrays, multiband combinations can be obtained that have improved gain and low-angle radiation compared with the element arrays. One of the best combinations is shown in Fig. 8-8. By using tuned feeders, this bidirectional beam will work well on 14, 21 and 28 Mc. Its construction is simply that of two W8JK beams (Fig. 8-4A), supported one above the other. For best results, the bottom section should be at least 30 or 35 feet above ground, but the system will work well with the bottom element only 15 or 20 feet high. For single-band operation, the beam can be fed with a flat line through a matching transformer at x and y, and the height of the bottom section should be a half wavelength. All similar sections of the system (i.e., the wires running from points x and y, the four active antenna wires, etc.) should be of equal length, but the numerical value of length is unimportant. The larger dimensions will give several db. greater gain than the minimum lengths.

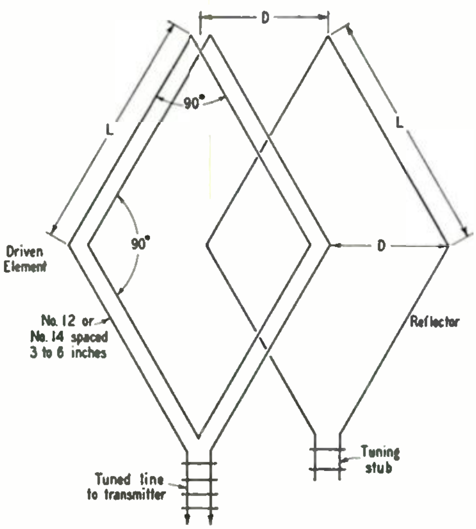


Fig. 8-9 — The "quad" antenna system consists of two square loops of wire, one acting as a reflector for the two-wire driven element. The spacing D is 0.15 wavelength: 3½ feet for 21.2 Mc., 2½ feet for 29 Mc. The length of one side, L, is a quarter wavelength: 11 feet 4 inches for 21.2 Mc., 8 feet 3 inches for 29 Mc. Any suitable wooden framework can be used for support — the insulation between the two wires of the driven element is not very important, but the side corners of the loops should be well insulated from ground. The tuning stub may be several feet long at these frequencies, and can be two parallel lengths of bare wire spaced 3 to 6 inches.

THE "QUAD" ANTENNA

The "quad" antenna shown in Fig. 8-9 is a square loop antenna that is compact enough to be practical for the 21- and 28-Mc. bands. Used with a reflector, it has a gain of about 7 db. over a dipole, provided the centers of the loops are at least ¾ wavelength (24 feet at 28 Mc., 33 feet at 21 Mc.) above ground.

The antenna and reflector are made of No. 12 or No. 14 wire on a suitable supporting frame of light wood. If the dimensions given in the illustration are followed, the only adjustment that is necessary after installation is to find the correct position of the shorting bar on the stub that tunes the reflector. This can be done in the usual way by exciting the antenna (constant input) and measuring the field strength at a distance of ten wavelengths or so away, at the same height as the antenna. The driven element can be made of a single loop of wire instead of the two wires shown in Fig. 8-9, but the impedance offered to the feed-line will be lower and the standing-wave ratio higher. Doubling the wire in the reflector adds nothing to the performance and is likely to make the system more frequency-sensitive.

The quad can be rotated 45° and fed at the center of the bottom side, if desired. In this case, the reflector should also be rotated by the same amount, and the tuning stub connected in the center of the bottom side.

Using the driven element without the parasitic reflector shows so little gain over an ordinary dipole that its use is not recommended.

VEES AND RHOMBICS

On 14, 21 and 28 Mc., the long-wire beams (vees and rhombics) are quite effective, and a properly-designed rhombic will give a gain that is not likely to be equaled by any other type of antenna. The pattern of the antenna radiation becomes rather sharp, though, and hence the really high gain is effective over only a relatively narrow angle of perhaps 20 or 25 degrees. However, minor lobes of considerable amplitude help to fill in some of the other directions.

The design of Vs and rhombics is treated in Chapter Five, but it can be repeated here that the larger and higher the rhombic is made, the higher will be the gain and the lower will be the wave angle. If the rhombic is terminated to give a unidirectional characteristic, one should be sure before terminating the beam that he knows over which of the two possible paths the signals generally travel. An 800-ohm noninductive resistor, rated at one-half the transmitter output power or more, should be used.

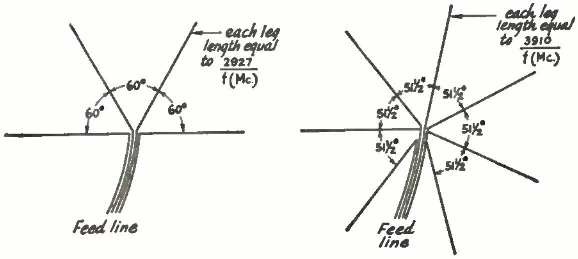


Fig. 8-11 — Horizontal vee beams can be arranged around a central support to give switchable directivity. The two examples are for 3- and 4-wavelength legs respectively. The system at A is not repeated for another 180° because any one vee beam is bidirectional. Two legs could be omitted from the system at B with no appreciable loss of coverage.

Unfortunately, the excellent results obtained with the long-wire beams do not maintain as the antenna is made smaller and smaller (for the same frequency), and so it is wiser to use some other form of beam if space is limited. The minimum rhombic dimensions shown in Fig. 8-10 are about as small as it is advisable to go. However, the investment in parts and labor in even the smallest practical rhombic is a good one, since the antenna will put out a good signal on bands from 3.5 to 30 Mc., even though the design center is one particular band and the directivity is optimum only for that frequency.

If room is available, a very effective system can be made by using a number of vee beams with their vertices at a common support (which is generally made higher than the outlying supports, so that the antennas will slope), and running feeder lines from each antenna element. Then any vee beam in the system can be selected by switching. Such an antenna is shown in Fig. 8-11, for four and seven wires. Other suitable combinations are five 5-wavelength elements arranged radially at 45 degrees, and nine 6-wavelength elements arranged radially at 40 degrees. If the elements are designed for 21 Mc., the antennas will work well from 14 to 30 Mc., and will of course have useful radiation (but not as much directivity) on the lower frequencies. The higher the wires can be strung, the greater will be the gain of the beams, and the antennas should be at least 35 feet high for any real effectiveness. The feeder wires should be spaced from 3 to 6 inches around the periphery of an imaginary circle, and all feed wires should be made the same length. The feeders in use at any time are coupled to the transmitter through a tuned coupling system, and the unused feed wires should be grounded or left floating, whichever condition shows less r.f. in the unused wires and gives the greater directional effect on receiving.

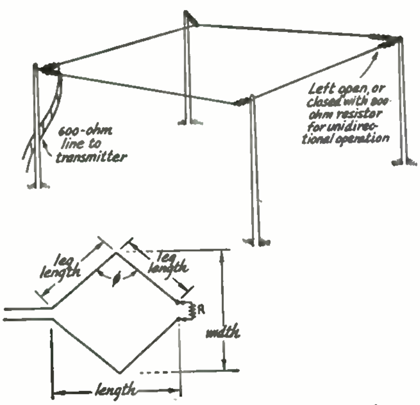


Fig. 8-10 — The rhombic antenna requires four supports, but it has excellent gain and directivity over a wide frequency range. If the rhombic is terminated in a resistor, for unidirectional work, it can be fed by a 600-ohm line and a low standing-wave ratio will be obtained. An unterminated rhombic should also be fed with an open-wire 600-ohm line, but the s.w.r. will be higher. A few recommended dimensions are given below for use on 14, 21 and 28 Mc. The height should be at least 35 feet — greater heights will result in better low-angle radiation. These dimensions are compromise figures for multiband work. For peak performance on one band, the design should be obtained from Fig. 5-18.

Leg Length (feet)	137	171	205
φ	107°	116°	123°
Length (feet)	220	290	360
Width (feet)	163	181	195

ROTARY BEAMS

Since the average operator on 14, 21 and 28 Mc. is not likely to have room for an "antenna farm" or rhombics or other arrays pointing in all directions, his best alternative is to rotate the most effective type of antenna he can install. While rotating a simple half-wavelength antenna at these frequencies gives some advantage, most operators prefer something more elaborate once they have gone to the trouble of acquiring a rotation mechanism. To minimize the height of the rotated structure, most rotary beams for the 14-, 21- and 28-Mc. bands are of the "parasitic-array" type, using a driven element, a reflector, and one or two directors. Because all of the elements are in one plane, the construction of the antenna is not a difficult job. Two-, three- or four-element beams of this type can be mounted on a common rotator and stacked one above the other, thus giving a separate antenna for each band. In general, the spacing between such stacked antennas should be as great as possible, to avoid interaction, although separate three-element beams have been mounted in the same plane or separated by only a few feet. A separation of one-eighth wavelength (figured at the lower frequency) results in but little interaction, however.

The design of the rotary beam depends to a great extent upon the facilities available to the constructor, and the allowable room and the supporting structure will do much toward determining what is built. It is vitally important that the beam and supporting structure be capable of withstanding severe wind and ice loads, to avoid possible damage to surrounding buildings in the event of a structural failure. Several practical designs are given in Chapter Twelve.

The length of the elements in a two- or three-element beam does not vary greatly with the element spacing, but the gain of the array, its bandwidth and the impedance of the driven element all do. For this reason, it is advisable to make the element spacing in a 3-element beam as great as possible (up to 0.2 wavelength), and to provide for adjusting the matching of the antenna to the feedline after the beam is installed on its support. A 2-element beam should have a spacing of 0.1 to 0.15 wavelength, depending on whether the parasitic element is to be used as a director or reflector. The parasitic director spaced 0.1 gives slightly more gain than the parasitic reflector spaced 0.15 wavelength. Three-element beams for 14 Mc. are usually spaced only 0.1 wavelength because of physical limitations. In any case, however, the adjustable matching requirement can be met by using a "T"-match, as shown in Fig. 8-12. Feeding the beam with either the two-wire line or the pair of coaxial cables, the feedline should be run away at right angles to the antenna for as far as possible, to avoid parallel components on the line. The match should be adjusted for minimum standing-wave ratio on the line after the antenna is installed on its support, using one of the s.w.r. indicators described in Chapter Three.

Although the direct feed with line and "T"-match is the most simple and straightforward approach to feeding the beam, many operators do not like the slight inconvenience of not being able to use continuous rotation of the beam. One way to avoid twisting the feedline is to install limit switches that warn the operator when he should turn the beam in the opposite directions. However, for continuous rotation of the beam, either inductive coupling or a pair of slip rings is used. Samples of these methods are shown in Chapter Twelve.

As mentioned before, the length of the elements does not affect the gain of the three-element beam to the extent that the spacing does. A gain of 6 or 6.5 db. can be expected from a three-element beam using 0.1-wavelength spacing for both director and reflector, and it will increase to over 9 db. with spacings of 0.2 wavelength. Intermediate gains and spacings are: 0.15D-0.1R, 7.8 db.; 0.2D-0.1R, 8.3 db.; and 0.2D-0.15R, 8.7 db. Element lengths for any of these spacings can be found in the charts of Fig. 8-13. If the lengths are taken from the graphs, very little improvement in forward gain will be obtained by adjustment of the elements after the antenna system is in place, but the front-to-back ratio can be increased by careful tuning of the reflector.

The "T"-match should be adjusted with the antenna in place and the elements set to the lengths shown in Fig. 8-13. A "twin-lamp" or "Micro-Match" at the transmitter will show when the best match is obtained.

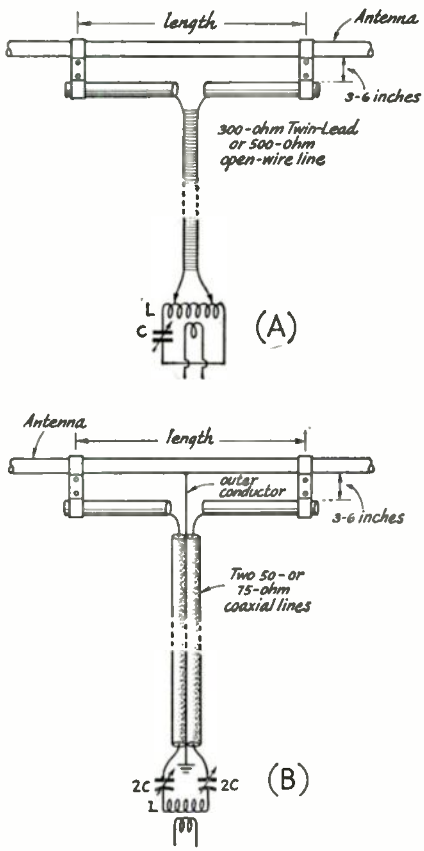
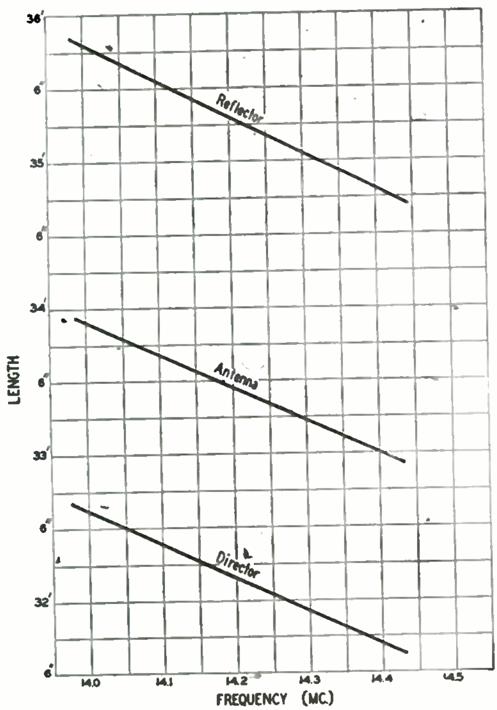
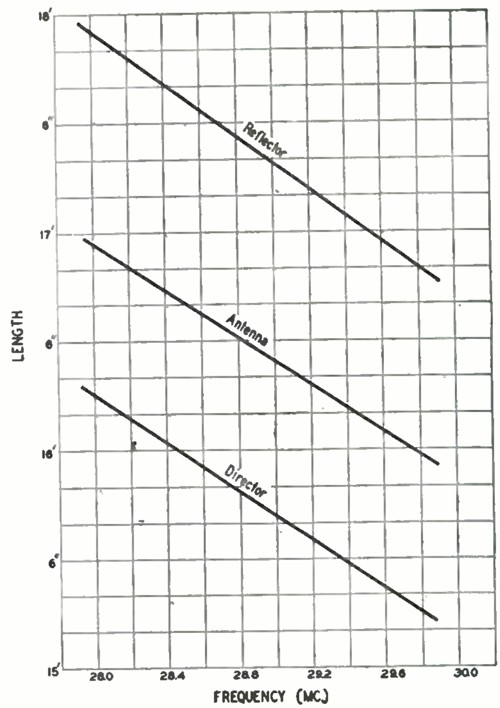
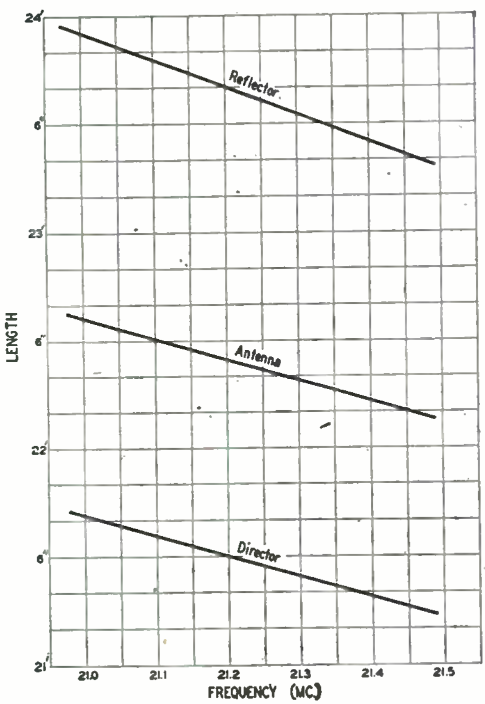


Fig. 8-12 — Two recommended forms of feedline match for a three-element beam. Both use the "T"-match principle, and the length of the matching section must be found by experiment, since it will vary with element spacing, height above ground, and nominal line impedance. For a 14-Mc. 3-element beam, it may be on the order of 6 to 9 feet, and correspondingly less at the higher-frequency bands.







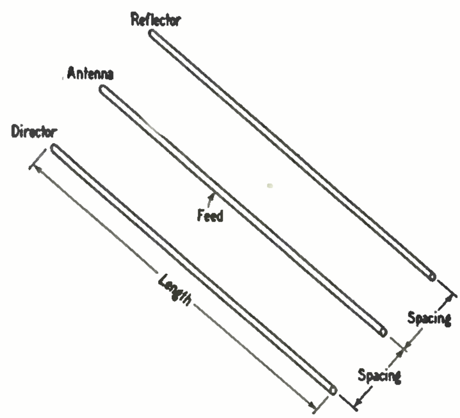


Fig. 8-13 — Director, antenna and reflector lengths for three-element beams, for element spacing of 0.1 to 0.2 wavelength. The greater spacing will result in slightly higher gain. These curves represent average maximum-gain values obtained with tubing elements supported only at the center, and are likely to vary slightly with the installation. For best results (and maximum front-to-back ratio) the beam should be tuned in position, on the usual operating frequency.

ANTENNAS FOR MOBILE WORK

For mobile operation on 14 and 21 Mc., the antennas should be top-loaded in the fashion shown in Fig. 7-7. However, at these frequencies the necessary loading will not be as great as for the low frequencies, and the inductance L_1 will become smaller. The feed circuit, L_2C, should resonate to the operating frequency with the antenna disconnected (with the components connected in parallel), and then L_1 can be trimmed (with the antenna connected properly) until the same setting of C is obtained. Another procedure is to prune L_1 until maximum signal strength, as observed several miles away on a receiver using a vertical antenna, is obtained, with constant transmitter input.

On 28 Mc., top loading the antenna will give a better signal with an 8-foot antenna than if it is not loaded, but many operators prefer the convenience of the simple quarter-wavelength whip fed at the base. In this case, a neon bulb fastened or touched to the top of the whip will indicate maximum r.f. in the antenna. The top-loaded 28-Mc. antenna will indicate maximum r.f. (glow in neon bulb) also at the base of the antenna, and the coil L_1 can be pruned as for the lower-frequency antennas mentioned above. The two types of 28-Mc. antennas are depicted in Fig. 8-14.

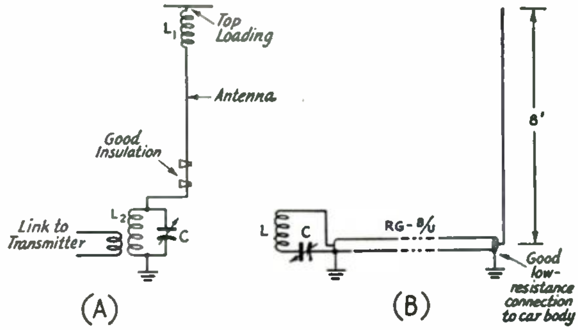


Fig. 8-14 — Two antennas for 28-Mc. mobile work. The top-loaded antenna of A is built as described in Fig. 7-7, and the height (length) of the antenna should be from 5 to 8 feet. Slightly better performance can be expected from the greater heights.

The 8-foot whip antenna shown at B can be fed very conveniently with any length of RG-8/U (52-ohm) coaxial line, and it is most readily coupled to the transmitter by series-tuning the coupling coil, L in B. At 29 Mc., the coil L should be one that tunes to the 10-meter band with about 10 or 15 $\mu\mu$fd. across it. C should be a 20- or 25-$\mu\mu$fd. variable.

CHAPTER 9

V.H.F. and U.H.F. Antenna Systems

While the basic principles of antenna design are essentially the same for all communication frequencies, certain factors peculiar to v.h.f. and u.h.f. work call for changes in amateur antenna technique for the frequencies above 50 megacycles. Here the physical size of multielement arrays is reduced to the point where an antenna system having some gain over a simple dipole is possible in almost every location, and the more progressive stations may employ arrays having much higher gain than is possible on lower frequencies.

The importance of high-gain antennas in v.h.f. work cannot be overemphasized. The reliable working range of a station operating on 144 Mc., for instance, may be only 30 miles or so when a simple dipole antenna is used, yet this same fellow may increase his working radius to 100 miles or more by the installation of a high-gain array. The directive system introduces other advantages also. By restricting the field covered at any one position the beam antenna helps to reduce pick-up of man-made noises, and it may be instrumental in correcting interference to broadcast and television reception, by permitting communication in directions not coinciding with near-by antennas used on receivers for these services. A good antenna system is often the difference between routine operation and outstanding success in the v.h.f. field, and it is safe to say that by no other means can so large a return be obtained from a small investment as results from the erection of a high-gain antenna system.

Design Considerations

Antenna systems for the v.h.f. range are usually called upon to work over a wider frequency range than those used on lower bands; thus maximum frequency response becomes an important consideration in the design of a v.h.f. array. It may be necessary, in some instances, to include this characteristic at the expense of other attributes which might be considered desirable, such as high front-to-back ratio.

A properly-matched line is of utmost importance in the proper functioning of the v.h.f. antenna system. As the frequency is increased, line losses increase sharply and it becomes well-nigh impossible to use tuned feeders of any appreciable length in v.h.f. work. Because any v.h.f. transmission line is long in terms of wavelength, it may be more effective to use a high-gain array at relatively low height, rather than a simpler array at great height above ground, particularly if the antenna location is not shielded by buildings or heavy foliage in the immediate vicinity.

Height above ground is helpful, especially in cases where added height increases the distance to the visible horizon appreciably, but great height is by no means so all-important as it was once thought to be. Outstanding results have been obtained, particularly on 50 Mc., with relatively low antennas, and many 144-Mc. stations are working out successfully with arrays not more than 25 to 40 feet above ground.

The effectiveness of a v.h.f. antenna system can be increased markedly by stacking half-wave elements one above the other and feeding them in phase. Such stacking helps to lower the radiation angle, an important factor in extending v.h.f. coverage. Several examples of outstandingly-successful stacked arrays are shown in the following pages.

Polarization

Most early work on the frequencies above 30 Mc. was done with vertical antennas, probably because the simple vertical dipole avoided the directivity problem, then considered to be a liability. In operation over almost pure line-of-sight ranges, vertical antennas provided better coverage than did the same dipoles turned over to a horizontal position. However, with the advent of high-gain directional arrays and extended operating ranges, horizontal systems began to assume importance in v.h.f. work, particularly in areas where extensive activity had not already been established with verticals.

Experience has shown that there is no marked difference in the effective working range with either polarization, *if* the same is used at both ends of the path. The horizontal array has some advantages, particularly for 50-Mc. work. It is somewhat easier to build a rotatable array for this band if horizontal polarization is employed, and the signal-to-noise ratio tends to be better with horizontal systems in locations where man-made noise is a problem. Simple 2-, 3- or

4-element arrays have proven extremely effective in 50-Mc. work, and the use of such arrays has reached the point of standardization on horizontal polarization for that band.

The picture is somewhat different when one goes to 144 Mc. and higher bands. Here the most effective vertical systems (those having two or more half-wave elements vertically stacked) are more easily constructed than on 50 Mc. The large number of mobile stations on 144 Mc., almost all using vertical antennas, are at a disadvantage in an area where horizontal polarization is used. Much of the country's 2-meter activity is in urban areas, where quite a few stations do not have space or facilities for the remotely-controlled rotary arrays which are a necessity for effective operation with horizontal systems. Television and f.m. receivers, both highly sensitive to interference from v.h.f. stations, use horizontal polarization, and it has been demonstrated that interference to near-by receivers can be reduced with vertical antennas on the amateur end.

Still, horizontal polarization is finding increasing 144-Mc. acceptance in many areas. At this writing the Middle West is almost completely horizontal, with the East and West Coasts largely vertical. Many of the more advanced stations employ both, but for the fellow who must be satisfied with one array the best procedure is to determine what polarization is in use in the areas he wishes to work, and then go on with that. In some instances, particularly where a station is directly in the shadow of a hill, there may be a considerable degree of polarization shift, permitting successful work with cross-polarization, but ordinarily it may be assumed that only by matching the polarization of other stations can work be carried on over extensive areas.

IMPEDANCE-MATCHING METHODS

Because line losses tend to be much higher in v.h.f. antenna systems, it becomes increasingly important that feedlines be made as "flat" as possible. Transmission lines normally used for v.h.f. work include the open-wire line of 450 to 600 ohms impedance, usually spaced about two inches; the polyethylene-insulated flexible lines, available in various impedances from 72 to 300 ohms; and coaxial lines of 50 to 90 ohms impedance. Occasionally transmission lines may be made up of two coaxial lines side by side, using the inner conductors as the line, with the outer conductors grounded. Such a line has approximately twice the impedance of one coaxial component. Any of these transmission lines may be matched to dipole or multielement antennas by the arrangements described below:

The "J"

Used principally as a means of feeding a stationary vertical radiator, around which parasitic elements are rotated, the "J" consists of a half-wave vertical radiator fed by a quarter-wave matching section, as shown at A, Fig. 9-1. The spacing between the two sides of the matching section should be two inches or less, and the point of attachment of the feedline will depend on the impedance of the line used. The feeder should be

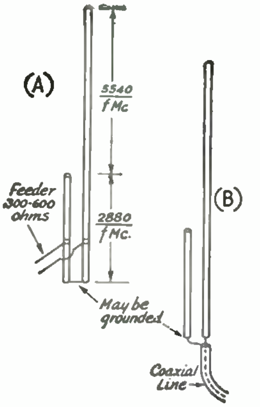


Fig. 9-1—Two versions of the "J" antenna, used in mobile applications, or in vertical arrays where parasitic elements are rotated around a fixed radiator.

slid along the matching section until the point is found that gives the best operation. The bottom of the matching section may be grounded for lightning protection. A variation of the "J" for use with coaxial-line feed is shown at B in Fig. 9-1. The "J" is also useful in mobile applications, though a simple quarter-wave whip will usually suffice for mobile work.

The Delta or "Y" Match

Probably the simplest arrangement for feeding a dipole or parasitic array is the familiar delta, or "Y" match, in which the feeder system is fanned out and attached to the radiator at a point where the impedance along the element is the same as that of the line used. Information on figuring the dimensions of the delta may be found in Chapter Three. Chief weakness of the delta is the likelihood of radiation from the matching section, which may interfere with the effectiveness of a multielement array. It is also somewhat unstable mechanically, and quite critical in adjustment.

The "T" Match

The principal disadvantages of the delta system can be overcome through the use of the arrangement shown in Fig. 9-2, commonly called the "T" match. It has the advantage of providing a means of adjustment (by sliding the clips along the parallel conductors), yet the radiation from the matching arrangement is lower than with the delta, and its rigid construction is more suitable for rotatable arrays. It may be used with coaxial lines of any impedance, or with the various other forms of transmission lines up to

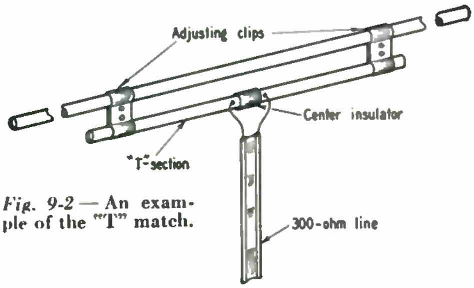


Fig. 9-2 — An example of the "T" match.

300 ohms. The position of the clips should, of course, be adjusted for maximum loading and minimum standing-wave ratio, the latter being most important as an indication of proper setting. The "T" system is particularly well suited for use in all-metal "plumbing" arrays.

The "Q" Section

A widely-used arrangement for matching an open-wire line to a dipole, or to the driven element in a 2- or 3-element array having wide (0.25 wavelength or greater) spacing, is the "Q" section (Chapter Three). This consists of a quarter-wave line, usually of ½-inch or larger tubing, the spacing of which is determined by the impedance at the center of the array. The parallel-pipe "Q" section is not practical for matching multielement arrays to lines of lower impedances than about 600 ohms, nor can it be used effectively with close-spaced parasitic arrays. The impedance of the "Q" section required in these cases is lower than can be obtained with parallel sections of tubing of practical dimensions. A quarter-wave section of coaxial or other low-impedance line is a commonly-used means of matching a line of 300 to 600 ohms impedance to the low center impedance of a 3- or 4- element array. The length of such a "Q" section will depend on the velocity of propagation (propagation factor) of the line used. The propagation factors of all the commonly-used lines are given in table form in Chapter Three.

In some installations it may be more convenient to use a line of greater length than a single quarter wave for matching purposes, in which case any odd multiple of a quarter wavelength may be used. The exact length required may be determined experimentally by shorting one end of the line and coupling it to a source of r.f., and trimming the line length until maximum loading is obtained at the center frequency of the operating range.

The Folded Dipole

Probably the most effective means of matching various lines to the wide range of antenna impedances encountered in v.h.f. antenna work is the folded dipole, shown in its simplest form in Fig. 9-3. When all portions of the dipole are of the same conductor size, the impedance at the feed point is approximately equal to the *square* of the number of elements in the folded dipole times the normal center impedance which would be present if only a conventional split half-wave radiator were used. Thus, the simple folded dipole of Fig. 9-3 has a feed-point impedance of 4×72, or approximately 288 ohms. It may be fed with the popular 300-ohm line without appreciable mismatch. If a three-wire dipole were used, the step-up in impedance would be *nine* times. Note that this step-up occurs *only* if all portions of the folded dipole are the same conductor size.

The impedance at the feed point of a folded dipole may also be raised by making the fed portion of the dipole smaller than the parallel section. Thus, in the 50-Mc. array shown in

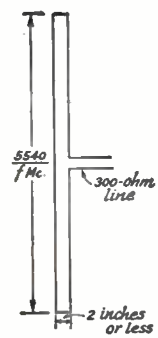


Fig. 9-3 — Details of a folded dipole for v.h.f. use.

Fig. 9-6, the relatively-low center impedance of a 4-element array is raised to a point where it may be fed directly with 300-ohm line by making the fed portion of the dipole of ¼-inch tubing, and the parallel section of 1-inch. A 3-element array of similar dimensions could be matched by substituting ¾-inch tubing in the unbroken section. Conductor ratios and spacings for other applications may be obtained from the folded-dipole nomogram in Fig. 3-51.

Using Coaxial Lines — The "Bazooka"

Flexible coaxial line has many advantages. It is completely weatherproof and may be run underground or inside a metallic mast or tower without harmful effects. However, unless it is properly used, losses may be excessive, especially when the line is long in terms of wavelength. Coaxial line is often used to feed the driven element of v.h.f. arrays, but it can operate with maximum effectiveness in such cases only if current flow on the outside of the outer conductor is prevented by the installation of a detuning sleeve or "bazooka" at the point where the line connects to the array.

Examples of bazooka construction are shown in Fig. 9-4. At B is shown the sleeve type which is connected to the outer conductor at the bottom end and insulated from it at the antenna end.

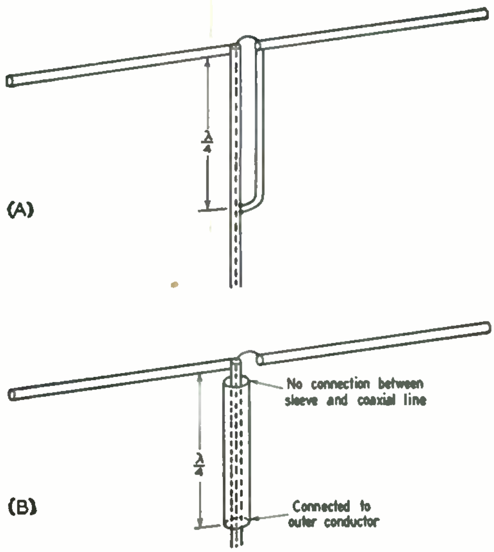


Fig. 9-4 — A "bazooka" line balancer is used to feed a balanced center-fed dipole with a coaxial line. In A it takes the form of a quarter-wave section of tubing the same size as the coaxial line. In B it is a metal sleeve connected to the outer conductor at the far end from the antenna.

A form which may be easier to use in most installations is shown at A. In this instance the quarter-wave section may be made from a piece of copper tubing of about the same diameter as the coaxial line, or it may even be another piece of the coaxial line with its outer and inner conductors connected together. In either case, the bazooka, being a quarter-wave line shorted at the far end, presents a high impedance at the antenna end, preventing a flow of current (and consequently radiation) down the outside of the outer conductor. Its length should be approximately a full quarter wavelength at the operating frequency.

ELEMENT LENGTHS

Since most arrays for v.h.f. use will be designed for wide frequency response, precise adjustment of the element lengths will not ordinarily be necessary. In fact, if very small changes in element length do have a considerable effect on the system's performance that, in itself, is evidence that the array will not be a particularly satisfactory one. The careful adjustment of element lengths and spacings to achieve a desired result, whether it be high front-to-back ratio, maximum forward gain, or any other objective, is a most interesting process, however, and many amateurs, especially those who are of an experimental turn of mind, will want to carry the process through with utmost care. But it should be emphasized that, if the array is properly designed, or the details have been worked out by careful experimentation, the array may be duplicated in other locations than the original with good results.

The arrays described in these pages do not represent all the designs which may be used successfully; they are simply models which have been tried in actual operation and have been duplicated satisfactorily under widely-varying conditions in many sections of the country. There is nothing mysterious about the principles upon which they operate, and the amateur who wants good antenna performance with a minimum of difficulty may use any of them with good results.

The Driven Element

Whenever a "half-wave antenna" is prescribed for v.h.f. use its length may be determined by the formula given below:

$$\textit{Length}\text{ (inches)} = \frac{5540}{\textit{Freq.}\text{ (Mc.)}}$$

This formula applies whether the dipole is to be used as the complete antenna system or as a driven element in a directional array. Table 9-I provides driven-element lengths for various frequencies in the 50- and 144-Mc. bands, based on the above formula. Ordinarily it is not desirable to tune an array for the edge of a v.h.f. band. Even if one expects to transmit only near the band edge, the array should be cut for a frequency well inside, in order to make it effective in reception over as much of the band as possible.

Parasitic Elements

The exact length of a reflector or director will depend upon the spacing between the elements. In general, however, with the spacings of 0.2 wavelength or more customarily used in v.h.f. arrays, the reflector and director in a v.h.f. system will be approximately 5 per cent longer and 5 per cent shorter, respectively, than the driven element. Where an additional director is used, as in a 4-element array, the forward director will be about 6 per cent shorter than the driven element. Parasitic-element lengths are also given in Table 9-I.

Phasing Sections

Where transmission-line lengths must be figured, as for phasing sections, "Q"-bar lengths, etc., the length of a half-wave section is determined from the formula:

$$\textit{Length}\text{ (inches)} = \frac{5760}{\textit{Freq.}\text{ (Mc.)}}$$

This applies only to air-dielectric lines. When Twin-Lead or coaxial-line sections are used the propagation factor of the line (see Chapter Three) must be applied to the above figure. Element spacings and stacking dimensions are full-length figures, a wavelength being twice the half-wave phasing-section figure, of course.

Antenna Systems for 50 Mc.

As the same basic principles apply to all antennas regardless of frequency, little discussion is given here of the various simple dipoles that may be used on any of the v.h.f. bands when nondirectional operation is desired. More complete information on half-wave antennas may be found in earlier chapters, and the only modification required for use of such standard half-wave systems on 50 Mc. is the reduction in element length (see Table 9-I) for increased diameter-length ratio at this frequency.

A Simple 2-Element Array

Though a 2-element array is very easy to build, it provides a worth-while improvement in working range over that obtained with a simple half-wave dipole. Such a 2-element array is shown in Fig. 9-5. Its design takes into account the small drop in the center impedance of a half-wave radiator when a parasitic element is placed a quarter wavelength away. This parasitic element may be

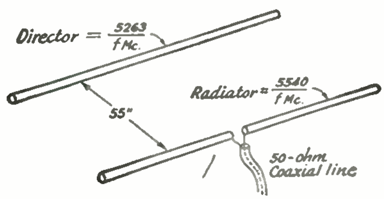


Fig. 9-5 — A simple 2-element array for 50 Mc. Though the parasitic element is shown as a director a reflector could be substituted.

either a director or reflector, though the former is shown in the sketch. Quite a good match is obtained when such an array is fed with 50-ohm coaxial line. For best results a bazooka (see section on matching devices) should be added at the feed point, though reasonably good results will be obtained with the array as shown in the sketch. It should provide approximately 5 db. gain over a dipole.

A 4-Element Array[1]

The importance of broad frequency response in any antenna designed for v.h.f. work cannot be overlooked. The disadvantage of all parasitic systems is that they tend to tune quite sharply, and thus are often effective over only a small portion of a given band. One way in which the response of a system can be broadened out is to increase the spacing between the parasitic elements to somewhat more than the 0.1 or 0.15 wavelength normally considered to provide optimum front-to-back ratio. Some broadening may also be obtained by making the directors slightly shorter and the reflector slightly longer than the optimum value. The folded dipole is useful as the radiator in such an array, as its over-all frequency response is somewhat broader than other types of driven elements.

[1] "World Above 50 Mc.," June, 1946, *QST*, page 60.

A 4-element array for 50 Mc. having an effective operating range of about 2 Mc. is shown in Figs. 9-6 and 9-8. It employs a folded dipole hav-

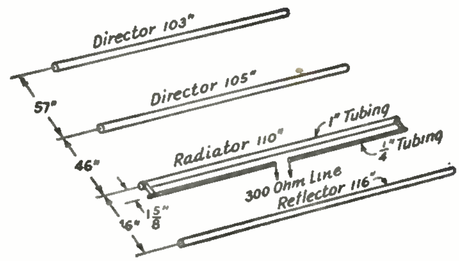


Fig. 9-6 — Dimensional drawing of the 4-element 50-Mc. array shown in Fig. 9-8. Element lengths and spacing were derived experimentally for maximum forward gain at 50.5 Mc.

ing nonuniform conductor size. Reflector and first director are spaced 0.2 wavelength from the driven element, while the forward director is spaced 0.25 wavelength. The spacing and element lengths given were derived experimentally, and are those that give optimum forward gain at the expense of some front-to-back ratio. As the latter quality is not of great value in 50-Mc. work, it can be neglected entirely in the tuning procedure for such an array.

The dimensions given in the sketch are for peak performance at 50.5 Mc. For higher frequencies, see Table 9-I. Peaked at 50.5 Mc., this array shows no appreciable change in performance from 50 to above 51 Mc. Above 51.5 Mc. the gain begins to drop, though it is still usable at 52 Mc. From 52.3 to 52.7 Mc. the array is no good at all, and at 53 Mc. the directivity reverses itself, giving a gain of 3 or 4 db. in the opposite direction, as the directors become long enough to act as reflectors. A slight broadening of the frequency response (at some expense of forward gain) may be obtained by adding to the reflector length and shortening the directors. For those interested in experimenting with element lengths a method of providing adjustable elements is shown in Fig. 9-7.

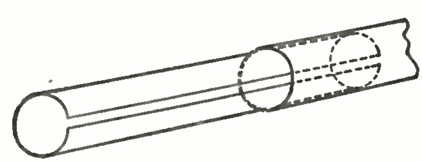


Fig. 9-7 — A simple method of providing for adjustment of element lengths. The insert is made of the same size tubing as the element, but slotted and compressed to permit insertion into the ends of the elements.

TABLE 9–I
Element Lengths and Spacings for 50- and 144-Mc. Antenna Systems, in Inches.

Freq. (Mc.)	*Driven Element*	*Re-flector*	*First Di-rector*	*Second Di-rector*	*Phasing Section*	*0.25 Wave-length*	*0.2 Wave-length*
50.5	110	116	105	103	114	57	46
51	108½	114	103	102	113	56½	45
52	106½	112	101	100	110½	55	44
53	104½	110	99½	98½	108½	54½	43½
145	38¼	40	36¼	36	39¾	20	16
146	38	40	36	35¾	39½	19¾	15¾
147	37¾	39½	35¾	35½	39¼	19½	15½

Dimensions for 50 Mc. are to nearest half inch; those for 144 Mc. to nearest quarter inch.

A Dual Array for 28 and 50 Mc. [2]

As many 50-Mc. enthusiasts also operate on 28 Mc., it is often desirable to stack arrays for the two bands on a common tower and rotating device. Such a dual array, combining a 4-element system for 50 Mc. with a 3-element array for 28 Mc., is shown in Fig. 9-8.

If space limitations make it absolutely necessary, the two arrays may be mounted with but a few inches separating them, but experience has shown that some effectiveness is sacrificed, particularly in the array for the higher frequency. A separation of at least three feet is recommended as the minimum for avoiding harmful interaction. In the example shown the separation is six feet, at which distance each array performs equally as well as it would if mounted alone.

In this dual array all-metal construction is employed, doing away with the use of insulators in mounting the elements. The booms are made of two pieces of 1-inch angle stock

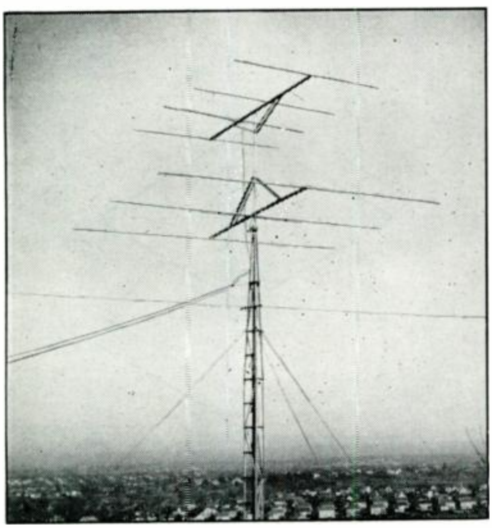


Fig. 9-8 — An example of stacking two arrays for different bands on a common support. All-metal construction is employed in this dual array for 10 and 6 meters.

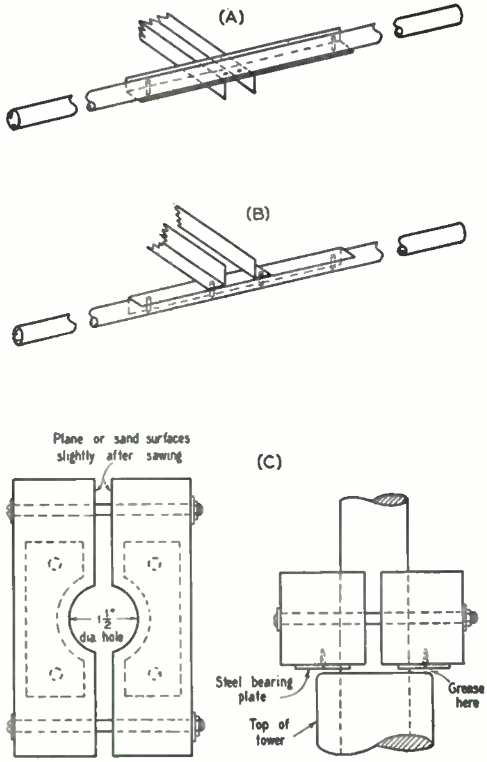


Fig. 9-9 — Detail sketches of portions of the dual array for 10 and 6 meters. A — The 50-Mc. boom is made of two pieces of angle stock mounted edge to edge to form a channel. Elements are cradled in another piece of angle stock. B — The two sides of the 10-meter boom are separated and mount on either side of the vertical support. The elements and their supporting crossarms are attached to the lower surface of the boom. C — The bearing for the array is made from a block of wood, drilled to the pipe size, and then sawed lengthwise. It is faced with two steel plates where it rests on the top of the tower.

(24ST aluminum), with supporting braces of the same material. The method of assembling the booms and mounting the elements is shown in Fig. 9-9. The booms are 150 inches and 160 inches in length for the 6- and 10-meter arrays respectively. To prevent swaying of the 10-meter elements, they are braced with guy wires, which are broken up with small insulators. These sway-brace wires are attached to the elements at approximately the midpoint between the boom and the outer end, and are brought up to the vertical support at the point of attachment of the horizontal fore-and-aft braces.

The 50-Mc. portion of the array is similar in element length and spacing to the 4-element array already described. The element spacing

[2] "An All-Metal Array for 6 and 10," July, 1947, *QST*, page 52.

for the 10-meter array is 0.2 wavelength, or 80 inches. The driven element is 198 inches long, the director 188 inches, and the reflector 208 inches. It is fed by means of a "T" match and a 300-ohm line. These dimensions give quite uniform performance and low standing-wave ratio over the range from 28 to 29.1 Mc., and the array will take power and show appreciable gain over a half wave from 27.2 to 29.7 Mc.

Stacked and Phased Arrays

A worth-while gain and lowered radiation angle result from the stacking of two or more parasitic arrays one above the other at intervals of one-half wavelength and feeding them in phase. Stacking two arrays results in a theoretical gain of about 4 db. over a single array, but the lowered radiation angle may make possible a greater gain in actual performance, particularly at times when the radiation angle is a critical factor, as in borderline DX hops.

Both vertical and horizontal stacking is an even more advantageous arrangement. An "H" array composed of four half-wave elements, with reflectors added, provides a gain somewhat in excess of that obtainable with a 4-element parasitic array, yet it is more tolerant as to frequency response.

When large numbers of elements are used in stacked and phased arrays, it is desirable to feed the system at its electrical center, as the current distribution is apt to be more uniform when this is done. When the point of connection of the feedline is midway between two elements (or two sets of elements) the impedance at the feedpoint is half that of a single element or set of elements. Thus it is often possible to make the phasing section also serve as a double "Q" section, as is done in the array shown in Fig. 9-10. If it is necessary to feed a stacked array at the bottom of the system, the feed impedance at the bottom element will be approximately the same as for a single element, provided the other elements are connected at half-wave intervals. The phasing sections are all transposed in such an array, whereas no transposition is made between two elements (or sets of elements) when the feeder is connected at the midpoint between the two.

The use of reflectors with a phased array such as the "H" may be handled in two ways. They may be made the same length as the driven elements, and spaced one-quarter wavelength behind them, in which case there is only a slight drop in feed impedance below the value of the system without reflectors. Slightly more gain may be obtained by making the reflectors longer (see Table 9-I) and placing them at 0.15 or 0.2 wavelength in back of the driven elements. There is a greater drop in feed impedance in this case. More information on phased and stacked arrays may be found in the sections devoted to antennas for 144 Mc.

Working with Long-Wire Arrays on 50 and 144 Mc.

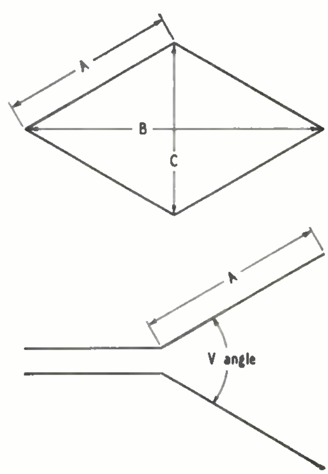
Where long-wire antenna systems such as the V or the rhombic are available they can usually be used on 50 or even 144 Mc. with good results, even though they were erected with lower-frequency operation in mind. The included angles in such arrays will not be optimum for v.h.f., but the arrays will be so large, in terms of v.h.f. wavelengths, that they will work well, particularly if the feeder systems are not too long. They will show little frequency discrimination over an entire v.h.f. band.

Long-wire arrays may be constructed according to the principles given in Chapter Five, designing them specifically for v.h.f. use. In such instances an effective rhombic array assumes proportions that make it usable in many amateur locations where a similar array for 7 or 14 Mc. would be out of the question because of its great size. By proper choice of leg lengths a V or rhombic can be made to work well on both 28 and 50 Mc., making it a highly useful system where the necessary space is available for its erection. Several examples are

TABLE 9–II

Dimensions of V and Rhombic Antennas for V.H.F. Use. Columns 1 and 2 Are for V Designs. For Rhombics Use 1, 3 and 4.

Freq. (Mc.)	*Side Length "A" in Feet*	*V Angle*	*Over-all Length "B" in Feet*	*Width "C" in Feet*
50.5	58	60°	96.5	65.5
145	58	35°	109	39
28.7	68	70°	101.5	84
50.5	68	55°	106.5	70.6
145	68	35°	129	41
50.5	106	42°	192.5	91.5
145	106	35°	205	47.5
28.7	136	52°	237.5	133
50.5	136	37°	252.5	102



given in Table 9-II. Other combinations may be worked out using the formula

$$\textit{Length} \text{ (feet)} = \frac{492\,(N - 0.05)}{\textit{Freq.} \text{ (Mc.)}}$$

where N is the number of *half waves* on the leg.

The above formula need be used only where the leg length is short in terms of wavelength. For longer dimensions the standard half-wave formula may be used

$$\textit{Length} \text{ (feet)} = \frac{492\,N}{\textit{Freq.} \text{ (Mc.)}}$$

Long-wire systems for combining operation on 50 and 144 Mc. are even more attractive as to size. Because of the nearness to third-harmonic relationship which exists between these two bands, the same matching section and feedline may be used to feed a terminated rhombic for both bands with a flat line. Since a "Q" section can be any odd multiple of a quarter wavelength, the matching section for a two-band v.h.f. rhombic can be a quarter wavelength long at 50 Mc., in which case it will be approximately three quarter waves long at 144 Mc. The feed impedance of a terminated rhombic is about 800 ohms; thus a 490-ohm "Q" section is required to match this impedance to a 300-ohm line. Such a matching section could be made of No. 14 wire spaced 1¾ inches, about 53 inches long, as a compromise for the two bands. The array could be fed directly with a 600-ohm line, without appreciable mismatch. Preferably such a line would be of small wire, in order to keep the spacing to relatively small dimensions. See Fig. 3-24 for wire sizes and line impedances.

Laying out a rhombic antenna for the v.h.f. bands is somewhat less complicated than for lower frequencies, because it is usually possible to have the v.h.f. array high enough (in terms of wavelength) so that the effect of ground is a minor consideration. The dimensions given in Table 9-II are based on the assumption that the lowest possible radiation angle is desired, in which case one side should be a half wave longer than half the over-all length. Using the terms of Table 9-II:

$$A = \frac{B}{2} + \frac{480}{\textit{Freq.} \text{ (Mc.)}}$$

The shape of a multiband V or rhombic may be set up according to Table 9-II with its width, C, at the optimum value for the band where highest efficiency is desired. It will be noted that the larger the array the less difference there is in the included angles for adjacent bands. In other words, the larger the array the better will be its capabilities for multiband operation.

Antenna Systems for 144 Mc.

Though the arrays already described may, of course, be used for 144 as well as 50 Mc., the smaller physical size of directional antennas for the higher band makes it possible to use multielement systems which would be excessively cumbersome for 50-Mc. work. Also, the still considerable use of vertical polarization on 144 Mc. requires a somewhat different technique than is embodied in the horizontal arrays previously described. With either polarization, the stacking of elements vertically lowers the radiation angle and extends the operating range, and such stacking is easily incorporated in 144-Mc. systems. Arrays using several elements fed in phase, with or without reflectors, provide a broader frequency response, for a given gain, than those having a single driven element with several parasitic elements to build up the gain. Broad response is even more desirable on 144 than on 50 Mc., at present occupancy levels on the two bands.

Plane-reflector systems are usable on 144 Mc., their size being within reason at this frequency. They provide excellent front-to-back ratio and broad frequency characteristics, and both vertical and horizontal polarization may be combined in one array, by placing two oppositely-polarized sets of elements on either side of a plane reflector. Arrays for two v.h.f. bands may be combined in one by a similar approach. The screen need not be a solid sheet of metal, or even a close mesh. Chicken wire, or even sets of rods or wires arranged in back of the driven elements, will work almost equally well.

A 6-Element Array

In designing directional arrays having more than one driven element it is advisable to arrange for feeding the array at a central point. A simple 6-element array of high performance,

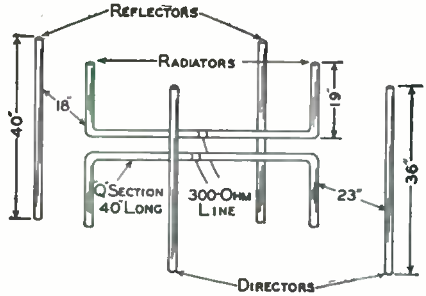


Fig. 9-10 — A double "Q" array for 144 Mc. The horizontal portion of the "H" acts as a "Q" section, matching the antenna impedance to the 300-ohm line. The pattern of this array is such that it works well in either vertical or horizontal positions.

incorporating this principle, is shown in Fig. 9-10. All the elements may be made of soft-copper tubing, ¼ inch in diameter. They may, of course, be made of any material of similar size that can be bent at right angles without breaking. The driven elements are comprised of two pieces which are bent into two "U"-shaped sections and arranged in the form of a half-wave "H." The horizontal portion of the "H" is then a double quarter-wave "Q" section, matching the impedance of the two radiators to that of the feedline. With the wide spacing used, the position of the parasitic elements is not particularly critical, except as it affects the impedance of the system, and the spacing of the elements may be varied to provide the best match. The spacing of the horizontal section may be varied for the same purpose. With the dimensions given, a spacing of one inch between centers is about right for feeding with a 300-ohm line. The pattern of this array is similar in both horizontal and vertical planes.

A 12-Element Horizontal Array

A 12-element array of excellent mechanical design and proven performance is shown in Fig. 9-11. It was in use at W3GV when the 660-mile two-way 144-Mc. record was established with WØWGZ, and has been responsible for many long-distance contacts before and since. It consists of three bays of four elements each, stacked one-half wavelength apart. The feed system is a "T" match and 50-ohm coaxial line, with a bazooka incorporated at the feed point.

The array is entirely of metal construction. Details of the feed system are given in Fig. 9-12. Element lengths are similar for each bay, except for the driven element in the center section. Reflectors are 40⅞ inches, first directors 38¹⁄₁₆ inches, and second directors 35⅞ inches. In the top and bottom bays the driven elements are 38¾ inches, while the center-bay driven element is 35¾ inches in length. The dimensions given are for maximum effectiveness at 145 Mc. As the

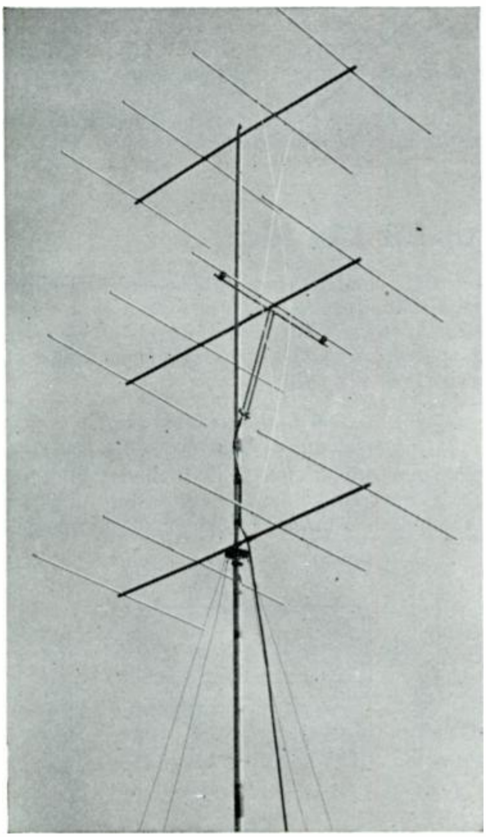


Fig. 9-11 — The 12-element 144-Mc. array at W3GV uses 3 bays of 4 elements each, stacked a half wave apart. Phasing sections are crossed No. 8 wires. The array is fed at the middle bay, by means of a "T" match, a bazooka, and coaxial line.

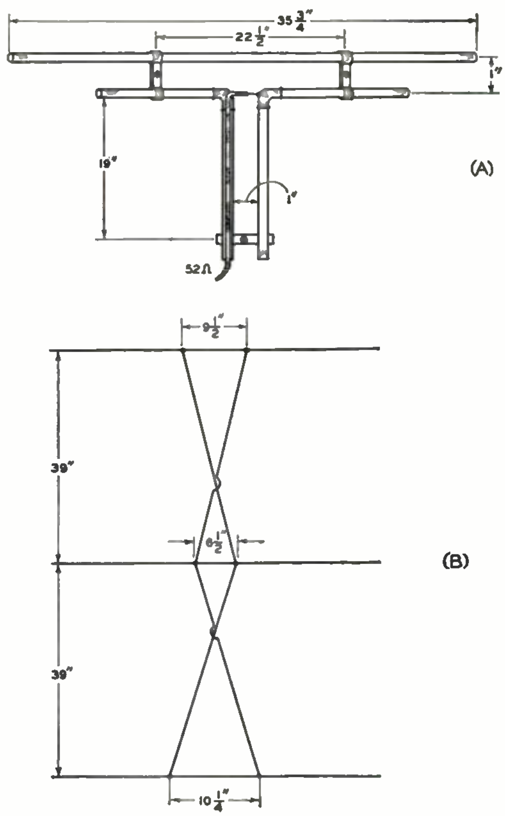


Fig. 9-12 — Details of the feed system used in the W3GV 2-meter array. The method of feeding the driven element in the middle bay is shown at A. Coaxial line is used, necessitating the inclusion of a bazooka at the feed point. All elements are ⅜-inch tubing. The phasing arrangement is shown at B.

parasitic-element spacing is only 0.2 wavelength, frequency response is relatively sharp, the array being most effective from 144 to 146 Mc. This necessary sharpness is the principal failing inherent in any array having a large number of parasitic elements in proportion to driven members. Gain at 145 Mc. has been measured at 15 db. over a half-wave dipole.

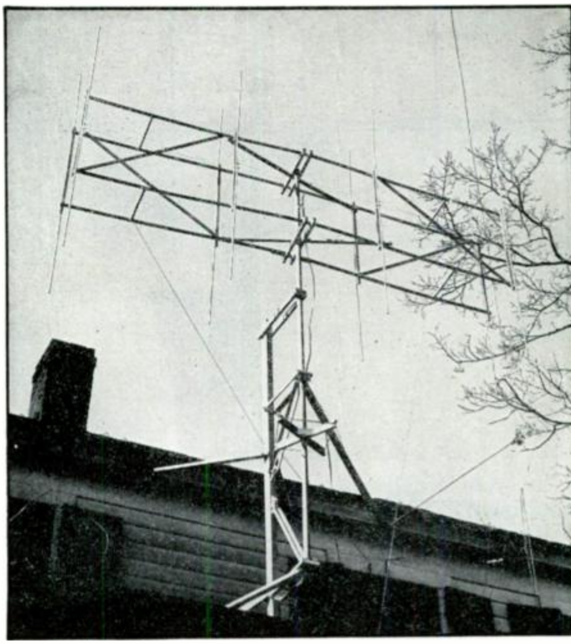


Fig. 9-14 — A 16-element array for 144 Mc., showing the supporting structure and "rotating mechanism." Sash cord wrapped three times around the crisscross pulley permits 360-degree rotation. The main vertical member is a 10-foot rug pole, 1½ inches in diameter.

8 Half Waves in Phase, with Reflectors [3]

By using a curtain of eight half-wave elements, arranged as shown in Fig. 9-13, backed up by reflectors, a gain of as much as 15 db. can be realized, but with a less-critical frequency response than in the 12-element array just described. The 16-element array shown in Fig. 9-14 has been duplicated by scores of 144-Mc. operators with excellent results. It is effective with either horizontal or vertical polarization. It was in use, as a horizontal array, at WØWGZ when the record-breaking contact with W3GV was made.

The cumbersome nature of such an array for any lower frequency would make it out of the question for most amateurs, but for 144 Mc. it is neither difficult nor expensive to construct. The outside dimensions of the array are only 1½ by 7 by 10 feet, and with proper design the supporting frame can be made quite light. Though the

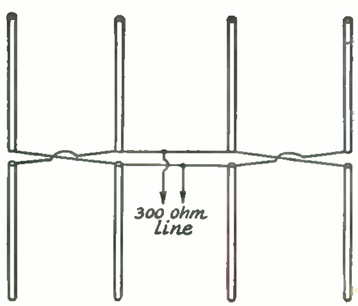


Fig. 9-13 — Schematic of the driven-element portion of the 16-element array of Fig. 9-14.

array shown in Fig. 9-14 has its elements mounted on stand-off insulators these may be done away with by incorporating the method of mounting shown in Fig. 9-17. Here the elements are forced through holes in the wooden supporting members, the latter being positioned so that they are at the midpoint (low-voltage point) of the elements. The element material should be fairly stiff when this method of construction is used in a 144-array, otherwise the phasing wires will tend to pull the elements out of alignment at the middle of the array. The insulators into which the elements are inserted should be made of some high-grade dielectric material, such as polystyrene. If large-diameter tubing is used these insulators may take the form of polystyrene rods turned down in a lathe to fit snugly inside the ends of the elements.

Phasing sections are constructed of No. 16 wire spaced about 1½ inches. The 300-ohm feeder is connected at the center of the system, at which point quite a good match is provided without the necessity for any special matching devices. The feed impedance is somewhat under 300 ohms, actually, and a "Q" matching section may be employed if desired. This should be made conveniently adjustable (see Fig. 9-16) so that the spacing of the "Q" bars can be varied for lowest standing-wave ratio on the line. Unless the feed-line is to be very long, however, the losses will not be excessive with the direct connection, as shown in Fig. 9-13.

The center section of the array described above may be used without the outside 8 elements, if space is limited and a simpler array of good performance is desired. Such an "H"-type array, with reflectors, may also be fed at the center point with 300-ohm line (Twin-Lead) without the necessity for additional impedance-matching arrangements.

[3] "World Above 50 Mc.," May, 1946, *QST*, page 56.

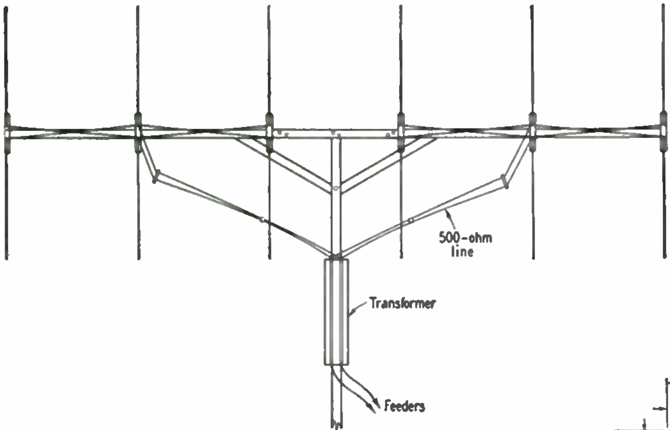
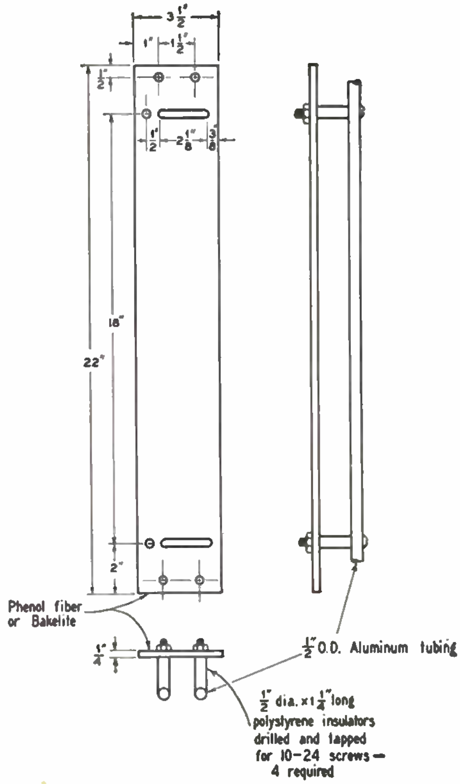


Fig. 9-15 — Detail drawing of the W2NLY 24-element array, showing the method of feed employed to insure uniform current distribution. Only the driven elements are shown. Reflectors are spaced one-quarter wavelength behind them.

Fig. 9-16 — An adjustable "Q" section for matching the impedance of the W2NLY array to a wide variety of feedlines.



A 24-Element Array[4]

An array having a gain of as much as 17 db. is used by W2NLY on 144 Mc. It consists of 12 half-wave elements fed in phase, backed up by 12 reflectors spaced a quarter wavelength behind the driven elements. Originally this array was fed at its center, but the unequal current distribution resulted in several large minor lobes, limiting the array's effectiveness. Substituting the split feed shown in Fig. 9-15 corrected this trouble.

With this arrangement the array is divided into two portions, and phasing sections, each one wavelength long, are connected to the center pair of elements in each half of the system. These sections are then connected in parallel and the common point fed with a transmission line, through a suitable matching device, in this case a parallel-pipe "Q" section, details of which are found in Fig. 9-16. The phasing sections are made of No. 14 wire, with a spacing of 2 inches, or about 500 ohms impedance. They are connected in parallel at the feed point, halving the impedance to which the line must be matched. A "Q" section such as that shown in Fig. 9-16 is adjustable over a sufficient range to match this 250-ohm feed impedance to lines of 52 to 300 ohms impedance.

[4] "An Antenna That Multiplies by 50," Kmosko, Sept., 1947, *QST*, page 50.

Arrays for 220 and 420 Mc.

In general the antenna problems at 220 and 420 Mc. are not greatly different than at 144 Mc., and the arrays already described may be adapted to use on these higher frequencies by suitable modification of their dimensions. The use of high-gain arrays becomes increasingly important as the frequency is raised, and practically all amateur work on these higher bands is done with directional systems. The smaller physical size of arrays for 220 or 420 Mc. makes possible the use of quite high-gain systems in almost any location.

A 32-Element Array[5]

A 32-element array for either of these bands is not excessively large. A method of building and feeding such an array, suggested by W1CTW, is shown in Fig. 9-17. Dimensions for use in this and other arrays for 220 and 420 Mc. will be found in Table 9-III. The elements are mounted without the use of insulators, the wooden frame-

[5] "Let's Start Right on 1¼," Hadlock, Dec., 1947, *QST*, page 22.

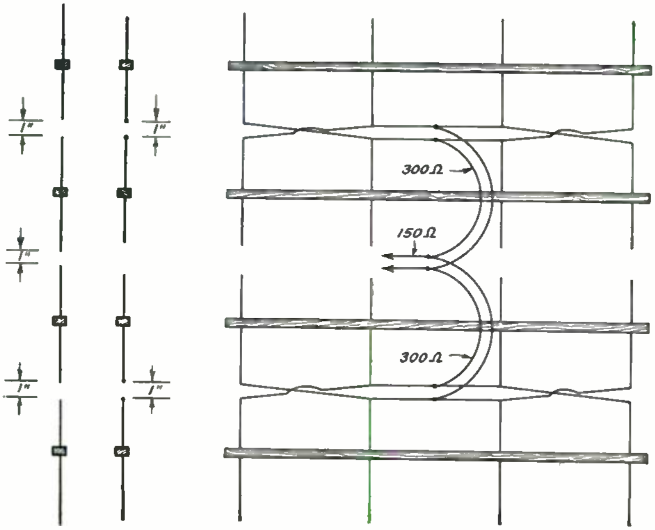


Fig. 9-17 — Design for a 32-element 220-Mc. array suggested by W1CTW. Elements are supported without the use of stand-off insulators, by designing the wooden support so that the elements are mounted at the low-voltage point. Dimensions for this array, for the 220- and 420-Mc. bands, may be found in Table 9-III.

work providing support at the low-voltage points throughout the system.

To insure balanced current distribution the system is fed as if it were two separate 16-element arrays. The 300-ohm lines feeding the two sections of the array are joined together and the whole

TABLE 9–III

Element Lengths and Spacings for 220- and 420-Mc. Antenna Systems, in Inches.

Freq. (Mc.)	*Driven Element*	*Re-flector*	*First Di-rector*	*Second Di-rector*	*Phasing Section*	*0.25 Wave-length*	*0.2 Wave-length*
221	25⅛	26⅜	23⅞	23⅝	26⅛	13	10½
222.5	24⅞	26⅛	23⅝	23⅜	25⅞	13	10⅜
224	24¾	26	23½	23¼	25¾	12⅞	10¼
425	13	13⅝	12⅜	12¼	13½	6¾	5½
435	12¾	13⅜	12⅛	12	13¼	6⅝	5⅜
445	12½	13⅛	11⅞	11¾	13	6½	5¼

Dimensions are given to the nearest ⅛ inch.

system is fed with a 150-ohm line. If some other form of transmission line is desired, a "Q" section may be used at the junction of the two lines. The two sections of 300-ohm line should be 0.82 wavelength, a figure which takes the propagation factor of polyethylene-insulated 300-ohm line into account. If open-wire line is used the sections should be a full wavelength long.

The Screen Reflector

A flat metallic screen or grid may be used to replace the parasitic elements in an array such as that shown in Fig. 9-17. For best results the area of the screen should be sufficient to project beyond the area covered by the driven elements by at least a quarter wavelength. Such an ideal reflector will have approximately similar performance to that obtainable with parasitic reflectors, but with a broader frequency response and improved front-to-back ratio.

Plane reflectors may be used with any number of driven elements, and a gain of about 7 db. can be obtained with a plane reflector and a single driven element. Optimum spacing, in this case, is about 0.1 wavelength, with good performance being obtained out to about 0.3 wavelength. The 180° curve in Fig. 9-19 shows the variation in center impedance of a dipole used with a plane reflector, at spacings from 0.1 to 0.5 wavelength.

Corner Reflectors[6]

When a single driven element is employed, the plane reflector may be bent to form an angle, giving an improvement in the radiation pattern and gain. The corner reflector, Fig. 9-19, becomes a practical size for 220- or 420-Mc. arrays, and a high degree of performance can be attained with a relatively small array. Corner angles of 60 or 90 degrees provide about optimum performance, though an angle of 45 degrees may be used, if the side lengths are extended sufficiently. The dipole is placed on a line bisecting

[6] "The Square-Corner Reflector for Ultra High Frequencies," Kraus, Nov., 1940, *QST*, page 18. "The Corner-Reflector Antenna," Kraus, *Proc. I.R.E.*, Nov., 1940, page 513.

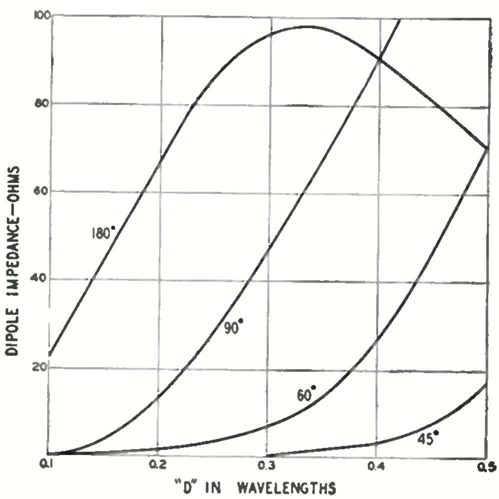


Fig. 9-18 — Feed impedance of the driven element in a corner-reflector array, for various corner angles of 180 (flat sheet), 90, 60 and 45 degrees.

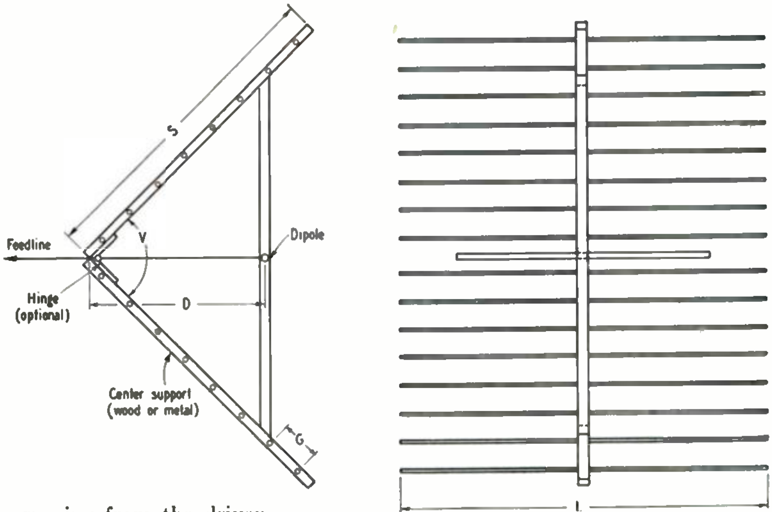


Fig. 9-19—The square-corner array using a grid-type reflector. Elements are stiff wire or tubing. The frame may be either metal or wood. Suitable dimensions may be found in Table 9-IV.

the corner angle. The spacing from the driven element to the corner may be anything from 0.25 to 0.7 wavelength for a 90-degree corner, 0.35 to 0.75 wavelength for a 60-degree corner, and 0.5 to 1.0 wavelength for a 45-degree corner angle.

The center impedance of the driven element in a corner array varies with the corner angle and the position of the radiator with respect to the vertex. From Fig. 9-18 it may be seen that, with a 90-degree corner angle, the center impedance of the driven element approximates the value for a dipole alone, at spacings around 0.35 wavelength, rising to about 150 ohms at 0.5 wavelength. With a 60-degree corner the impedance is about 70 ohms at a spacing of 0.5 wavelength.

Generally speaking, the dimensions of a corner-reflector array are not at all critical, and the frequency characteristics are much better than for a simple parasitic array of anything like the same gain. A gain of about 12 db. can be obtained with a 60-degree corner reflector, whose sides are about two wavelengths long. Approximately 10 db. gain can be realized from 60- or 90-degree reflectors with a side length of one wavelength. The reflector may be made of sheet metal, wire netting, or a series of spines, there being very little difference in performance, provided the spacing of the spines is kept 0.06 wavelength. This spacing may even be increased to 0.1 wavelength with only a very slight drop in effectiveness. They should be about 0.6 wavelength long, though the minimum dimension is the important one.

A single corner reflector may be used for several bands, provided the spacing of the spines is set up for the highest frequency, and the side and spine lengths for the lowest. Separate dipoles should be used, of course, each being set at the optimum distance from the vertex for the band in question. The corner array is equally effective for either vertical or horizontal polarization. As suggested by W8JK, who developed the corner array for amateur use, the design may be adapted for portable use, by hinging the sides at the vertex.

For increased gain, the radiator may be two half waves fed in phase, with a corresponding increase in reflector size, or two or more corner-reflector arrays may be stacked vertically or horizontally and fed in phase. The dimensions for corner arrays are available in convenient form in Table 9-IV.

The corner reflector is probably the most effective means of developing high gain when only a single dipole is used in the driven portion of the array. The only superior arrangement would involve the use of parabolic reflectors of very large dimensions.

TABLE 9-IV

Dimensions of Corner-Reflector Arrays for 144, 220 and 420 Mc.

Band (Mc.)	*Side Length "S" (Inches)*	*Dipole to Vertex "D" (Inches)*	*Reflector Length "L" (Inches)*	*Reflector Spacing "G" (Inches)*	*Corner Angle "V" (Degrees)*	*Feed Impedance (Ohms)*
144*	65	27.5	48	7¾	90	70
144	80	40	48	4	90	150
220*	42	18	30	5	90	70
220	52	25	30	3	90	150
220	100	25	30	screen	60	70
420	27	8¾	16¼	2⅝	90	70
420	54	13½	16¼	screen	60	70

* Side length and number of reflector elements somewhat below optimum—slight reduction in gain.

The large designs for 220 and 420 Mc. have a gain in excess of 12 db.—intermediate sizes (1 wavelength on each side) about 10 db.

Mobile and Portable Antennas

The most convenient type of antenna for mobile v.h.f. work is the quarter-wave vertical radiator, fed with 50-ohm coaxial line. The antenna may be a telescoping whip if it is to be used for more than one band, or, in the case of 144 Mc. and higher bands, it may be simply a small rod cut to length. It may be mounted in any of several positions on the car, although the preferred spot is on the car top. This ideal may be possible only

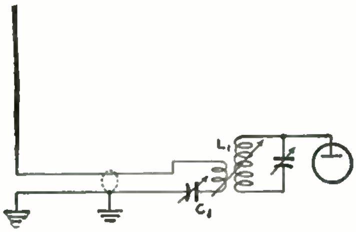


Fig. 9-20 — Method of feeding quarter-wave mobile whip antennas with coaxial line. C_1 should have a maximum capacity of 75 μμfd. for 28- and 50-Mc. work. L_1 is an adjustable link.

for the higher bands, however. Where the whip type of antenna is mounted in any position below the top of the car, marked directional characteristics are always in evidence; thus it is nearly always desirable to use the same antenna for transmission and reception.

When coaxial feed is used in the mobile installation the coupling circuit should be arranged as shown in Fig. 9-20. The coupling should be adjustable, the optimum setting of C_1 being that which allows the loosest coupling. Inclusion of such a variable tuned coupling arrangement will make it possible to load the antenna at various lengths, a desirable feature if multi-band operation is contemplated.

Horizontal Polarization — The "Halo"[7]

Obviously the whip antenna is at a disadvantage in work with fixed stations using horizontal polarization, except possibly in contacts where ionospheric propagation is involved. A conventional form of horizontal array is applicable to mobile operation on 144 Mc. and higher, but a half-wave antenna for 50 Mc. becomes somewhat cumbersome. A solution was found by W1MUX in the form of a circular folded dipole, or "halo," shown in Fig. 9-21. It is nothing more than a folded dipole bent into a circle, with the ends capacitively loaded to reduce the over-all length. When such a dipole is mounted above the top of the car an almost perfectly-circular radiation pattern results, and the performance is a vast improvement over the vertical whip, in working with horizontally-polarized stations.

For 50-Mc. operation the antenna is a 20-inch diameter circle composed of ⅞-inch and ⅜-inch tubing, 2½ inches apart, center to center. The condenser plates can be any convenient shape, with an average diameter of about 5 inches. Spacing of these plates is critical, and they should be mounted securely. Details of the condenser assembly, the dipole, and the mounting arrangement are shown in Fig. 9-22. The spacing

[7] "A 'Halo' for 6 Meters," Stites, Oct., 1947, *QST*, page 24.

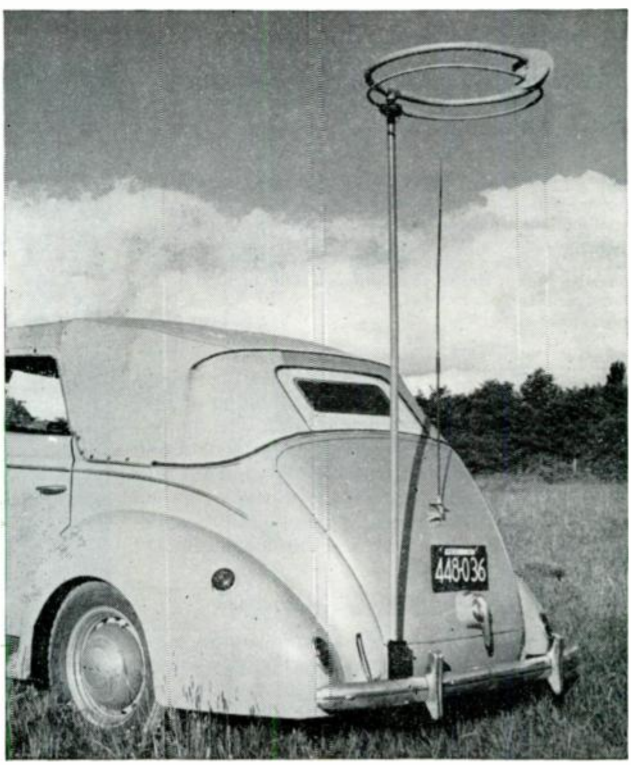


Fig. 9-21 — The "halo" antenna, as used by W1MUX for nondirectional horizontal polarization in 50-Mc. mobile work.

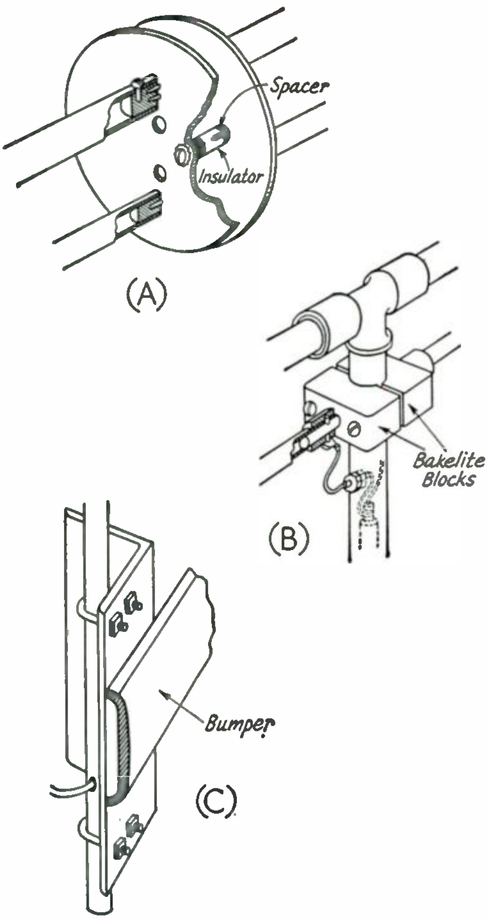


Fig. 9-22 — **Details of the various assemblies used in the halo-antenna installation. A is a cutaway view of the end-loading capacitor assembly. B shows the method of attaching the coaxial feedline and insulating the center of the fed section of the folded dipole. C is the assembly used to mount the vertical support. The angle plate is welded to the car bumper.**

of the condenser plates should be set at the value which resonates the system at the operating frequency. It may be adjusted by adding or removing washers, as seen in Fig. 9-22. The dipole dimensions given result in a feed impedance of 58 ohms, providing a good match for 50-ohm coaxial cable. The construction of the dipole may be modified for other tubing dimensions or cable impedances by referring to the folded-dipole nomogram, Fig. 3-51.

A Collapsible Array for 50 Mc.

The best antenna possible for operation under mobile conditions is not particularly effective, as compared with antenna systems normally used in fixed-station work. To make the most of the fine opportunities for DX work afforded by countless high-altitude locations which are accessible by car, it is helpful to have some sort of collapsible antenna array which can be assembled "on the spot." Even a simple array like the one shown in Figs. 9-23 and 9-24 will effect a great improvement in the operating range of the low-powered gear normally used for mobile operation. This one is designed for 50-Mc. use, but similar arrangements can be made for operation on other frequencies.

The array shown is a 2-element system, comprised of a radiator which is fed with coaxial line by means of a "T" match, and a reflector which is spaced 0.15 wavelength in back of the driven element. It is made entirely of ¾-inch dural tubing, except for the vertical support, which is 1-inch tubing of the same material. A suggested method of mounting is shown in Fig. 9-23. A short length of 1 × 2-inch or larger wood is bolted to the car bumper. A piece of ¾-inch dural tubing is bolted to this upright, and the 1-inch vertical section of the array slips over the top of the ¾-inch section. The array is turned by means of ropes attached to the reflector element. Height of the array may be increased over that shown by using a longer wooden support, in which case it is desirable to use a 2 × 2 for greater strength. An anchoring pin made from a spike inserted in the bottom end of the wooden support is helpful to prevent tilting of the array. With such a device

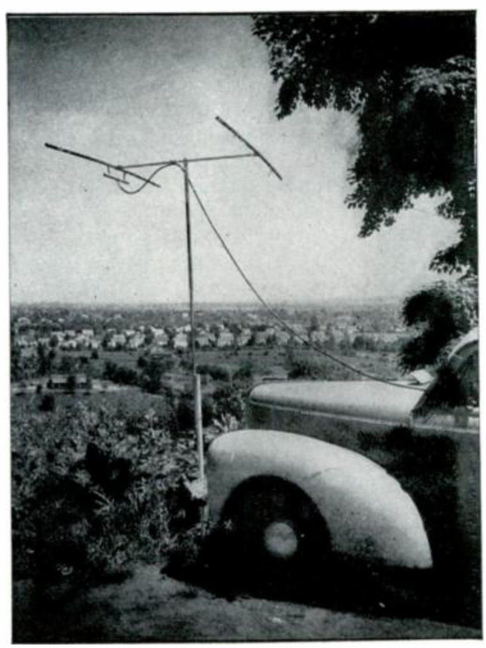


Fig. 9-23 — **A collapsible array for 50-Mc. portable use.**

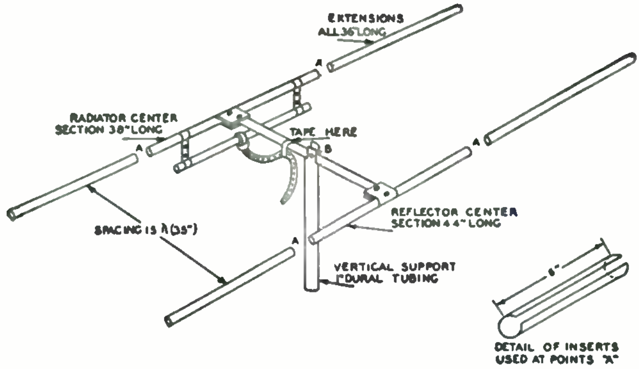


Fig. 9-24 — **The collapsible array for 50 Mc. is made of ¾-inch duralumin tubing, except for the vertical support, which is 1-inch. For carrying purposes it is taken apart at points *A* and *B*, inserts of slotted dural tubing being used at points *A* to hold the sections together. All extensions are the same length, the difference in element length being made up in the center sections.**

embedded in the ground, the whole assembly will remain rigid, which is helpful in the high winds usually encountered in mountain-top locations. Portability is provided by making the elements in three sections, with the end sections all the same length. The center section of the radiator is 6 inches shorter than that of the reflector.

The fed section of the "T" matching device is composed of two pieces of ¾-inch dural tubing about 14 inches long. The two sections are held together mechanically, but insulated electrically, by a piece of polystyrene rod which is turned down just enough to make a tight fit in the tubing. The inner and outer conductors of the coaxial line are fastened to the two inside ends of the matching section. Clips made of spring bronze are used for connection between the radiator and the "T." The position of these should be adjusted for maximum loading and minimum standing-wave ratio on the line.

This antenna system may be used as a dipole on 29 Mc. by plugging the reflector sections into the driven element, thus bringing its over-all length to approximately that of a half wave for the high end of the 10-meter band.

Miscellaneous V.H.F. and U.H.F. Antennas

The Coaxial Antenna

End-fed vertical radiators such as the "J" are relatively ineffective because of the tendency of the transmission line to radiate. This condition becomes worse as the frequency increases, becoming quite bad in the v.h.f. range. Radiation from the line tends to combine with that from the antenna itself in such a way as to raise the radiation angle. To eliminate this difficulty the coaxial antenna (Fig. 9-25) was developed. It has enjoyed wide acceptance in applications where a nondirectional vertical radiator is required.

The center conductor of a 70-ohm concentric transmission line is extended one-quarter wave beyond the end of the line, to act as the upper half of a half-wave antenna. The lower half is provided by the quarter-wave sleeve, the upper end of which is connected to the outer conductor of the concentric line. The sleeve acts as a shield about the transmission line and very little current is induced on the outside of the line by the antenna field. The line is nonresonant, since its characteristic impedance is the same as the center impedance of the half-wave antenna. The sleeve may be made of copper or brass tubing of suitable diameter to clear the transmission line. The coaxial antenna is somewhat difficult to construct, but is superior to simpler systems in its performance at low radiation angles.

Cylindrical and Cone Antennas

An ordinary antenna made of wire is purely resistive only over a very small frequency range. Its *Q*, and therefore its selectivity, is sufficient to limit its optimum performance to a very narrow frequency range, and readjustment of length will be required to maintain efficient operation over the width of an amateur band. A properly-designed wide-band system will, however, exhibit nearly constant performance over a considerable band of frequencies.

The simplest method of obtaining broad-band char-

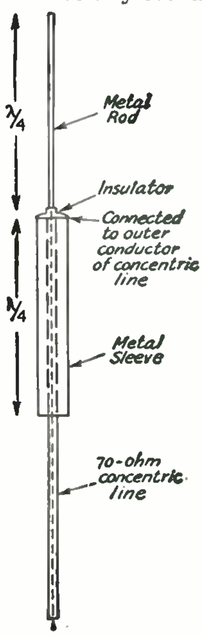


Fig. 9-25 — **The coaxial antenna. The center conductor of the 70-ohm coaxial line is connected to the rod which forms the upper portion of the antenna, and the outer conductor to the sleeve.**

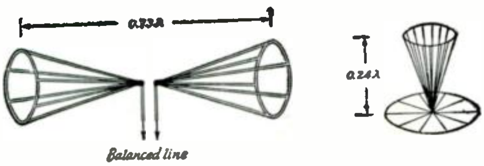


Fig. 9-26 — **Conical broadband antennas have relatively constant impedance over a wide frequency range. The three-quarter wavelength dipole at the left and the quarter-wave vertical with ground plane at the right have the same input impedance — approximately 65 ohms. Sheet-metal or spine-type construction may be used.**

acteristics in the v.h.f. range is the use of what might be termed a "cylindrical antenna" — a conventional doublet, but with its elements made of large-diameter tubing. The larger elements have lower Q, and consequently broader frequency response. As the diameter is increased the antenna must be made shorter than a thin-wire antenna resonating at the same frequency.

From the cylindrical antenna various specialized forms of broadly-resonant radiators have been evolved, including the ellipsoid, spheroid, cone, diamond and double diamond. Of these, the conical antenna is perhaps the most interesting. With large angles of revolution the characteristic impedance can be reduced to a very low value suitable for extremely wide-band operation. The cone may be made up either of sheet metal or of multiple wire spines, as in Fig. 9-26.

The Fan Antenna

A simple broadband antenna may be made in the form of two flat fan-shaped sheets of metal, as shown in Fig. 9-27. Such a fan antenna will have a sufficiently-broad frequency response to permit its use on two adjacent amateur v.h.f. bands with fairly-good results. The dimensions in Fig. 9-27 are for operation on 220 and 420 Mc., with a 300-ohm feedline. A plane reflector may be used with such a dipole, without a great change in its frequency characteristics. This might take the form of a curtain of wire netting, or even a sheet

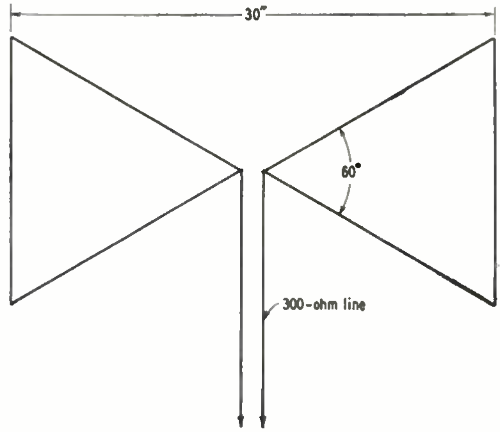


Fig. 9-27 — **A fan antenna for operation on 220 and 420 Mc. A plane reflector may be added for gain and directivity.**

of plywood or cardboard coated with a metallic paint, if the array is to be used indoors. The spacing would be about 10 inches, though this dimension would have to be varied for maximum effectiveness on both bands, the optimum dimension being about 0.3 wavelength. The reflector could be any size larger than about 30 by 50 inches.

Antennas for 160 Meters

With respect to sky-wave transmission, the requirements that the antenna system must meet on 160 meters do not differ materially from those which hold on the high-frequency bands. Of course, waves entering the ionosphere even vertically are reflected back to earth so that there is no such phenomenon as skip distance on these frequencies. However, it is still true that to cover the greatest possible distance the waves must enter the ionosphere at low angles. Although a given distance may be covered by multiple hops when the radiation angle is high, there will be less absorption, and hence the signal strength will be greater, at the same point when the wave reaches it by only one hop.

On the "160-meter" band the ground wave assumes considerable importance for transmission over short distances. The useful range of the ground wave will depend upon the transmitter power, the background noise at the receiver, and the type of soil over which the wave must travel. If the antenna system radiates most of the transmitter power at relatively low angles, particularly along the ground, the ground wave will give reliable communication over distances from 50 to considerably over 100 miles, the latter distances applying where conditions are particularly favorable, as when the path is mostly over sea water.

● POLARIZATION

It was mentioned in Chapter One that a ground wave must be vertically polarized, so that the radiation from an antenna which is to produce a good ground wave likewise must be vertically polarized. This dictates the use of an antenna system the radiating part of which is mostly vertical. A horizontally-polarized antenna will produce practically no ground wave, and it is to be expected that such an antenna will be ineffective for daytime communication. This is because absorption in the ionosphere in the daytime is so high at these frequencies that the reflected wave is too weak to be useful. At night a horizontal antenna will give better results since nighttime ionosphere conditions permit the reflected wave to return to earth without excessive attenuation. The difference between daytime and nighttime conditions is similar to that existing on the broadcast band, where distant stations can be heard well at night but not at all in the day.

There is still another reason why a vertical antenna is better than the horizontal for 160-meter work. Comparison of the ground-reflection factors in Figs. 2-26 and 2-27, Chapter Two, for horizontal antennas at heights of ¼ and ⅛ wave will show that at the lower height the ground is less effective in reinforcing radiation. At 160 meters even a height of ⅛ wave, about 65 feet, is not easy for all amateurs to attain, while a height of ¼ wave is out of the question for nearly everyone. Any reasonable height is small in terms of wavelength, so that a horizontal antenna on 160 meters is a poor radiator at angles useful for long distances ("long," that is, for this band). Its chief field of usefulness is for communication over relatively short distances at night.

The chief disadvantage of vertical polarization is the fact that the stronger ground wave is more likely to cause interference with near-by broadcast receivers.

Vertical-Antenna Design Considerations

For good night coverage at distances toward the limit of the ground wave it is desirable to use an antenna that will give comparatively little radiation at angles above about 45 degrees. This is because the high-angle radiation returns to earth within the useful range of the ground wave, and in the outer part of this range may have intensity comparable with that of the ground wave itself. The sky waves arrive at the ground in random phase with respect to the ground wave, giving rise to severe fading in this area. The antenna should, therefore, confine its radiation to angles sufficiently low so that the nearest point to the transmitter at which sky waves return to earth is just beyond the limits of the ground wave.

The various conditions can be met by the use of an antenna a half wave high, but this is impractical since a height of over 250 feet would be required. Fortunately it is possible to approach the effect of a half-wave antenna by suitable treatment of a much lower structure.

A vertical antenna will be most effective when it can be erected in a fairly clear spot so that the ground wave is not absorbed in near-by buildings. Frame buildings are not likely to cause much trouble, but it is best to keep clear of steel structures by at least a wavelength or two.

● GROUNDED ANTENNAS

It was explained in Chapter Two that a quarter-wavelength grounded antenna is resonant, and that a still smaller one can be made resonant by "loading" it at the base. However, as pointed out in that chapter, it is far better to do the loading at the top of the antenna, to raise the point of current maximum in the antenna and thus to increase the radiating efficiency of the system. Several methods for top loading an antenna were described.

Bent Antennas

Perhaps the simplest method of meeting the fundamental requirement of keeping the current loop high is to use a bent antenna such as is shown in Fig. 10-1A, with part of the antenna vertical and part horizontal. The horizontal part should be one-quarter wave in length so that the current loop will appear at the top of the vertical portion. The current distribution will be as shown in the drawing, assuming that the vertical portion is ¼ wave high. If smaller heights are used, the horizontal portion should still be ¼ wave in length. Since the most useful radiation is from the vertical part, it is of course desirable to make the antenna as high as possible.

The length of the horizontal top portion can be calculated from the formula

$$\textit{Length of quarter wave} \text{ (ft.)} = \frac{234}{f\text{ (Mc.)}}.$$

There is no need for excessive accuracy in determining this length, since a discrepancy of 5 or 10 feet will make comparatively little difference in the performance of the antenna.

The lower end of the antenna is grounded through a loading circuit that tunes the system to resonance and also provides a means for coupling power from the transmitter into the antenna. The constants of the loading circuit will depend upon the total length of the antenna system, and therefore depend upon the antenna height. For heights between 40 and 70 feet a circuit of the type shown in Fig. 10-2 will be suitable, provided the leads between the bottom of the antenna and the coupling circuit, and between the loading circuit and the effective ground, are only a few feet in length. These leads are part of the effective length of the antenna, and must be added to the antenna length in determining the actual constants required in the loading circuit.

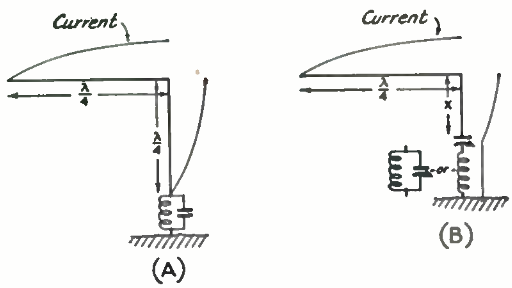


Fig. 10-1 — Bent antennas using a quarter-wave horizontal section to bring a current loop at the top of the vertical wire. A quarter-wave vertical section is shown at A; at B the height X is made as great as the circumstances permit. Series tuning may be used for lengths of X up to about ⅛ wavelength; parallel tuning for greater lengths of X.

For maximum effectiveness, the vertical part of the antenna should actually be vertical, and not simply run off at some convenient angle from the operating room to the top of the pole. The wire

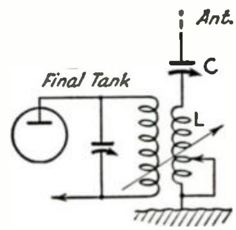


Fig. 10-2 — A practical loading and coupling circuit for antennas of the type shown in Fig. 10-1B when the height X is ⅛ wavelength or less (up to 65 feet approximately). The series tuning condenser C should be 250 to 500 μμfd.; receiving-type condensers will suffice for moderate powers. Coil L may consist of 20 turns of No. 12 wire space-wound (6 turns per inch) to a diameter of 3 inches, arranged so that it can be tapped conveniently at least every few turns. Tuning procedure is that for series tuning as described in Chapter Three. An r.f. ammeter may be connected in series with the antenna where it joins C. A 2.5-ampere instrument will suffice for powers up to a few hundred watts.

may come down the pole on stand-off insulators, or may be pulled down vertically from the horizontal strain insulator after the fashion shown in Fig. 10-3. Wire guys on the pole should be broken up at intervals of 25 feet or so with egg-type insulators to prevent pick-up of r.f. energy from the antenna.

Antennas of this type offer an opportunity for use of a rather simple feeder system that permits installing the antenna at some distance from the transmitter. If the antenna height is ⅛ wave, for example, the total length is ⅜ wave including the horizontal part. An additional ⅛ wave wire may be added to the antenna, as shown in Fig. 10-4, to make the total length ½ wave. This extra section is connected to the bottom of the vertical wire and is used as a feeder. It should run parallel to and fairly close to the ground for as much of its length as possible (a height of 7 feet is permissible so that it will not be a hazard to walkers) and terminate at the transmitter in a parallel-tuned circuit, the other end of which is grounded. (The length of the ground lead should be included in the "feeder" length.) At this point the impedance looking into the feeder and antenna has its highest value so that losses in the ground connection are relatively low. There will be very little current in the ground lead under these conditions, but an ammeter inserted at the base of the vertical portion will read about 70 per cent of the current at the top.

Such a "feeder" does comparatively little radiating because it is parallel to and close to the ground and because it represents the section of the antenna which carries the least current. In cases where the antenna height is not an eighth wavelength the "feeder" length, including the ground lead, should be ½ wavelength less the actual length of the antenna from the base to the far end of the horizontal portion. The length of a half wave is given closely enough by the formula

$$\textit{Length of half wave (ft.)} = \frac{468}{f\,(\text{Mc.})}.$$

The feeder may be made longer or shorter than the exact length necessary to make the whole system a half wave long, if more convenient, provided the whole system is brought to resonance by means of the coupling system. However, excessive length in a feeder of this type is not desirable. Also, it is preferable to have the length to the ground connection a half wave so that the current in the ground lead will be minimum, which means lowest loss in the ground connection.

Grounds

One of the chief problems of obtaining optimum performance on 160 meters is that of getting a good low-resistance ground. The old stand-by connection to a water pipe may serve in a pinch, but seldom results in the best possible antenna performance.

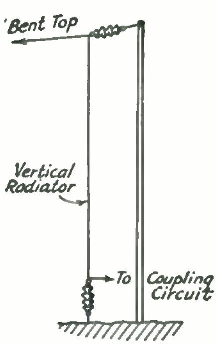


Fig. 10-3 — An arrangement for keeping the main radiating portion of the antenna vertical.

If circumstances make it necessary to use a water pipe for a ground connection, always select a cold-water pipe since it usually goes more directly to ground than the hot-water variety. Gas pipes never should be used because insulated joints are sometimes included in the piping. Wherever possible, the connection to the cold water piping should be made directly at the point where the pipe enters the ground; that is, on the street side of the water meter. The length of the ground lead necessarily must be taken into account in computing the total length of the antenna.

To make the connection, carefully clean the pipe by scraping and sandpapering. Fit on a clean ground clamp, make it good and tight, and make sure that the ground wire makes a good

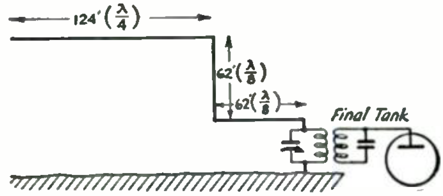


Fig. 10-4 — Bent antenna ⅛ wave high, with a "feeder" section making the total length ½ wavelength. The length figures are for 1900 kc., approximately, and the same antenna may be used over the whole band. The parallel-tuned coupling circuit should be capable of being tuned independently to the operating frequency, and the inductance of the coil preferably should be variable by means of taps so that the optimum L/C ratio can be secured.

electrical connection to the strap. Solder it if necessary. The assembly may be rubber-taped to prevent oxidation if there is considerable dampness.

If it is impossible to reach the pipe at the point where it enters the ground, a connection of the type described above may be made to any convenient cold-water pipe as a secondary resort. In such cases, estimation of the effective length of the ground lead is difficult, since piping systems sometimes are rather extensive and hence have considerable capacity to ground. The effective length usually will be appreciably less than the actual length of the shortest path which might be traced back to ground along the piping, and in the case of a ground to a heating system may be quite small because of the large masses of metal at the radiators. In such cases the amount of loading for bringing the system to resonance must be determined experimentally.

A simple outdoor ground may be made by driving a length (6 feet or more) of 1-inch pipe into the soil. If possible, pick a spot where there is considerable natural moisture; the resistance

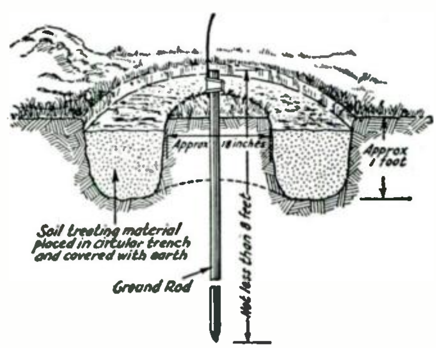


Fig. 10-5 — Ground system treated to increase conductivity. The circular trench is filled with rock salt, magnesium sulphate, or copper sulphate, put in dry and then flooded with water. After treatment, the trench is covered with earth. Fifty pounds of treating material so disposed will have a life of two or three years.

will be less under such conditions. Four pipes arranged at the corners of a 10-foot square, all connected together at the top, will be considerably better than one.

A quite good low-resistance ground connection can be made as shown in Fig. 10-5, if the space is available and some digging is permissible. The chemicals increase the conductivity of the ground in the vicinity of the grounding pipes or rods and thus reduce the losses from current flow.

Radial Grounds

The ideal form of ground is a series of conductors buried a foot or two beneath the surface, radiating like the spokes of a wheel from under the vertical part of the antenna, as shown in Fig. 10-6. Its construction is beyond most amateurs, but it is mentioned here for the benefit of those who may have the space and a plow to cut the furrows which contain the ground conductors. Such a ground system not only reduces I^2R losses at the ground connection but provided it is made extensive enough also greatly reduces power losses in the ground in the immediate vicinity of the antenna.

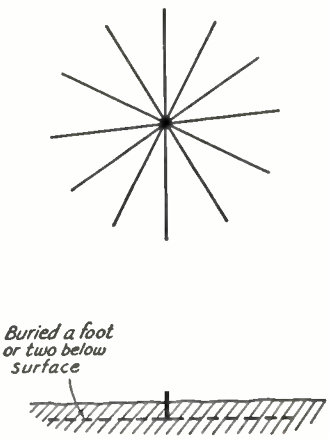


Fig. 10-6 — **The best ground is a radial system of buried copper strip or heavy bare copper wire.**

Better results can be expected as the length of the radial wires is increased. There is no necessity for a length greater than ½ wavelength, however, and even 1/10 wavelength will give satisfactory performance. This calls for a length of about 50 feet per radial, or a total diameter of about 100 feet for the ground system. As many radials as possible should be used.

The Counterpoise

The counterpoise is a form of capacity ground which is quite effective. Its use is particularly beneficial when an extensive buried system is not practicable, or when an ordinary pipe ground cannot be made to have sufficiently low resistance, as in rocky or sandy soils.

To work properly, a counterpoise must be large enough to have considerable capacity to ground, which means that it should cover as much ground area as the location will permit. No specific dimensions are necessary, nor is the number of wires particularly critical. A good form is an approximately circular arrangement using radial wires with cross-connectors joining them at intervals, as in Fig. 10-7. There is no particular necessity for extending the radius of a circular counterpoise beyond a half wavelength, nor is it desirable that the lengths of the individual wires bear any particular relation to the wavelength. Rather, the intention is to have the counterpoise act as a pure capacity instead of exhibiting resonance effects. The capacity of the counterpoise will be approximately equal to that of a condenser consisting of two plates each of the same area as that of the counterpoise, with spacing equal to the height of the counterpoise above the ground.

The shape of the counterpoise may be made anything convenient; square or oblong arrangements are usually relatively easy to construct and will work satisfactorily. There should not be too few wires, but on the other hand separations between wires up to 10 or 15 feet will do no harm on fairly large counterpoises, and 5 to 10 feet on smaller ones. It is a good plan to join adjacent wires with jumpers at intervals about equal to the wire separation so that resonance effects will be minimized.

The height of the counterpoise is not particularly critical. It is best to construct it high enough to be out of the way, which ordinarily means from 6 to 10 feet above the ground. Remember that the height of the antenna is reduced by the amount of counterpoise height.

Satisfactory results have been secured with counterpoises simply lying on the ground, or with large screens of chicken wire similarly laid under the antenna. However, the best performance will be secured, as a general rule, when the counterpoise is insulated from ground. When in contact with the ground surface, the losses are likely to be higher because the counterpoise tends to act either as a poorly-conducting direct ground or as a leaky-dielectric condenser.

ANTENNAS FOR SMALL SPACE OR HEIGHT

The antennas discussed thus far have been designed to take advantage of the transmission characteristics of the 160-meter band. A certain amount of height and ground space is essential for this purpose. Many amateurs, however, do not have the facilities for the construction of even these simple forms and — particularly the city "cliff dwellers" — must simply string up a wire where some space is available.

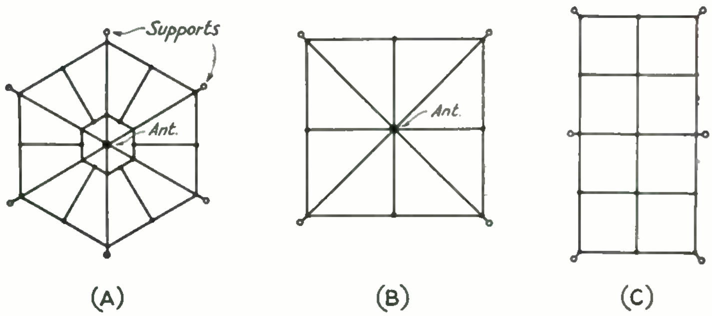


Fig. 10-7 — Some suggested forms of counterpoise. Perfect symmetry is not essential, but it is desirable to extend the counterpoise as nearly as possible for the same distance in each direction from the antenna.

A vertical antenna must be quite clear of surrounding buildings, particularly those of steel construction, if good results are to be secured. If the height required for this purpose is not obtainable, then a horizontal wire must suffice. No useful purpose is served in erecting a vertical antenna between buildings which are going to absorb most of the radiated energy, or which perhaps reradiate some of the energy to make the horizontal directional pattern of the antenna poor in the most desired directions of communication.

The fundamental requirement for an antenna which cannot be "designed" for 160-meter work is that it must be resonant at the operating frequency. That is, it must accept as much power as possible from the transmitter, even though the radiation of the power must be left more or less to chance. It is desirable to get the high-current portion of the antenna well away from buildings, if this is possible. The antenna may be bent, if necessary, to fit the available space, but the bends should be made with a view to their effect on the performance of the antenna as a radiator.

To make tuning easy, it is desirable that the antenna length be a multiple of a quarter wavelength, within reasonable limits. The ground lead should be short, although, as already explained, the "length" of this lead may depend upon the grounding system. An antenna length of about 125 feet is the smallest recommended for working over the band, although shorter lengths can be loaded to the proper electrical length with a series inductance as shown in Fig. 10-8B. Provided the effective length of the ground lead is not too great (up to perhaps 35 or 40 feet) the system may readily be tuned to resonance with an adjustable coil and series condenser.

If no space is available sufficient to allow the antenna to be installed in a straight line, it may be bent to fit. The far end may be bent down, as shown in Fig. 10-8C, or even back on the antenna as in Fig. 10-8D. In the latter case at least ⅛ wavelength of the near end (the high-current part) should be unparalleled by the bent wire, since there is partial cancellation of the radiation from the folded-back part. Bends in horizontal directions may be made at several points along the wire, in cases where this is necessary, provided the angle between the bent portions is as large as possible. Try not to have less than a right-angle bend, especially in the high-current portion of the antenna.

A disadvantage of the quarter-wave "random" antenna is the fact that the high-current end, which does the most radiating, is the end brought into the station. If there is at least one quite long straight stretch available for erecting the antenna, it is a better plan to make the antenna length such that the maximum current point comes at the middle of the straight section. This means that the wire should be a quarter wave long (125 feet will be satisfactory) from the middle of the span to the far end, the necessary bends or folds to make up any excess length being made at the far end. The

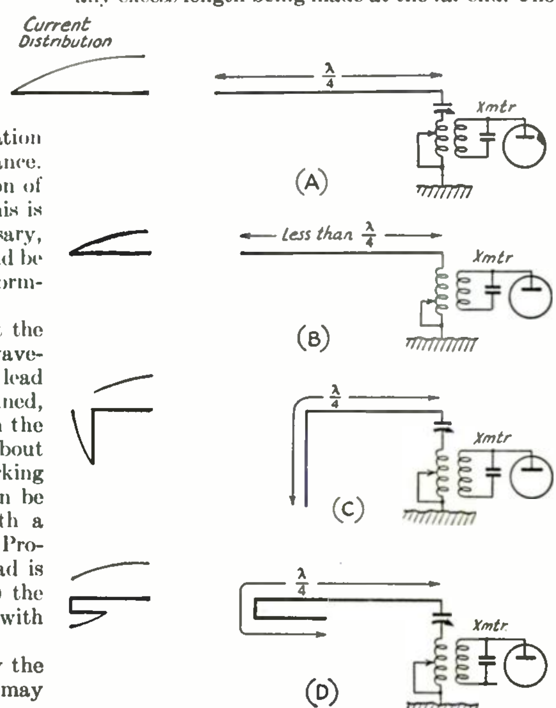


Fig. 10-8 — Typical arrangements of a quarter-wave horizontal antenna, for installation where height or space is limited. Current distribution is shown for each case. A length of about 125 feet will be satisfactory.

distance from the middle of the span to the transmitter can form the antenna length on that side or, alternatively, the wire length here may also be made a quarter wavelength, with bends or folds, to make a half-wave antenna and thus bring a voltage loop at the coupling point. The total length of the wire would be 250 feet in that case, and parallel tuning would be called for at the transmitter.

Antennas of this type will work quite well, especially for moderate distances at night, even though they are not capable of the type of performance to be expected from a good vertical antenna. The chief points to be remembered are these: it is easiest to make the antenna take power if its length is near a multiple of a quarter wavelength, and bends will not do a great deal of harm provided they are made in parts of the antenna where the current is small. The aim should be to obtain the longest possible straight stretch for the high-current part of the antenna.

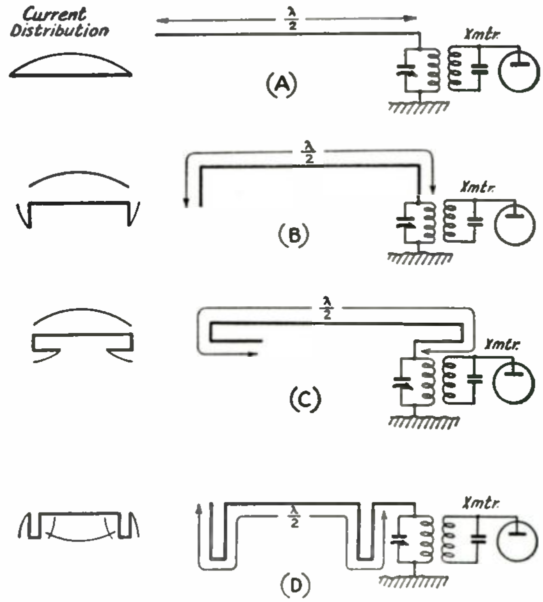


Fig. 10-9 — Suggested methods of bending half-wave antennas for installation where space is limited. The points to watch out for in making bends are discussed in the text. The total wire length is about 250 feet for an antenna which can be used over the 160-meter band.

Alternative Coupling Methods

Other types of coupling systems may be substituted for those shown in this chapter, and Chapter Three should be consulted for design and adjustment information on this subject. In the case of vertical antennas, particularly, where the base of the antenna may be some distance from the transmitter, it may be desirable to use a link line to a coupling circuit installed in a weatherproof box at the antenna. The feeder already described usually will be more convenient for this purpose, however, if its length fits in with the station and antenna layout.

The pi-section filter type of coupler is especially convenient with antennas shorter than a quarter wavelength, but should be tuned carefully to prevent harmonic radiation. A quarter-wave grounded antenna inherently discriminates against even harmonics, but this is not true of several of the systems described in this chapter since some of them approach or equal a half wavelength in total wire length.

When the antenna proper is located at some distance from the transmitter the two may be connected by means of a transmission line, either tuned or untuned, of one of the types described in Chapter Three. The principles of design and operation are exactly as set forth in that chapter. Probably in the majority of cases, however, the distance does not justify the use of such a line.

Supports and Construction

The problem of supporting the antenna at the requisite height above ground can be solved in a variety of ways, depending upon local conditions. Quite frequently at least one end of the antenna can be hooked to a convenient building or tree, thereby obviating the expense and trouble of erecting a pole or tower.

Anchoring one end of the wire to a building presents no particular problems. Some precautions must be taken, however, when the support is a tree, particularly if the antenna is fastened near its top. Trees sway considerably in a wind, so that some means must be taken to prevent the antenna wire from snapping in two, and at the same time to get rid of too much slack in calm weather. If the tree can be climbed to the point where the antenna is to be supported, a wire can be looped around the trunk or limb to hold a pulley as firmly as possible. The trunk or branch must be protected from cutting by the wire, and this is easily accomplished with several strips of wood between the wire and the tree. The antenna rope, which should run freely through the pulley, should not be tied at its free end but should be fastened to a counterweight heavy enough to keep the antenna taut.

A somewhat similar scheme can be used when it is necessary to throw a rope over a limb because the tree cannot be climbed to the height desired. In this case the pulley can be fastened to the end of the rope which goes through the tree, while the antenna rope goes through the pulley, somewhat as shown in Fig. 11-1. The tree rope is securely fastened at the bottom.

In using either of these systems it is essential, to prevent the rope from jamming in the pulley, to use a rope heavy enough so that it will not jump out of the pulley wheel and get caught between the wheel and frame. Some tension should be kept on the pulley, especially while the antenna is being raised, to keep it from twisting.

TELEPHONE POLES

Probably the most satisfactory type of pole is the kind used by utility companies to carry power and telephone wires. These poles are heavy enough to support most amateur antennas without being guyed, and can be provided with steps so that they can be climbed without difficulty. Their chief drawback is the fact that they are comparatively expensive.

Costs vary in different parts of the country, depending upon the distance to the source of the poles. In lengths of 30 to 60 feet, however, an average cost for a creosoted pole, installed, is a dollar or two per foot. In ordinary soil about one-tenth of the length is set in the ground, so that the pole height is about 90 per cent of its length.

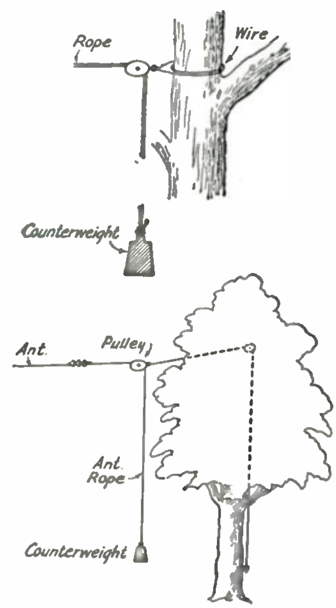


Fig. 11-1 — Fastening the antenna to a tree. The counterweight prevents the antenna wire from being snapped off when the tree sways in a wind.

Poles of this type usually can be purchased from the local electric light or telephone company and installed by their crews. In some localities the companies let out this work to local contractors, in which case the contractor can be approached directly. Use caution in picking up "bargains" in the pole line; a pole with an unsound center is not a good investment nor is it a safe thing to climb.

● SMALL MASTS

Where the height required is of the order of 20 feet or less (particularly when the mast is to be installed on the roof of a house or garage) a single piece of 2 × 3 lumber will make a good pole. A vertical antenna for 10 or 6 meters, for instance, can be constructed on a 20-foot 2 × 3 as shown in Fig. 11-2. Three guys spaced 120 degrees around the pole and fastened about half-way up will suffice to keep the pole erect. If the roof is flat, the bottom of the pole need not be fastened down since it will have no tendency to move once the guys are tightened. A small flat board may be placed under it to prevent damage to the roof. On peaked roofs, a small wooden inverted "V" can easily be nailed to the bottom of the pole to keep it from slipping.

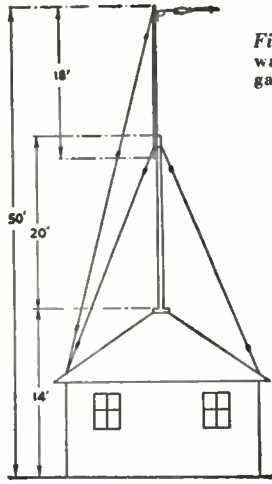
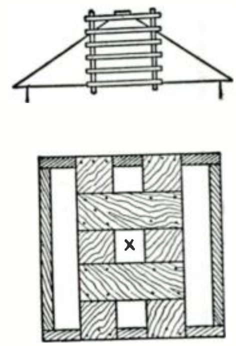


Fig. 11-3 — A socket and "cattle walk" for installing a mast on the garage roof.



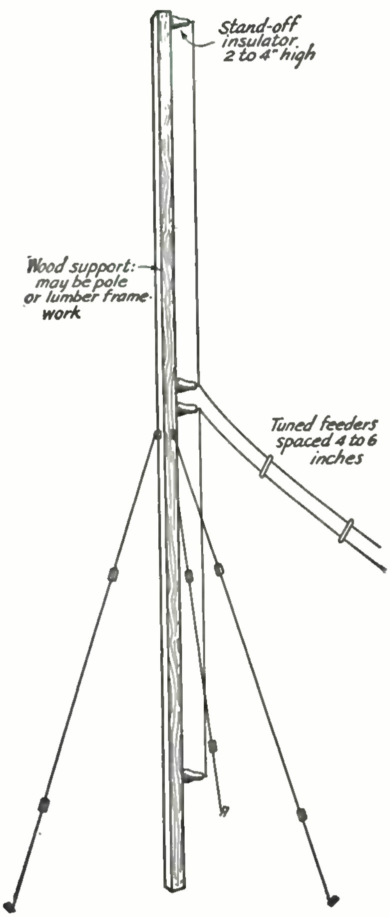


Fig. 11-2 — A single 2 × 3 will serve as a mast for heights up to about 20 feet. This drawing shows how a 28-Mc. vertical can be installed on such a pole. For horizontal antennas, the guys should go to the top.

To support a horizontal antenna, the guys should run to the top of the mast. Actually, only the two which pull away from the antenna need go to the top; the antenna itself provides the pull in the opposite direction. The third guy, in the direction of the antenna, is used only to keep the pole from falling over when the antenna is lowered, and may be fastened about halfway up where it will not interfere with the antenna pulley and rope.

Fig. 11-3 shows how a socket can be constructed to fit over the pointed roof of a typical two-car garage to seat a mast, which in this case consists of two pieces bolted together. A satisfactory method of guying also is shown. From the joint between the two sections of the pole four guys go to the corners of the garage, but only two at the back are used at the top of the mast. The pull of the antenna is ample to keep the upper section vertical. The lower section of a bolted mast of this type preferably should be 2 × 4, with the upper section 2 × 3.

Fig. 11-3 also shows a sketch of a "cattle walk" that is lashed in place on the roof to permit walking the mast up without danger of slipping. Such a gadget can be used on any sloping roof; even a ladder will suffice in many cases when its upper end is roped to a solid anchorage so that it cannot slide.

● THE "A-FRAME" MAST

A type of light and inexpensive mast that has been very popular is shown in Fig. 11-4. In lengths up to 40 feet it is very easy to erect and will stand without difficulty the pull of ordinary antenna systems. The lumber used is 2 × 2 straight-grained pine (which many lumber yards know as hemlock) or even fir stock. The uprights can be each as long as 22 feet (for a mast slightly over 40 feet high) and the crosspieces are cut to fit. Four pieces of 2 × 2, 22 feet long, will provide enough and to spare. The only other materials required are five ¼-inch carriage bolts 5½ inches long, a few spikes, about 300 feet of No. 12 galvanized iron wire for the guys or stays, enough glazed-porcelain compression insulators ("eggs") to break up the guys into sections, and the usual pulley and halyard rope. If the strain insulators are put in every 20 feet approximately 15 of them will be enough.

After selecting and purchasing the lumber — which should be straight-grained and knot-free — three sawhorses or boxes should be set up and the mast assembled in the manner indicated in Fig. 11-5. At this stage it is a good plan to give the mast two coats of "outside-white" house paint.

After the second coat of paint is dry, attach the guys and rig the pulley for the antenna halyard. The pulley anchorage should be at the point where the top stays are attached so that the back stay will assume the greater part of the load tension. It is better to use wire wrapping around the stick, with a small through-bolt to prevent sliding down, than to use eye bolts.

If the mast is to stand on the ground, a couple of stakes should be driven to keep the bottom from slipping. At this point the mast may be "walked up" by a pair of helpers. If it is to go on a roof, first stand it up against the side of the building and then hoist it, from the roof, keeping it vertical. The whole assembly is light enough for two men to perform the complete operation — lifting the mast, carrying it to its permanent berth, and fastening the guys — with the mast vertical all the while. It is therefore entirely practicable to put up this kind of mast on a small flat area of roof that would prohibit the erection of one that had to be raised vertical in its final location.

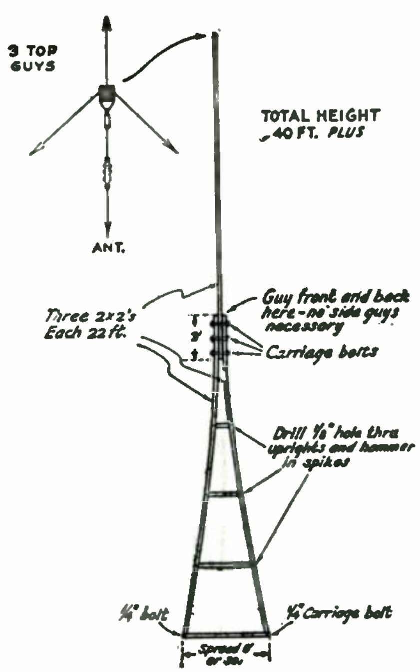


Fig. 11-4 — The "A-frame" mast, lightweight and easily constructed and erected.

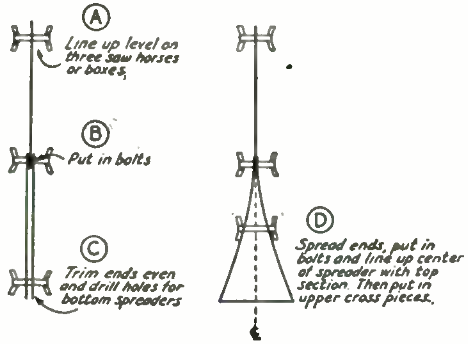


Fig. 11-5 — Method of assembling the "A-frame" mast.

● ANOTHER SIMPLE MAST

The mast shown in Fig. 11-6 is relatively strong, easy to construct, and costs very little. Like the "A frame," it is suitable for heights of the order of 40 feet. It is also easily dismantled in case it has to be moved.

The top section is a single 2 × 3, bolted at the bottom between a pair of 2 × 3s with an overlap of about 2 feet. The lower section thus has two legs spaced the width of the narrow side of a 2 × 3. At the ground, the two pieces are bolted to a 2 × 4 which is set in the ground. A short length of 2 × 3 is set between the two legs about halfway up the bottom section to maintain the spacing. Four ½ × 6 carriage bolts are needed, along with washers; this length is sufficient since the "2 × 3s" actually are about 1¾ × 2½ inches. All pieces of lumber are set so that the long axis faces the antenna direction.

It will be sufficient to guy the mast as shown in the drawing. The two back guys at the top pull against the antenna, while the three lower guys prevent any buckling at the center of the pole. The two sets of back guys may be anchored at the same point. For a height of about 40 feet, the guys should be anchored about 15 feet or more from the bottom of the pole.

The mast can easily be raised by two people; in fact, one man can manage the job without too much difficulty. The length of 2 × 4 which is set in the ground should be placed so that it faces the proper direction, and should be made vertical by lining it up with a plumb bob. The holes for the bolts should be drilled beforehand. The lower section is then laid on the ground so that bolt *A* can be slipped in place through the three pieces of wood and tightened just enough so that the section can turn freely on the bolt. Then the top section is bolted in place and the mast pushed up, using a ladder or another 20-foot 2 × 3 for the job. As the mast goes up the slack in the guys can be taken up so that the whole structure is in some measure continually supported. When the mast is vertical bolt *B* is slipped in place and both *A* and *B* tightened. The lower guys can next be given a final tightening, leaving those at the top a little slack until the antenna is pulled up, when they can be adjusted to pull the top section into line with the bottom.

The 2 × 4 should extend at least 3 feet into the ground, and should set solidly. Concrete is not necessary, but it will help to pack rocks in the hole to provide some bracing. The pole will stand without guying when the two bottom bolts are in, which does away with the necessity for having a helper on each guy, but, of course, will not stand much strain under those conditions.

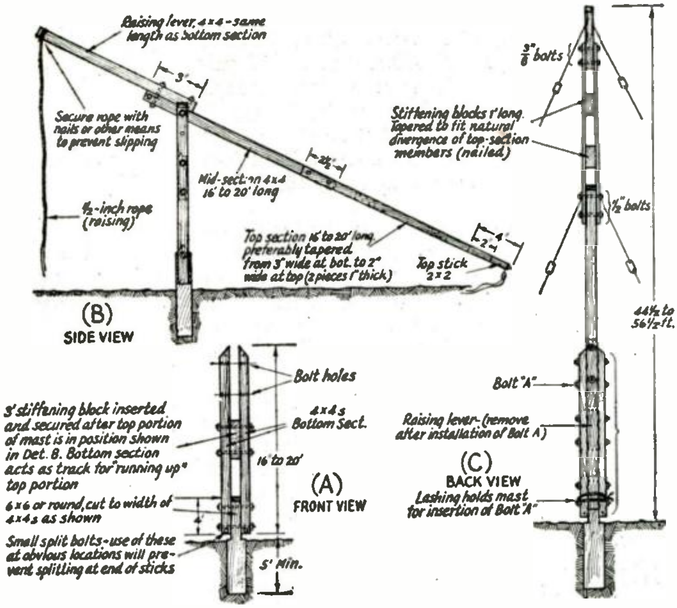


Fig. 11-7 — **Constructional and raising details of the 55-foot wooden mast.**

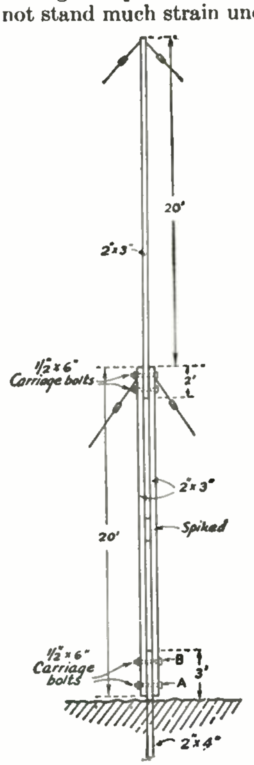


Fig. 11-6 — **A simple and sturdy mast for heights in the vicinity of 40 feet, pivoted at the base for easy erection. The height can be extended to 50 feet or more by using 2 × 4s instead of 2 × 3s.**

The mast shown in Fig. 11-7 is somewhat similar but can be used for heights up to 55 feet or so. The bottom section, minus the 3-foot stiffening block, should be set in place first, as shown at A. The top section is then slid between the two uprights, with the raising lever on the upper side. The bottom end of the top section then is pulled up and the two sections lined up so that the bolt can be put in place, as shown at B.

By pulling downward it is not difficult to swing the top section up into vertical position. Bolt *A* (Fig. 11-7C) is then slipped into place and the stiffening block inserted.

For masts up to 45 feet, only two guys at the top will suffice. For greater heights, two at the top and three at the midpoint in the top section are recommended.

A HOLLOW MAST

A different type of mast construction, suitable for heights of the order of 50 to 60 feet, is shown in Fig. 11-8. Although comparatively light in weight, it is practically as strong as a solid pole of the same cross section.

It is a square hollow pole, held together in the fashion of bamboo growth; that is, with a strengthening section spaced about every 2 feet.

The foundation is a 6 × 6 timber about 6 feet long. The next section is about 14 to 18 feet long, depending upon the availability of the lumber, which should be good smooth-finish, hard pine, cypress or spruce. Three pieces (assuming that 18-foot stock is available) are laid out. There will be two sides 6″ wide by 18′ long of ¾″ or ⅞″ stock, and the third will overlap on the edges so that it will be 8″ wide by 18′ by ¾″ or ⅞″. Insert the 6 × 6 about 2′ into the "U" formed by the three pieces and after lining up so that all edges are flush, commence to fasten together with nails, or preferably iron screws with flat heads, not forgetting to paint thoroughly every edge with a thick mixture of white lead and linseed oil. The next operation is to put in the braces every 2 feet. These are 6 × 6 × 1-inch thick, and should be put in with white lead between wood surfaces and thoroughly fastened with screws or finishing nails. At the open end of the "U" leave 18″ for the insertion of the next section of the mast which telescopes into the base section. Give the inside a thorough coat of white lead and linseed oil about the consistency of glue and let dry for a week or so. The cover for this section of the "U" is another 8-in. × 18-ft. × ¾-in. piece which is screwed down after the second coat of white lead and linseed oil is dry. Any irregularities in the lumber and joints should be smoothed down with a plane.

The second and third sections of the mast are constructed in a fashion similar to that of the base, except that they are progressively smaller. The second section should have outside dimensions of 6 by 6 inches to telescope inside the base section. The third section will be 4 by 4 inches to telescope into the second and the top section is a solid 2 × 2.

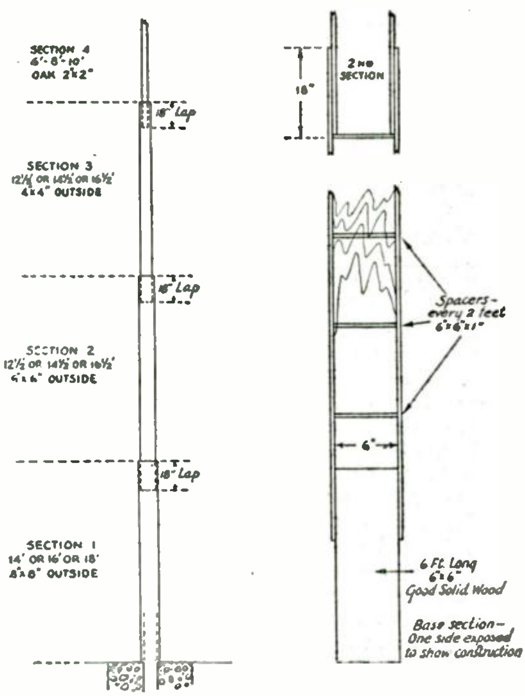


Fig. 11-8 — A hollow mast, lightweight but strong, constructed of 1-inch boards.

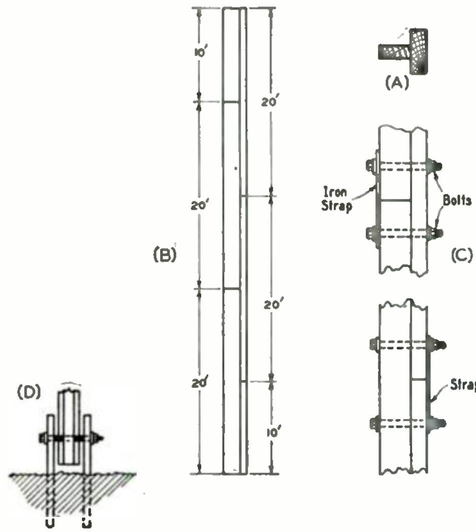


Fig. 11-9 — "T"-section mast made from overlapping 2 × 4s or 2 × 6s.

The complete list of material required is given below:

1st Section

1 — 6″ × 6″ × 6′
2 — 6″ × 18′ × ⅞″
2 — 8″ × 18′ × ⅞″
7 — Spacers 6″ × 6″ × 1″

2nd Section

2 — 4″ × 18′ × ⅞″
2 — 6″ × 18′ × ⅞″
8 — Spacers 4″ × 4″ × 1″

3rd Section

2 — 2″ × 18′ × ⅞″
2 — 4″ × 18′ × ⅞″
8 — Spacers 2″ × 2″ × 1″

Top Section

1 — 2″ × 2″ × 12′ (oak)
1 — Pulley (brass or bronze)
1 — Manila rope (necessary length)
4 — Guys (necessary length) No. 8 galvanized-iron wire
Strain insulators

MASTS WITH BUTTED TIMBERS

A type of mast construction suitable for heights up to about 80 feet is shown in Fig. 11-9. The mast is built up by butting 2 × 4 or 2 × 6 timbers edgewise against a second 2 × 4, as shown at A, with alternating joints in the edgewise and flatwise sections as shown at B. The construction can be carried out to greater lengths than shown simply by continuing the 20-foot sections. It will be noted that one or both ends must end with a 10-foot section on either the edgewise or flat timbers.

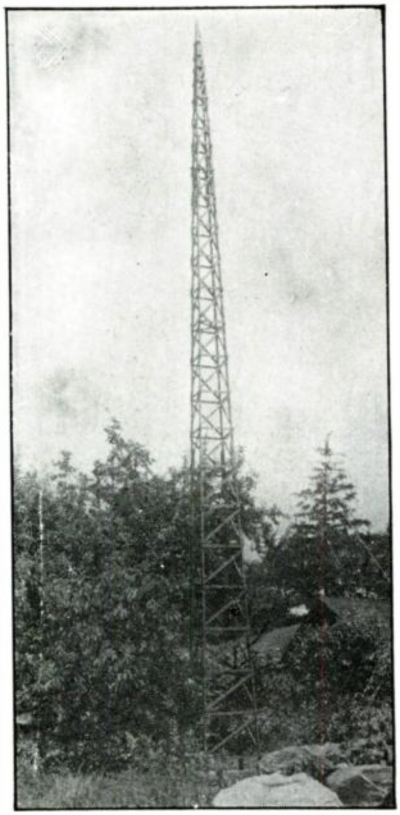


Fig. 11-10 — A lattice mast that is easy to build and erect, and can readily be taken down for repainting when necessary. It is strong enough to withstand practically any wind load that may be encountered.

Longer or shorter sections may be used if more convenient.

The method of making the joints is shown at C. Quarter-inch or 3⁄16-inch iron, 1½ to 2 inches wide, is recommended for the straps, with half-inch bolts to hold the pieces together. In addition, a bolt should be run through the pieces midway between joints to provide additional rigidity.

Although there are many ways in which such a mast can be secured at the base, the "cradle" illustrated at D has many advantages. Heavy timbers set firmly in the ground, just far enough apart so that the base of the mast will pass through them, hold a large carriage bolt or steel bar which serves as a bearing. This passes through a hole in the mast so that the latter is pivoted at the bottom. As the mast swings upward in an arc while being raised the bottom is free to pivot on the bearing.

The job of raising the mast can be simplified, when a bottom bearing of this nature is used, because half of the guys can be put in place and tightened up before the mast leaves the ground. Four sets of guys should be used, one in front, one directly in the rear, and one on each side at right angles to the direction in which the mast will face. Since the base position is fixed by the bearing, all the side guys can be put in place, anchored and tightened while the mast is lying on the ground. Thus there is no danger of side-sway or bending while the mast is going up, and a smaller crew can do the job. A set of guys should be used at each of the joints in the edgewise sections, the guy wires being wrapped around the pole rather than fastened to bolts or passed through holes in the pole, as either of the latter two methods tends to weaken the joint.

For heights up to 50 feet, 2 × 4s may be used throughout. For greater heights it is advisable to use 2 × 6s for the edgewise sections, although 2 × 4s will be satisfactory for the flat sections.

A LATTICE MAST

Another popular type of antenna support is the lattice mast, illustrated in Figs. 11-10 and 11-11. Such masts are built in many different ways, depending upon the ideas and ingenuity of the constructor, but the tower shown in the photograph will serve as a guide for any such construction work.

The mast requires 14 pieces of clear pine, 16 feet long and 1½ × 3 inches in cross section; 300 feet of 1 × "furring" strips, 75 ½-inch carriage bolts 2½-inches long, and a quantity of 2-inch wire nails. Each of the four 48-foot side pieces is made of three of the 16-foot lengths of 1 × 3. The two extra lengths are cut in 4-foot lengths and bolted to the long pieces, as shown in Fig. 11-11-D. Before bolting the pieces together, the facing wood surfaces should be given a liberal coating of pure white lead. The bolts, with iron washers, should be drawn up as tight as possible.

With the help of a carpenter's square and a long piece of cord, one side of the mast is laid out, running from the exact four-foot width of the base to the joining of the side pieces at the top. The first crosspiece of 1 × 2 is nailed on 1½ feet from the bottom, using white lead where wood meets wood. The second cross piece is nailed on 2 feet above the first. Each subsequent piece is nailed on with the spacing reduced one inch each time. Thus the first crosspiece is 18 inches from the bottom, the second is 24 inches from the first, the third is 23 inches from the second, and so on until the 32-foot level is reached. From there on to the 40-foot level the spacing is reduced 2 inches each time. Above 40 feet the spacing is reduced 3 inches each time. Then the angle crosspieces are installed between each horizontal member.

When two sides have been finished, they are turned on edge and the bottoms spaced exactly 4 feet. Cross and angle pieces are then installed from the bottom to the 24-foot level, after which the mast can be turned over and the other

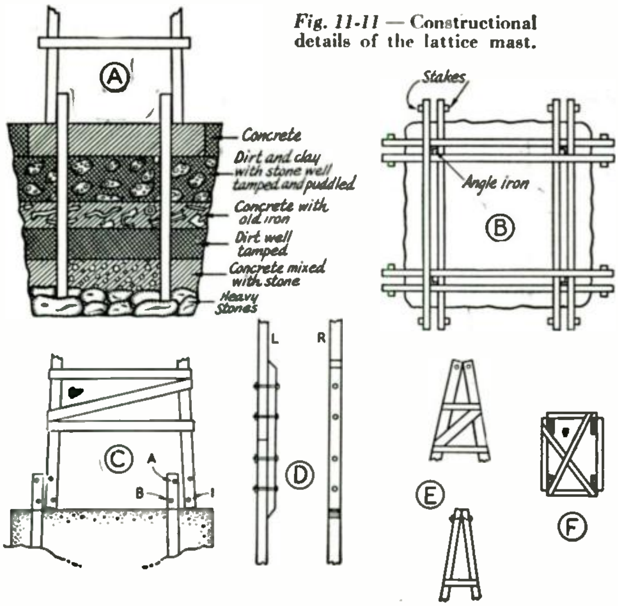


Fig. 11-11 — Constructional details of the lattice mast.

side completed. The third side is then completed to the top. The two sides are bolted together at the top, as shown in Fig. 11-11E.

The final construction step consists of nailing on the internal cross braces, at the 10-, 20- and 30-foot levels, as shown in Fig. 11-11F. It is convenient to cut the cross and angle braces after they have been nailed on.

The foundation for the mast requires a 6-foot cube hole, if the foundation details shown in Fig. 11-11A are followed. Four 6-foot lengths of angle iron are set in concrete and dirt as indicated, and are held in proper alignment during this process by the jig indicated in Fig. 11-11B. Most of the concrete mixture will require very little cement — only the top finish layer of concrete requires a good 3-to-1 mixture of cement and sand. While waiting for the various sections of the foundation to set, the mast can be painted, using two coats of good paint and a day's drying time between coats.

The mast can be raised easily with the help of a block and tackle, after bolting two legs to the iron angles. Hinging the mast in this fashion makes getting it into the air a relatively simple job. When the mast is upright, correct holes for bolting the mast to the iron angles can be drilled and the whole structure bolted firmly together.

Without any guy wires, a mast of this type has withstood the fury of a hurricane that was strong enough to tilt the 6-foot deep foundation.

WINDMILL TOWERS

In some parts of the country it is possible to buy old windmill towers from farmers who no longer have any need for them. While the tower can be disassembled by taking it apart piece by piece, it is a tedious process. The simpler scheme is to bolt a 10- or 12-foot piece of 2 × 6 timber across the base of the tower, as shown in Fig. 11-12, place bracing 2 × 4s between the legs as illustrated, and lower the whole thing in one piece. Use rope guys to steady the tower during the process. Once on the ground, a cold chisel can be used to cut off the bolts. Thoroughly buff off all rusty surfaces and apply a coat of aluminum paint *before* reassembling. Use new galvanized bolts. It is well also to add a drop of "No-oxide" to the threads, in case the tower is taken apart later on.

Raising the tower requires the same equipment as lowering it, as indicated in the sketch. Holes for the corner posts are dug before raising the tower, but the dirt is not tamped solidly around the corner posts until the tower is in place and has been made exactly vertical. It will be found much more convenient to install any beam antenna on the top of the tower at the time that the tower is going up than it will after the tower is in place. With the block and tackle, pull the tower up just far enough to clear the antenna array, install the antenna, and then finish the job.

In the event that lack of horizontal room prevents assembling the tower on the ground, it will have to be built up piece by piece. With a well-built and level foundation, this is a simple job for one man working on the tower and a partner on the ground to pass up the pieces.

RAISING THE MAST

Specific instructions have been given for raising the masts already described, in most cases. There are, however, a few kinks which will help in the

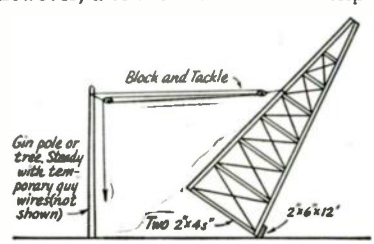


Fig. 11-12 — Suggested method for raising or lowering a steel windmill tower. The temporary 2″ × 6″ × 12′ timber bolted to the base serves to steady the tower, and the two lengths of 2 × 4 prevent any possible buckling of the base. Temporary rope guys on either side of the tower should be used to steady the structure while it is going up or coming down. After raising the tower, the two legs nearest the gin pole should be anchored first, after which the tower can be eased back and the other two legs fastened.

erection of almost any kind of antenna mast.

The "scissors" arrangement shown in Fig. 11-13, constructed of two 2 × 3s or 2 × 4s about 20 feet long, will be of considerable assistance in getting the mast off the ground. Starting out near the end, the mast is pushed up a little at a time and the scissors moved in each time to keep it from dropping back. With small masts (40 feet or so) it is the only auxiliary necessary, since by the time its length is too small to be of further service it should be possible to pull the mast up the rest of the way by the guy wires. A 20-foot ladder can be substituted for the scissors if one is available, but it does not possess the stability of the scissors arrangement and therefore cannot as readily prevent sidesway. In either case a short pole or ladder can be used to push up the mast while the scissors or ladder is being moved into a new position.

When the mast is 50 feet or more high, it will be easier to pull it up if an auxiliary mast or gin pole is used, as shown in Fig. 11-14. The gin pole should be ⅓ to ½ the height of the mast, and should be erected fairly close to the base of the mast so that the maximum possible leverage can be secured. The erection of a small auxiliary mast of this type should present no special problems. Provision should be made for keeping the guy wires of the main mast from slipping off the top of the gin pole as they are pulled back. All the back guys should pass over the gin pole and should be kept taut so that there will be no bending of the main mast as it goes up.

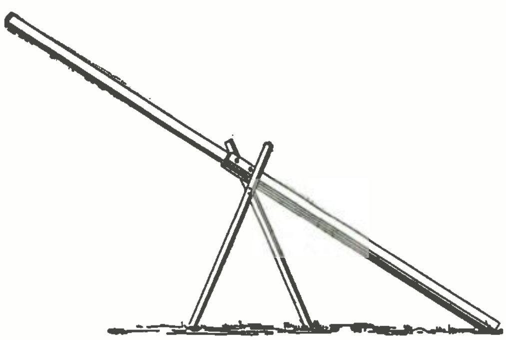


Fig. 11-13 — "Scissors" for putting up masts. As the mast goes up, the scissors are moved in to keep it from falling and to prevent sidesway.

As the mast is pulled up, the guys should be allowed just enough slack to permit the mast to move without the necessity for "pulling against the guys." With tall masts of the usual construction a too-slack guy may allow the pole to bend enough to get out of control and perhaps snap. It is advisable to take the pulling-up process slowly on that account.

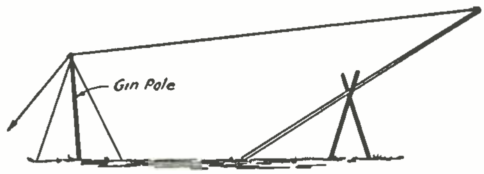


Fig. 11-14 — A gin pole is a useful accessory when a tall mast is to be raised, providing additional leverage when the mast is near the ground.

GUYS AND GUY ANCHORS

For poles up to about 50 feet, No. 12 iron wire makes a satisfactory guy (No. 12 in this wire is considerably heavier than in copper). A heavier size, or stranded cable, can be used for taller poles or poles installed in locations where the wind velocity is high.

Guy wires are normally broken up by strain insulators to avoid the possibility of their becoming resonant at the transmitting frequency. Common practice is to insert an insulator near the top of each guy within a few feet of the pole and then make each section of guy wire, between insulators, a length which will not be resonant in any amateur band to be used, either on its fundamental or harmonics. An insulator every 20 feet will be satisfactory for all bands up to and including the 28-Mc. band. The insulators should be of the "egg" type, with the insulating material under compression so that if the insulator breaks the guy will not come down.

Guy wires should *not* be fastened to the mast at equal intervals if resonance effects are to be avoided. For example, a 55-foot pole should be guyed at the top, 15 feet down, and 33 feet down.

Guy wires may be anchored in a variety of ways. Simplest of all is to anchor the wires to a tree or building, when they happen to be in convenient spots. For small poles a 6-foot length of pipe (about 1-inch diameter) driven into the ground at an angle, with the bottom of the pipe pointing to the base of the pole, will suffice. Additional bracing can be provided by using two pipes as shown in Fig. 11-15.

One form of "dead man" guy anchor is shown in Fig. 11-16. The "dead man" is a heavy plank (two 2 × 6s nailed together, for example) 5 or 6 feet long and buried about 3 or 4 feet in the

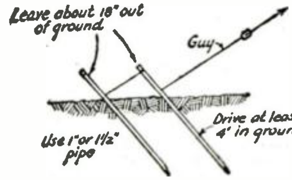


Fig. 11-15 — Pipe guy anchors. One will be sufficient for small masts, but the two installed as shown will provide additional strength for larger poles.

ground. A fair amount of surface is necessary to give maximum resistance to the pull of the guys. The wires are brought out at the proper angle to the pole so that they are, in effect, a continuation of the regular guy wires.

For heavy jobs, regular guy anchors can be secured through firms dealing in line materials, or through the local utility company.

When several guy wires are connected to a single "dead man," an open thimble (obtainable from marine supply stores) should be used to prevent sharp bends and possible breakage of the wires.

With large guy wires, it is difficult to make tight joints at the insulators even with pliers. A simple tool can be made for the purpose from a piece of heavy iron or steel with a single hole drilled about a half inch from one end. The wire is passed through the insulator, given a single turn by hand, and then held with a pair of pliers at the point shown in Fig. 11-17. By passing the wire through the hole in the iron and rotating the iron as shown, the wire can be twisted quickly and neatly.

Fig. 11-16 — The "dead man" guy anchor.

With high poles it may be advisable to use turnbuckles in the guy wires so that they can be tightened satisfactorily. With small poles this is usually an unnecessary refinement, since the wires can be pulled tight enough by hand.

HALYARDS AND PULLEYS

A free-running pulley and a long-lived halyard are definite assets to an antenna system. Common clothesline rope will be strong enough for small antennas, but does not stand the weather too well and should be renewed fairly frequently. Sash cord is a bit better, but still not weather-resistant. A satisfactory halyard is ⅜- or ½-inch waterproofed manila rope, the larger size being needed only to hold long stretches of wire.

Ordinary rope or cord can be waterproofed by soaking it a day or two in automobile top-dressing.

If it is feasible, duplicate pulleys and halyards should be installed at the top of the mast before raising the structure. This has the advantage that, if one rope breaks the antenna can be transferred to the other halyard without lowering the pole. Further, if at some future date the mast is used to support the end of another antenna, the new halyard is already available.

Halyards should always consist of a continuous loop of rope running through the pulley, because the half of the rope that has no normal strain is always available for pulling down the antenna end when the wire is lowered.

It will pay to purchase good-quality pulleys. A good grade of galvanized-iron pulley will be satisfactory in locations where the atmosphere is free from salt, but at seashore locations a pulley intended for marine use should be used. One of the best types is a hardwood block with bronze roller-bearing shaft, which will stand up well and resist corrosion under adverse conditions.

FEEDER CONSTRUCTION

Two-wire open transmission lines are readily constructed from materials available from all amateur radio supply houses. Rod-type spacers or spreaders are universally used to keep the separation between the wires constant, the usual size of spacer being 4 or 6 inches long. Six-inch spacing is quite satisfactory on the 3.5- and 7-Mc. bands, and 4 inches is good for 14 to 30 Mc. Still smaller spacing is desirable on the very-high frequencies, to minimize line radiation, and 2-inch spacers are available for the purpose.

If the line is to hang free, a 6-inch line should have a spacer every 6 feet or less, to reduce swinging of the wires with respect to each other during a wind. Correspondingly smaller separation between spreaders should be used with lines having closer spacing. Even though the wires do not swing enough to touch, the movement will vary the characteristic impedance of the line and thereby cause variable loading on the transmitter.

Manufactured spreaders, generally made of ceramic or plastic material, are inexpensive and stand up well in the weather, so that it is not usually worth while to make spreaders at home. However, there may be cases where home construction is desirable, as when special sizes are needed, and Fig. 11-18 shows how they can be made. Probably the best material for home use is polymethyl methacrylate (Lucite), because it is lightweight, easy to work, and stands up well under weathering. Ceramic materials are better able to withstand the effects of weather over long periods, however.

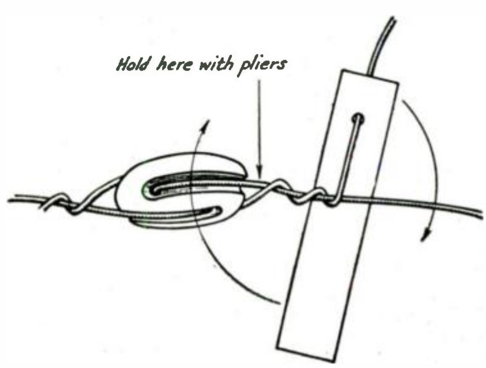


Fig. 11-17 — Simple lever for twisting guy wires.

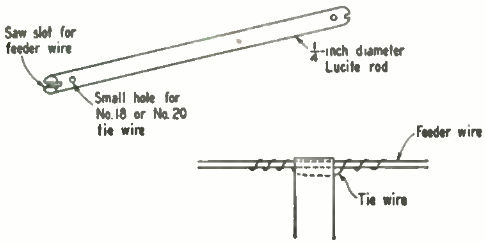


Fig. 11-18 — Details of feeder-spacer construction and installation of the spacer on the feedline.

The feedline should be made of soft-drawn wire and never hard-drawn or "copperweld," because the latter two will be found very difficult to work and straighten out without considerable tension. The tie wire that secures the feeder wire to the spacer should be wrapped tightly around the feeder wire, but it should not draw the feeder wire hard against the spacer. If the feeder wire is left loose in the groove of the spacer, it will be found that the feedline straightens out nicely under only slight tension, and there is little danger of breaking a spacer.

To keep feeders taut it is necessary to have the same tension on each wire. With an end-fed antenna this may be difficult to accomplish unless an arrangement similar to that shown in Fig. 11-19 is used. The single-point support permits pulling off the feeders at any convenient angle to the antenna. Where a feeder must be run a considerable distance so that simply letting it hang is impracticable, a support of the type shown in Fig. 11-20 may be used. A similar mechanical terminating arrangement at each end of the line will keep the spacing constant and the line shipshape.

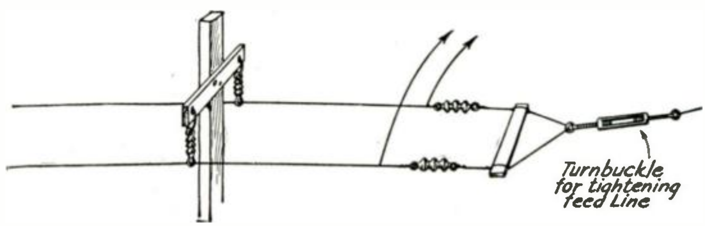


Fig. 11-20 — Method of supporting long feedlines. The wires are run through the insulators and tightened at the ends of the line. The wire is allowed to slip freely through the insulators.

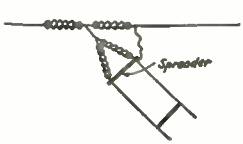


Fig. 11-19 — Method of pulling off Zepp feeders to keep the wires taut.

The popular 300-ohm "Twin-Lead" can be run along the side of a mast or building, supported at intervals by the ceramic or plastic supports that are designed for the service. It should always be supported so that it will not twist around or lay across guy wires and other metallic objects, and it should always clear a metallic object by an inch or so if an "impedance bump" is to be avoided. The impedance of the line will change in wet weather because water collects on the surface, although this condition can be minimized by coating the line with silicon greases (sold for the purpose) before the line is installed. At the point where the line joins the antenna, the line should be supported by the polyethylene dielectric rather than by the wires themselves, because when the line whips in the wind it will eventually break the wire. Small clamps of thin Lucite sheet, held together with small screws, can be used to hold the line and take the strain from the wires. Since the polyethylene dielectric will flow under the heat of an ordinary flatiron, joints in the line can be sealed over by patching with small bits of the insulating material.

Solid-dielectric coaxial line is very tough and requires no special care, except at the ends where the connections are made. Here it is wise to safeguard against moisture and condensation between the dielectric and the outer conductor, and suitable insulating tapes can be obtained in the amateur radio supply houses. These tapes are wrapped over the insulating material and the outer conductor after the braid has been stripped back on the dielectric. Solid coaxial cables can be supported where necessary by the ceramic or metal clamps designed for the purpose, or small "cable clamps" of the right size will do. Coaxial line can be run along metal surfaces without fear of affecting the impedance of the line. The only necessary consideration is to avoid bending the cable around too sharp a corner, which may cause slow mechanical flow inside the cable.

BRINGING THE ANTENNA OR TRANSMISSION LINE INTO THE STATION

In bringing the antenna or transmission line into the station, the line should first be anchored to the outside wall of the building, as shown in Fig. 11-21, to remove strain from lead-in insulators. When permissible, holes cut directly through the walls of the building and fitted with feed-through insulators of suitable size are undoubtedly the best means of bringing the feeders into the station, for the job can be done with little difficulty and can provide greater mechanical permanence than other schemes. It involves no interference to screening or storm windows. The holes should have plenty of air clearance about

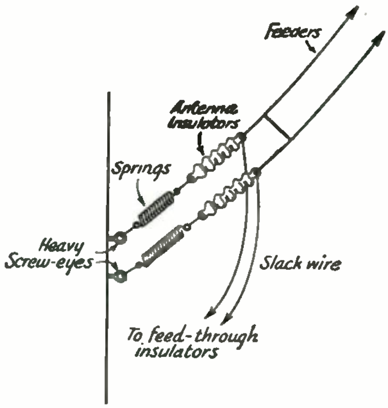


Fig. 11-21 — Feeders anchored to take the strain from feed-through insulators or window glass.

the conducting rod, especially when tuned lines, which develop high voltages, are employed. Probably the best place to go through the walls, from the standpoint of appearance, is the trimming board at the top or bottom of a window frame which provides flat surfaces for tightening lead-in insulators. Cement or rubber gaskets may be used successfully to waterproof the exposed joints.

Where such a procedure is not permissible, the window itself usually offers the best opportunity. One satisfactory method is to drill holes in the glass near the top of the upper sash. If the glass which is to be drilled is replaced by plate glass, a stronger job will result. Plate glass may be obtained reasonably from automobile junk yards and may be drilled before placing in the frame. The glass itself provides the necessary insulation and the transmission line may be fastened to bolts fitting the holes. Rubber gaskets cut from an inner tube will render the holes waterproof. The lower window sash should be provided with stops at a suitable height to prevent damage when it is raised.

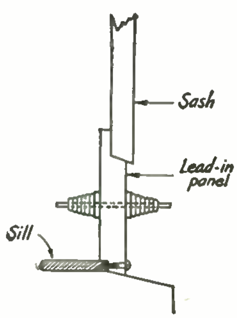


Fig. 11-22 — Antenna lead-in panel. It may be placed over the top sash or under the lower sash of window. The overlapping joint makes it weatherproof. The single thick board may be replaced by two thinner boards fastened together.

In a less permanent method, the window is raised from the bottom or lowered from the top to permit the insertion of a board three or four inches wide which carries the feed-through insulators. This arrangement may be made weatherproof by making an overlapping joint between the board and window sash, as shown in Fig. 11-22, and covering the opening between upper and lower sashes with a sheet of soft rubber cut from an inner tube.

Fig. 11-23 shows a simple and satisfactory method for keeping rain from running down the feeders and into the lead-in insulators. The short lengths of wire are pinched on the feeders far enough out to clear the building so that the water does not drip on the sill. These "drip wires" also help prevent ice formation around the lead-in insulators in winter.

Three hundred ohm Twin-Lead can be brought into the house through a window, between a sash and sill, by clamping it between two pieces of wood notched out to pass the line. The line should be clamped outside the window at another point as well, if the line makes a bend as it enters the house. Solid-dielectric coaxial line can be run into the house in any way that will be weatherproof, and here again clamping it between two pieces of wood is probably the simplest device.

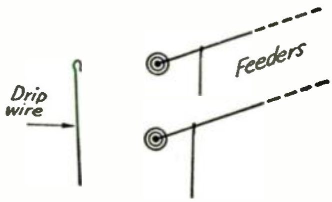


Fig. 11-23 — Drip wires for feeders.

DRILLING GLASS OR CERAMICS

Glass and ceramic materials may be drilled by several methods, one of which involves the use of a special drill of the type shown in Fig. 11-24, ground from a piece of ⅜-inch drill rod. To drill a hole, place the drill in a hand brace, engine lathe or slow-speed drill press. If the material to be drilled is flat, such as plate glass, make sure that the supporting surface is flat. Apply turpentine to the point of the drill and press firmly against the work in the desired location. Then turn the drill slowly and apply sufficient turpentine to keep the drill wet at all times. Use care when breaking the point through the work to avoid chipping. After the point has broken through, turn the work over and drill from the opposite side, repeating this operation as often as is necessary to keep the edges of the hole nearly parallel.

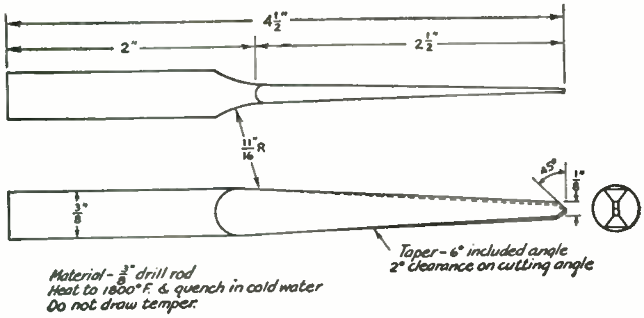


Fig. 11-24 — Drill for cutting glass.

LIGHTNING PROTECTION

Some form of protection from heavy induced charges on the antenna from lightning discharges in the vicinity always should be provided. The conductors in an ordinary antenna system are not heavy enough to handle a direct stroke, but fortunately these are rare.

One method satisfactory to the fire underwriters is the use of a special "lightning switch," somewhat larger than the ordinary porcelain-base knife switch, by means of which the feeders are grounded when the station is not in use. The ground wire should run directly to a good ground connection preferably, although not necessarily, outside the building. Alternatively, a similar ground wire with a clip which can be fastened on the feeders may be substituted for the switch.

Automatic protection can be secured by installing spark gaps between the feeders and a good ground. A good method of making such a gap is shown in Fig. 11-25. The gap electrodes are made of pieces of heavy wire. Rubber cones cut from an old inner tube, slipped over the upper electrodes, protect the gap from short circuit by snow or rain.

It is a good plan to get in touch with a fire insurance agent or the city inspection department to ascertain local requirements with respect to antenna lightning protection. *Safety Rules for Radio Installations*, handbook No. 9 of the Bureau of Standards also will provide useful information. It is available from the Superintendent of Documents, Government Printing Office, Washington, D. C., for ten cents.

REPLACING BROKEN ANTENNA HALYARDS

In the event that the antenna rope breaks, getting a new rope through the pulley may involve taking down the mast if the latter cannot be climbed. There are several ways in which a new pulley and rope can be installed at the top, however.

Some schemes make use of the top guy wires in coaxing a new pulley, fitted with a new halyard, to the top of the mast. If you have a second mast or can make use of a tree or housetop or temporarily-erected support, the method shown in Fig. 11-26A is probably one of the easiest to execute. One of the top guys is set free. The new pulley with halyard is fitted with a heavy metal ring or a loop of several turns of heavy wire and the loose end of the guy wire is passed through this loop. The loop should be large enough to pass easily over the guy-wire insulators. A light cord is tied to the loop and the free end of the guy wire is tied to the halyard from the second support, hoisted up and pulled tight. It should then be possible to make the loop slide along the guy wire toward the top of the mast by shaking the new halyard and pulling on the assister cord from a distance. In some cases, it may be possible to coax the loop up over the top of the mast if one has sufficient patience and the top guys are not fastened too far from the top of the mast, although this is not necessary. When the loop has been worked up close to the mast, it may be held there by the assister cord while the guy wire is lowered. While holding the free end of the guy wire, several turns about the mast should be made by walking around the mast outside all other guys. This will bind the loop securely to the mast. A sharp yank will break the assister cord after the job is finished.

Another scheme which may be tried is shown in Fig. 11-26B. A loop of wire, as previously described, is passed around the rear top guy wire. If the loop is covered with tape or a section of old bicycle tire or garden hose, it may slide more readily on the guy wire. The new halyard and the halyard from the second support are tied together and a large slipknot is tied in the other side of the new halyard to prevent the new halyard from

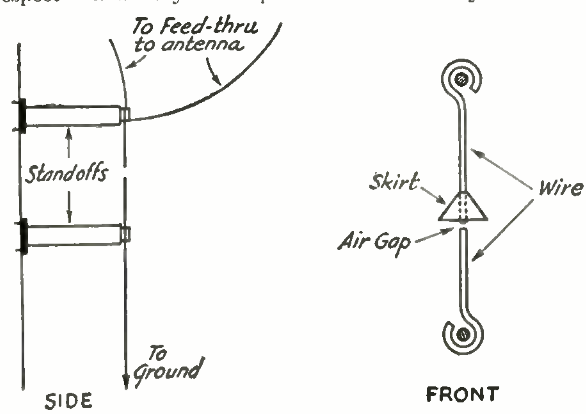


Fig. 11-25 — Spark gap for static discharges.

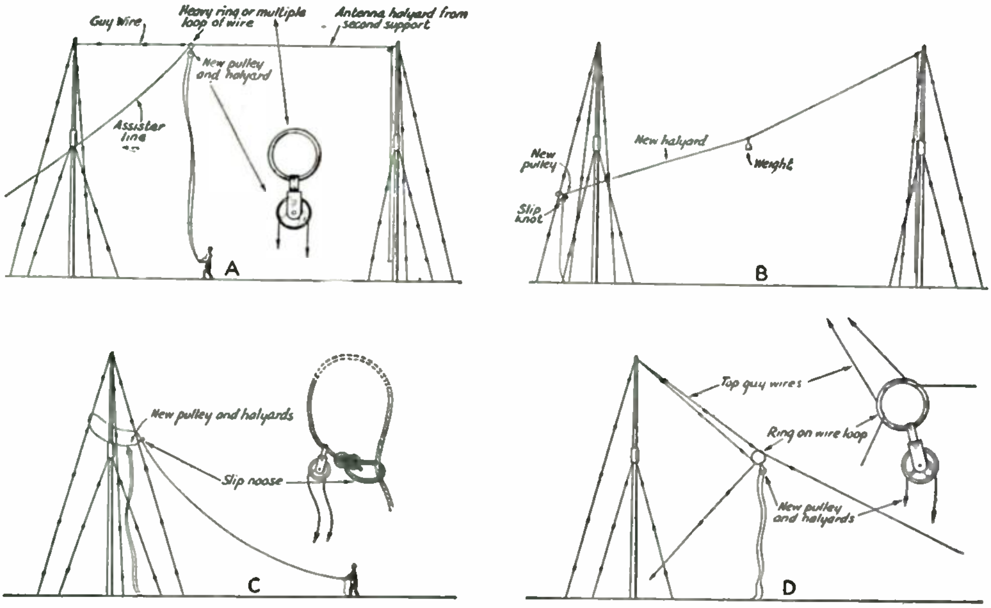


Fig. 11-26 — Schemes for replacing broken antenna halyards.

running through the pulley when it is pulled up the guy. Alternatively, the two ends of the new halyard may be tied together and then tied to the halyard from the second support. An assister cord tied to the wire loop might be helpful in getting the loop over insulators; shaking the guy wire should also help. When the pulley reaches the top of the mast, the guy wire is wrapped around the top of the mast as previously described. A sharp yank on the free end of the new halyard will take the slipknot out. It might be a good idea to tie a weight between the two halyards to make sure that they will fall to the ground when released.

If no second support is available there are other ways, one of which is shown in Fig. 11-26C. Pass a heavy rope around the outside of all top guy wires. Then pass the rope through the eye of the new pulley fitted with the new halyard and form a slip noose. By shaking and pulling the rope, it should be possible to work the loop up the guy wires to the top. Best results will be obtained by working the rope at a fairly good distance. If the loop becomes caught on an insulator, a friend can assist by sliding the pulley along the loop to a point near the insulator and whipping the halyard. When the noose reaches the top of the mast, its rope should be made fast to the pole.

Rotary-Beam Construction

The deserved popularity of rotary-beam antennas has stimulated interest in the purely mechanical problems of the construction of suitable supporting structures and their rotation. The schemes devised to rotate antennas vary from very simple and inexpensive ones to elaborate and costly arrangements. The methods shown in this chapter are presented chiefly to indicate general principles and not necessarily to give full descriptions suitable for exact duplication. In other words, the accent is on ideas and only secondarily on details. All of them have been in everyday use at various amateur stations, however, and are thoroughly practical.

Naturally the size of the structure will depend upon the frequency, and for this reason the use of rotating antennas is confined to 14 Mc. and the higher-frequency bands. Some of the structures shown here are for 14-Mc. and others for 28-Mc. antennas. Most of them readily can be adapted to different electrical systems which may be selected after consultation of Chapter Eight.

A number of 2-, 3- and 4-element beams are currently offered by manufacturers, and in many cases these beams will save the amateur time in acquiring materials and building a beam. Several rotating heads for beams are available from specialists in this field, and even metal towers can be obtained that are specially designed for amateur work.

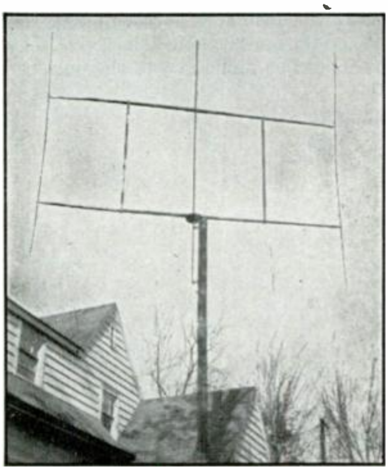


Fig. 12-1 — **In this vertically polarized system the antenna or driven element is fixed, while the parasitic elements rotate around it. Construction details are given in Fig. 12-2.**

Most amateurs want horizontal polarization on 14 Mc. and higher, to reduce the response on receiving man-made noise. However, if the location is fairly free from this type of interference the vertically-polarized antenna can be expected to be equally as effective as its horizontal counterpart. Further, in some instances TVI can be reduced by using vertical polarization, since TV antennas are horizontal and do not respond readily to other planes of polarization.

● A VERTICAL ROTARY ANTENNA

Vertical beam systems have a distinct advantage over horizontal antennas in that when parasitic director or reflector elements are used the antenna element itself need not rotate. It is only necessary to rotate the director or reflector *around the antenna*. There is, therefore, no necessity for special feeder contacts; the feeders may be installed and connected just as though the antenna were not a rotary.

Fig. 12-1 shows a photograph of a rotary beam using this principle, a 28-Mc. antenna used at W2BSF. The mast on which it is installed is a 20-foot 4 × 4. As shown in Fig. 12-2, the detailed drawing of the antenna, the antenna element is a section of pipe fitted with bearings on which the rotating assembly turns. The director and reflector are at the extremities of the wooden framework. The whole beam is rotated by means of ropes and pulleys which run to the station. For electric rotation it would be relatively easy to arrange a belt drive to replace the ropes, mounting the motor in a housing at the top of the pole.

● A TRANSPORTABLE 10-METER BEAM

The beam shown in Figs. 12-3, 12-4 and 12-5 weighs only 16 pounds (the boom alone weighs only 4½ pounds), and it can be dismantled, together with its mast, into lengths of not over 12 feet. Wire braces permit using light and cheap aluminum angles for the boom, and also provide an easy way to correct any horizontal misalignment of the elements. Element lengths and spacings are given in Fig. 12-4.

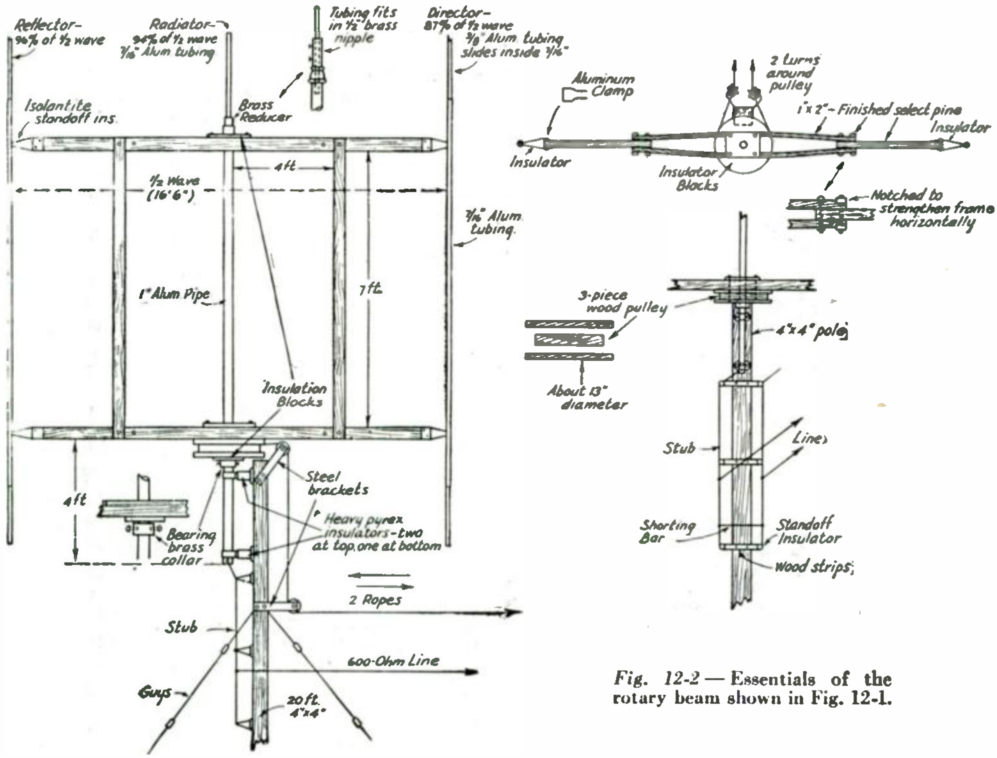


Fig. 12-2 — Essentials of the rotary beam shown in Fig. 12-1.

The elements are made of a center section of 1-inch o.d. aluminum tubing, $\frac{1}{16}$-inch wall, with ⅞-inch o.d. sections telescoping in the ends. The boom is made of two parallel aluminum angles 6 inches apart, screwed at both ends to a 2 × 2 wood block. The angles are 1 by 1 by $\frac{1}{16}$ inch, and come in 16-foot lengths. The excess lengths (about 4 feet each) are fastened alongside the regular lengths at the center of the boom to make it more rigid.

The center support is 6 by 11 by 1½ inches (two ¾-inch boards nailed together), with two ¾-inch pipe flanges bolted together through it from opposite sides. The exact location of this center support with respect to the boom should be determined by balancing the boom with the elements lashed temporarily in place; it will be about 5½ feet from the forward end of the boom. After the center support is attached with screws, the elements are attached by means of metal straps about 1 inch wide. The straps holding the director and reflector are made a little longer on the inner ends, and an extra hole is drilled in each strap to provide points for attaching the bracing wires. A 15-inch length of ¾-inch pipe is screwed into the upper flange and drilled at the top to provide a point for attaching the upper ends of the braces. Small turnbuckles are inserted in the bracing wires to permit correction of minor misalignment of the elements.

The mast is made of water pipe supported by three guy wires, and is 35 feet high. The lengths are 12 feet or less and all joints are treated with pipe-thread compound so that the sections can be disassembled. The lower sections are 2-inch pipe, tapering to 1½-inch pipe about halfway up.

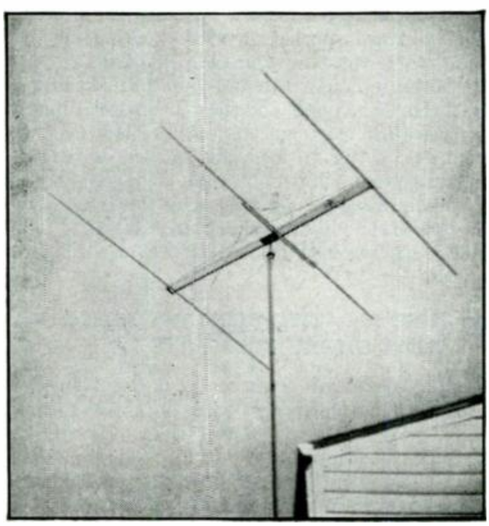


Fig. 12-3 — The mast and beam of the transportable beam rotate as a unit in a guyed bearing at the top.

The final foot or so is ¾-inch pipe, the reducer at this joint also serving as a bearing surface for the guy-wire bearing.

The guy-wire bearing is an iron disk with a hole in the center to pass the ¾-inch pipe, and has three holes 120 degrees apart near the rim to take the guys. The edges of these holes will shear

Fig. 12-4 — **Construction of the transportable 28-Mc. beam. The reflector spacing is 6 feet 10 inches, and the director spacing is 5 feet 2 inches. The over-all length of the "T" match (between clamps) is 24 inches for 300-ohm line, with 2-inch separation between matching line and driven element. Element lengths can be found from Fig. 8-13.**

through the guys unless thimbles, sister hooks, or chain links are used. The disk rotates smoothly on the reducer between the ¾-inch and the 1½-inch pipe. The guys are made of cheap No. 12 galvanized-iron clothesline wire, and are broken at intervals by porcelain strain insulators.

A lower bearing in the form of a strong wooden box nailed to the side of the house can be used halfway down the mast. The box is notched to pass the pipe, and a metal strip screwed across the notch after the mast is in place. An alternative is to use another guy-wire bearing, at the point where the mast diameter increases.

The beam is fed with 300-ohm Twin-Lead led up through the mast, and the "T"-match dimensions are for this type of feedline. However, it would be better to feed the antenna with two equal lengths of RG-11/U cable, tying the outer conductors together and running the inner conductors to the "T" match. In this case, the "T"-match spacing would be reduced over the figure shown.

A 28-MC. TWO-ELEMENT BEAM WITH SHORT ELEMENTS

The two-element beam shown in Fig. 12-6 uses wire-wrapped plastic fish poles for the elements, and the over-all length of each element is only 13 feet. The plastic fish poles (sold under the trade name "Conolon") are 6½ feet long. Two are required for an element, and each pole is wrapped with a spiral of No. 14 wire. A 7-foot length of wire is required to resonate at 29 Mc. Two wire-wrapped poles placed butt-to-butt form an element equivalent to the normal 16-foot tubing element.

To keep the losses down in the antenna, no attempt should be made to use close spacing between driven element and reflector, and a spacing of 8 feet should be used. The driven element can be fed at the center through 72-ohm line, with no matching device required.

The light weight of the elements makes it possible to use a very simple supporting structure, and the boom and braces can be made of 2 × 2 lumber. The poles are stronger than aluminum tubes of equivalent weight, and they do not sag or take a permanent set.

A 4-ELEMENT 10-METER BEAM

The beam shown in Fig. 12-7 illustrates several principles of beam construction that are worth considering. Its dimensions were obtained by a series of tests on beams for 144 Mc., after which the same proportions were applied to the lower-frequency band. The boom material is 2-inch square 24S or 61S aluminum, through which are mounted the ¾-inch diameter elements. The elements can be secured in place with "J" hooks through the bottom of the boom. The beam can be fed with 300-ohm line, but the dimensions given for the "T"-match are for use with 52-ohm coaxial cable. Only a single piece of coaxial line is used to feed the antenna, because the extra half wavelength of coaxial cable results in a balanced system. This construction and feed can, of course, be applied to other frequencies and other combinations of element length and spacing. The extra piece of coaxial line is reduced in length over a half wavelength in air by the velocity factor of the line (0.66 for RG-8/U 52-ohm line).

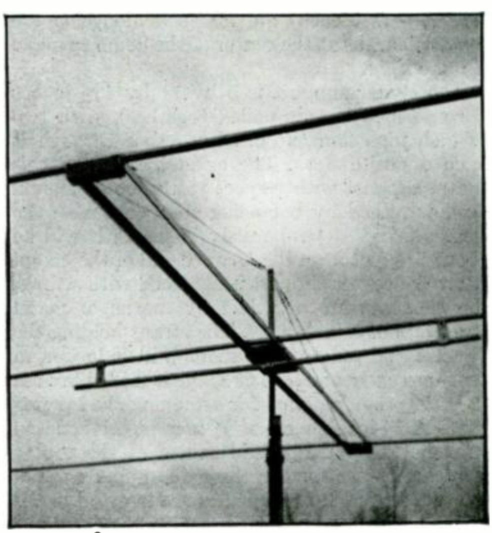


Fig. 12-5 — **Another view of the aluminum-angle boom, showing how the guy wires are terminated.**

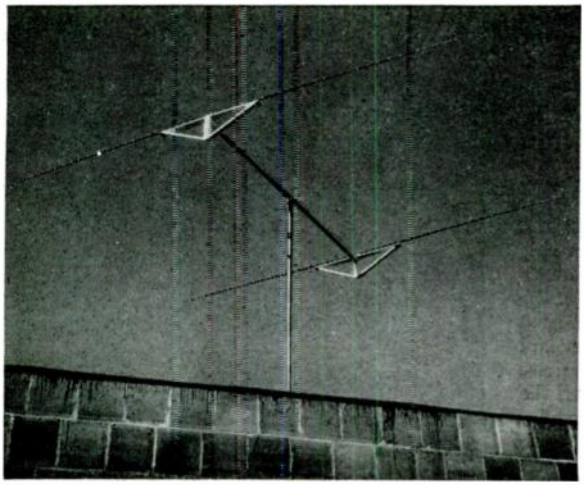


Fig. 12-6 — A lightweight 28-Mc. 2-element beam using wire-wound fish-pole elements.

A MULTIWIRE-DRIVEN-ELEMENT 28-MC. BEAM

To avoid the need for "T"-matches and other devices between the driven element and the feed-line, the impedance of the driven element can be raised by making the element of several conductors (see Chapter Four). As a result, the antenna offers a better match over a wider frequency range, and the standing-wave ratio on the feedline remains more nearly constant. The construction of such a driven element is not difficult for 28 Mc.

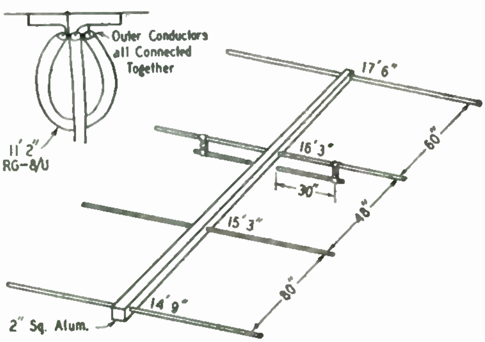


Fig. 12-7 — A 4-element "plumber's delight" antenna featuring a square-cross-section boom and a half-wave-length balancing section for the coaxial feed line. The dimensions are for a 29-Mc. design center.

One such antenna is shown in Figs. 12-8 and 12-9. The radiator consists of ½-inch aluminum tubing, paralleled with six No. 12 wires spaced 4 inches from the center of the tubing, arranged symmetrically around it and connected only at the ends, as shown in Fig. 12-9. The tubing is separated 2 inches in the center for connection to a 2-inch 500-ohm line of No. 14 wire. (If 300-ohm line feed is used, only five wires should be used around the center conductor.) The size of the tubing and wire and the spacing between wire and tubing do not appear to be critical, judging from the results obtained by several experimenters. Various construction methods may be used to furnish support for the wires. One way is to use small rods, 8 inches in length, arranged like the spokes of a wheel, at each end of the tubing. A disk or pointed star welded to the ends of the tubing is equally satisfactory. Some means of support should be used about one-third of the distance in from each end, and polystyrene or Lucite is a good lightweight material for this purpose.

An antenna of this type does not lend itself well to the all-metal "plumber's delight" type of construction, and a wooden boom is suggested.

DIRECT DRIVE OF SMALL BEAMS

In cases where the radio shack is mounted directly under the roof (as in an attic or top floor), it will often be found convenient and practical to use the roof for supporting the beam antenna, instead of going to the complication of a separate mast or tower. If desired, the antenna can be rotated by hand, and thus the attendant problems of drive motors and direction indicators can be ignored.

While there are many methods of running the antenna drive shaft through the roof, Figs. 12-10, 12-11 and 12-12 show how the problem was solved in one instance. A length of 1-inch black iron water pipe was used for the antenna mast, and a piece of 1¼-inch pipe served as the housing. Other diameters can be used, of course, depending upon the magnitude of the load to be carried. With a 1-inch pipe mast, it is not advisable to extend it more than about 4 feet beyond the housing, unless the antenna system is quite light and has low wind resistance.

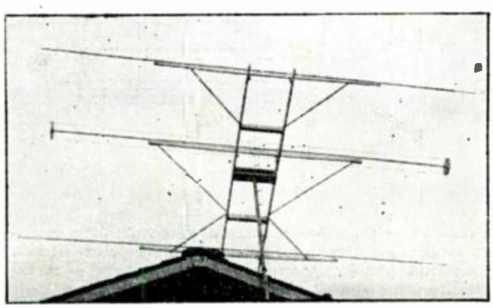


Fig. 12-8 — A 3-element rotatable antenna with a multiconductor driven element.

The block *B* in Fig. 12-11 is made from an 8-inch cube of fir or hard pine. A hole to clear the housing pipe is bored about ⅔ through the block, and then the block is cut at an angle to match the pitch of the roof. The clearance hole in the roof should be cut from the inside, to insure missing ceiling and roof rafters and, once spotted and started with a small bit, it can be brought up to proper size with a keyhole saw.

The lower brace, *C* in Fig. 12-11, is a 2 × 3 long enough to straddle two rafters. Bore a hole in it, at the point directly under the roof hole, to pass the mast housing. Place the upper end of the housing (threaded to accommodate two flanges) through the hole in the roof, and the lower end through the brace *C*. Take a two-foot square of zinc or copper sheeting and form it around the roof block to make flashing. This is added insurance against leaks and also will protect the block from rot. Leave plenty sticking out on all four sides for tucking under the shingles. A pair of heavy shears, a chisel and a hammer are all the tools needed for working on the flashing.

To install the block *B* and the flashing, first clear away any shingle debris from around the hole in the roof, to leave a flat surface for the block. Smear the surface liberally with heavy roofing cement, place the block down over the housing, and screw it to the roof, using 3-inch wood screws. Smear all the joining surfaces between the block and the roof with cement and put on the prepared flashing. Tuck the side and top edges under the adjoining shingles and nail them down with galvanized shingle nails. Put a gob of cement over all nail heads and joints in the flashing. Do the same around the housing where

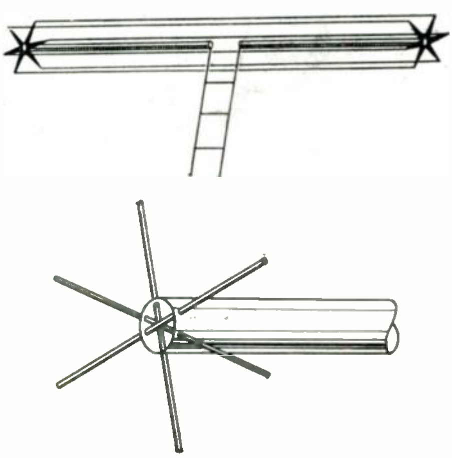


Fig. 12-9 — The essential plan of constructing a multi-element cage antenna for use as the driven element in a beam antenna (top), and details of one type of metal wire-element spacer that is fitted on the outer ends of the tubing element. The length of the driven element, as obtained by formula or chart, should be decreased by twice the diameter of the cage.

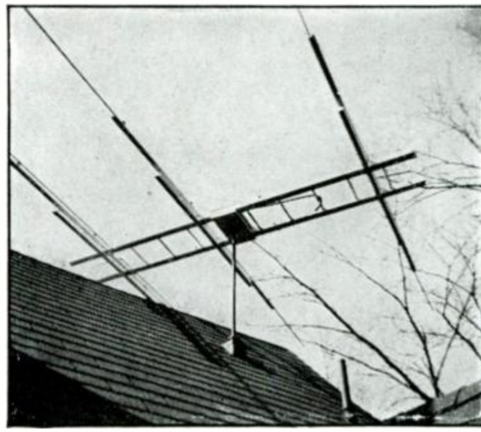


Fig. 12-10 — A roof-supported 28-Mc. beam. The pipe mast projects through the roof and rests on a bearing made from pipe flanges. A housing of larger pipe, inside the attic, serves as a sleeve in which the mast can rotate.

it projects through the block and then screw on the first flange. You may need a helper inside the attic to hold the housing with a pipe wrench. Make small punch holes through the flashing and screw the flange to the block with 1½-inch screws. (Both bottom flanges, where opposite flanges are called for, must be tapped out to remove the taper. A plumber or trade school can do this by running a tap through from the flat side of the flange.)

Screw the bearing flange on with the flat side up. If any of the housing projects above the top of the flange it must be removed with a hack saw and filed to leave a smooth surface. A little care in judging the length of thread required, when cutting, will make this step unnecessary. Return to the attic and "true" the housing to perpendicular and nail down the bottom brace.

To make the mast, measure down four feet, or the height selected for the antenna above the bearing block, and drill opposing holes, as shown in Fig. 12-12. Tap these holes for ⅜-24 screws. Screw the bolts in slightly more than the thickness of the pipe. Take a 1-inch bearing flange that has previously been bored out to slide over the pipe and notched as shown, and slide it on to the bottom of the mast, with the notched side up to accept the bolts. The heads of the bolts will come close enough to the upper surface of the flange, when it is in position, to keep them from turning out. Drop the mast into the housing and let it rest on the top surface of the ceiling. Mark and cut a hole through the ceiling just big enough to let the mast slide through. A nickeled floor plate on the pipe inside your shack will dress it up and keep the hole from showing. Fiber wheel grease should be smeared on the lower bearing flange before the mast is finally dropped in place. It keeps out moisture and allows easier turning.

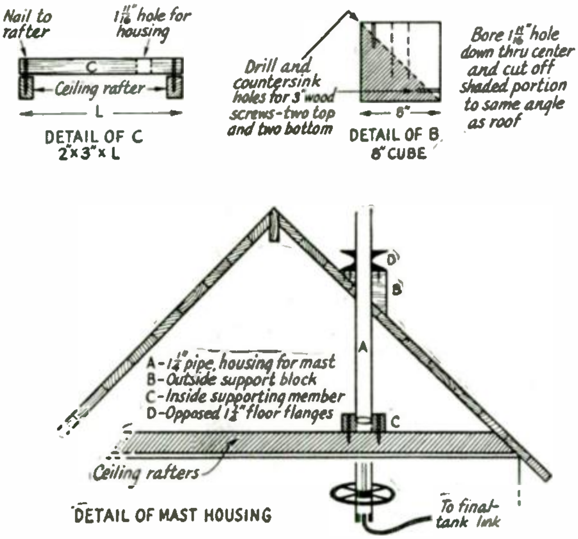


Fig. 12-11 — Showing the attic housing, with the mast protruding through the operating-room ceiling and automobile steering wheel for turning the antenna. Dimensions of the housing and mast will depend on the individual location.

A steering wheel procured from the local junk-yard can be put on the bottom of the mast for easy turning. The hub should be bored out to pass the mast, and should also have a tapped hole for a $\frac{3}{8}$-inch setscrew to hold it tightly. A "dimple" drilled into the mast where the setscrew is to go will insure a nonslip fit. Paint the section of mast which comes inside the room, as well as the wheel, with enamel. A strip of white tape around the wheel, at the point corresponding to the direction in which the array is pointed, effectively completes the job. The wheel should be about 2 feet down from the ceiling, if it is going to be over the operating position, where it can't be walked into.

Before drilling holes to pass the two sizes of pipe used, be sure to check the outside diameters as there may be some variation in size with different manufacturers. Bore the holes at least $\frac{1}{32}$-inch larger than the outside diameter of the pipe to avoid troubles due to a too-close fit. When larger sizes of pipe are used, the space between the outside diameter of the smaller pipe and the inside diameter of the larger may be greater than can be tolerated. If this is the case it would be possible to make a good fit by taking two reducing fittings (larger to smaller) and boring out the smaller ends to pass *over* the smaller pipe. Cotton waste soaked with grease should be packed in before screwing on the reducers at either end of the housing. This makes a water-tight packing similar to that used in centrifugal water pumps. See Fig. 12-13.

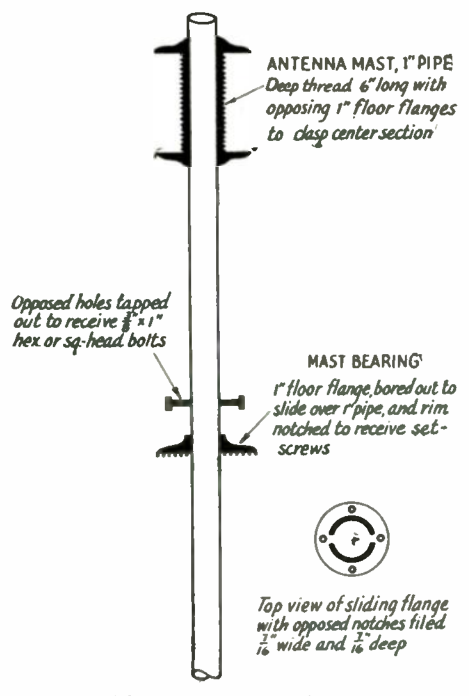


Fig. 12-12 — The rotatable mast. The two pipe flanges at the top clamp the antenna boom, while the bolts ride in notches cut in the third flange. The latter rests on assembly *D* in Fig. 12-11.

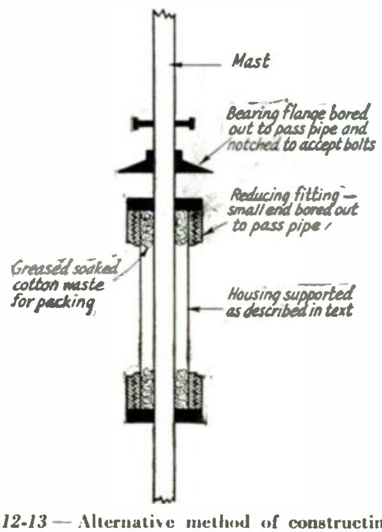


Fig. 12-13 — Alternative method of constructing the housing when larger pipe sizes than those used in the author's beam are necessary.

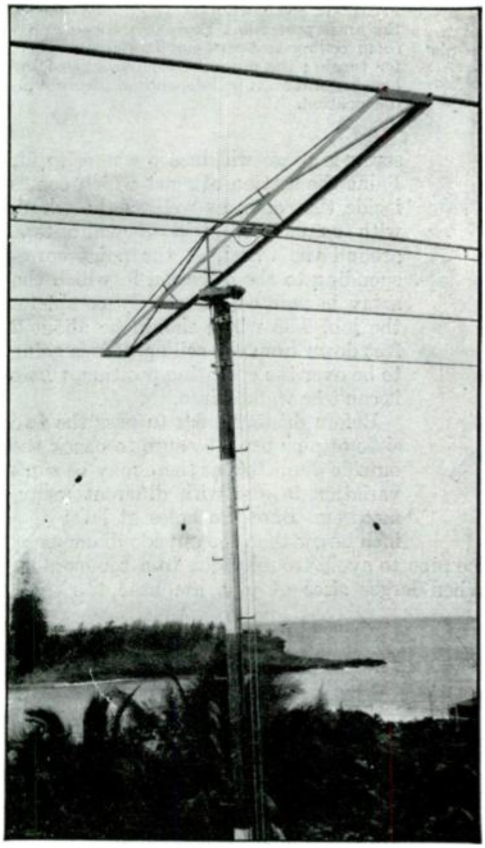


Fig. 12-14 — A four-element 14-Mc. beam of lightweight all-metal construction. Fed by coaxial cable and hand-rotated, the antenna and boom assembly weighs only 40 pounds. (KH6IJ, Dec., 1947, *QST*.)

A LIGHTWEIGHT 14-MC. BEAM

By making use of tubing or duralumin angle, a lightweight boom for a 20-meter antenna can be built. The four-element beam shown in Figs. 12-14, 12-15 and 12-16 is an example. It uses 1¾-inch angle for the main pieces and ¾-inch angle for the other members, and the entire framework plus elements weighs only forty pounds. This simplifies considerably the problem of supporting the beam.

The following aluminum pieces are required:

4 — 1-inch diameter tubing, 12 feet long, 1/16-inch wall

8 — ⅞-inch diameter tubing, 12 feet long, 1/32-inch wall. Must fit snugly into 1-inch tubing

2 — 1¾-inch angle, 21 feet long

2 — ¾-inch angle, 21 feet long

4 — ¾-inch angle, 1 foot long

2 — ½-inch diameter tubing, 6 feet long

Aluminum tubing and angle corresponding to the above sizes can possibly be bought from scrap dealers at reasonable prices, if not directly from the manufacturer. If the sections of the elements do not fit snugly, insert shims or make some other provision for a tight fit, since the appearance of the beam will be spoiled by sagging elements. Some amateurs reinforce their beam elements with copper-clad steel wire supported a foot above the elements at the boom and tied to the extreme ends of the elements.

As shown in Fig. 12-15A, two 1¾-inch aluminum angles 21 feet long serve as the main members of the boom. They are spaced one foot apart. The elements are spaced 7 feet apart. Wooden spacers of 2 × 2 are placed at the ends of the boom and screwed on with brass screws. These spacers are also placed under each element where it crosses the boom. These spacers may be unnecessary if the elements are bolted to the boom, but if the construction is as in Fig. 12-15B the spacers are recommended.

The cross braces shown in Fig. 12-16 are put into position at the very last, after the beam is hung in position on the central pivot, since they offer a means for truing up minor sag in the elements.

The central pivot consists of a structure made from ¾-inch angle iron and ½-inch pipe, as shown in Fig. 12-15C. It has to be brazed. The crossbar rest is made separate from the boom and central pivot, and affords a means for tilting the beam when unbolted from these structures. The ½-inch pipe is drilled for the coaxial line that is fed through this pipe. The pinion gear on the ½-inch pipe should be brazed on.

A washing-machine gear train is well suited for this type of beam. Another possibility (used in this instance) is a discarded forge blower. It was fitted with a ½-inch pipe which serves as the central pivot. The gear train ends up in a "V" pulley, and the beam is easily rotated by a system of ropes and pulleys that ends up in an automobile steering wheel at the operating position. A plumb bob attached to the shaft of the steering wheel serves as a direction indicator. A small cardboard scale mounted along the line of plumb-bob travel can be readily calibrated to show the direction of the beam.

The supporting structure for this beam consists of a 4 × 4 pole 30 feet long, with ten-foot extensions of 2 × 4 bolted to both sides of the bottom, making the total length about 36 feet. Two sets of guy wires should be used, approximately 2 feet and 15 feet from the top. As an alternative, the pole can be set against the side of the house, and only the top set of guys used to provide additional support.

With all-metal construction, delta match and "T" match are the only practical matching methods to use to the line, since anything else requires opening the driven element at the center, and this complicates the support problem for that element.

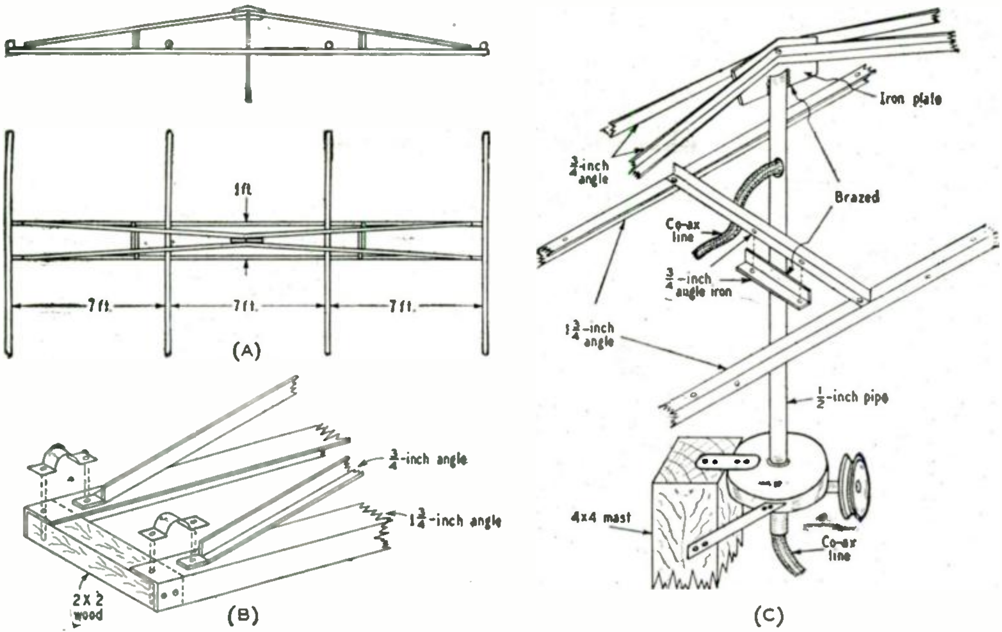


Fig. 12-15 — **Details of the 4-element beam construction. The general dimensions and arrangement of the beam are given in A, the detail of the ends of the boom is shown at B, and C shows the construction of the central pivot. A discarded forge-blower gear train is used to drive the assembly.**

"PLUMBER'S DELIGHT" BEAM FOR 14 MC.

The all-metal "plumber's delight" type of construction is attractive for beams because, by using a minimum amount of material, it can be made to have low wind resistance and it doesn't appear as bulky as other types when it is up in the air. The beam shown in Fig. 12-17 is a typical example of clean-cut beam construction, and some constructional hints and details are given in Fig. 12-18.

The supporting boom is 22 feet long, permitting 12-foot spacing of the director and 10-foot spacing on the reflector side. Moving the driven element a little to one side of center is an aid in mounting the beam. The boom is made of two 10-foot lengths of 3-inch diameter 24ST dural tubing of 0.072-inch wall thickness. The two sections are spliced together with a 3-foot length of 6 × 6-inch oak that has been turned down at each end to fit inside the tubing. The center of the block is left square to provide a flat surface for attaching to the vertical rotating pipe. At each extremity of this boom is cut a horizontal hole the exact diameter of the parasitic elements. A Greenlee socket punch of the right diameter can be used to cut the holes, if the work is not done in a machine shop. A square should be used to align the center holes, so that the elements will be at right angles to the boom.

A two-foot length of ¾-inch pipe, complete with flange mounting plate, is bolted to the top surface of the oak center block, and a single umbrella guy is run to each end of the boom. An egg insulator and a turnbuckle are placed in each guy. The 'buckles should be tightened until there is no sag in the boom when it is supported at the center (see Fig. 12-18A, and then safety-wired. Finally the center block should be given a good coat of paint or varnish to prevent it from splitting.

Each of the three elements is composed of a 12-foot length of 1⅝-inch-diameter 0.050-inch-wall 24ST dural tube, with each end slotted for about four inches. This slotting operation can be done easily with a hack saw. Into each end of this tube is pressed a 12-foot length of 1½-inch-diameter 0.032-inch-wall 52ST tubing, as shown

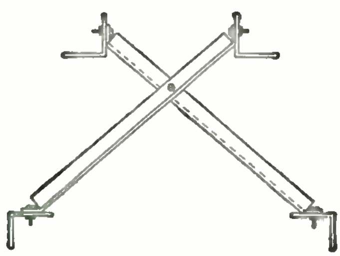


Fig. 12-16 — **The boom for the 4-element beam is cross-braced at two points, about 6½ feet in from the ends.**

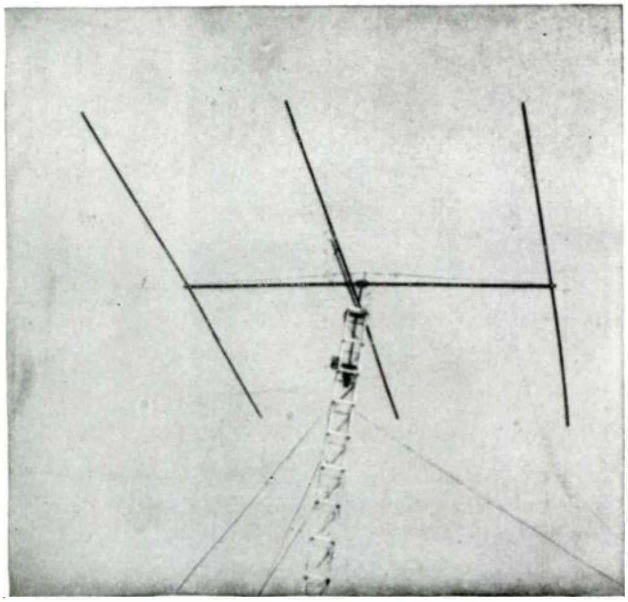


Fig. 12-17 — A 14-Mc. "plumber's delight" beam. (W6TEZ, Feb., 1949, *QST*.)

inserted in the boom hole. The end tips may now be inserted in the reflector and director and the clamps tightened.

The radiator is placed atop the center boom, a little off center in order that it will clear the center stay. The radiator is attached to the boom by a special clamp, constructed as illustrated in Fig. 12-18E. Two pieces of iron pipe a foot long each are obtained. These should be of proper inside diameters to slip tightly over the boom and radiator, respectively. These pipes are then cut lengthwise into two pieces and two of the halves welded to each other back-to-back at right angles to form a mounting that will sit astride the boom and provide a cradle for the radiator. This mounting should be bolted to the boom by means of three ¼-inch bolts 4 inches long. The mounting should be placed as close as possible to the center of the beam, so that at least two of the mounting bolts can pass through the oak block. The radiator is seated in the cradle and held in place by two adjustable aircraft-tubing clamps.

The "T"-match section is made of two 4-foot pieces of 1-inch diameter dural tubing joined together by a 1-foot piece of oak dowel rod (broomstick). The tubes are driven onto the rod until they are spaced two inches apart. Holes are then drilled through the tubes on each side of the joint, and two machine screws are inserted for the connection of the transmission line. The "T" is connected to the antenna by two brass clamps, fashioned of 1-inch brass strip and formed as shown in Fig. 12-18D.

Power for rotation of the beam is the surplus "prop-pitch" motor available from many sources for a modest sum. A pipe flange is welded to the spline gear, and a threaded section of 1⅝-inch iron pipe is used as a supporting and rotating member. To prevent slipping of the threaded joint, it is pinned by a ¼-28 bolt after assembly. It is a good idea not to let the pipe exceed twelve feet in length or else it will develop axial twist in a heavy wind and allow the beam to whip about.

Several different systems have been developed for mounting the beam atop the pipe. The completed beam is light enough to be pulled up a tower by a rope or passed up hand-over-hand, and when it arrives at the top it can then be swung

in Fig. 12-18B. The correct element length is set by changing the overlap of the tubes. To prevent oxidizing at the joints, a special compound that is used in the aircraft industry to seal aluminum joints against oxidization can be used. One source of this compound, or paste, is in large aluminum electrical lugs that may be obtained at an aircraft-surplus supply store. These lugs are filled with the paste and are capped with a red plastic cover. The paste is a mixture of grease and metal filings — the grease to keep the air from the joint and the metal filings to pass the current through the joint. A dozen of these electrical lugs can be bought and the paste extracted and smeared inside both ends of the three center tubing sections. As an added precaution after assembly, an expandable aircraft-tubing clamp is slipped over each joint and tightened.

Before the elements are assembled, the center-element sections of the reflector and director should be inserted in their respective holes in the ends of the boom and accurately centered. It is a good idea to slot these center sections after they have been passed through the boom holes instead of before, as the tubing expands slightly after it is slotted and it may be quite a job to compress it enough to get it through the boom holes. With the center sections aligned with respect to the boom, a ¼-inch hole is drilled and a ¼-28 machine screw is run through the top wall of the boom and through both walls of the element, as in Fig. 12-18C. When this joint is tightened the element will be firmly anchored to the boom. Any play at this joint will lead to bad element vibration in a wind, so any slippage here should be shimmed out with thin brass strips

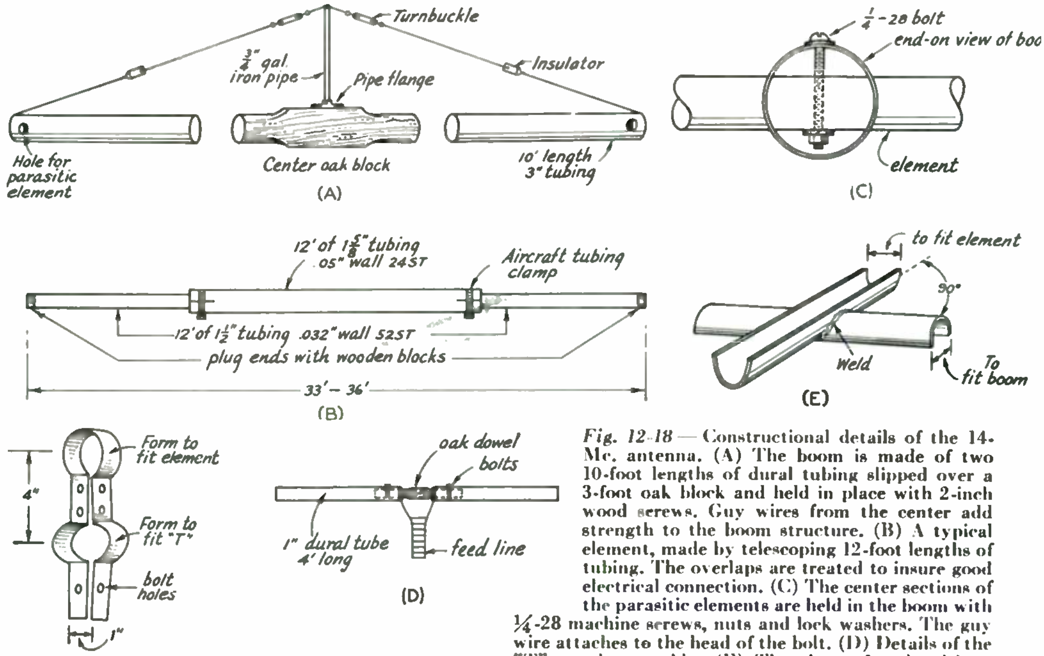


Fig. 12-18 — Constructional details of the 14-Mc. antenna. (A) The boom is made of two 10-foot lengths of dural tubing slipped over a 3-foot oak block and held in place with 2-inch wood screws. Guy wires from the center add strength to the boom structure. (B) A typical element, made by telescoping 12-foot lengths of tubing. The overlaps are treated to insure good electrical connection. (C) The center sections of the parasitic elements are held in the boom with ¼-28 machine screws, nuts and lock washers. The guy wire attaches to the head of the bolt. (D) Details of the "T"-match assembly. (E) The clamp for the driven element is made by splitting 1-foot lengths of iron pipe and welding them together as shown.

into a horizontal position and dropped into some kind of a cradle at the top of the pipe. However, one "brute-force" method is to bolt to the center oak block a large pipe flange that can be threaded on to the top end of the vertical supporting pipe. The beam is pulled up the side of the tower and swung up and on top of the vertical pipe. Turning on the rotation motor will screw the pipe into the flange, after which the joint should be pinned. All this takes a lot of muscle, but it has been done, as in the case of Fig. 12-17.

A tiltable iron mounting bracket, as shown in Fig. 12-19A, can be welded at a local machine shop. It is welded to the supporting pipe and provides a flat tiltable metal plate the exact size of the bottom of the oak center block. The plate is drilled to correspond to bolt holes in the block. The beam can be passed up the mast hand-over-hand until the oak block is centered with the mounting plate. It is a simple job to bolt the block to the plate and then swing the beam up into a horizontal position. It is held horizontally by two short guys running between the boom and the vertical pipe.

A "U" channel as shown in Fig. 12-19B can be constructed, into which the boom will fit and which has extended side flanges at each end, drilled to fit corresponding holes in the boom. The boom is hoisted atop the tower and positioned between the two flanges and a bolt run through the flanges and the boom. The boom is then swung up to a horizontal position and the second bolt put in place.

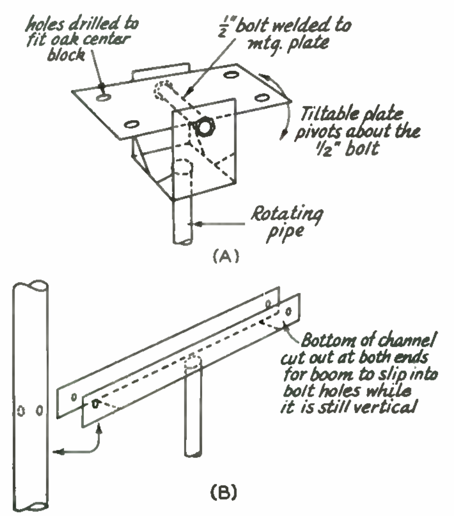


Fig. 12-19 — Two types of mounting plates for tubing booms. The one shown at A, since it is supported at only one point, requires that the boom be guyed to the rotating vertical member. See Fig. 12-20.

The mounting plate shown at B is made of a length of "U"-channel iron cut and drilled as shown. The boom is raised vertically until one set of bolt holes is in a line and a bolt is slipped through. The boom is then swung into its horizontal position and the other bolt is put in place.

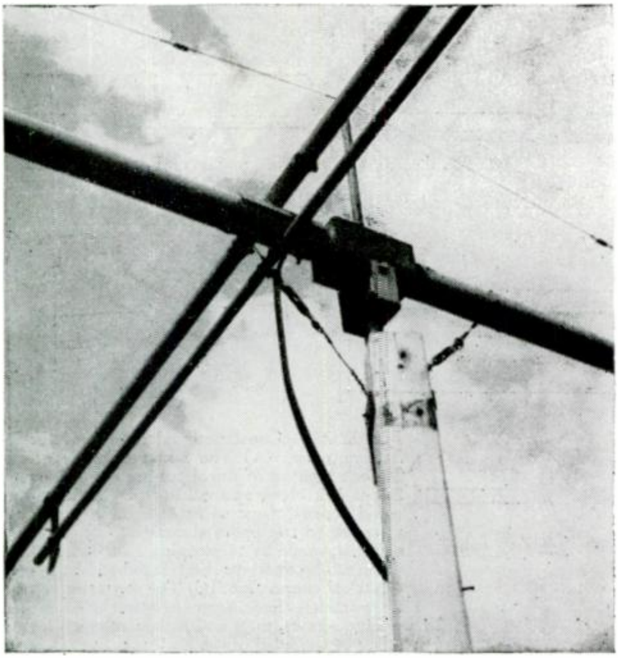


Fig. 12-20 — A close-up of the tiltable mounting plate of Fig. 12-19A. The two short lower guys go to the rotating pipe and keep the boom horizontal. The saddle mounting for the driven element (Fig. 12-18E) is also visible. (W6SAI, Feb., 1949, *QST*.)

Upon completion it is a good idea to paint the whole beam with aluminum paint to prevent corrosion of the dural.

The boom can be constructed with rectangular dural tubing instead of round. In some cases square tubing is more easily obtainable. Tubing about 2 by 4 inches on edge is satisfactory.

It is permissible to replace the oak center block with a splicing piece of dural tubing. This makes the beam slightly more difficult to mount and a little more floppy. The flop may be taken out by the umbrella guys, however. This substitution works better with square tubing as it still provides a flat mounting surface.

In case of vibration in the element tips, it can be damped out by plugging the tips with wood.

● WOODEN-BOOM CONSTRUCTION AND SLIP-RING FEED

Many amateurs prefer to build their beam booms from standard pieces of lumber, and the beam shown in Figs. 12-21 and 12-22 is an example of excellent design in wooden-boom construction. The boom members are two 20-foot 2 × 4s fastened to the 4 × 12 × 24-inch center block with six lag screws. The two center screws serve as the axis for tilting — the other four lock the boom in position after final assembly and adjustment have been completed. The blocks midway from each end are 2 × 4s spaced about six inches apart, with a long bolt between them. When this bolt is drawn tight, a very sturdy box brace is formed. The crossarms are 3 × 3s twelve feet long, bolted to the boom with carriage bolts.

The umbrella guys should have turnbuckles in them, and the guys are fastened to the center support after the beam has been permanently locked in its horizontal position. With the turnbuckles properly adjusted, there will be no sag in the boom, the elements will be parallel and neat, and weaving in the wind will be eliminated.

The elements are 1⅜- and 1½-inch diameter duralumin tubing, supported by 1½-inch stand-off insulators. Hose clamps are used to hold the elements on the insulators. Final adjustment of element lengths is possible through "hairpin" loops.

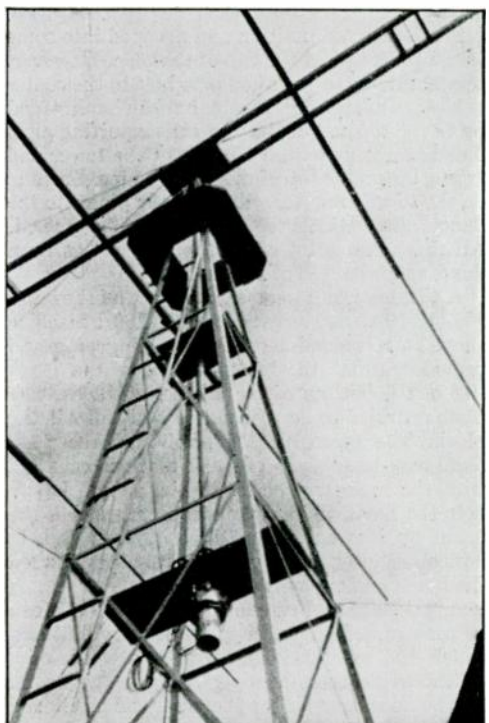


Fig. 12-21 — A wooden boom for a 4-element 14-Mc. boom can be made quite strong by judicious use of guy wires. The drive motor is mounted halfway down on the tower. (W6MJB, Nov., 1947, *QST*.)

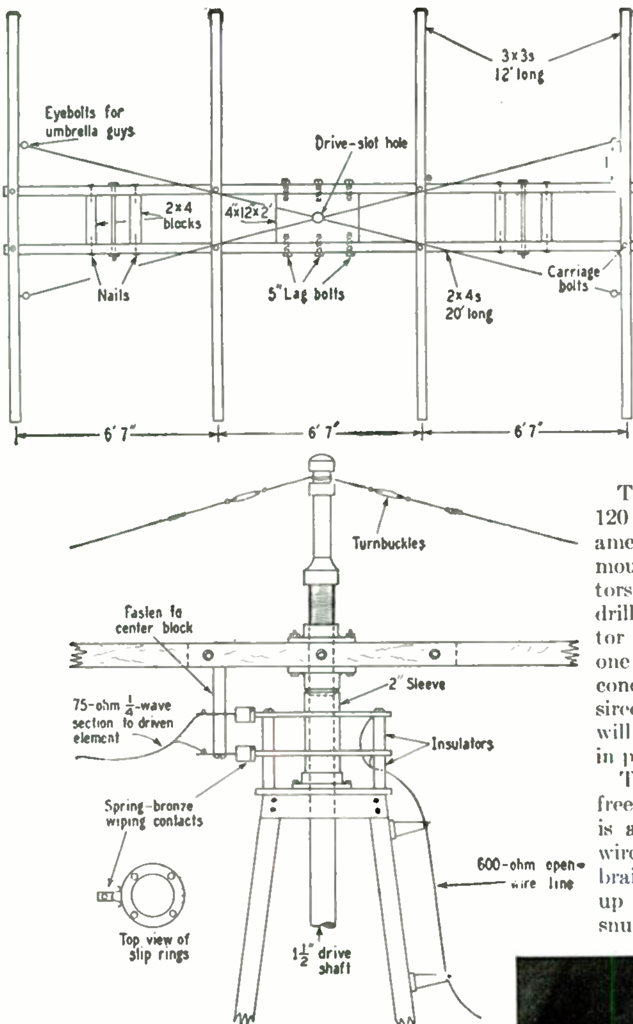


Fig. 12-22 — **Details of the wooden boom, its method of support, and the construction of the slip rings.**

The support for the beam shown in Fig. 12-21 was a Sears-Roebuck windmill tower. The driving motor for the beam was located halfway down the tower, the torque being transmitted through a length of 1½-inch drive shaft. A pipe flange is welded to the drive shaft and bolted to the center block. A cone bearing is obtained by turning both the flange and a sleeve of 2-inch pipe to match, as shown in Fig. 12-22.

One method of matching the line to the antenna is to use a quarter wavelength of 75-ohm Twin-Lead between the radiator and the slip-ring contacts, to match a 600-ohm line from the slip rings to the transmitter.

A 600-ohm open-wire line is run to a point about halfway up on the tower, then up the side of the tower to the slip rings. The slip rings are mounted on the top of the tower, directly under the center block. A quarter-wave-length matching section of transmitting-type 75-ohm Amphenol Twin-Lead hangs in a loop between the driven element and the slip-ring contacts.

An alternative method of slip-ring construction is shown in Fig. 12-23. In this case, the 75-ohm line would run down through the central pipe to the two rings, and the open-wire line would connect to the sliding contacts. Two brass disks ¼-inch thick are turned out on a lathe from a piece of 3-inch brass stock, and then grooved to form a shallow pulley.

Three holes are drilled in the disks, 120 degrees apart, on a circle whose diameter is two inches; these are used for mounting the isolantite spacing insulators. The lower ring has a clearance hole drilled in its center for the inner conductor of a concentric line, while the upper one is drilled for clearance of the outer conductor or pipe. A press fit is to be desired. After being sure that the assembly will turn true, the rings may be soldered in place.

The important work of getting a noise-free low-resistance contact to the rings is accomplished by covering a flexible wire with several layers of tinned copper braid. The outside diameter of the built-up braid should be such that it will fit snugly into the grooves in the rings. The

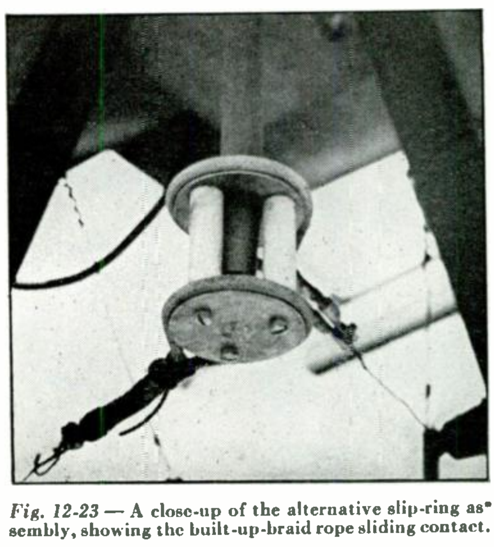


Fig. 12-23 — **A close-up of the alternative slip-ring assembly, showing the built-up-braid rope sliding contact.**

braided wire is looped around the ring and both ends soldered into heavy-duty lugs, which may be used for the connection and also as a point to attach a medium-stiff spring which will assure a positive wiping contact at all times. A little vaseline or other petroleum lubricant on this unit is of considerable help. If the antenna is to be fed with an open-wire line it is convenient to have the springs 180 degrees apart, thus neutralizing the pull that would occur were they on the same side.

INDUCTIVE COUPLING TO THE FEEDLINE

Two common methods of connecting the feedline to the antenna have already been illustrated in this chapter. Direct coupling is perfectly satisfactory, if the feedline is flexible enough and if some kind of stops are provided to avoid rotating the antenna too far in the same direction. The slip-ring contacts will give good service if reasonable care is taken in their construction, although their use is generally confined to high-impedance points where the current will not be excessive. The contact surfaces should be wide enough to take care of wobble in the rotating shaft, and these surfaces should be kept clean. Spring contacts are essential, and an "umbrella" or other scheme for keeping rain off the contacts is a desirable addition.

Still another way to obtain continuous rotation of the beam antenna is to use inductiye coupling between the driven element and the transmission line (see Fig. 3-61). This is an excellent method for coupling into an open-wire or other high-impedance line, although the precision of the mechanical work must be high if the coupling is to be constant at all bearings of the antenna. Typical of such devices is the unit shown in Figs. 12-24 and 12-25. The antenna tuning condensers (not shown) that are used to tune out the reactance of the antenna loop can be mounted at the driven element or nearby. The driven-element length is made the same as in other beams, and can be obtained from Fig. 8-13. The antenna coupling loop is made of ¼-inch diameter soft copper tubing, and is 11 inches in diameter. It is mounted on three or more stand-off insulators, and it must be aligned accurately to be exactly at right angles to the axis of the antenna shaft.

The transmission-line loop is similar in dimensions, and it should be mounted on three or more insulators in such a manner that the clearance between it and the antenna loop can be varied from about 1 inch to ⅛ inch. This can be done by mounting the stand-off insulators on an auxiliary flat plate (with a clearance hole in the center for the antenna drive shaft) that can be raised or lowered on three screws. It has only to be adjusted during the tune-up process, so the emphasis can be put on sturdy construction rather than on ease of adjustment. The coupling loops should be mounted at least 6 inches away from any metal, and the drive shaft should be kept as small as possible, so that too much metal is not introduced into the coil fields.

In constructing and locating the loops, make sure they are true round and that the rotating loop turns about a true center with reference to the fixed loop. If the rotating loop has a wobble or changes its vertical or horizontal plane with reference to the fixed loop, the loading on the final amplifier will vary with rotation of the beam.

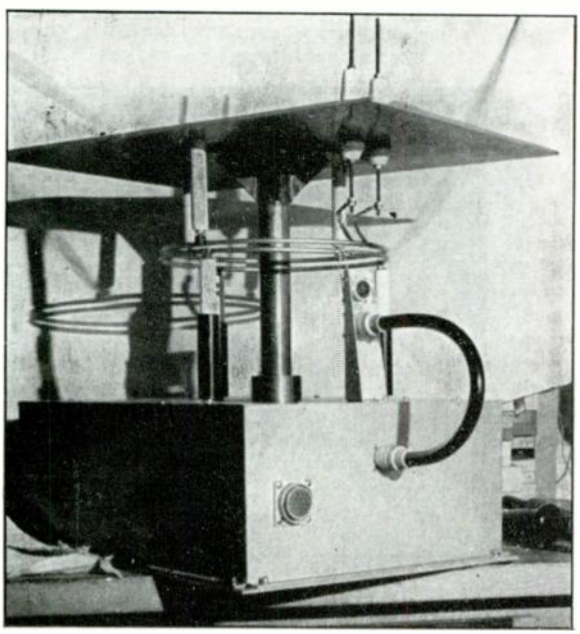


Fig. 12-24 — A rotator assembly with inductive-coupling rings. The box contains the drive motor, gear train, control relays, and the line tuning condenser. The short length of coaxial cable connects the fixed loop to the tuning condenser.

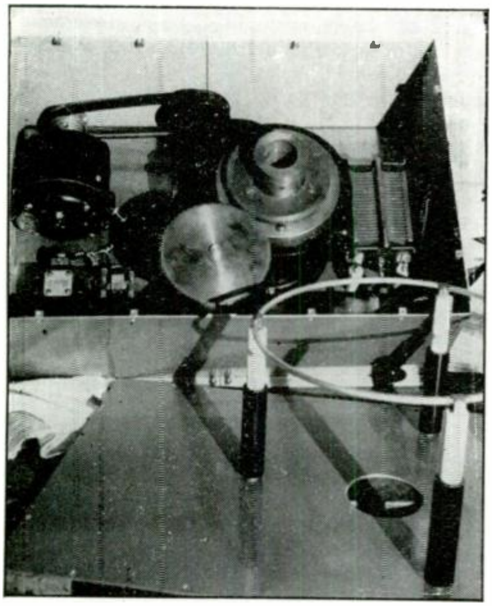


Fig. 12-25 — Inside the rotator-assembly box. The tuning condenser is at the right. The pipe support for the boom plate sets in the socket of the gear-train assembly.

It is an excellent idea to weatherproof the tuning condensers because smoky areas, salt atmosphere or places where high humidity prevails decrease the life expectancy of the unit. The condensers should be placed as close as possible to the junction of the transmission line and the fixed ring, and the junction of the driven element and the line leading to the rotating ring.

Weatherproofing the whole unit, by mounting it within a watertight box, is almost a practical impossibility because of the relative motion and consequent need for a watertight bearing on the housing. The best alternative is to protect the coupling loops from the direct effects of rain, snow and ice by using a protective housing, and to mount the condensers, drive motor and direction-indicating devices in a watertight box. An alternative to the watertight box for the condensers and motors is to provide a box with one or two drain holes and to cover everything with waterproofing compound (Dow-Corning No. 4 Ignition Sealing Compound, for example).

When adjusting the array with inductive coupling, the antenna condensers are tuned to resonance once and left there. The transmission-line condenser and the coupling (distance between loops) is then varied until minimum standing-wave ratio is obtained on the line. Any modification of element lengths over the original will probably require readjustment of the line condenser and the loop coupling to restore the low standing-wave ratio.

A TWO-BAND 4-ELEMENT BEAM

One disadvantage of the conventional 2-, 3- or 4-element beam is the fact that it is a one-band affair. This means that, where rotary beams are to be used on several bands, most operators provide separate supports and rotating mechanisms for each antenna or they "stack" the beams one above the other on the same support. In some cases the two beams are mounted in the same plane, although the beams generally show some interaction that is hard to predict or evaluate.

One solution to the problem has been to mount relays in the center of each element, so that tuning devices can be switched in by the relays. While this is an effective solution, it is rather complicated and has not found too much favor. The best answer to date is the one presented in this section.

The antenna shown in Fig. 12-26 will operate

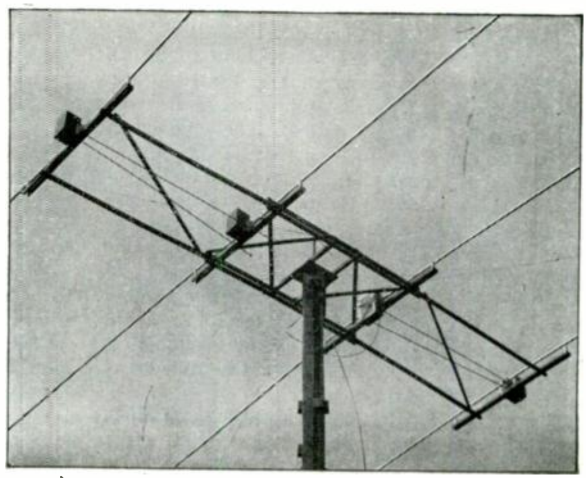


Fig. 12-26 — Close-up view of the two-band antenna, showing the weatherproof enclosures for the networks and the extension rods for tuning the networks associated with the outside elements. The loop of 300-ohm line is also visible in this photograph. (W3NJE, Oct., 1948, *QST*.)

on both 14 and 28 Mc. without the need for switching or other adjustment of the elements. The principle can be applied to 2- and 3-element antennas, as well as 4, for any two bands in harmonic relation.

Basically, the array consists of a parasitic-type four-element antenna — driven element, two directors and a reflector — cut for 14 Mc. and using 0.1λ spacing between all elements. All elements are insulated and open in the center.

The driven element is energized as a half-wave

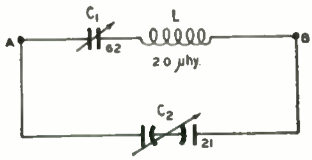


Fig. 12-27 — Circuit of network used at the center of the parasitic elements.
C_1 — 116-μμfd. variable, 3800 volts peak.
C_2 — 50 μμfd. per section (same as C_1 with center stator plate removed).
(Numbers near C_1 and C_2 indicate calculated values to effect resonance at 14.25 and 28.5 Mc.)
L — 2.0 μhy.; 7 turns No. 8 wire, diam. 3½ inches, length 4¾ inches.

dipole on 14 Mc. and as two half waves collinear, in phase, on 28 Mc. The parasitic elements must be shorted at the center on 20 and opened at the center on 10 meters. This is done to provide the usual half-wave parasitic elements on 14 Mc. to work in conjunction with the half-wave driven element, and to provide half-wave collinear parasitic elements that are properly phased to work with the 28-Mc. collinear driven element.

Thus the antenna may be described as a four-element close-spaced (0.1λ) beam on 20 meters and as eight elements wide-spaced (0.2λ) on 10, each of the collinear half waves of the driven element being headed by two directors and backed by a reflector.

The parasitic elements are automatically short-circuited on 14 Mc. and open-circuited on 28 Mc. by the circuit shown in Fig. 12-27. At the lower frequency, C_1 and L provide a series-resonant circuit (low impedance). The condenser, C_2, has no effect at this frequency. At the higher frequency, C_2 tunes the circuit to parallel resonance, and a very high impedance is developed across *A-B*. Capacitor C_2 has no effect upon 14-Mc. tuning as it is effectively shorted at this frequency. Capacitor C_1 does affect tuning at 28 Mc. because C_2 must be varied to effect parallel resonance with each change of capacitance of C_1. Therefore, to avoid interaction in tuning, it is necessary that the antenna system be tuned first at 14 Mc.

The element lengths are the usual ones, obtained from Fig. 8-13. The element-supporting structure is constructed of wood, similar to that shown in Fig. 12-22. The center spacing between all elements is 0.1λ (free-space value) at 14.25 Mc., which is equal to 6 feet 11 inches. It is recommended that at least 4-inch insulators be used throughout because on 28 Mc. the center of each conductor is at a high-voltage point.

The circuit of the two-band matching network for the driven element is shown in Fig. 12-28, and the completed network is shown in Fig. 12-29. The network is built in a wooden box, 13 by 10½ by 5 inches, which serves as a weatherproof enclosure. The capacitors and inductors are insulated from the box and the input and output terminals of the network are brought out through feed-through insulators.

The coil-winding information is supplied only as a reference; the coils should be checked by means of a *Q*-meter or some other method that will verify the resultant inductance. The detailed design of this type network is covered in a *QST* article[1] so that those desiring to use different transmission lines and parameters (e.g., 2- or 3-element basic antennas) may do so.

The circuit constants shown are for use with a 300-ohm transmission line. This is done so the

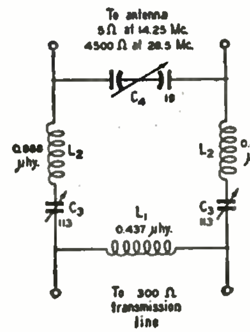


Fig. 12-28 — Circuit of two-band matching network to couple a 300-ohm line to a 5-ohm load at 14.25 Mc. and to a 4500-ohm load at 28.5 Mc.
C_3 — 116-μμfd. variable, 3800 volts peak.
C_4 — 35 μμfd. per section, 4200 volts peak (Cardwell MT-35-GD).
L_1 — 0.437 μhy.; 4 turns No. 8 wire, diam. 1 9/16 inches, length 2½ inches.
L_2 — 0.888 μhy.; 6 turns No. 8 wire, diam. 1 9/16 inches, length 1 9/16 inches.

ribbon-type line can loop downward and swing with the rotating structure, thereby eliminating slip rings with attendant maintenance and impedance-discontinuity problems. Higher efficiency would be achieved by using a 4-conductor 300-ohm open-wire line from the transmitter to the top of the supporting mast, with a short length of 300-ohm ribbon for the downward-hanging loop. However, the de luxe type of 300-ohm line has reasonably low attenuation, is unaffected by moisture, and gives excellent results.

[1] J. G. Marshall, "Matching the Antenna for Two-Band Operation," *QST*, September, 1945.

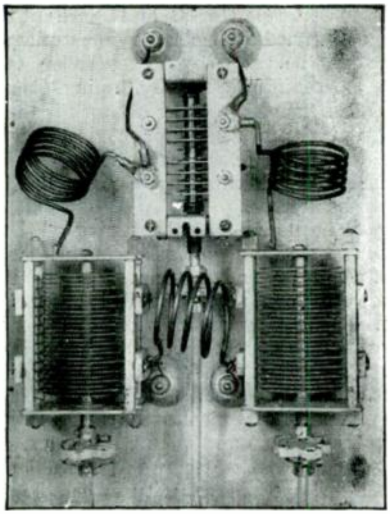


Fig. 12-29 — The two-band matching network. The coils shown are not the same as those specified under Fig. 12-28. The latter values should be used.

The parasitic networks are identical for each parasitic element. A completed network is shown in Fig. 12-30. The networks are enclosed in wooden boxes 9⅞ by 7⅝ by 5¾ inches for weatherproofing as in the case of the matching network. The capacitors and inductor are insulated from the box, with the capacitor shafts extended (by use of flexible insulated couplers and ½-inch aluminum tubing) so that they can be rotated while standing at the top of the supporting mast.

A hole 2 inches in diameter is drilled in the bottoms of the boxes to provide adequate "breathing." This helps to prevent moisture condensation. Without this "breathing" hole, condensation, particularly on the capacitors, may cause frequent flash-overs after periods of high humidity.

Two feed-through insulators are used to connect the network to the elements. These insulators are mounted on 2½-inch centers and are connected to the elements by means of wide aluminum straps as shown in Fig. 12-31. Fig. 12-32 illustrates the method of mounting the parasitic network boxes by means of aluminum angle and the method of attaching the extension shafts to the capacitors. It can be seen that the two center-support insulators must be separated sufficiently to provide clearance for the extension shafts.

All inductors are wound of No. 8 solid copper wire and are self-supporting. The parasitic-network inductors can be wound using the specified turns and sizes as they can be pulled apart or pushed together if necessary. However, the matching-network inductors should be accurately measured, because even though the variable capacitors will tune the network through a wide range the standing-wave ratio will not be as favorable if the inductors differ from those specified.

Tuning Procedures

There are two methods by which this antenna system may be tuned. One is the conventional method of supplying power to the antenna and making adjustments in accordance with readings taken by a remotely-located field-intensity meter. The other is to tune the system by observing the S-meter of a receiver while tuning to a remotely-located station or signal. The second method has two importance advantages. Fewer men are required. If the transmission line is properly terminated in its characteristic impedance, there will be no standing waves on the line and changes in S-meter readings will be the result of changes in tuning of the antenna system proper with no transmission line pick-up (assuming a balanced line). In making direct field-strength readings it is difficult to tell to what degree the antenna proper or the feedline contribute to the meter readings. In addition, the readings may be erroneous because of the close proximity of power and telephone lines, or other conductors, to the field-strength measuring apparatus.

In using Method 1 the transmission line is coupled to the transmitter by means of link coupling, with enough power supplied to the antenna to obtain a satisfactory reading on the field-intensity meter. The power coupled to the antenna should be no more than is necessary for adequate meter reading. The antenna is first tuned on 14 Mc. The series-network capacitors, C_1, should be set to approximately one-half capacity. The matching-network series capacitors, C_3, should be tuned together for minimum stand-

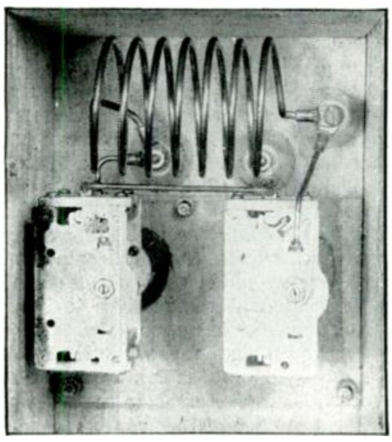


Fig. 12-30 — A network of this type is used at the center of each parasitic element. C_1 and C_2 are mounted on a sheet of Lucite fastened to the side of the enclosure by ½-inch stand-off insulators. C_1 is at the right and C_2 at the left.

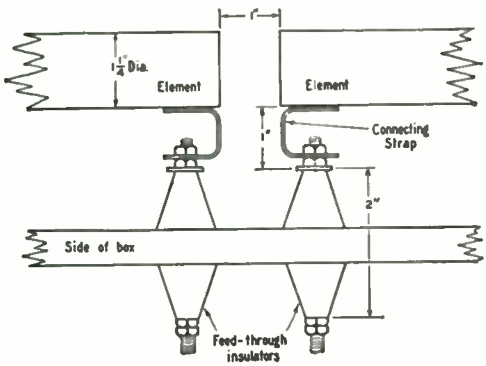


Fig. 12-31 — Method of connecting networks to beam elements.

ing-wave ratio. It will be noted that these capacitors are interrelated to the extent that numerous settings will provide proper loading. However, the correct position is that at which both capacitors are at approximately the same capacity. This maintains proper balance.

The parasitic elements may now be tuned in the following order: first director, reflector and second director.

The tuning is done in this order because the elements have the greatest effect upon field intensity and antenna impedance in that order. Each time that the various elements are brought into approximate tune, the antenna-matching network series capacitors, C_3, should be retuned a number of times, because of the normal interaction between elements, to find the point where maximum field intensity or maximum front-to-back ratio is achieved as is normally done with the conventional parasitic-type array. The power input to the transmitter must be maintained constant and checked after each adjustment.

The second method of tuning the array requires running two conductors from the S-meter in the receiver to the top of the antenna mast, where a second S-meter is located so that the man tuning the system can observe it while making adjustments. This line should be run to the base of the mast and vertically up the mast, and should be by-passed at each end with a 1000-$\mu\mu$fd. mica capacitor to eliminate as much as possible any detrimental r.f. effects. A signal located at least five miles away should be used to tune on. A man should be located at the receiver to keep the signal tuned continuously, in case of receiver drift. The man tuning the system at the top of the mast tunes capacitors C_3 for maximum S-meter reading and also capacitors C_1 in the order previously specified. Again the various capacitors should be tuned a number of times to overcome the effects of interaction and to achieve best peaking.

When tuning the system on 28 Mc. the same procedure is followed in the two methods with the exception that capacitors C_1 and C_3 are not tuned. If they are tuned or displaced while tuning on 28 Mc. it will be necessary to retune on 14 Mc. before proceeding with 28-Mc. tuning. Capacitor C_4 of the antenna-matching network is tuned to provide maximum loading or maximum S-meter indication on 28-Mc. and capacitors C_2 are tuned for maximum field intensity, S-meter reading or front-to-back ratio.

It will be noted that the tuning of the first director (or directors) on both frequencies is quite sharp. The reflector is moderately sharp and the second director is comparatively broad. The parasitic-network series capacitors should be at approximately one-half capacity when the antenna is peaked on 14 Mc. If this is not the case, when the antenna is peaked the inductor L should be compressed or extended to change the resultant inductance and shift the amount of capacity required. If capacitors C_1 are not at approximately one-half capacity it will be found that capacitor C_2 will not have sufficient range to tune to 28 Mc.

● ROTATING THE ANTENNA

Many methods have been devised for rotating beam antennas, but they all have to be designed, to a great extent, to fit the particular conditions at the ham shack in question. Several ideas have already been shown earlier in this chapter. Where possible, the direct approach shown in Fig. 12-11 is a good one. An alternative to this is to mount the rotating mast in two or more bearings on the side of the house, in which case it can be rotated by hand-driven chains or cables running in to the shack, by bevel-gear drive at the bottom of the mast, or by a motor at the base of the mast. A stronger support for the antenna, where greater height is desired, can be obtained by building a lattice tower that is either partially supported

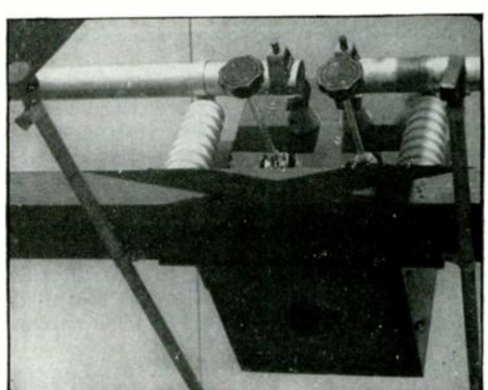


Fig. 12-32 — The method of mounting the network enclosure, in this case for the first director. Note the aluminum angle, at the bottom, holding the box to the crossarm. The shafts at the extreme left and right are the extensions of the tuning controls from the second-director network.

by the side of the house or is completely self-supporting, and running the drive shaft for the antenna up the center of the tower. If the tower is strong enough to support the weight of the motor, the motor can be mounted close to the antenna, as shown in Figs. 12-17 and 12-21. When the weight of the motor would be undesirable at the top of the mast, a cable-drive system can be used, as in Fig. 12-14.

One disadvantage of mounting the motor too far from the antenna boom and transmitting the torque through a small-diameter shaft is the backlash that will be encountered if the shaft is more than 15 or 20 feet long. In this case, either the shaft diameter must be made greater or the whole mast can be rotated. The whole mast can be rotated if suitable collar bearings are used at each point where the guy wires tie into the mast. The antenna of Fig. 12-3 shows how this can be done on a small scale.

The bearings for the rotating shaft that supports the antenna boom should be good but they do not have to be superlative. Since any hand- or motor-drive of the system will involve a speed-reduction system, enough torque will be developed to overcome minor bearing frictions. A speed-reduction system is mandatory for any 14-Mc. beam and for all but the lightest-weight 28-Mc. systems. The speed of rotation of the antenna should be about 1 or 1½ r.p.m. for a 14-Mc. beam, but it can be increased over this if the supporting structure can stand it. Not a great deal of power is needed: a motor capable of delivering ⅛ horsepower is ample for the average rotary, and a ¼-horsepower motor will take care of the largest. The reduction gears will also act as a brake for the system so that it does not "coast" too far past the position at which the power was shut off.

A reversible motor should be used. If the motor

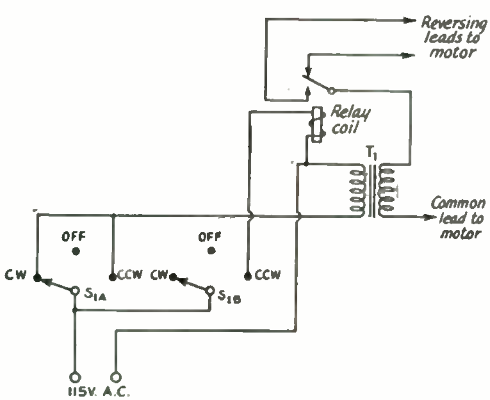


Fig. 12-33 — A method for using only three wires to control a reversible propeller-pitch motor used as a rotary-beam drive unit. The d.p.d.t. switch has a neutral, or off position. The relay is a 115-volt a.c. s.p.s.t. type.

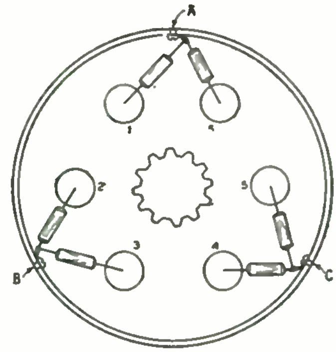


Fig. 12-34 — Noise created by propeller-pitch beam rotators can be eliminated by by-passing the brush holders to the case of the motor as shown. Points *A*, *B* and *C* are grounds made by drilling and tapping the rim of the motor case for 6-32 screws.

is a low-voltage high-current unit, as are the very popular "propeller-pitch" motors to be found in war surplus, the step-down transformer should be mounted near the motor (if a long run of power line is required), to avoid the use of unnecessarily heavy lines. Control switching can then be done with a heavy-duty relay at the transformer, as shown in Fig. 12-33.

A beam is generally positioned by rotating it to the bearing that gives the loudest received signal. To do this, the rotating motor must radiate little or no electrical noise, of course, and this requires some type of filtering at the brushes of the motor. Shielding the leads to the motor and grounding its frame are also helpful in some cases. The noise generated by the propeller-pitch motors can be eliminated by using six small condensers, as shown in Fig. 12-34. First remove the thin aluminum motor cover and then take out the motor itself. After cleaning the six copper-surfaced brush holders and drilling and tapping three holes for 6-32 screws, solder a condenser from each brush holder to a ground lug at the tapped holes. Mica or ceramic condensers can be used, in the range 0.002 to 0.01 μfd. The "Hy-Kap" ceramic type is good for the job because of its small size.

INDICATING BEAM DIRECTION

Most users of beam antennas like to know which way the beam is pointing without having to resort to telescopes, periscopes and other optical devices. If the beam is manually operated with a 1-to-1 drive (as in Fig. 12-1), it is a simple matter to attach a small pointer to the shaft and paste a compass card or suitable map (see Chapter Thirteen) on the wall or ceiling. If this simple solution is not feasible, a string- or cable-drive indicator can be constructed, along the lines of the schemes shown in Fig. 12-35. However, these indicators are useful only when the beam is lo-

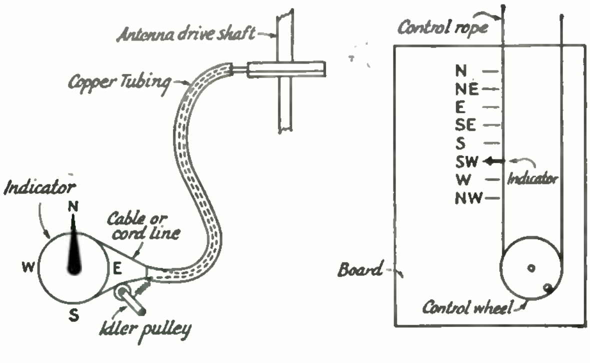


Fig. 12-35 — Mechanical direction-indicator systems.

cated close to the radio shack, where the cable run is not too great.

Remote-indicating devices generally depend on electrical circuits for their operation. At one time many amateurs used multiposition switches mounted on the drive shaft of the beam, and each switch contact was wired to individual dial lamps properly spaced around the periphery of a compass card. As the beam rotated, the light that was burning indicated the direction. The only disadvantage to this system is that suitable switches are not always easy to find, and the more precise the indication (large number of lamps) the larger is the required number of wires. This indicating method has been largely superseded by the circuit of Fig. 12-36. The potentiometer is a special one made for the purpose that has no stops, and the meter has a new scale reading "N-E-S-W" instead of its usual number scale. The potentiometer is driven by the antenna shaft (1-to-1 ratio), and it is only necessary to align the potentiometer properly during the installation to have the beam-direction indication available by closing switch S_1. Once R_2 has been set to the correct value, as found by experiment, the indication will be accurate so long as the voltage of the battery remains constant. However, as the battery voltage drops, the setting of R_2 can be shifted accordingly, to make the indication correct.

A very popular system of beam-direction indication uses "synchro" (Selsyn) generators

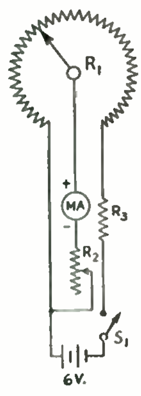


Fig. 12-36 — A direction indicator using a potentiometer and a milliammeter.

R_1 — Special potentiometer (Ohmite RB-2).
R_2 — 7500-ohm potentiometer, calibration set.
R_3 — 220 ohms, 1 watt.
MA — 0-1 milliammeter.
S_1 — S.p.s.t. toggle.

and motors, as shown in Fig. 12-37. These a.c.-operated devices are available for 60-cycle 115-volt operation and for 400-cycle 115-volt operation. The 400-cycle units are less expensive, as a rule, and can be used with a 60-cycle supply at reduced voltage. A suitable resistor or transformer can be used to drop the voltage to a value that will give sufficient torque without heating of the synchro. From 25 to 50 volts is the usual operating value. There is no difference in operation in the two systems in Fig. 12-37 except that Circuit B requires one less wire. In Circuit A or B the relative directions of rotation can be reversed by interchanging the connections of two of the stator (S) connections at one unit.

The generator synchro should be coupled to

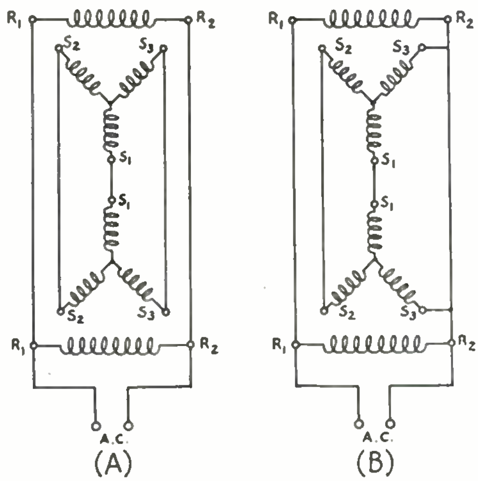


Fig. 12-37 — Interconnections for synchros used for beam-direction indicators. The system at B requires only four wires instead of the usual five. With either circuit, the relative direction of rotation can be reversed by interchanging the leads to S_1 and S_2 at one of the units.

the antenna through a 1-to-1 ratio unless a duplicate gear system is used at the motor (shack) end. The coupling can be through gears, pulleys and belts, or friction drive. Many different ways have been developed for indicating the beam direction at the operating position, ranging from simple pointers moving over a compass chart to back-lighted world maps and rotating globes. Examples of the latter systems are shown in Figs. 12-38 and 12-39.

The rotating-globe indicator of Fig. 12-38 is attractive and not difficult to build. A 7-inch or larger globe is purchased, and the hometown and a point directly opposite on the globe are located. Then drill ⅛- or ¼-inch holes at these points and pass a shaft through the globe, fastening the shaft to the globe with glue or suitable cement.

Next cut a strip of Lucite or Plexiglas to the shape shown and drill the two holes that pass the shaft. At the same time, the hairline can be scribed along the Lucite, on both sides if parallax is to be avoided. The Lucite is then heated in an oven until it can be bent to the shape shown in the drawing. It is then assembled with two Lucite supports on the small box that houses the synchro (Selsyn).

To align the indicator, point the beam due north and energize the synchro indicators. Align the globe so that the hairline is directly over the North Pole on the globe, and tighten the coupling screws. The globe will then rotate in synchronism with the beam, and the line of maximum strength is always right under the hairline. If desired, the top plate of the box can be made from a mirror or other reflector, to permit viewing the bottom of the globe.

The indicator of Fig. 12-39 shows a great-circle map and a compass card superimposed, but the construction will suffice for either one or both. The map or card is framed and mounted in front of two 15-watt lamps. A synchro motor, coupled to its mate at the antenna, is mounted between the bulbs, with its shaft extending through a small hole in the map or card. A piece of glass or Lucite, with a suitable hole to pass the shaft, is mounted over the map, and the entire assembly is supported within a sloping-front box of wood or metal. The pointer that is fastened to the shaft should be very lightweight or it should be counterbalanced, to eliminate any indication error introduced by unbalance. As an added refinement, the map itself can be "plasticized" for a small additional cost. In the larger cities there are many companies doing such work for discharge papers, courtesy cards and the like. If the map is not plasticized, it can be backed up by glass or Lucite.

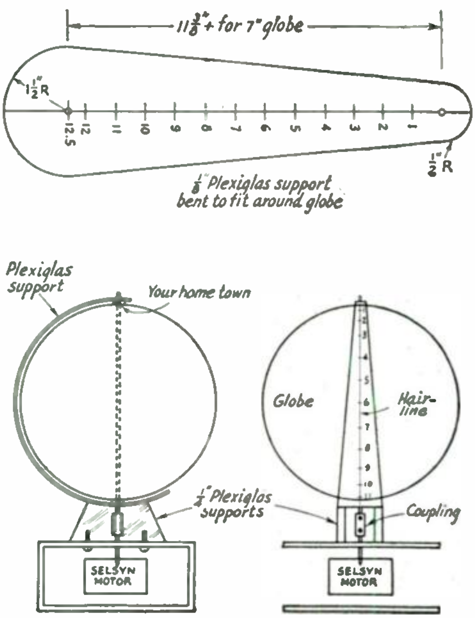


Fig. 12-38 — **Constructional details of the globe direction indicator. The dimensions shown are for a 7-inch globe, and should be modified for other sizes. The curved piece of Lucite is cemented to a base that is then bolted to the top of a box that houses the synchro motor.**

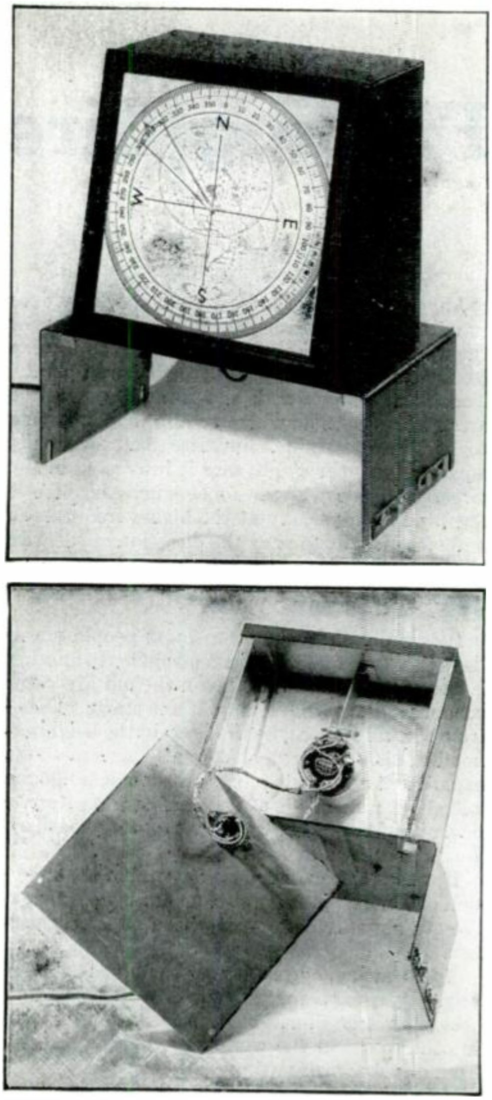


Fig. 12-39 — **Two views of the indirectly-lighted beam indicator. The rear view shows the synchro and the lamps mounted inside the box. Control wires are cabled and run out of the box through a 5-prong terminal mounted on the rear cover plate.**

CHAPTER 13

Finding Directions

Anyone laying out a fixed directive array does so in order to put his signal into certain parts of the world; in such cases, it is essential to be able to determine the bearings of the desired points. Too, the amateur with the rotatable directive array likes to know where to aim if he is trying to pick up certain countries. And even the amateur with the single wire is interested in the directive pattern of the lobes when the wire is operated harmonically at the higher frequencies, and often is able to vary the direction of the wire to take advantage of the lobe pattern.

FINDING DIRECTION

It is probably no news to most people nowadays that true direction from one place to another is not what it appears to be on the old Mercator school map. On such a map, if one starts "East" from central Kansas, he winds up in the neighborhood of Lisbon, Portugal. Actually, as a minute's experiment with a strip of paper on a small globe will show, a signal starting due East from Kansas never hits Europe at all but goes into the southern part of Portuguese West Africa.

If, therefore, we want to determine the direction of some distant point from our own location, the ordinary Mercator projection is utterly useless.

True bearing, however, may be found in several ways: by using a special type of world map that *does* show true direction from a specific location to other parts of the world; by working directly from a globe; or by using mathematics.

AZIMUTHAL MAPS

While the Mercator projection does not show true directions, it is possible to make up a map which will show true bearings for all parts of the world from any single point. Three such maps are reproduced in this chapter. Fig. 13-1 shows directions from Washington, D. C., Fig. 13-2 gives directions from San Francisco and Fig. 13-3 (a simplified version of the ARRL 30″ × 40″ amateur radio map of the world) gives directions from the approximate center of the United States — Wichita, Kansas.

For anyone living in the immediate vicinity (within 150 miles) of any of these three reference points, the directions as taken from the maps will have a high degree of accuracy. However, one or the other of the three maps will suffice for any location in the United States for all except the most accurate work; simply pick the map whose reference point is nearest you. Greatest errors will arise when your location is to one side or the other of a line between the reference point and the destination point; if your location is near or on the resulting line, there will be little or no error.

By tracing the directional pattern of the antenna system on a sheet of tissue paper, then placing the paper over the azimuthal map with the origin of the pattern at one's location, the "coverage" of the antenna will be readily evident. This is a particularly useful stunt when a multilobed antenna, such as any of the long single-wire systems, is to be laid out so that the main lobes cover as many desirable directions as possible. Often a set of such patterns will be of considerable assistance in determining what length antenna to put up, as well as the direction in which it should run.

Anyone who uses the CRPL-D series, "Basic Radio Propagation Predictions," issued by the National Bureau of Standards, can get the bearing from his hometown to any spot in the world by using the special maps used in the propagation predictions. By using the base map (a modified cylindrical projection) in conjunction with the great-circle chart, a map can be made to show the path of radio waves for every 30° (or less) around the compass from one's location. Anyone not subscribing to the service can obtain the map and chart from Circular 465 of the National Bureau of Standards, entitled "Instructions for the Use of Basic Radio Propagation Predictions," available from the Superintendent of Documents, U. S. Government Printing Office, Washington 25, D. C., price 25 cents (foreign, 35 cents). It is a simple matter to trace the necessary great-circle paths on a piece of transparent paper and then transfer these lines to the world map. Photographic methods can be used to enlarge the map and adapt it to direction-finding or beam-bearing indication.

WORKING FROM A GLOBE

Entirely satisfactory bearings for beam purposes can be taken from an ordinary globe with nothing more complicated than a small school

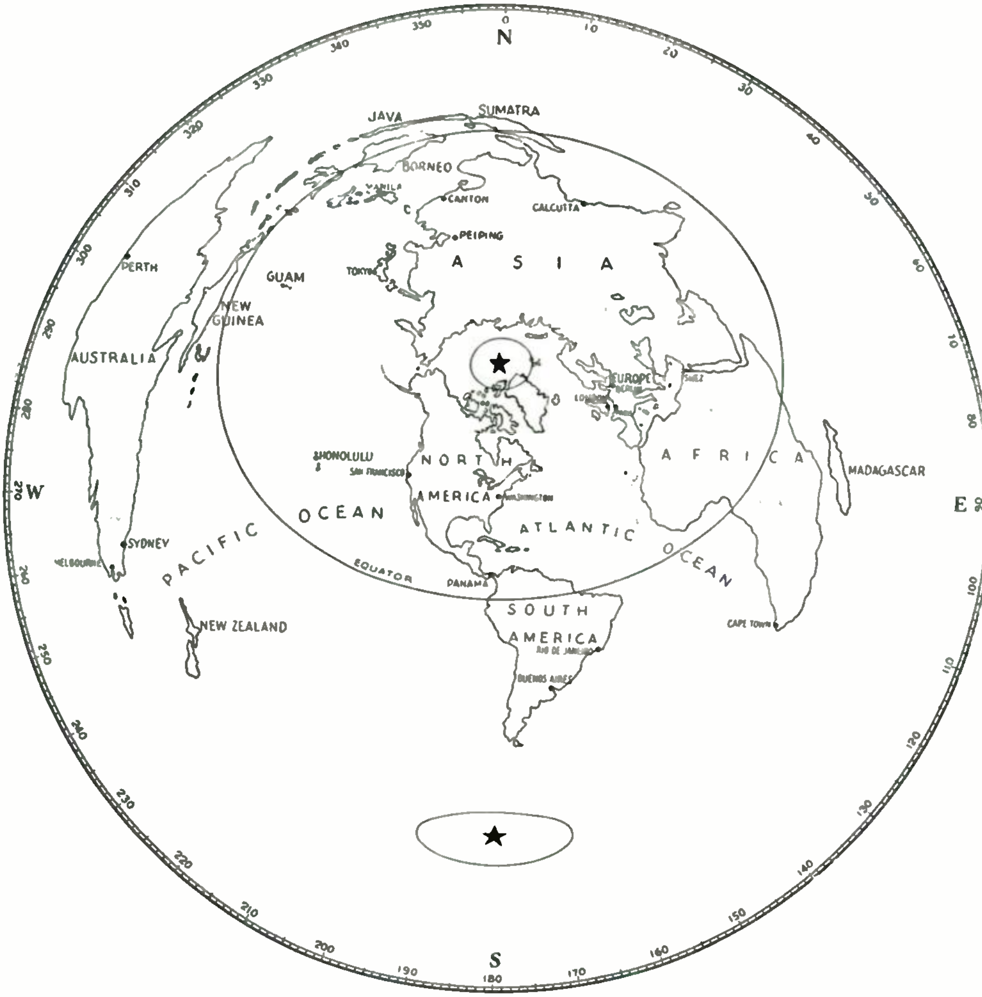


Fig. 13-1 — Azimuthal map centered on Washington, D. C.

protractor of the type available in any school-supply or stationery store, as illustrated in Fig. 13-4. For best results, however, the globe should be at least eight inches in diameter.

From a piece of thin paper, cut out a small circle — something like a three-inch circle for use with an eight-inch globe. Put a pin through the center and draw a straight line from the center to any point on the circumference. Now, put the paper circle on the globe, sticking the center pin into your location. Using the edge of a sheet of plain paper as a straightedge, line up the straight line on your paper circle so that it points North; this is done by laying the straightedge against the center pin and running it up to the North Pole at the top of the globe, then turning the paper circle until the straight line on it coincides with the straightedge. When you have done this, stick another pin through the paper circle into the globe to hold it in position with this line pointing North.

Now all you have to do is to use your paper straightedge from the center pin to such points as you wish, drawing short lines on your paper circle and labeling them as required. These lines may be extended later to the periphery of the circle.

With your protractor it is now a simple matter to determine the bearing, in degrees from North, of any of the points.

If your problem is to lay out a long wire to best advantage, make a diagram from the data in Chapter Two, showing the angular direction of the lobes, and superimpose this on your direction chart, adjusting it until the theoretical power lobes seem to take in the points in which you are

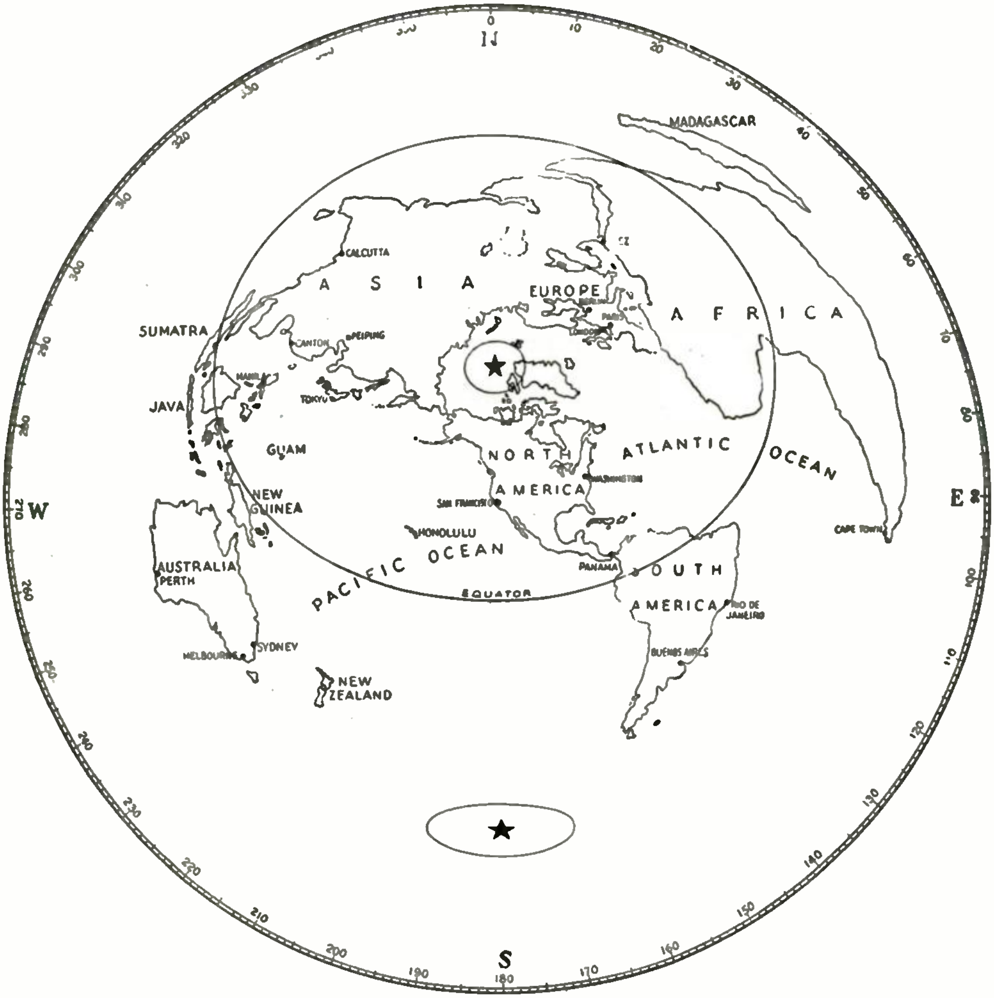


Fig. 13-2 — Azimuthal map centered on San Francisco, Calif.

interested. The direction of the wire can then be determined with the protractor.

DIRECTION AND DISTANCE BY TRIGONOMETRY

The method to be described will give the bearing (and distance) as accurately as one cares to compute them. All that is required is a table of latitude and longitude, such as may be found in the *World Almanac*, and a set of trigonometry tables. For most purposes, the latitude and longitude can be taken from maps of the areas in question. Two formulae are used:

$\cos D = \sin A \sin B + \cos A \cos B \cos L \ldots (1)$

and, $\sin C = \cos B \csc D \sin L \ldots\ldots\ldots (2)$

where
A = *your* latitude in degrees
B = the other location in degrees (positive for N latitude, negative for S latitude)
L = longitude difference between you and the other location
C = the direction of the other location from yours, in degrees East or West from North
D = distance along path, in nautical miles or minutes of arc (1 min. = 1 nautical mile = 1.152 statute miles)

The following example will show how the formulae are used. To find the bearing and distance of Cairo from Chicago:

From the *Almanac* or map:
Chicago = 41° 52′ N, 87° 38′ W
Cairo = 30° 00′ N, 31° 14′ E
$A = 41° 52'$
$B = 30°$
$L = 87° 38' + 31° 14' = 118° 52'$

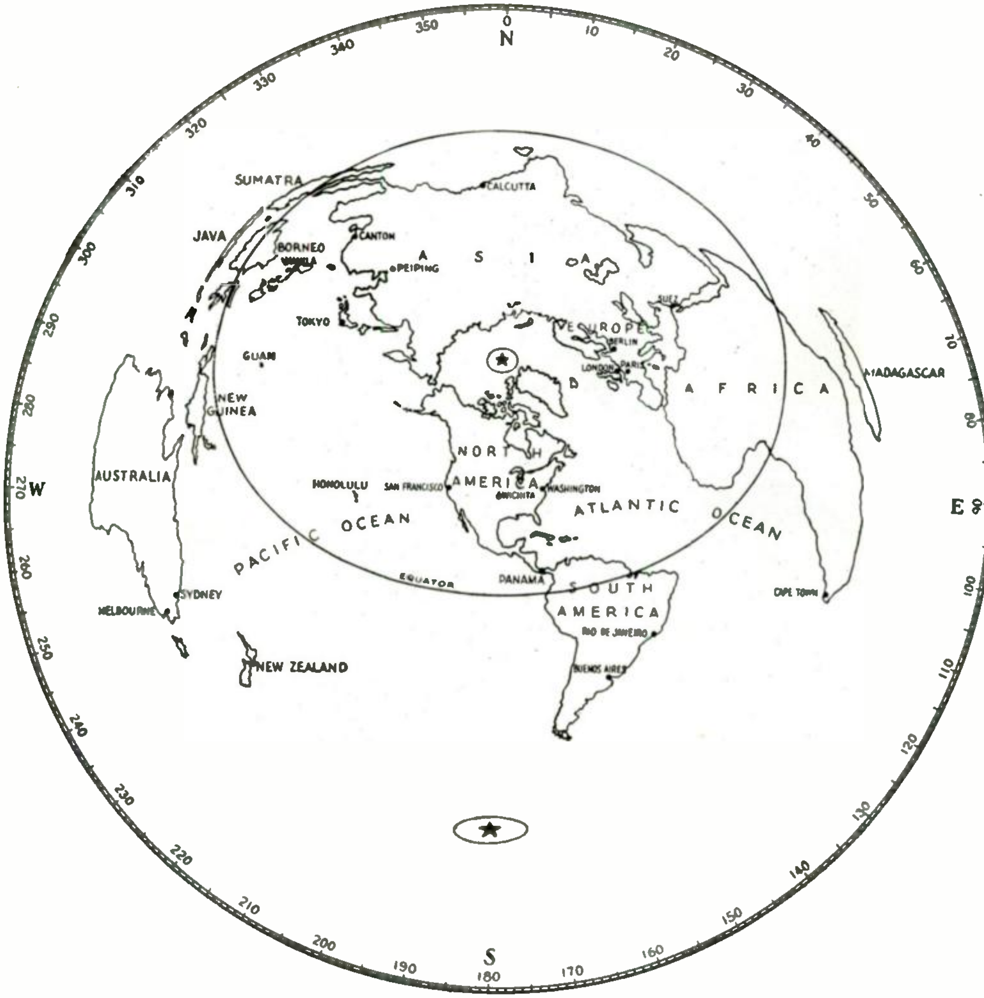


Fig. 13-3 — Azimuthal map centered on Wichita, Kansas.

Substituting in (1):

$$\cos D = \sin(41°\ 52') \times \sin(30°) + \cos(41°\ 52') \times \cos(30°) \times \cos(118°\ 52')$$
$$= .6774 \times .5 + .7447 \times .8660 \times (-.4828)$$
$$= .3387 - .3141$$
$$= .0246$$
$$D = 88°\ 35' = 35 + 88 \times 60 = 35 + 5280$$
$$= 5315 \text{ nautical miles}$$
$$= 5315 \times 1.152 = 6123 \text{ statute miles}$$

Substituting in (2):

$$\sin C = \cos(30°) \times \csc(88°\ 35') \times \sin(118°\ 52')$$
$$= .8660 \times \frac{1}{\sin(88°\ 35')} \times .8758 = \frac{.7584}{.9997}$$
$$= .7585$$
$$C = 49°\ 20'$$

Thus the true bearing of Cairo is 49° 20′ from Chicago and, by inspection of a globe, this direction is seen to be *East* of North. With the antenna in Chicago pointed 49° 20′ East of North, a maximum signal should be pumped into Cairo, 6123 miles away. Similarly, the direction and range between any two points on the earth can be computed.

DETERMINING TRUE NORTH

Determining the direction of distant points is of little use to the amateur erecting a directive array unless he can put up the array itself in the desired direction. This, in turn, demands a knowledge of the direction of *true* North (as against magnetic North), since all our directions from globe or map are worked in terms of true North.

A number of ways may be available to the

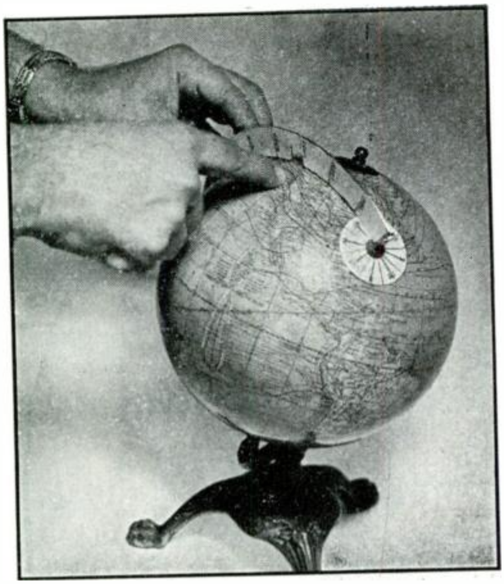


Fig. 13-4 — A direction indicator made from a semicircle of thin metal can be fitted easily to a small globe. Pins at the ends permit fastening one end to the home location, the other to the antipodes. The paper scale is marked in miles to show approximate distances (12,500 miles to the semicircle).

amateur for determining true North from his location. Frequently, the streets of a city or town are laid out, quite accurately, in North-South and East-West directions. A visit to the office of your city engineer will enable you to determine whether or not this is the case for the street in front of or paralleling your own lot. Or from such a visit it is often possible to locate some landmark, such as factory chimney or church spire, which lies true North with respect to your house.

If you cannot get true North by such means, three other methods are available: compass, Pole Star and Sun.

By Compass: Get as large a compass as you can; it is difficult, though not impossible, to get satisfactory results with the "pocket" type. In any event, the compass *must* have not more than two degrees per division for satisfactory results.

It must be remembered that the compass points to *magnetic* North, not true North. The amount by which magnetic North differs from true North in a particular location is known as *variation.* Your local weather bureau or city engineer's office can tell you the magnetic variation for your locality. When correcting your "compass North," do so *opposite* to the direction of the variation. For instance, if the variation for your locality is 12 degrees West (meaning that the compass points 12 degrees West of North) then true North is found by counting 12 degrees *East* of North as shown on the compass.

When taking the bearing, make sure that the compass is located well away from ironwork, fencing, pipes, etc. Place the instrument on a wooden tripod or support of some sort, at a convenient height as near eye level as possible. Make yourself a sighting stick from a flat stick about two feet long with a nail driven upright in each end (for use as "sights") and then, after the needle of the compass has settled down, carefully lay this stick across the face of the compass — with the necessary allowance for variation — to line it up on true North. *Be sure you apply the variation correctly.*

This same sighting-stick-and-compass rig can also be used in laying out directions for supporting poles for antennas in other directions — provided, of course, that the compass dial is graduated in degrees.

By the Pole Star: Many amateurs use the Pole Star in determining the direction of true North. An advantage is that the Pole Star bears true North, so that no corrections are necessary. Disadvantages are that some people have difficulty identifying the Pole Star, which is none too bright at best, and that because of its comparatively high angle above the horizon it is not always easy to "sight" on it accurately.

In any event, it is a handy check on the direction secured by other means.

By the Sun: With some slight preparation, the sun can easily be used for determination of true North. One of the most satisfactory methods is described below. The method is based on the fact that exactly at noon, local time, the Sun bears due South, so that at that time the shadow of a vertical stick or rod will bear North. The resulting

TABLE 13-I

Apply to Clock Time as Indicated by the Sign, To Get Time of True Noon

Jan. 1	+ 4 min.	July 10	+ 5 min.
10	+ 8 "	20	+ 6 "
20	+ 11 "	30	+ 6 "
30	+ 13 "		
Feb. 10	+ 14 "	Aug. 10	+ 5 "
20	+ 14 "	20	+ 3 "
28	+ 13 "	30	+ 1 "
Mar. 10	+ 10 "	Sept. 10	− 3 "
20	+ 8 "	20	− 6 "
30	+ 5 "	30	− 10 "
Apr. 10	+ 1 "	Oct. 10	− 13 "
20	− 1 "	20	− 15 "
30	− 3 "	30	− 16 "
May 10	− 4 "	Nov. 10	− 16 "
20	− 4 "	20	− 14 "
30	− 3 "	30	− 11 "
June 10	− 1 "	Dec. 10	− 7 "
20	+ 1 "	20	− 2 "
30	+ 3 "	30	+ 2 "

shadow direction, incidentally, is *true* North.

Two corrections to your standard time must be made to determine the exact moment of local true noon.

The first is a longitude correction. Standard Time is time at some particular meridian of longitude: EST is based on the 75th meridian; CST on the 90th meridian; MST on the 105th meridian; and PST on the 120th meridian. From an atlas, determine the difference between your own longitude and the longitude of your time meridian. Getting this to the nearest 15 minutes of longitude is close enough. Example: West Hartford, which runs on 75th meridian time (EST) is at 72° 45′ longitude, or a difference of 2° 15′. Now, for each 15′ of longitude, figure 1 minute of time; thus 2° 15′ is equivalent to 9 minutes of time (there are 60 "angle" minutes to a degree, so that each degree of longitude equals 4 minutes of time). *Subtract* this correction from noon if you are *East* of your time meridian; *add* it if you are *West*.

To the resulting time, apply a further correction for the date from Table 13-I. The resulting time is the time, by Standard Time, when it will be true noon at your location. Put up your vertical stick (use a plumb bob to make sure it is actually vertical), check your watch with Standard Time and, at the time indicated from your calculations, mark the position of the shadow That is true North.

(In the case of West Hartford, if we wanted correct time for true noon on October 20th: First, subtracting the longitude correction — because we are East of the time meridian — we get 11:51 A.M.; then, applying the further correction of −15 minutes, we get 11:36 A.M. as the time of true noon at West Hartford on October 20th.)

Receiving Antennas

For the amateur engaged in traffic work or general rag-chewing, where operating convenience overshadows any other factor, a separate straight-wire antenna of random length is normally quite satisfactory for reception. This should preferably be as far removed from the transmitting antenna as possible, so that a minimum of energy from the transmitter will be picked up. If a short wire picks up too much "man-made noise" in the form of hash and interference from motors and other electrical appliances, it is advisable to use a half-wave antenna cut for the frequency band most often used and fed by low-impedance twisted pair such as is sold for "all-wave" broadcast antennas and television receivers. The folded dipole with 300-ohm feeders is a very good receiving antenna. By placing the antenna at a point removed from the source of noise, the noise will be reduced and, since the line is balanced, the low-impedance line will run back through the noisy area but will not pick up any noise. This is shown in Fig. 14-1. Low-impedance line will not in itself reduce noise, but it allows the antenna to be placed at a distance and still feed its signal back to the receiver. If a half-wave antenna is used, it should be placed at right angles to the transmitting antenna and as far away as possible.

RECEIVING WITH DIRECTIVE ANTENNAS

The amateur fortunate enough to own a directive antenna, or even the one who is content to work without break-in, should always use his transmitting antenna for receiving. The reason is obvious: signals that are loud when the transmitting antenna is used for receiving indicate that the antenna is favorable for that direction and consequently will put a signal there. If separate antennas are used, a signal might be heard for a direction where the transmitting antenna has a null, and all the calling one could muster probably wouldn't result in a QSO. With rotary antennas, one can rotate the array until the signal peaks up and the operator can then feel assured that his antenna is aimed correctly at the station. Unfortunately, no simple broad-band system has yet been devised to permit using the same antenna system for transmitting and receiving during break-in operation, and still retain the directional characteristics of the antenna while receiving.

If two or more directive systems are available, some provision should be made for quick switching from one to the other or others, so that when listening one can identify the direction of the signal even before the station signs. Even with antennas with no marked directional characteristics, switching will show which one yields the loudest signal and hence should be used for transmitting as well. The coupling to the receiver should duplicate the transmitter coupling, however, to insure that the antenna system works the same for transmitting and receiving.

Many stations operate with a d.p.d.t. switch hand-operated by the operator to switch the antenna from the transmitter to the receiver and back again. However, it is far more satisfactory to use one of the antenna change-over relays now available on the market. They are made with good insulation and work from the 115-volt a.c. line, so that it is only necessary to connect them across the primary of a plate transformer and every time the transmitter is switched on the antenna relay operates and switches the antenna from the receiver to the transmitter. The relay should preferably be mounted on the wall where the antenna feeders come in the shack. If untuned-line feed is used to the antenna, the line

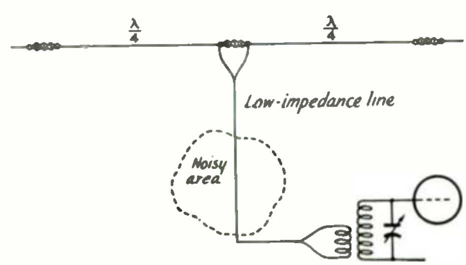


Fig. 14-1 — The doublet antenna with low-impedance line is used for noise reduction by placing the antenna outside or away from the noisy area and running a low-impedance line to it. The line itself doesn't reduce the noise but, because it has no pick-up itself, it enables the antenna to be fed through an interference-generating area.

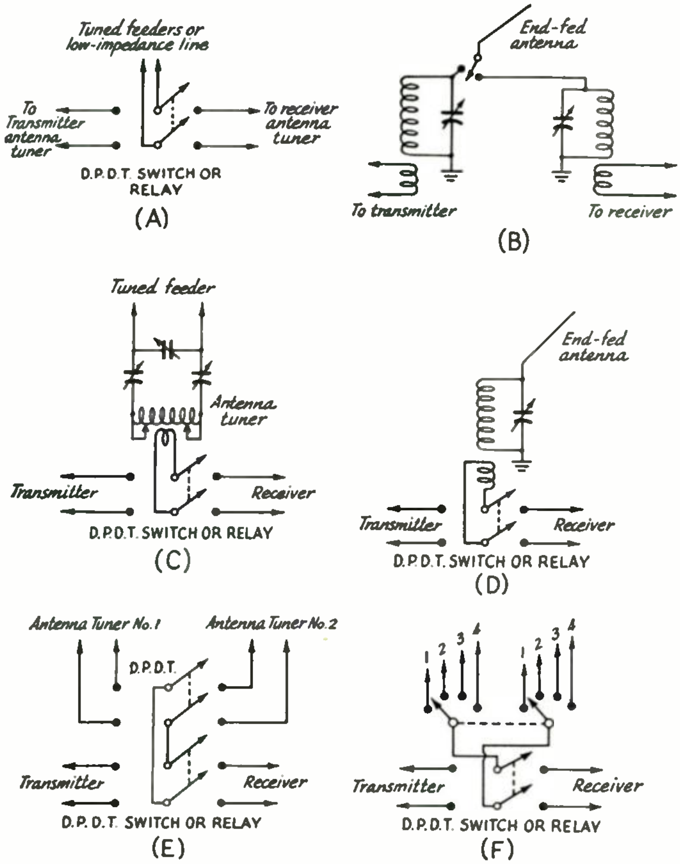


Fig. 14-2 — Antenna switching systems. A — For tuned or untuned lines with separate antenna tuners. B — For voltage-fed antenna. A series-tuned circuit will be required with some lengths. See Chapter Six. C — For tuned line with single tuner. D — For voltage-fed antenna with single tuner. See B above. E — For two tuned-line antennas with tuner for each antenna or for low-impedance lines. F — For several two-wire lines.

inside the shack should be made of the same impedance as the line that runs outside the station to the antenna. The line inside the shack doesn't have to have the same dimensions, but the wire size and spacing should be such that the impedance of the line is the same as that outside the station. The same applies to the line running to the receiver, but this is not too important and can be disregarded if it seems to be too much trouble. However, for maximum signal into the receiver, the line impedance should remain as nearly constant as possible and the line should be matched at the receiver.

The change-over relay (or switch) does not necessarily have to be connected in the antenna feedline. If an antenna tuning unit is used, link-coupled to the transmitter, the relay can be placed in the link line, and the receiver will then be able to take advantage of the signal build-up in the tuning unit. The various ways in which the change-over relay (or switch) can be used are shown in Fig. 14-2.

Using the same antenna tuning unit for both transmitting and receiving is advantageous in that it reduces the amount of equipment needed; furthermore, the greatest response to incoming signals is automatically secured at and near the frequency to which the transmitter is tuned. However, the separate tuner is more convenient in that it is possible to listen on other frequencies, and on other bands, without loss of signal strength and without disturbing the tuning for transmission.

The circuits of Figs. 14-2E and 14-2F show no coupling device to the receiver. While in many cases this is perfectly satisfactory, and particularly at frequencies below 10 Mc., additional signal strength at the receiver can often be obtained by matching the receiver to the line. Several methods of doing this are shown in Fig. 14-3. The networks should be located fairly close to the receiver, in a position convenient for occasional adjustment by the operator.

In some instances it will be found that a rotary beam used for receiving does not show the rejection to unwanted signals off the side that might be expected from such an antenna. In most of these cases, it will be found that the receiver is coupling in some of the parallel components on the transmission line, and the line itself is acting as a vertical or semivertical antenna. To make sure that the receiver is not at fault, connect a 3- or 4-foot length of the feedline to the receiver, to see if signals can be picked up on the feedline alone. If a balanced line such as 300-ohm Twin-Lead is used, connect it to the two antenna terminals of the receiver, leaving the ground post open. If a shielded two-wire line, or two pieces of coaxial line, are used, connect the inner conductors to the antenna terminals and the outer conductor(s) to the ground post. With single coaxial line, connect the inner conductor to one antenna terminal and the outer conductor to ground *and* the other antenna terminal. No signals should be heard, even with the receiver running at maximum gain.

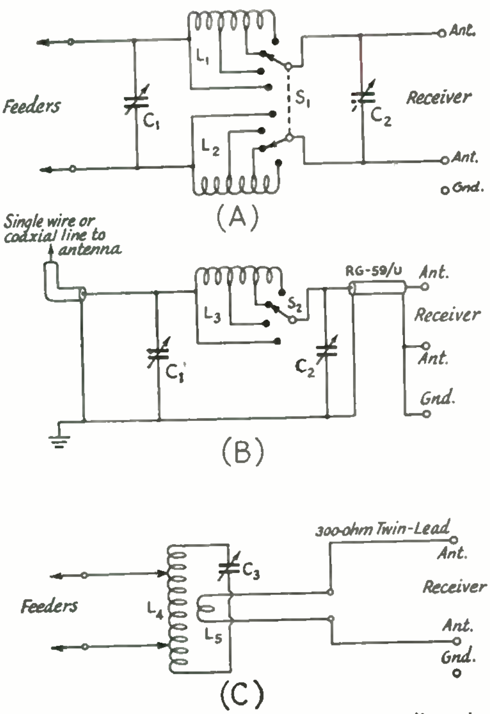


Fig. 14-3 — Three types of circuits for coupling the antenna to the receiver. A — Balanced pi-section network for use with balanced lines of any kind. B — Single-ended pi-section network for use with single wires or coaxial line. C — Tuned circuit for use with any balanced line.

C_1, C_2, C_3 — 100-$\mu\mu$fd. midget variable.
L_1, L_2, L_3 — 30 turns No. 28 d.c.c. close-wound on ½-inch diameter form, tapped at 2½, 6½ and 14½ turns.
L_4 — Proportioned to resonate with C_3 in the desired band.
L_5 — 2- or 3-turn swinging link at center of L_4.
S_1 — 2-circuit 5-position single-section ceramic wafer switch.
S_2 — 1-circuit 5-position single-section ceramic wafer switch.

If signals are heard, an antenna coupler as shown in Fig. 14-3C will help to eliminate them if a single length of coaxial line is run from L_5 to the receiver and connected as described above for the test with single coaxial line. However, the receiver must be capable of passing the test for single coaxial line. If it won't, but is satisfactory with shielded two-wire line, use shielded two-wire line from L_5 to the antenna terminals on the receiver.

GROUNDS

Most modern receivers do not require an external ground, since they are grounded through the power supply by the capacity of the transformer windings. However, in some instances a direct ground to the receiver will boost the signal pick-up and, in the case of regenerative receivers, reduce the body capacity. The direct ground can be made by running a wire to a radiator or water pipe or directly to a pipe in the ground.

LOOP ANTENNAS

Loop antennas can be used on any amateur band, but they find their greatest usefulness below 4 Mc. In the usual location, the pick-up inside a house at the high frequencies is not too good, and metallic objects in the vicinity, such as house wiring and transmitter control wires, tend to confuse the directional characteristics of the loop. However, it is sometimes very useful in rejecting unwanted signals, and two types of loops are shown in Fig. 14-4. The loop at A is a little less elaborate than that of B and will not give quite as good results, but it is still useful in rejecting signals.

In either case, the loop should be made from not more than about 20 feet of flexible insulated wire (stranded No. 18 aircraft wire or equivalent), formed in a square of from 12 to 20 inches on a side or a circle of from 12 to 20 inches in diameter. It can be mounted on the face of a large sheet of plywood, and it should be hinged in both the vertical and horizontal planes, so that it can be pointed in any direction. The tuning condenser

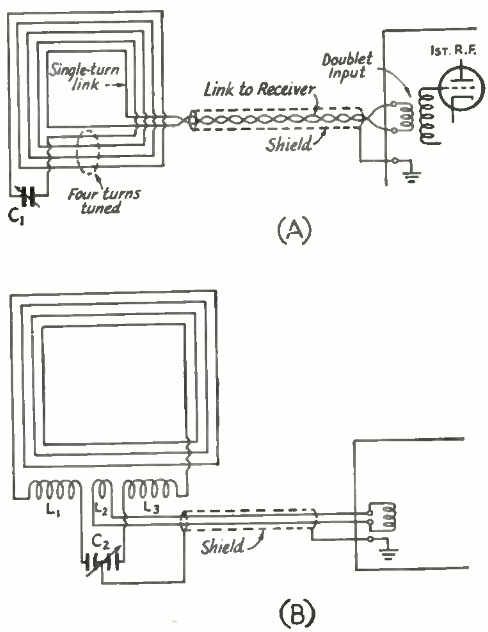


Fig. 14-4 — Two types of loop antennas for 3.5-Mc. reception. The loops are made of not more than about 20 feet of No. 18 stranded wire formed in a circle or square of 3, 4 or 5 turns.

C_1 — 150-$\mu\mu$fd. variable.
C_2 — 100 $\mu\mu$fd. each section.
L_1, L_3 — 17 turns No. 20, wound on 1½-inch diameter form and spaced to occupy ¾-inch winding length.
L_2 — 4-turn link winding, 1½-inch diameter, movable between L_1 and L_3.

is adjusted for maximum signal and, in the case of the loop of Fig. 14-4B, the coupling between L_2 and L_1L_3 should also be adjusted for best results.

Although the signal strength with a loop antenna is smaller than from a regular antenna, the sharp null of a properly-constructed loop often makes possible reception through noise or interference which could not be carried on with a conventional antenna. Except for the null, the directional characteristics are not very marked, so that the response to an interfering signal or source of man-made noise may be reduced to a very low value without affecting materially the response to the desired signal, providing only that the desired signal is not in exactly the same direction as the QRM.

In using any loop, it is essential to make sure that other antennas in the vicinity are detuned from the receiving frequency, since reradiation from regular antennas may be as strong or stronger than the direct signal received by the loop. In such case, of course, good nulls cannot be secured because of the supplementary fields of the reradiating wire or wires.

When a good null cannot be obtained even under ideal conditions, it indicates pick-up on the line from the loop to the receiver, and the shielding of this line and the symmetry of the loop should be investigated.

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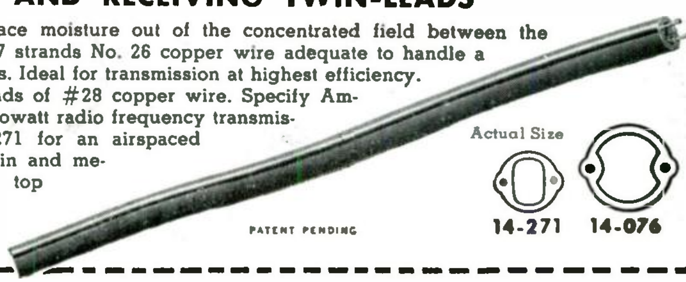

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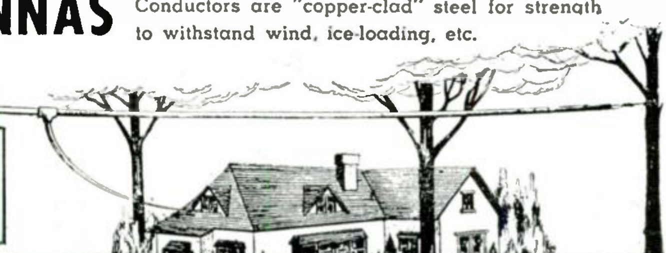

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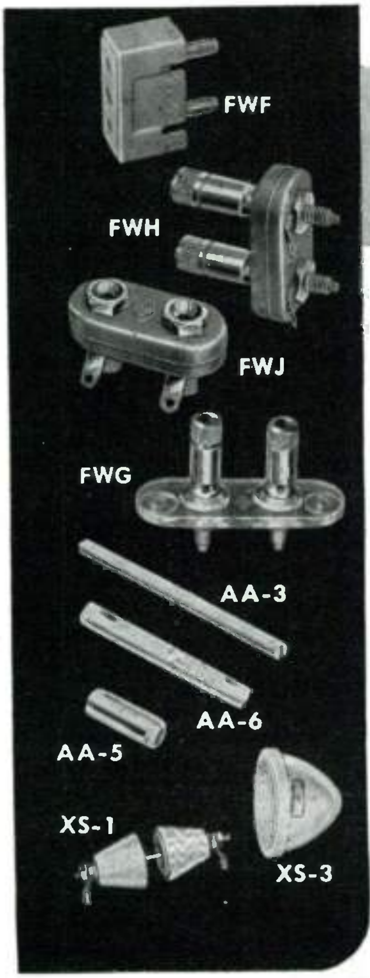

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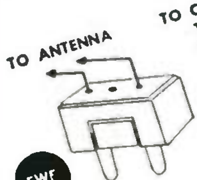

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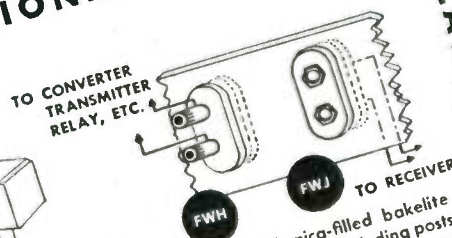

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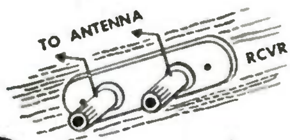

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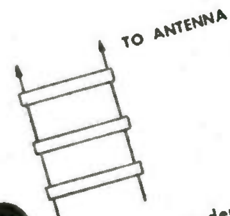

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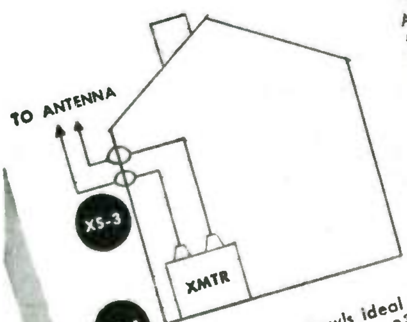

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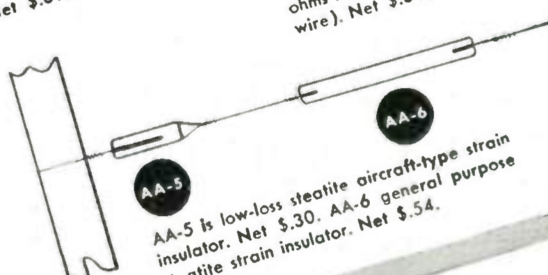

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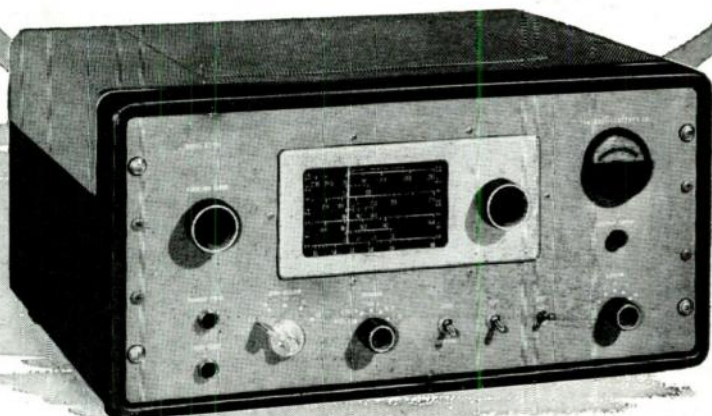

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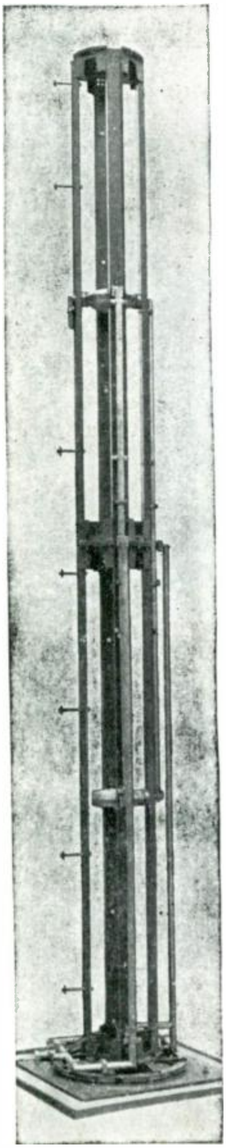

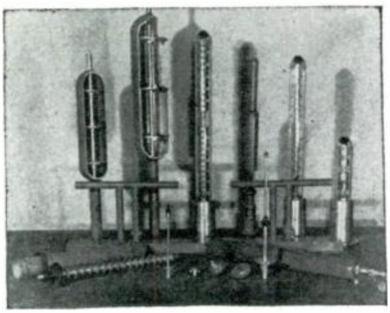

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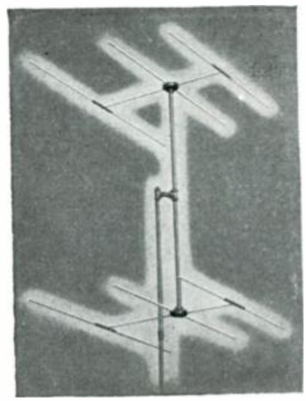

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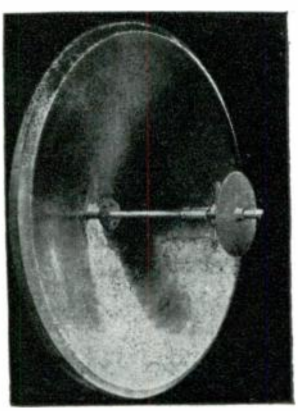

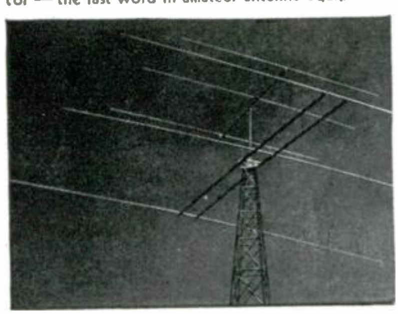

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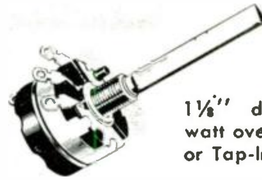

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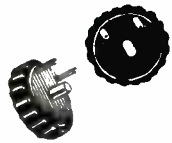

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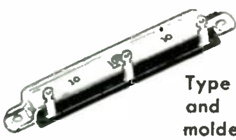

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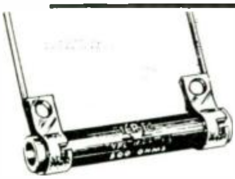

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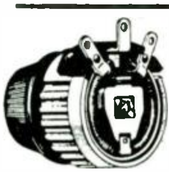

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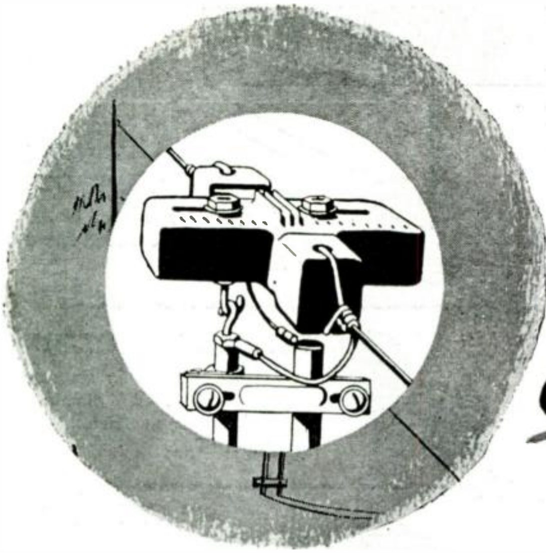

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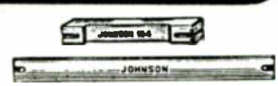

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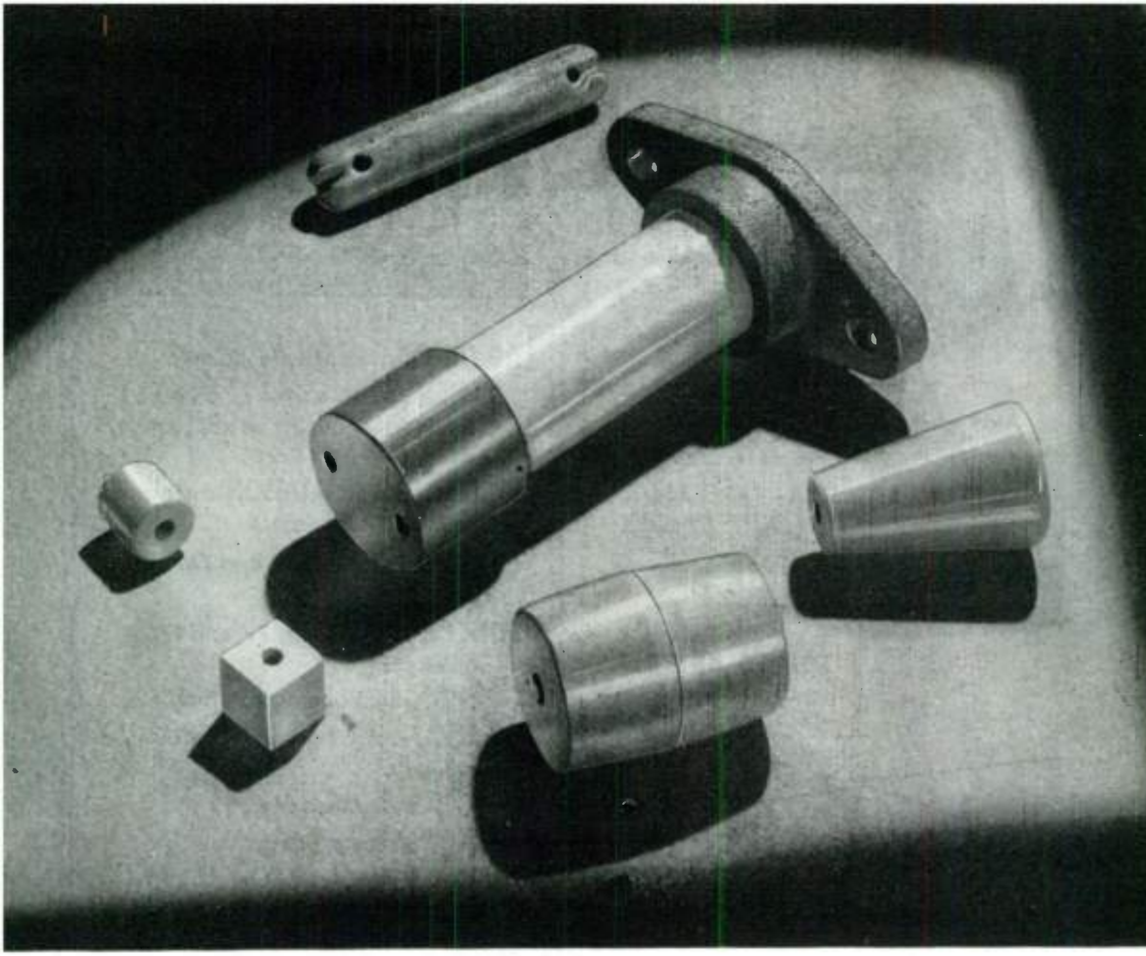

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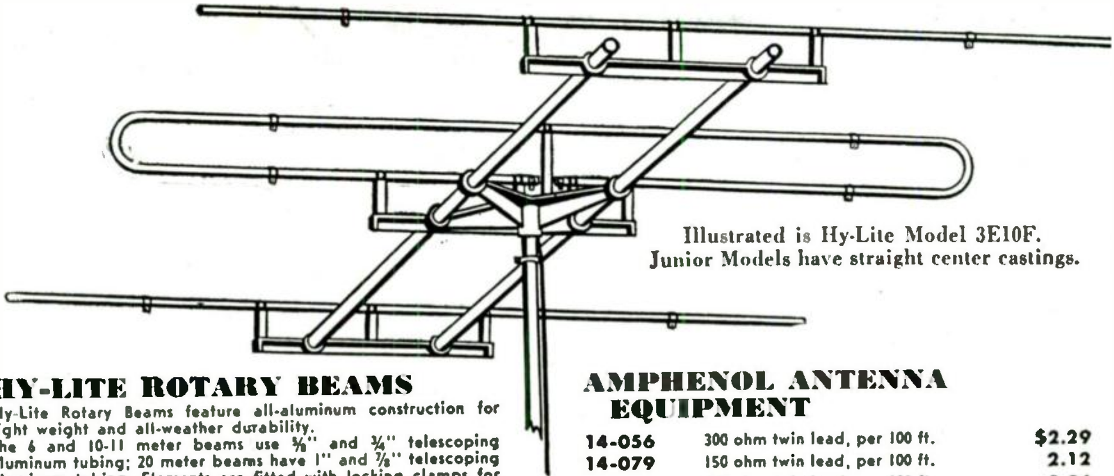
ANTENNAS



Illustrated is Hy-Lite Model 3E10F.
Junior Models have straight center castings.

HY-LITE ROTARY BEAMS

Hy-Lite Rotary Beams feature all-aluminum construction for light weight and all-weather durability.
The 6 and 10-11 meter beams use ⅝" and ¾" telescoping aluminum tubing; 20 meter beams have 1" and ⅞" telescoping aluminum tubing. Elements are fitted with locking clamps for securing the setting after tuning. Heavy steatite insulators with heavy rubber-covered clamps hold elements firm and straight. Strong aluminum castings support reflectors, directors and elements. Each antenna is supplied with a 6 foot mast. These antennas are designed with the future in mind. Any beam can be purchased and later a beam of a different frequency added to the original. Stacked arrays can be had in any combination on order.

Type No.	Style	Band	Shpg. Wt. Lbs.	Net Price
3E6	3 Element	6 Meter	18	$27.60
4E6	4 Element	6 Meter	21	33.35
3E6 JR	3 Element	6 Meter	14	23.30
4E6 JR	4 Element	6 Meter	16	27.60
3E10	3 Element	10-11	22	36.40
4E10	4 Element	10-11	26	45.95
3E10 JR	3 Element	10-11	17	28.60
4E10 JR	4 Element	10-11	20	34.95
2E20T	2 El. (incl. T Match) 20 M.		22	47.95
3E20T	3 El. (incl. T Match) 20 M.		28	64.95
4E10-20T	2 El. 10-11 and 2 El. 20 Meter Stacked with 2 T Matches		42	74.95
6E10-20T	3 El. 10-11 and 3 El. 20 Meter Stacked with 2 T Matches		48	98.95
6E10S	Two 3 El. 10-11 Meter Stacked		42	69.95
3E10-2E20T	3 El. 10-11 and 2 El. 20 Meter Stacked with 2 T Matches		45	84.90

FOLDED DIPOLE for 6 and 10-11 meter beams, add 6.00
T MATCH FOR 6, 10-11 and 20 meter beams, add 5.40

All above beams are close spaced (.1 Director—.15 Reflector) Except 3 element 20 meters (.075 Director, .1 Reflector)

Wide spaced beams (.15 Director—.2 Reflector) available on 3 element 6 and 10-11 meter beam only, additional cost. 4.50

FAMOUS "SIGNAL SQUIRTER"

AMPHENOL 140-005—The ultimate in deluxe rotary beam antennas is Amphenol's famous Mims Signal Squirter, dual 3-element beam for 10 and 20 meters. Kit includes Rotator (117 volts, 50/60 cycles) with double mounted pair Inducto-stubs assembly, Direction Indicator, center section, elements, insulators and all hardware ready for installation. **358.00**

AMPHENOL ANTENNA EQUIPMENT

14-056	300 ohm twin lead, per 100 ft.	$2.29
14-079	150 ohm twin lead, per 100 ft.	2.12
14-080	75 ohm twin lead, per 100 ft.	1.96
10-076	300 ohm kilowatt twin lead, tubular, constant impedance, per ft.	.07
14-023	75 ohm kilowatt twin lead, per ft.	.07
FOR TWIN LEAD,	screw eye standoff	.06
FOR COAXIAL,	screw eye standoff	.06
FOR TWIN LEAD,	polystyrene clip standoff	.06

Polystyrene Feeder Spreaders
2" — 9¢ 4" — 12¢ 6" — 15¢

SOLID ENAMELED Antenna Wire, per 100 ft.
No. 14 — 95¢ No. 12 — 1.35 No. 10 — 2.10

ELINCOR 200-EA 5 ELEMENT 2 METER BEAM

High gain broad band parasitic array uses folded dipole driver element, 3 directors and 1 reflector. Directors spaced .1 wavelength, reflector spaced .15 wavelength. This all-aluminum beam is pre-tuned to the center of the 2 meter band. Furnished with swivel bracket for horizontal or vertical plane mounting. May be fed with RG8/U 52 ohm cable. Complete with hardware and instructions **8.23**

ROTENNA Remote controlled beam rotator, 2 RPM, with automatic stop to prevent twisting of antenna feed line. Requires 115 volts AC to rotation control box which supplies 24 volts to antenna rotator. Positive, powerful — ideal for television and not-too-heavy ham beams. Complete **41.70**

AMPHENOL TWIN-LEAD *Folded Dipole Antennas*

Folded dipole section twin-lead conductors are copper-clad steel for withstanding severe stress of high winds, ice-loading, etc. The 75 ft. lead-in (300 ohm Twin-Lead) is joined to the antenna with a weatherproof molded polyethylene "T" junction. Simply trim to your operating frequency in accordance with the simple instructions furnished, add end insulators and go on the air with one of the best broad band antennas ever developed!

Band	Antenna Length	Net
10 meters	18 feet	4.53
20 meters	35 feet	5.64
40 meters	70 feet	7.94
80 meters	135 feet	12.20

NEW! Amphenol 300 ohm Twin-Lead with Copperweld conductors, #14-022. Per foot......... .07

Prices subject to change

Let our antenna specialists help you. Ask for Frank Miller, W2BUS and Ed Mandelberg, W2OTV. Call or write for detailed literature of any antenna — we stock all well-known lines.

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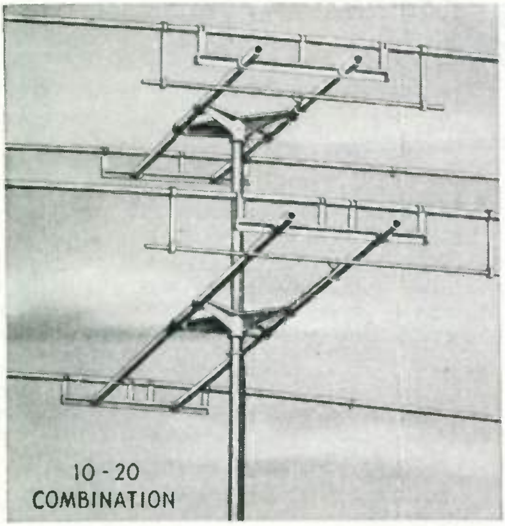
Many hams are TV conscious today. HY-LITE has a complete line of TV antennas built of the same high grade material and strong aluminum castings found in their very popular Ham beams. HY-LITE TV antennas come in many designs and combinations, complete with 6 ft. mast, guy ring, hinge base, lugs and complete instructions for assembly. WRITE FOR OUR NEW TV CATALOG for determining the antenna or combination best suited for your needs. Price list on request. JOBBERS: Many choice locations still open. ORDER NOW.

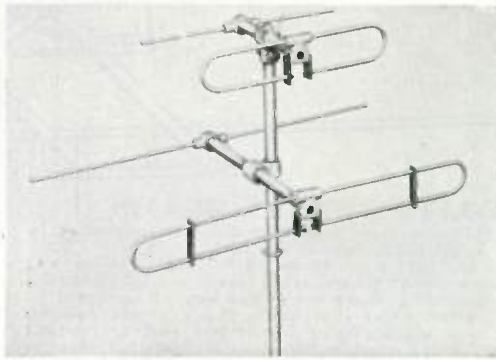
ONLY ONE OF THE MANY TV COMBINATIONS

GET OUR NEW CATALOG AH-4 NOW!



10 - 20 COMBINATION

WRITE FOR LITERATURE



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Collins 32V	*475.00*
Collins 310B-1	*190.00*
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National NC-57	*89.50*
National NC-173	*189.50*
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National HRO-7T	*292.50*
National HRO-7C	*372.45*
National HFS	*142.00*
National NC240D	*236.25*
Hallicrafters S38	*49.95*
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Hallicrafters SX43	*189.50*
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RME HF-10-20	*77.00*
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Gon-Set 10-11 converter	*39.95*
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