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February 2012

**Navy Electricity and
Electronics Training
Series**

Module 11—Microwave Principles

NAVEDTRA 14183A

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PREFACE

By enrolling in this self-study course, you have demonstrated a desire to improve yourself and the Navy. Remember, however, this self-study course is only one part of the total Navy training program. Practical experience, schools, selected reading, and your desire to succeed are also necessary to successfully round out a fully meaningful training program.

THE COURSE: This self-study course is organized into subject matter areas, each containing learning objectives to help you determine what you should learn along with text and illustrations to help you understand the information. The subject matter reflects day-to-day requirements and experiences of personnel in the rating or skill area. It also reflects guidance provided by Enlisted Community Managers (ECMs) and other senior personnel, technical references, instructions, etc., and either the occupational or naval standards, which are listed in the *Manual of Navy Enlisted Manpower Personnel Classifications and Occupational Standards*, NAVPERS 18068.

THE QUESTIONS: The questions that appear in this course are designed to help you understand the material in the text.

VALUE: In completing this course, you will improve your military and professional knowledge. Importantly, it can also help you study for the Navy-wide advancement in rate examination. If you are studying and discover a reference in the text to another publication for further information, look it up.

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Sailor's Creed

“I am a United States Sailor.

I will support and defend the
Constitution of the United States of
America and I will obey the orders of those
appointed over me.

I represent the fighting spirit of the Navy
and those who have gone before me to
defend freedom and democracy around the
world.

I proudly serve my country's Navy combat
team with honor, courage and commitment.

I am committed to excellence and the fair
treatment of all.”

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NAVY ELECTRICITY AND ELECTRONICS TRAINING SERIES

The Navy Electricity and Electronics Training Series (NEETS) was developed for use by personnel in many electrical and electronic-related Navy ratings. Written by, and with the advice of, senior technicians in these ratings, this series provides beginners with fundamental electrical and electronic concepts through self-study. The presentation of this series is not oriented to any specific rating structure, but is divided into modules containing related information organized into traditional paths of instruction.

The series is designed to give small amounts of information that can be easily digested before advancing further into the more complex material. For a student just becoming acquainted with electricity or electronics, it is highly recommended that the modules be studied in their suggested sequence.

Considerable emphasis has been placed on illustrations to provide a maximum amount of information. In some instances, knowledge of basic algebra may be required.

Course descriptions and ordering information may be found at <https://www.netc.navy.mil> then click on the Programs tab, then select the Nonresident Training Courses from the list.

Throughout the text of this course and while using technical manuals associated with the equipment you will be working on, you will find the below notations at the end of some paragraphs. The notations are used to emphasize that safety hazards exist and care must be taken or observed.

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AN OPERATING PROCEDURE, PRACTICE, OR CONDITION, ETC., WHICH MAY RESULT IN INJURY OR DEATH IF NOT CAREFULLY OBSERVED OR FOLLOWED.

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NOTE

An operating procedure, practice, or condition, etc., which is essential to emphasize.

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We value your suggestions, questions, and criticisms on our courses. If you would like to communicate with us regarding this course, we encourage you, if possible, to use e-mail or to post your comments on the Community of Practice (COP) page located at <https://wwwa.nko.navy.mil/portal/home/>. If you write or fax, please use a copy of the Student Comment form that follows this page.

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1 WAVEGUIDE THEORY AND APPLICATION

LEARNING OBJECTIVES

Upon completion of this chapter the student will be able to:

1. Describe the development of the various types of waveguides in terms of their advantages and disadvantages.
2. Describe the physical dimensions of the various types of waveguides and explain the effects of those dimensions on power and frequency.
3. Explain the propagation of energy in waveguides in terms of electromagnetic field theory.
4. Identify the modes of operation in waveguides.
5. Explain the basic input/output methods used in waveguides.
6. Describe the basic principles of waveguide plumbing.
7. Explain the reasons for and the methods of terminating waveguides.
8. Explain the basic theory of operation and applications of directional couplers.
9. Describe the basic theory of operation, construction, and applications of cavity resonators.
10. Describe the basic theory of operation of waveguide junctions.
11. Explain the operation of ferrite devices in terms of their applications.

1.1 INTRODUCTION TO WAVEGUIDE THEORY AND APPLICATION

That portion of the electromagnetic spectrum which falls between 1000 megahertz and 100,000 megahertz is referred to as the MICROWAVE region. Before discussing the principles and applications of microwave frequencies, the meaning of the term microwave as it is used in this module must be established. On the surface, the definition of a microwave would appear to be simple because, in electronics, the prefix "micro" normally means a millionth part of a unit. Micro also means small, which is a relative term, and it is used in that sense in this module. Microwave is a term loosely applied to identify electromagnetic waves above 1000 megahertz in frequency because of the short physical wavelengths of these frequencies. Short wavelength energy offers distinct advantages in many applications. For instance, excellent directivity can be obtained using relatively small antennas and low-power transmitters. These features are ideal for use in both military and civilian radar and communication applications. Small antennas and other small components are made possible by microwave frequency applications. This is an important consideration in shipboard equipment planning where space and weight are major problems. Microwave frequency usage is especially important in the design of shipboard radar because it makes possible the detection of smaller targets.

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Microwave frequencies present special problems in transmission, generation, and circuit design that are not encountered at lower frequencies. Conventional circuit theory is based on voltages and currents while microwave theory is based on electromagnetic fields. The concept of electromagnetic field interaction is not entirely new, since electromagnetic fields form the basis of all antenna theory. However, many students of electronics find electromagnetic field theory very difficult to visualize and understand. This module will present the principles of microwave theory in the simplest terms possible but many of the concepts are still somewhat difficult to thoroughly understand. Therefore, you must realize that this module will require very careful study for you to properly understand microwave theory. Antenna fundamentals were covered in *NEETS*, Module 10, *Introduction to Wave Propagation, Transmission Lines, and Antennas*. This module will show you the solutions to problems encountered at microwave frequencies, beginning with the transmission of microwave energy and continuing through to waveguides in chapter 1. Later chapters will cover the theory of operation of microwave components, circuits, and antennas. The application of these concepts will be discussed more thoroughly in later *NEETS* modules on radar and communications.

Q-1. What is the region of the frequency spectrum from 1000 MHz to 100,000 MHz called?

Q-2. Microwave theory is based upon what concept

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1.2 WAVEGUIDE THEORY

The two-wire transmission line used in conventional circuits is inefficient for transferring electromagnetic energy at microwave frequencies. At these frequencies, energy escapes by radiation because the fields are not confined in all directions, as illustrated in figure 1-1. Coaxial lines are more efficient than two-wire lines for transferring electromagnetic energy because the fields are completely confined by the conductors, as illustrated in figure 1-2.

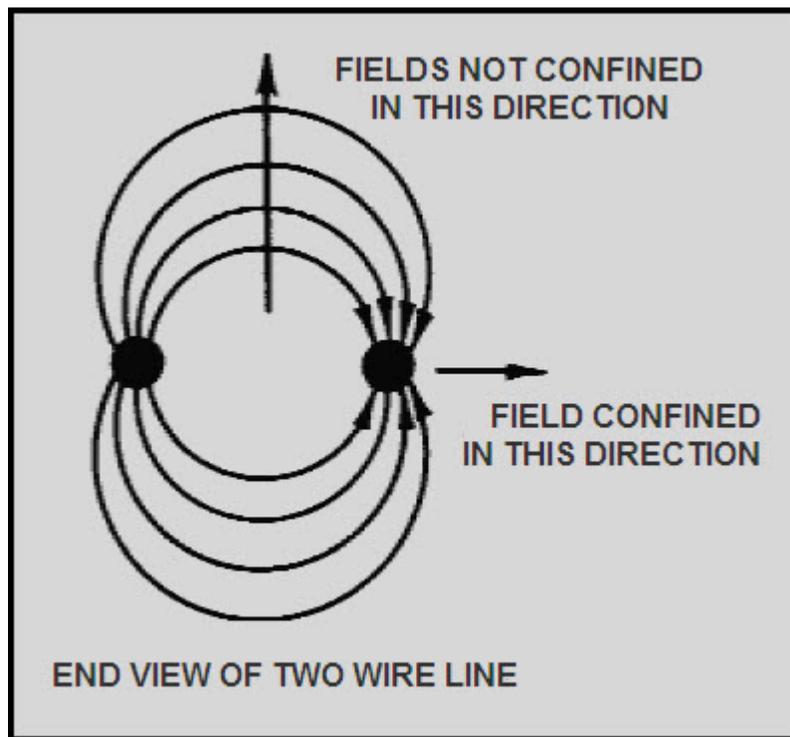


Figure 1-1 Fields confined in two directions only

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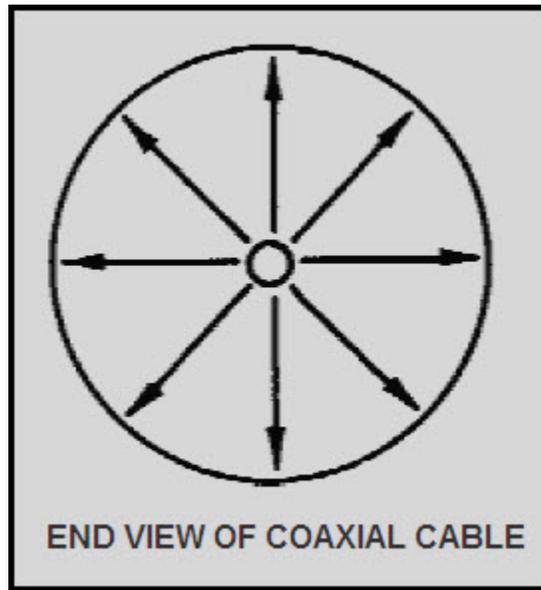


Figure 1-2 Fields confined in all directions

Waveguides are the most efficient way to transfer electromagnetic energy. WAVEGUIDES are essentially coaxial lines without center conductors. They are constructed from conductive material and may be rectangular, circular, or elliptical in shape, as shown in figure 1-3.

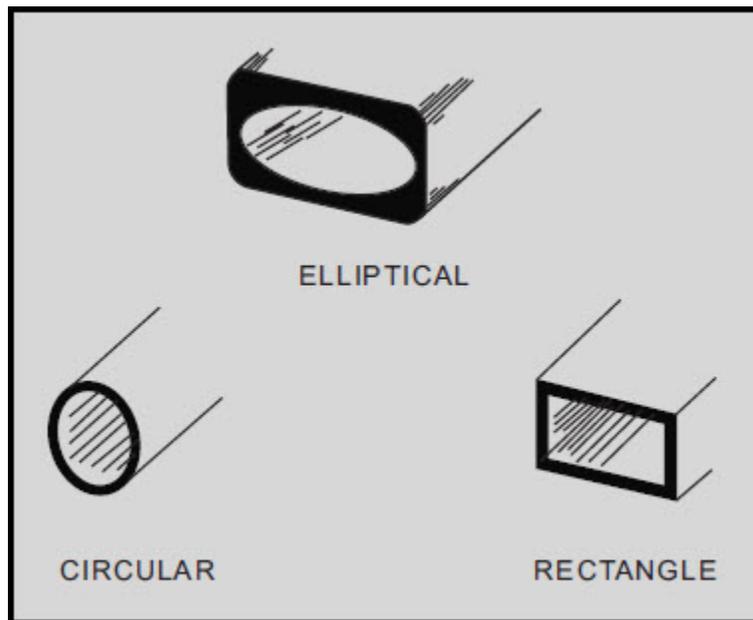


Figure 1-3 Waveguide shapes

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1.2.1 Waveguide Advantages

Waveguides have several advantages over two-wire and coaxial transmission lines. For example, the large surface area of waveguides greatly reduces COPPER (I^2R) LOSSES. Two-wire transmission lines have large copper losses because they have a relatively small surface area. The surface area of the outer conductor of a coaxial cable is large, but the surface area of the inner conductor is relatively small. At microwave frequencies, the current-carrying area of the inner conductor is restricted to a very small layer at the surface of the conductor by an action called SKIN EFFECT.

Skin effect was discussed in NEETS, Module 10, Introduction to Wave Propagation, Transmission Lines, and Antennas, Chapter 3. Skin effect tends to increase the effective resistance of the conductor. Although energy transfer in coaxial cable is caused by electromagnetic field motion, the magnitude of the field is limited by the size of the current-carrying area of the inner conductor. The small size of the center conductor is even further reduced by skin effect and energy transmission by coaxial cable becomes less efficient than by waveguides. DIELECTRIC LOSSES are also lower in waveguides than in two-wire and coaxial transmission lines. Dielectric losses in two-wire and coaxial lines are caused by the heating of the insulation between the conductors. The insulation behaves as the dielectric of a capacitor formed by the two wires of the transmission line. A voltage potential across the two wires causes heating of the dielectric and results in a power loss. In practical applications, the actual breakdown of the insulation between the conductors of a transmission line is more frequently a problem than is the dielectric loss.

This breakdown is usually caused by stationary voltage spikes or "nodes" which are caused by standing waves. Standing waves are stationary and occur when part of the energy traveling down the line is reflected by an impedance mismatch with the load. The voltage potential of the standing waves at the points of greatest magnitude can become large enough to break down the insulation between transmission line conductors.

The dielectric in waveguides is air, which has a much lower dielectric loss than conventional insulating materials. However, waveguides are also subject to dielectric breakdown caused by standing waves. Standing waves in waveguides cause arcing which decreases the efficiency of energy transfer and can severely damage the waveguide. Also since the electromagnetic fields are completely contained within the waveguide, radiation losses are kept very low.

Power-handling capability is another advantage of waveguides. Waveguides can handle more power than coaxial lines of the same size because power-handling capability is directly related to the distance between conductors. Figure 1-4 illustrates the greater distance between conductors in a waveguide.

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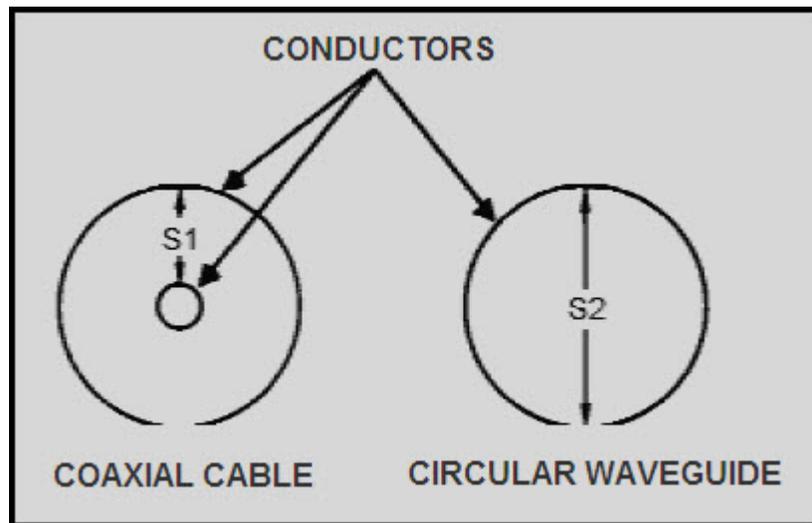


Figure 1-4 Comparison of spacing in coaxial cable and a circular waveguide

In view of the advantages of waveguides, you would think that waveguides should be the only type of transmission lines used. However, waveguides have certain disadvantages that make them practical for use only at microwave frequencies.

1.2.2 Waveguide Disadvantages

Physical size is the primary lower-frequency limitation of waveguides. The width of a waveguide must be approximately a half wavelength at the frequency of the wave to be transported. For example, a waveguide for use at 1 megahertz would be about 500 feet wide. This makes the use of waveguides at frequencies below 1000 megahertz increasingly impractical. The lower frequency range of any system using waveguides is limited by the physical dimensions of the waveguides.

Waveguides are difficult to install because of their rigid, hollow-pipe shape. Special couplings at the joints are required to assure proper operation. Also, the inside surfaces of waveguides are often plated with silver or gold to reduce skin effect losses. These requirements increase the costs and decrease the practicality of waveguide systems at any other than microwave frequencies.

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Q-3. Why are coaxial lines more efficient at microwave frequencies than two-wire transmission lines?

Q-4. What kind of material must be used in the construction of waveguides?

Q-5. The large surface area of a waveguide greatly reduces what type of loss that is common in two-wire and coaxial lines?

Q-6. What causes the current-carrying area at the center conductor of a coaxial line to be restricted to a small layer at the surface?

Q-7. What is used as a dielectric in waveguides?

Q-8. What is the primary lower-frequency limitation of waveguides?

1.2.3 Developing the Waveguide from Parallel Lines

You may better understand the transition from ordinary transmission line concepts to waveguide theories by considering the development of a waveguide from a two-wire transmission line. Figure 1-5 shows a section of two-wire transmission line supported on two insulators. At the junction with the line, the insulators must present a very high impedance to ground for proper operation of the line. A low impedance insulator would obviously short-circuit the line to ground, and this is what happens at very high frequencies. Ordinary insulators display the characteristics of the dielectric of a capacitor formed by the wire and ground. As the frequency increases, the overall impedance decreases. A better high-frequency insulator is a quarter-wave section of transmission line shorted at one end. Such an insulator is shown in figure 1-6. The impedance of a shorted quarter-wave section is very high at the open-end junction with the two-wire transmission line. This type of insulator is known as a METALLIC INSULATOR and may be placed anywhere along a two-wire line. Note that quarter-wave sections are insulators at only one frequency. This severely limits the bandwidth, efficiency, and application of this type of two-wire line.

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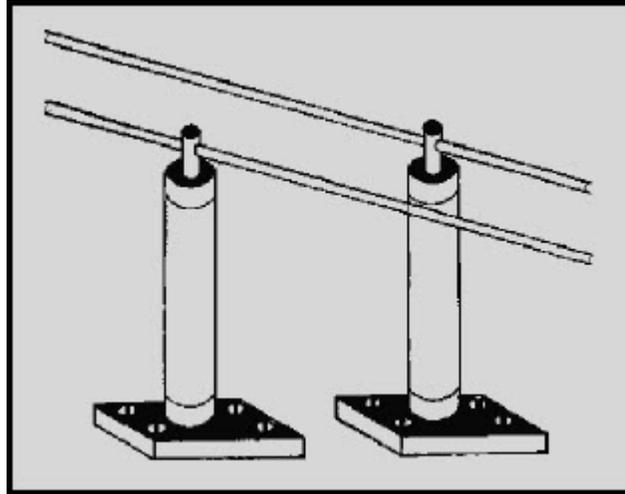


Figure 1-5 Two-wire transmission line using ordinary insulators

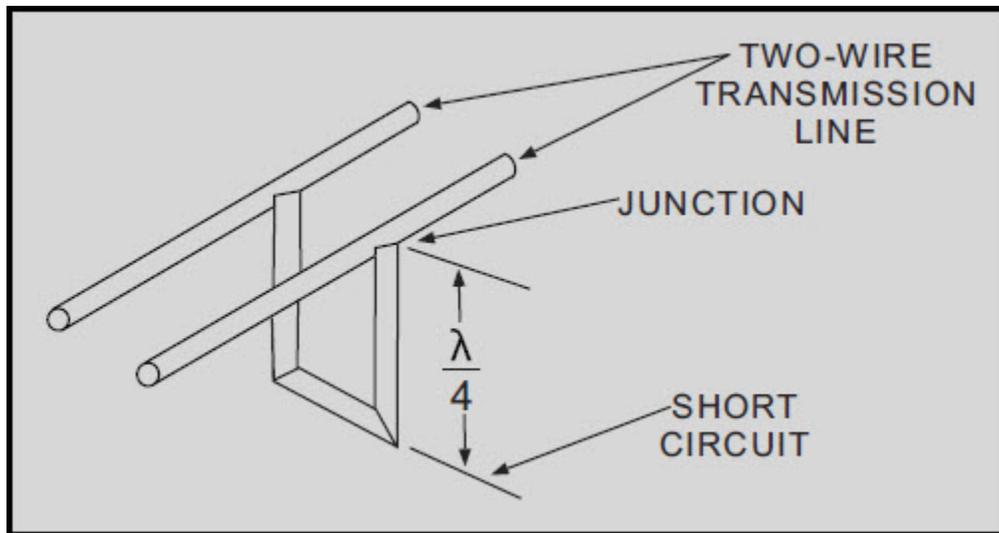


Figure 1-6 Quarter-wave section of transmission line shorted at one end

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Figure 1-7 shows several metallic insulators on each side of a two-wire transmission line. As more insulators are added, each section makes contact with the next, and a rectangular waveguide is formed. The lines become part of the walls of the waveguide, as illustrated in figure 1-8. The energy is then conducted within the hollow waveguide instead of along the two-wire transmission line.

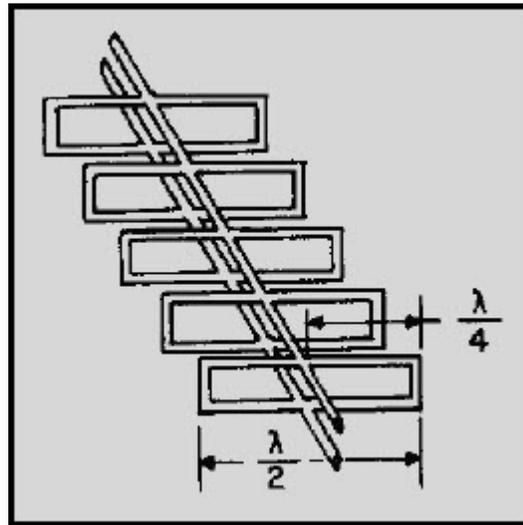


Figure 1-7 Metallic insulators on each side of a two-wire line

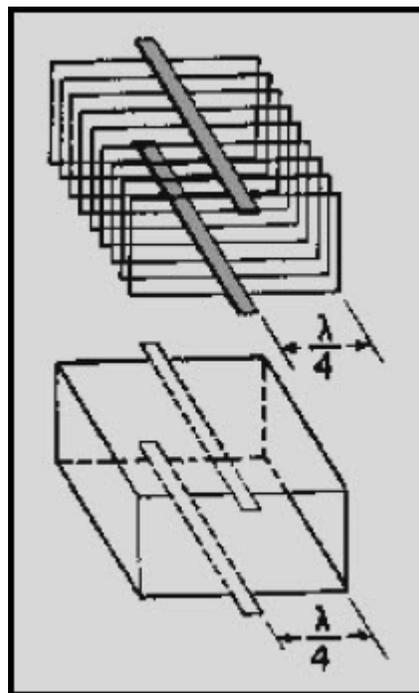


Figure 1-8 Forming a waveguide by adding quarter-wave sections

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The comparison of the way electromagnetic fields work on a transmission line and in a waveguide is not exact. During the change from a two-wire line to a waveguide, the electromagnetic field configurations also undergo many changes. These will be discussed later in this chapter. As a result of these changes, the waveguide does not actually operate like a two-wire line that is completely shunted by quarter-wave sections. If it did, the use of a waveguide would be limited to a single-frequency wavelength that was four times the length of the quarter-wave sections. In fact, waves of this length cannot pass efficiently through waveguides. Only a small range of frequencies of somewhat shorter wavelength (higher frequency) can pass efficiently.

As shown in figure 1-9, the widest dimension of a waveguide is called the "a" dimension and determines the range of operating frequencies. The narrowest dimension determines the power-handling capability of the waveguide and is called the "b" dimension.

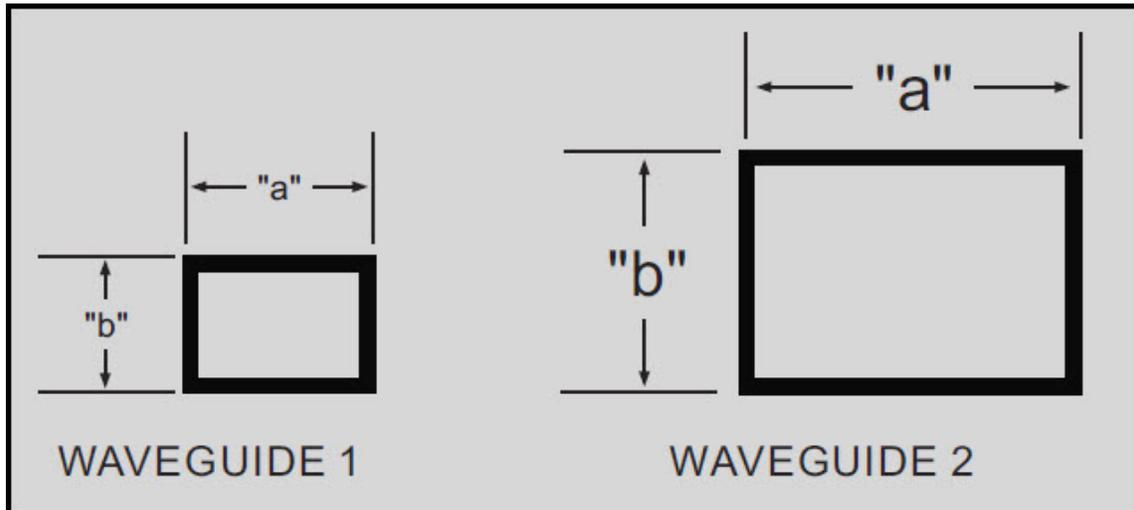


Figure 1-9 Labeling waveguide dimensions

NOTE

This method of labeling waveguides is not standard in all texts. Different methods may be used in other texts on microwave principles, but this method is in accordance with Navy Military Standards (MIL-STDS).

The ability of a waveguide of a given dimension to transport more than one frequency may be better understood by analyzing the actions illustrated in figure 1-10A, B, and C. A waveguide may be considered as having upper and lower quarter-wave sections and a central section which is a solid conductor called a BUS BAR. In figure 1-10A, distance mn is equal to distance pq, and both are equal to one quarter-wavelength ($\lambda/4$).

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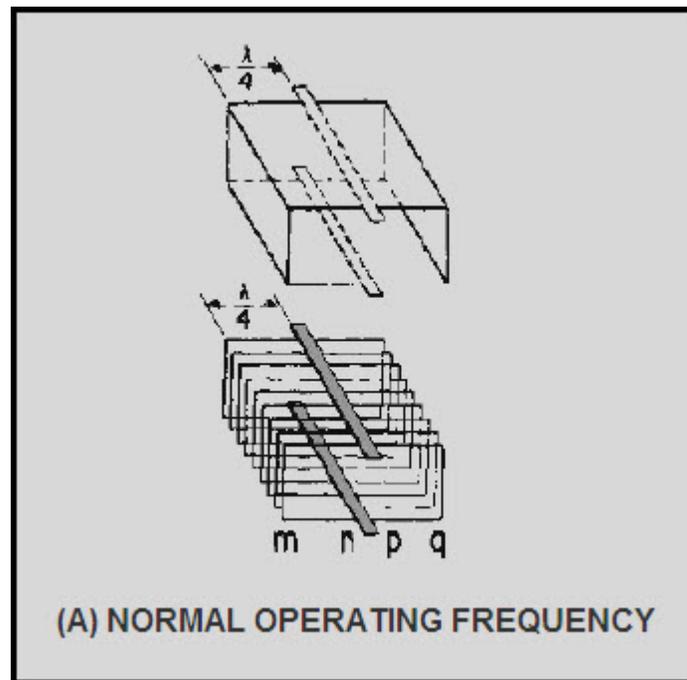


Figure 1-10A Frequency effects on a waveguide. NORMAL OPERATING FREQUENCY

NOTE

Throughout *NEETS*, $1/4\lambda$ and $\lambda/4$ are both used to represent one quarter-wavelength and are used interchangeably. Also, $\lambda/2$ and $3/2\lambda$ will be used to represent one half-wavelength and $1\ 1/2$ wavelengths, respectively. Distance np is the width of the bus bar. If the overall dimensions of the waveguide are held constant, the required length of the quarter-wave sections DECREASES as the frequency increases. As illustrated in figure 1-10B, this causes the width of the bus bar to INCREASE. In theory the waveguide could function at an infinite number of frequencies higher than the designed frequency; as the length of each quarter-wave section approaches zero, the bus bar continues to widen to fill the available space. However, in practice, an upper frequency limit is caused by modes of operation, which will be discussed later.

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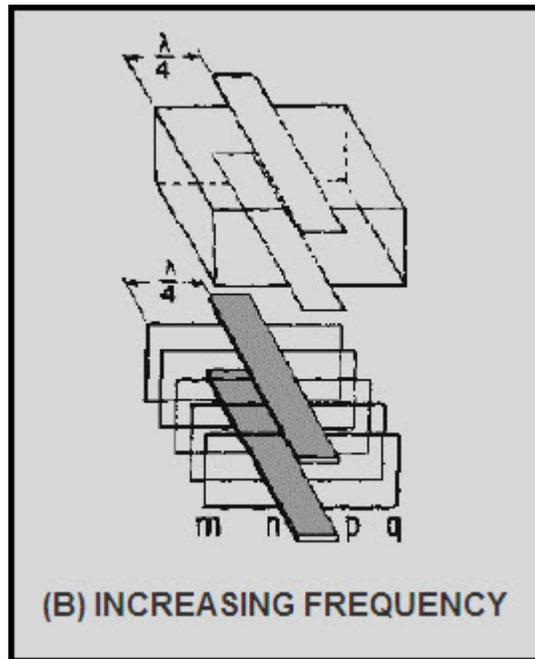


Figure 1-10B Frequency effects on a waveguide. INCREASING FREQUENCY

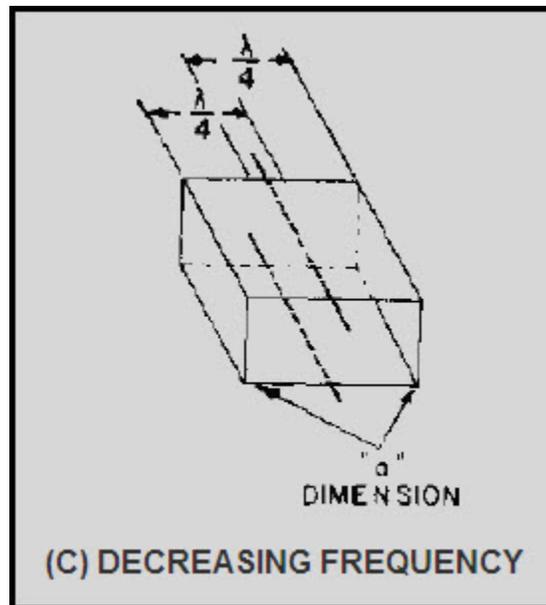


Figure 1-10C Frequency effects on a waveguide. DECREASING FREQUENCY

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If the frequency of a signal is decreased so much that two quarter-wavelengths are longer than the wide dimension of a waveguide, energy will no longer pass through the waveguide. This is the lower frequency limit, or CUT-OFF FREQUENCY, of a given waveguide. In practical applications, the wide dimension of a waveguide is usually 0.7 wavelength at the operating frequency. This allows the waveguide to handle a small range of frequencies both above and below the operating frequency. The "b" dimension is governed by the breakdown potential of the dielectric, which is usually air. Dimensions ranging from 0.2 to 0.5 wavelength are common for the "b" sides of a waveguide.

Q-9. At very high frequencies, what characteristics are displayed by ordinary insulators?

Q-10. What type of insulator works well at very high frequencies?

Q-11. The frequency range of a waveguide is determined by what dimension?

Q-12. What happens to the bus bar dimensions of the waveguide when the frequency is increased?

Q-13. When the frequency is decreased so that two quarter-wavelengths are longer than the "a" (wide) dimension of the waveguide, what will happen?

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1.2.4 Energy Propagation in Waveguides

Since energy is transferred through waveguides by electromagnetic fields, you need a basic understanding of field theory. Both magnetic (H FIELD) and electric field (E FIELD) are present in waveguides, and the interaction of these fields causes energy to travel through the waveguide. This action is best understood by first looking at the properties of the two individual fields.

E FIELD - An electric field exists when a difference of potential causes a stress in the dielectric between two points. The simplest electric field is one that forms between the plates of a capacitor when one plate is made positive compared to the other, as shown in figure 1-11A. The stress created in the dielectric is an electric field.

Electric fields are represented by arrows that point from the positive toward the negative potential. The number of arrows shows the relative strength of the field. In figure 1-11A, for example, evenly spaced arrows indicate the field is evenly distributed. For ease of explanation, the electric field is abbreviated E field, and the lines of stress are called E lines.

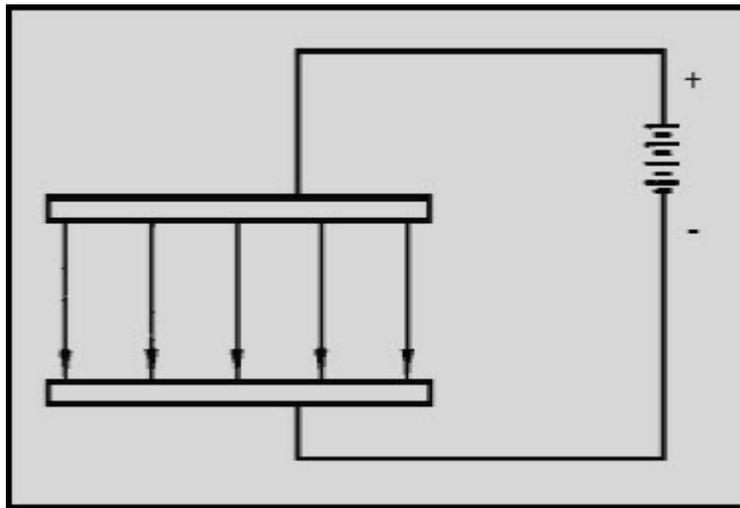


Figure 1-11A Simple electric fields. CAPACITOR

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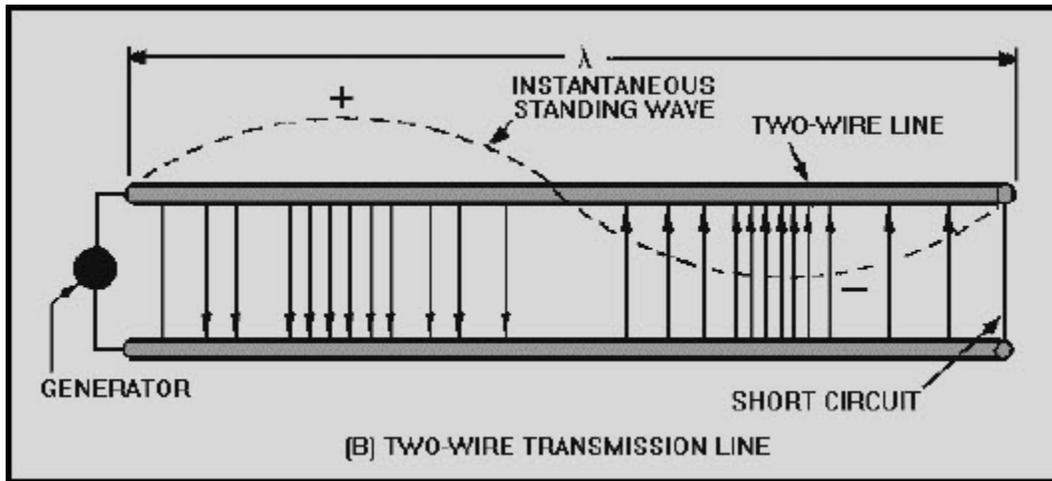


Figure 1-11B Simple electric fields. TWO-WIRE TRANSMISSION LINE

The two-wire transmission line, illustrated in figure 1-11B, has an instantaneous standing wave of voltage applied to it by the generator. The line is short-circuited at one-wavelength, at the positive and negative voltage peaks, but the arrows, representing each field, point in opposite directions. The voltage across the line varies sinusoidally. Therefore, the density of the E-lines varies sinusoidally.

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The development of the E field in a waveguide can be illustrated by a two-wire transmission line separated by several, double quarter-wave sections, called half-wave frames, as illustrated in figure 1-12. As shown, the voltage across the two-wire line varies in a sine-wave pattern and the density of the E field also varies in a sine-wave pattern. The half-wave frames located at high-voltage points (1) and (3) have a strong E field. The frames at the zero-voltage points (2) have no E fields present. Frame (4) has a weak E field and is located at a point between maximum and minimum voltage. This illustration is a buildup to the three-dimensional aspect of the full E field in a waveguide.

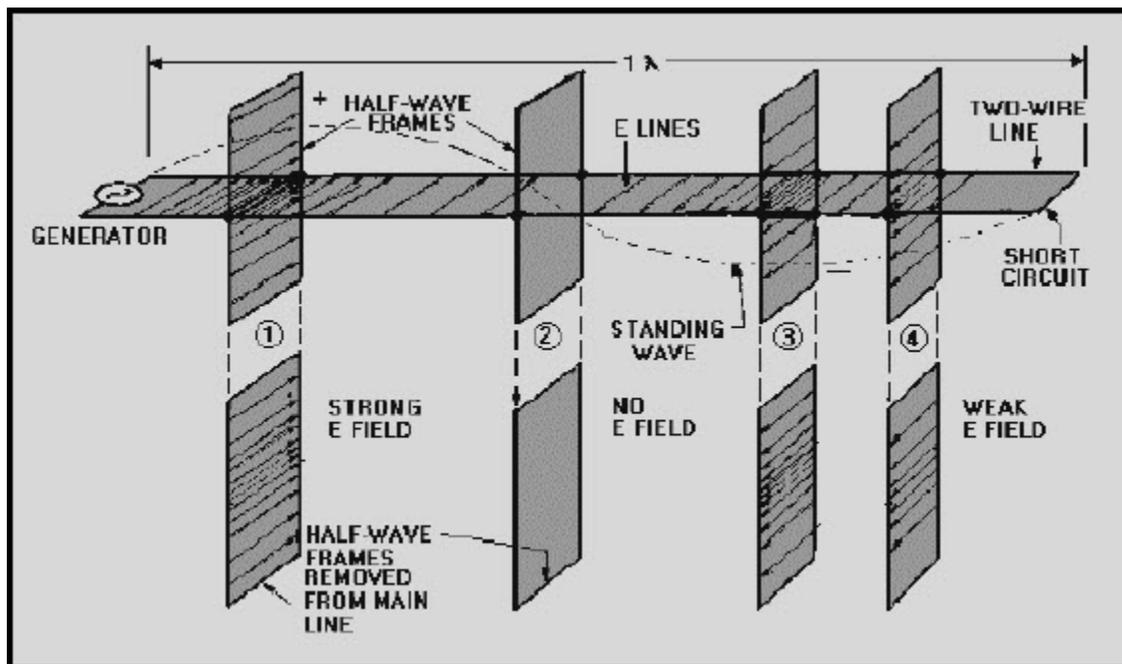


Figure 1-12 E fields on a two-wire line with half-wave frames

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Figure 1-13, view (A), shows the E-field pattern created by a voltage sine wave applied to a one-wavelength section of waveguide shorted at one end. The electric fields are represented by the arrows shown in views (B) and (C). In the top view of view (A), the tip of each arrow is represented by a dot and the tail of each arrow is represented by an X. The E field varies in density at the same sine-wave rate as the applied voltage. This illustration represents the instant that the applied voltage wave is at its peak. At other times, the voltage and the E field in the waveguide vary continuously from zero to the peak value. Voltage and E-field polarity reverse with every reversal of the input. Note that the end view shown in view (B) shows the E field is maximum at the center and minimum near the walls of the waveguide. View (C) shows the arrangement of electromagnetic fields within a three-dimensional waveguide.

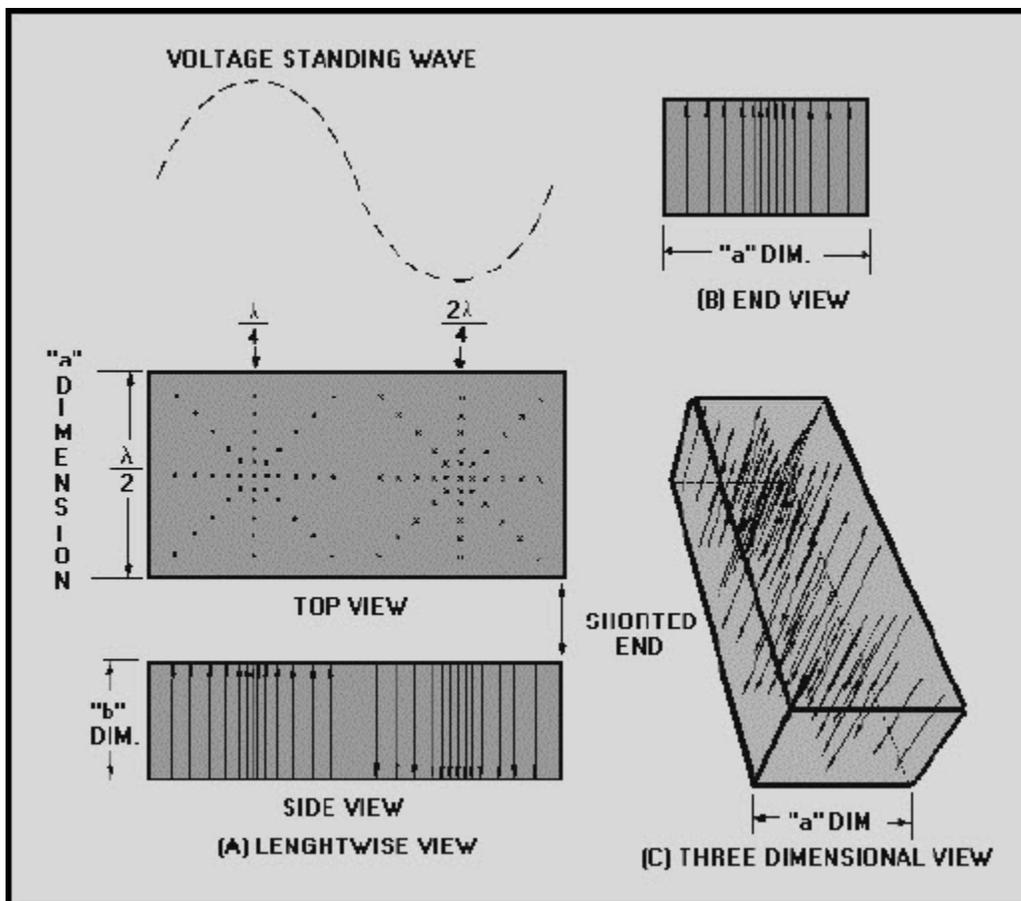


Figure 1-13 E field of a voltage standing wave across a 1-wavelength section of a waveguide

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H FIELD - The magnetic field in a waveguide is made up of magnetic lines of force that are caused by current flow through the conductive material of the waveguide. Magnetic lines of force, called H lines, are continuous closed loops, as shown in figure 1-14. All of the H lines associated with current are collectively called a magnetic field or H field. The strength of the H field, indicated by the number of H lines in a given area, varies directly with the amount of current.

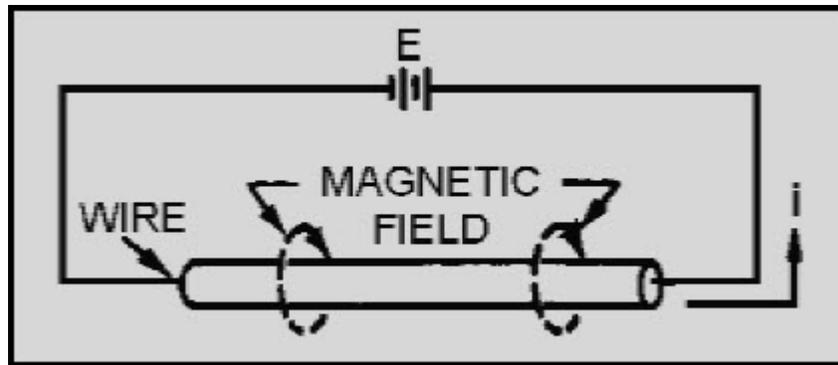


Figure 1-14 Magnetic field on a single wire

Although H lines encircle a single, straight wire, they behave differently when the wire is formed into a coil, as shown in figure 1-15. In a coil the individual H lines tend to form around each turn of wire. Since the H lines take opposite directions between adjacent turns, the field between the turns is cancelled. Inside and outside the coil, where the direction of each H field is the same, the fields join and form continuous H lines around the entire coil.

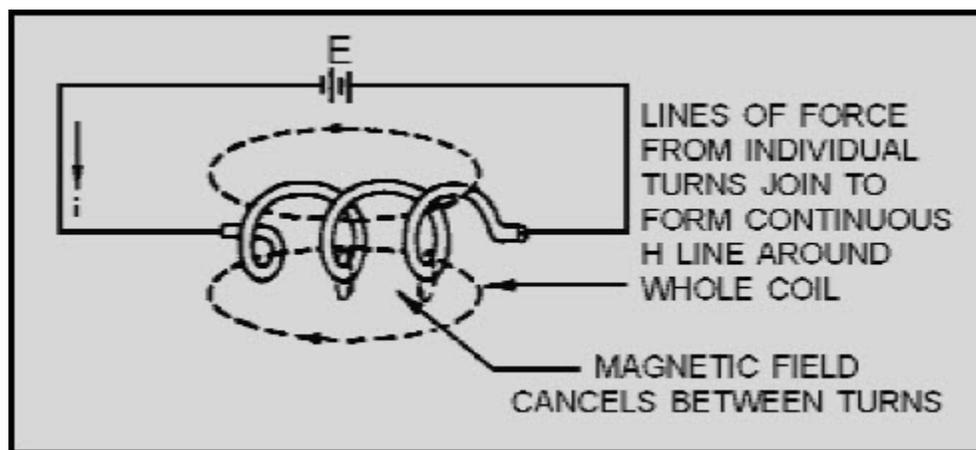


Figure 1-15 Magnetic field on a coil

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A similar action takes place in a waveguide. In figure 1-16A, a two-wire line with quarter-wave sections is shown. Currents flow in the main line and in the quarter-wave sections. The current direction produces the individual H lines around each conductor as shown. When a large number of sections exist, the fields cancel between the sections, but the directions are the same both inside and outside the waveguide. At half-wave intervals on the main line, current will flow in opposite directions. This produces H-line loops having opposite directions. In figure 1-16A, current at the left end is opposite to the current at the right end. The individual loops on the main line are opposite in direction. All around the framework they join so that the long loop shown in figure 1-16B is formed. Outside the waveguide the individual loops cannot join to form a continuous loop. Thus, no magnetic field exists outside a waveguide.

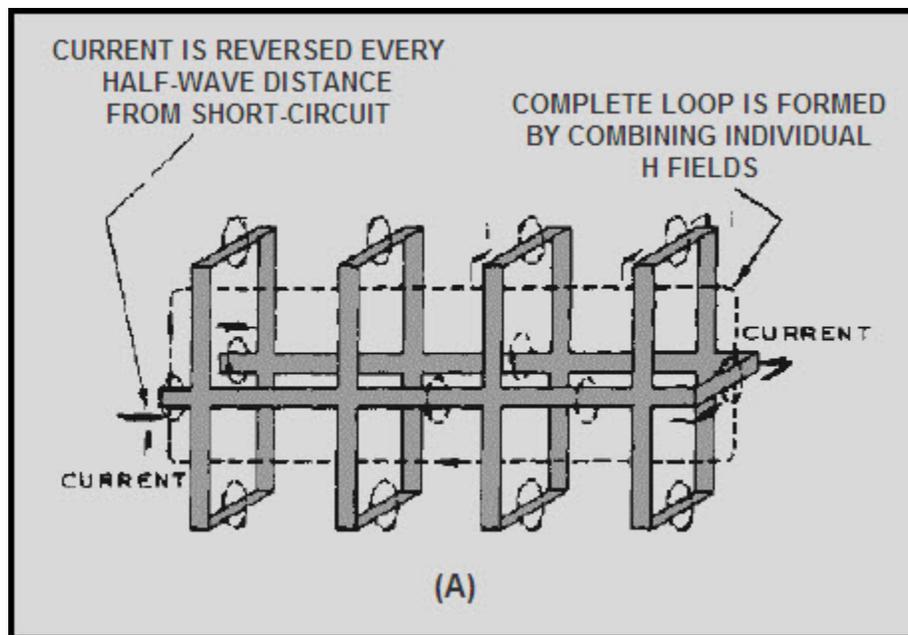


Figure 1-16A Magnetic fields on a two-wire line with half-wave frames

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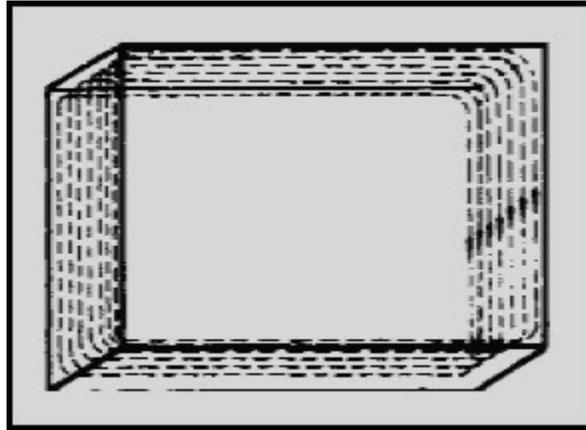


Figure 1-16B Magnetic fields on a two-wire line with half-wave frames

If the two-wire line and the half-wave frames are developed into a waveguide that is closed at both ends (as shown in figure 1-16B), the distribution of H lines will be as shown in figure 1-17. If the waveguide is extended to $1\frac{1}{2}\lambda$, these H lines form complete loops at half-wave intervals with each group reversed in direction. Again, no H lines can form outside the waveguide as long as it is completely enclosed.

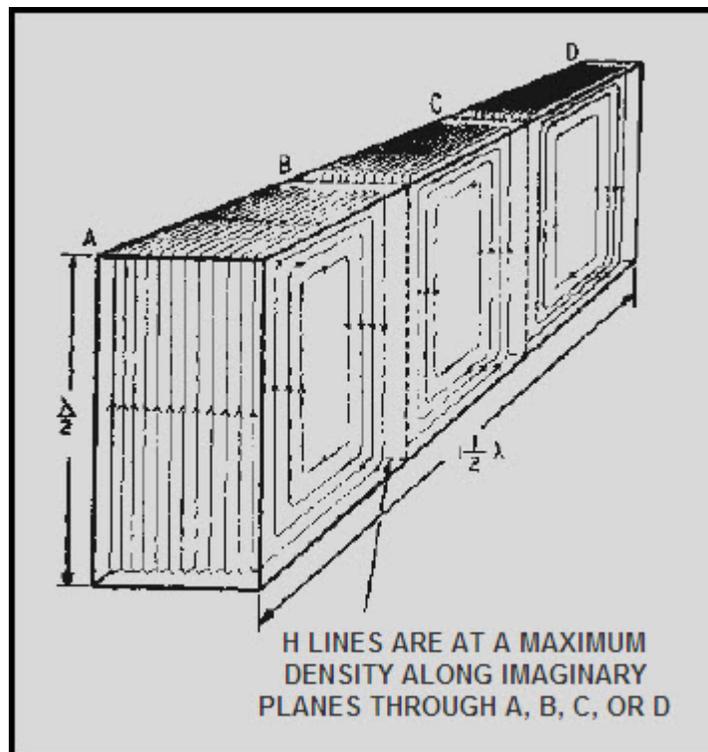


Figure 1-17 Magnetic field pattern in a waveguide

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Figure 1-18 shows a cross-sectional view of the magnetic field pattern illustrated in figure 1-17. Note in view (A) that the field is strongest at the edges of the waveguide where the current is highest. The minimum field strength occurs at the zero-current points. View (B) shows the field pattern as it appears $\lambda/4$ from the end view of the waveguide. As with the previously discussed E fields, the H fields shown in figures 1-17 and 1-18 represent a condition that exists at only one instant in time. During the peak of the next half cycle of the input current, all field directions are reversed and the field will continue to change with changes in the input.

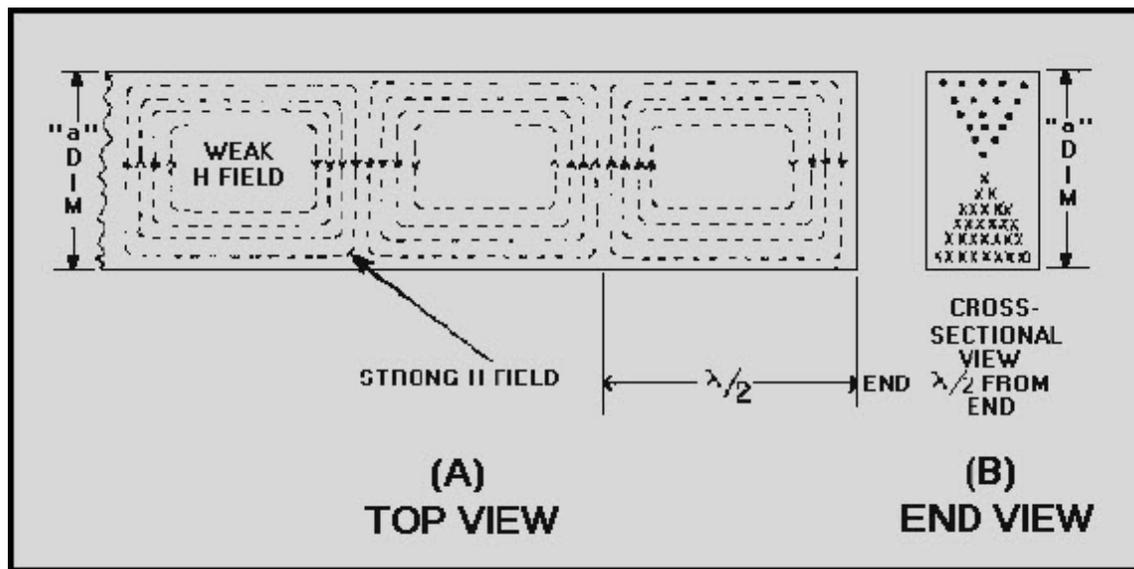


Figure 1-18 Magnetic field in a waveguide three half-wavelengths long

BOUNDARY CONDITIONS IN A WAVEGUIDE - The travel of energy down a waveguide is similar, but not identical, to the travel of electromagnetic waves in free space. The difference is that the energy in a waveguide is confined to the physical limits of the guide. Two conditions, known as BOUNDARY CONDITIONS, must be satisfied for energy to travel through a waveguide.

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The first boundary condition (illustrated in figure 1-19A) can be stated as follows:

For an electric field to exist at the surface of a conductor it must be perpendicular to the conductor.

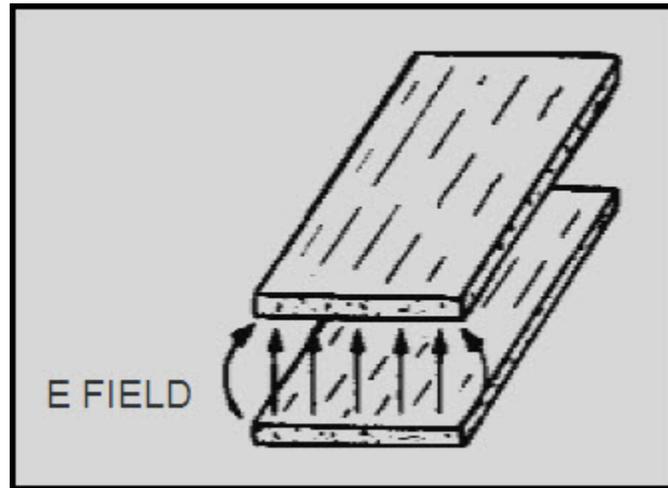


Figure 1-19A E field boundary condition. MEETS BOUNDARY CONDITIONS

The opposite of this boundary condition, shown in figure 1-19B, is also true. An electric field CANNOT exist parallel to a perfect conductor.

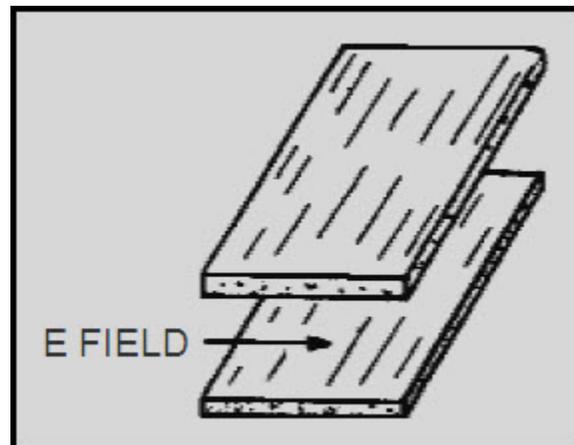


Figure 1-19B E field boundary condition. DOES NOT MEET BOUNDARY

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The second boundary condition, which is illustrated in figure 1-20, can be stated as follows:

For a varying magnetic field to exist, it must form closed loops in parallel with the conductors and be perpendicular to the electric field.

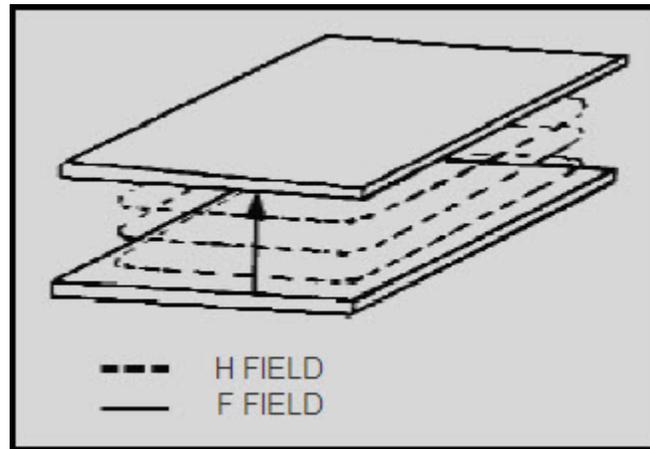


Figure 1-20 H field boundary condition

Since an E field causes a current flow that in turn produces an H field, both fields always exist at the same time in a waveguide. If a system satisfies one of these boundary conditions, it must also satisfy the other since neither field can exist alone. You should briefly review the principles of electromagnetic propagation in free space (NEETS, Module 10, Introduction to Wave Propagation, Transmission Lines, and Antennas). This review will help you understand how a waveguide satisfies the two boundary conditions necessary for energy propagation in a waveguide.

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WAVEFRONTS WITHIN A WAVEGUIDE..Electromagnetic energy transmitted into space consists of electric and magnetic fields that are at right angles (90 degrees) to each other and at right angles to the direction of propagation. A simple analogy to establish this relationship is by use of the right-hand rule for electromagnetic energy, based on the POYNTING VECTOR. It indicates that a screw (right-hand thread) with its axis perpendicular to the electric and magnetic fields will advance in the direction of propagation if the E field is rotated to the right (toward the H field). This rule is illustrated in figure 1-21.

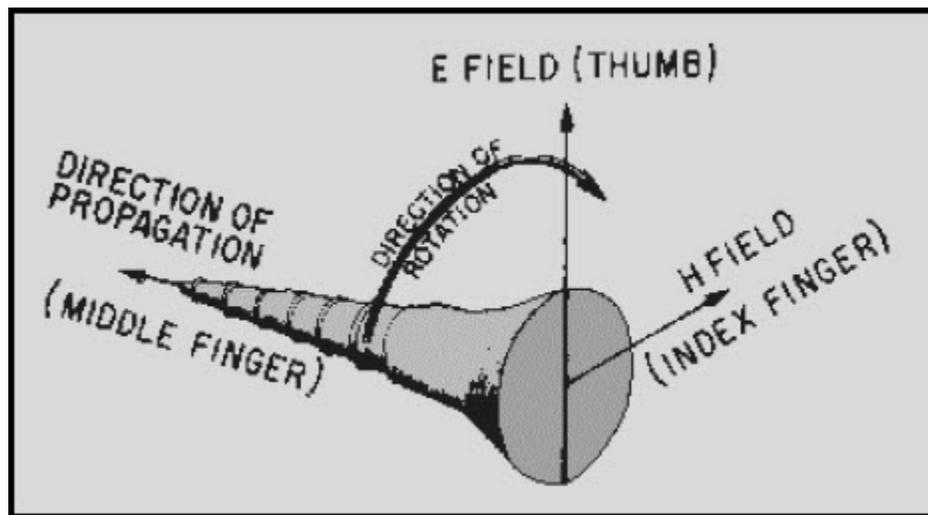


Figure 1-21 The Poynting vector

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The combined electric and magnetic fields form a wavefront that can be represented by alternate negative and positive peaks at half-wavelength intervals, as illustrated in figure 1-22. Angle θ is the direction of travel of the wave with respect to some reference axis.

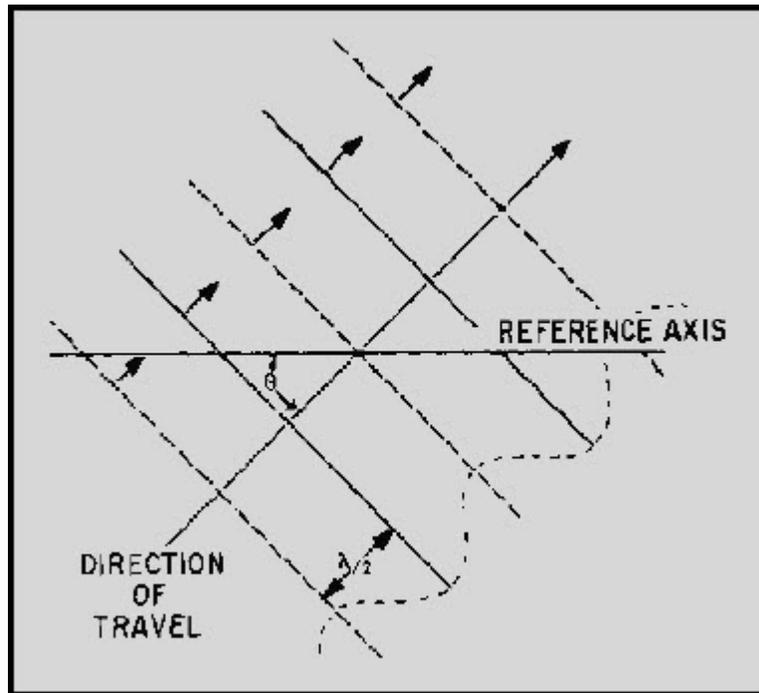


Figure 1-22 Wavefronts in space

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If a second wavefront, differing only in the direction of travel, is present at the same time, a resultant is formed. The resultant is illustrated in figure 1-23, and a close inspection reveals important characteristics of combined wavefronts. Both wavefronts add at all points on the reference axis and cancel at half-wavelength intervals from the reference axis. Therefore, alternate additions and cancellations of the two wavefronts occur at progressive half-wavelength increments from the reference axis. In figure 1-23, the lines labeled A, C, F, and H are addition points, and those labeled B, D, E, and G are cancellation points.

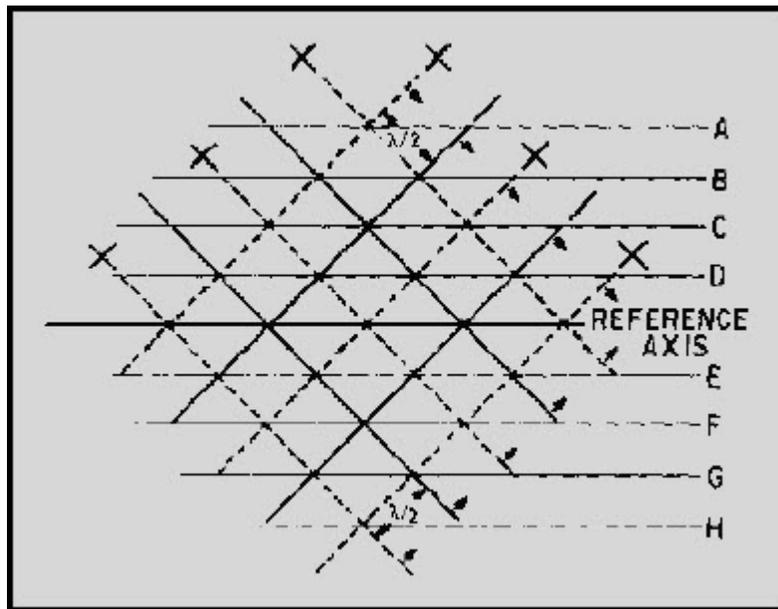


Figure 1-23 Combined wavefronts

If two conductive plates are placed along cancellation lines D and E or cancellation lines B and G, the first boundary condition for waveguides will be satisfied; that is, the E fields will be zero at the surface of the conductive plates. The second boundary condition is, therefore, automatically satisfied. Since these plates serve the same purpose as the "b" dimension walls of a waveguide, the "a" dimension walls can be added without affecting the magnetic or electric fields.

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When a quarter-wavelength probe is inserted into a waveguide and supplied with microwave energy, it will act as a quarter-wave vertical antenna. Positive and negative wavefronts will be radiated, as shown in figure 1-24. Any portion of the wavefront traveling in the direction of arrow C will rapidly decrease to zero because it does not fulfill either of the required boundary conditions. The parts of the wavefronts that travel in the directions of arrows A and B will reflect from the walls and form reverse phase wavefronts. These two wavefronts, and those that follow, are illustrated in figure 1-25. Notice that the wavefronts crisscross down the center of the waveguide and produce the same resultant field pattern that was shown in figure 1-23.

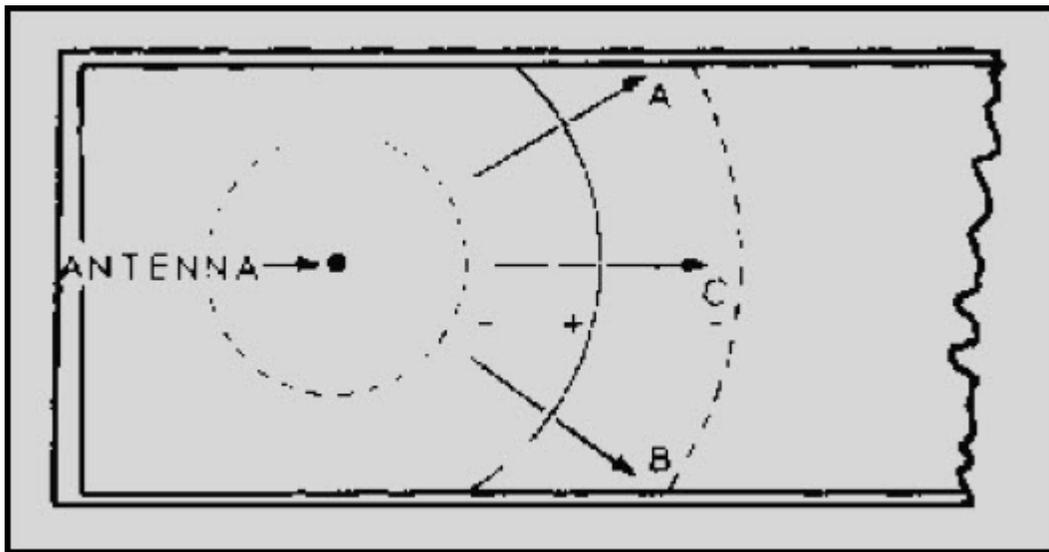


Figure 1-24 Radiation from probe placed in a waveguide

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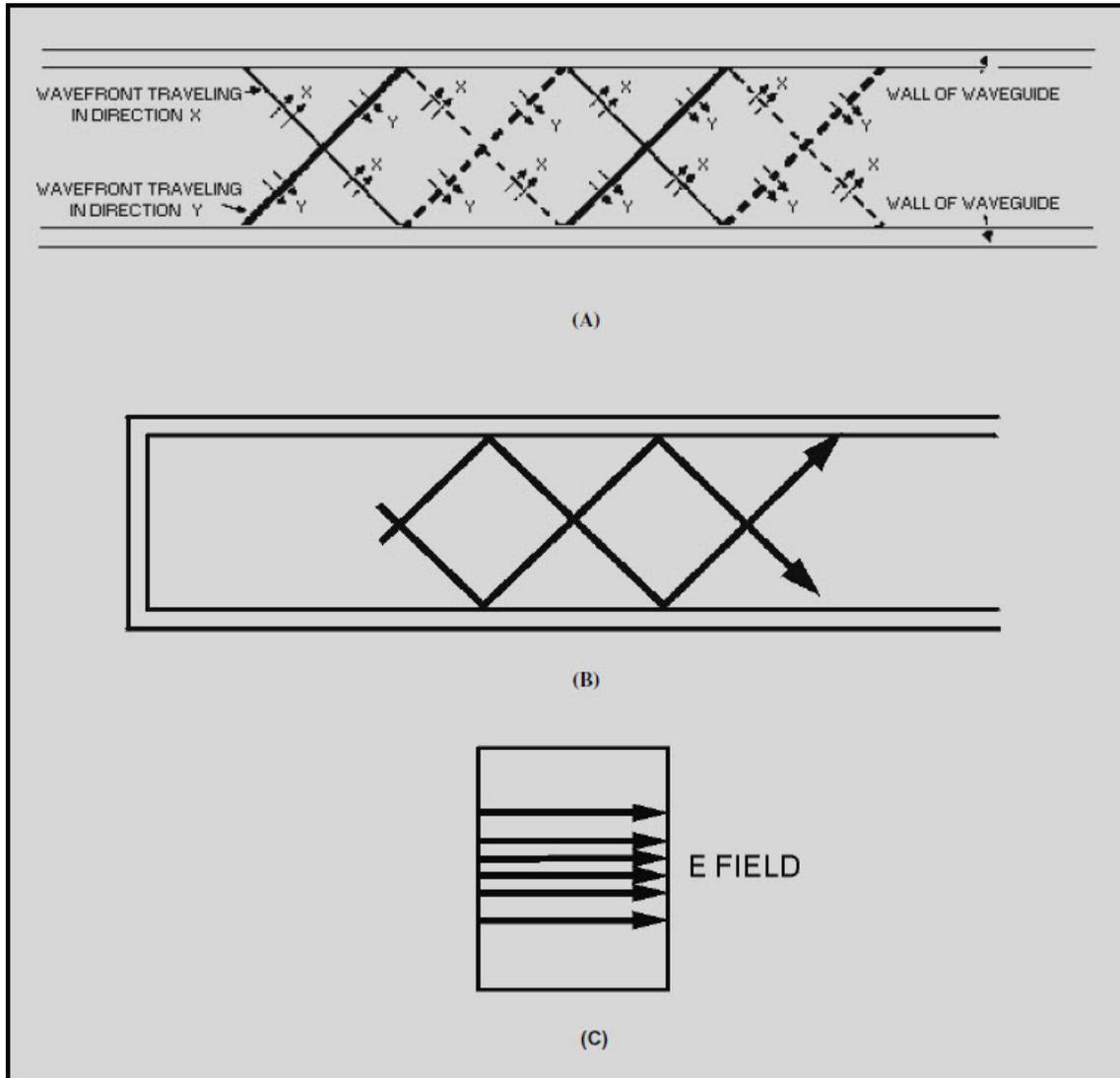


Figure 1-25 Wavefronts in a waveguide

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The reflection of a single wavefront off the "b" wall of a waveguide is shown in figure 1-26. The wavefront is shown in view (A) as small particles. In views (B) and (C) particle 1 strikes the wall and is bounced back from the wall without losing velocity. If the wall is perfectly flat, the angle at which it strikes the wall, known as the angle of incidence (θ), is the same as the angle of reflection (θ) and are measured perpendicular to the waveguide surface. An instant after particle 1 strikes the wall, particle 2 strikes the wall, as shown in view (C), and reflects in the same manner. Because all the particles are traveling at the same velocity, particles 1 and 2 do not change their relative position with respect to each other. Therefore, the reflected wave has the same shape as the original. The remaining particles as shown in views (D), (E) and (F) reflect in the same manner. This process results in a reflected wavefront identical in shape, but opposite in polarity, to the incident wave.

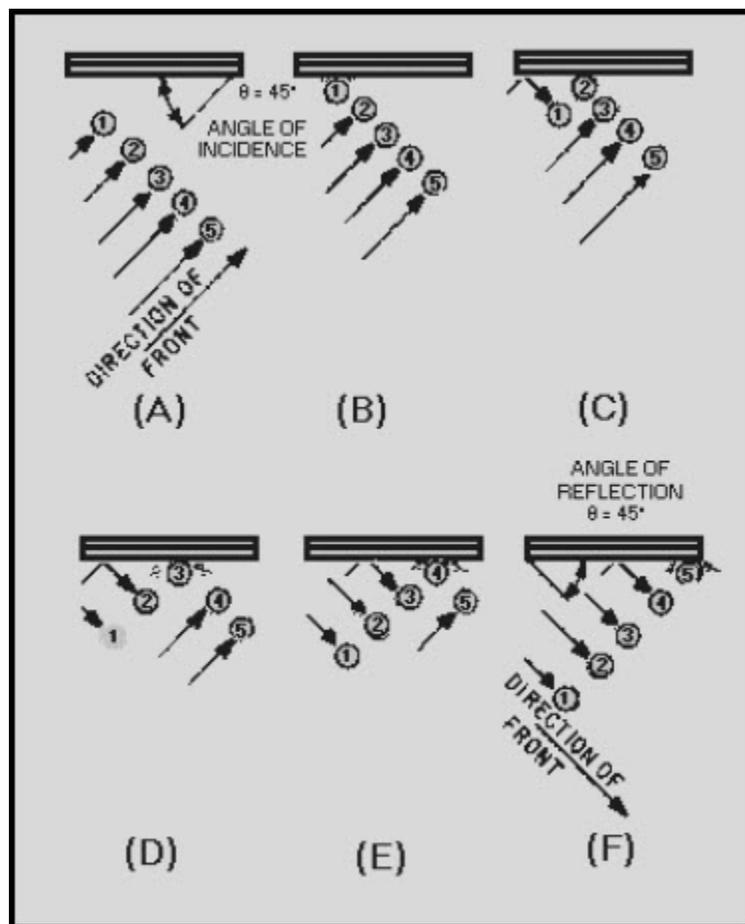
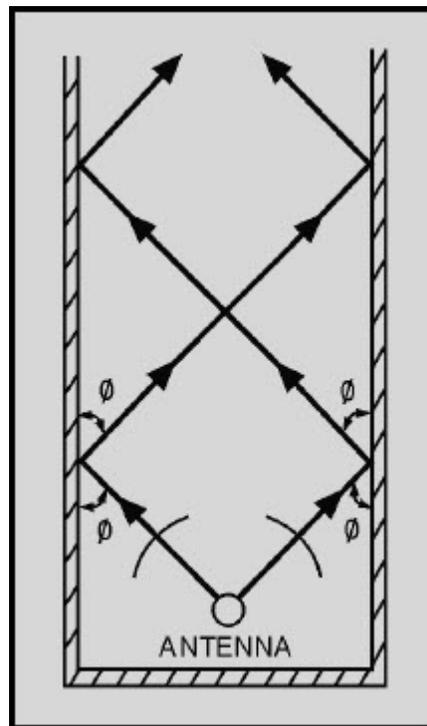


Figure 1-26 Reflection of a single wavefront

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Figures 1-27A and 1-27B, each illustrate the direction of propagation of two different electromagnetic wavefronts of different frequencies being radiated into a waveguide by a probe. Note that only the direction of propagation is indicated by the lines and arrowheads. The wavefronts are at right angles to the direction of propagation. The angle of incidence (θ) and the angle of reflection (ϕ) of the wavefronts vary in size with the frequency of the input energy, but the angles of reflection are equal to each other in a waveguide. The CUTOFF FREQUENCY in a waveguide is a frequency that would cause angles of incidence and reflection to be zero degrees. At any frequency below the cutoff frequency, the wavefronts will be reflected back and forth across the guide (setting up standing waves) and no energy will be conducted down the waveguide.



1-27 Different frequencies in a waveguide

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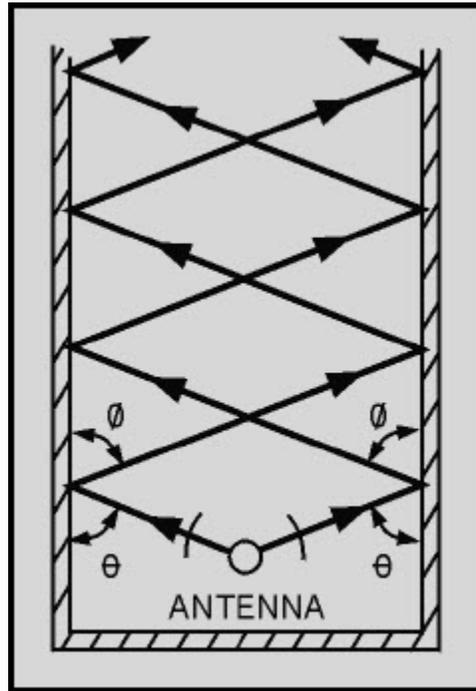


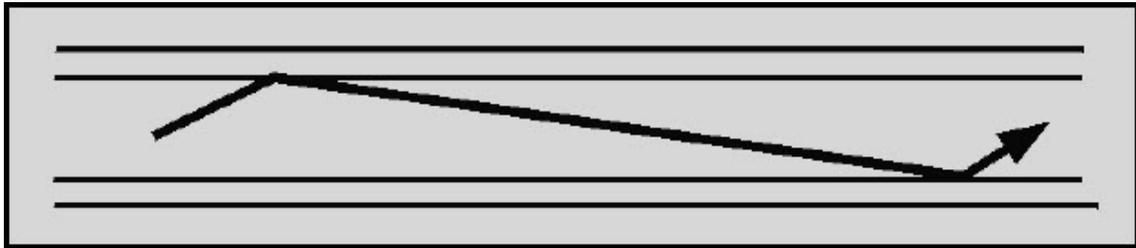
Figure 1-27B Different frequencies in a waveguide

The velocity of propagation of a wave along a waveguide is less than its velocity through free space (speed of light). This lower velocity is caused by the zigzag path taken by the wavefront. The forward-progress velocity of the wavefront in a waveguide is called **GROUP VELOCITY** and is somewhat slower than the speed of light.

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The group velocity of energy in a waveguide is determined by the reflection angle of the wavefronts off the "b" walls. The reflection angle is determined by the frequency of the input energy. This basic principle is illustrated in figures 1-28A, 1-28B, and 1-28C. As frequency is decreased, the reflection angle decreases causing the group velocity to decrease. The opposite is also true; increasing frequency increases the group velocity.



1-28A Reflection angle at various frequencies. **LOW FREQUENCY**

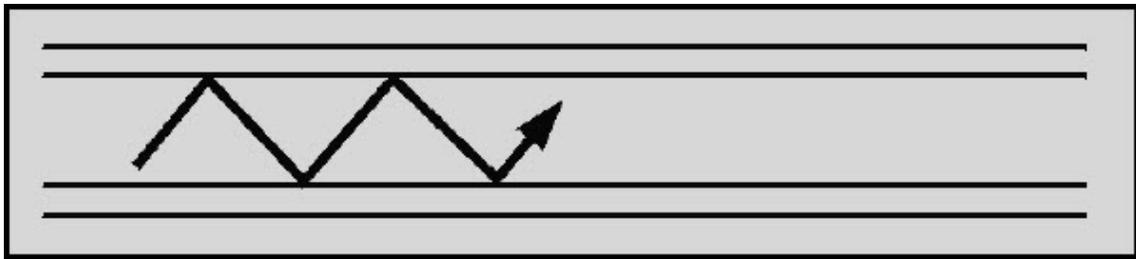


Figure 1-28B Reflection angle at various frequencies. **MEDIUM FREQUENCY**

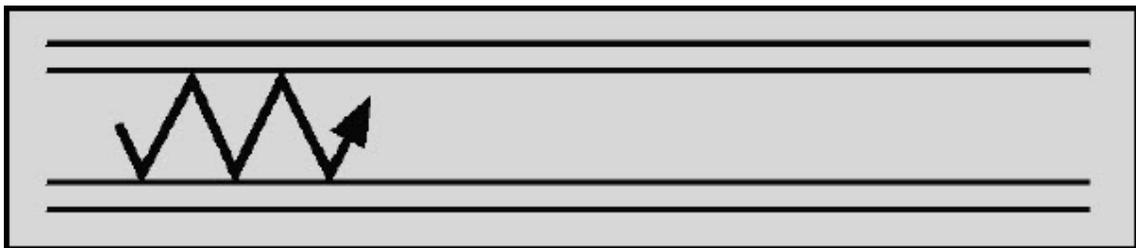


Figure 1-28C Reflection angle at various frequencies. **HIGH FREQUENCY**

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Q-14. What interaction causes energy to travel down a waveguide?

Q-15. What is indicated by the number of arrows (closeness of spacing) used to represent an electric field?

Q-16. What primary condition must magnetic lines of force meet in order to exist?

Q-17. What happens to the H lines between the conductors of a coil when the conductors are close together?

Q-18. For an electric field to exist at the surface of a conductor, the field must have what angular relationship to the conductor?

Q-19. When a wavefront is radiated into a waveguide, what happens to the portions of the wavefront that do not satisfy the boundary conditions?

Q-20. Assuming the wall of a waveguide is perfectly flat, what is the angular relationship between the angle of incidence and the angle of reflection?

Q-21. What is the frequency called that produces angles of incidence and reflection that are perpendicular to the waveguide walls?

Q-22. Compared to the velocity of propagation of waves in air, what is the velocity of propagation of waves in waveguides?

Q-23. What term is used to identify the forward progress velocity of wavefronts in a waveguide?

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1.2.5 Waveguide Modes of Operation

The waveguide analyzed in the previous paragraphs yields an electric field configuration known as the half-sine electric distribution. This configuration, called a MODE OF OPERATION, is shown in figure 1-29. Recall that the strength of the field is indicated by the spacing of the lines; that is, the closer the lines, the stronger the field. The regions of maximum voltage in this field move continuously down the waveguide in a sine-wave pattern. To meet boundary conditions, the field must always be zero at the "b" walls.

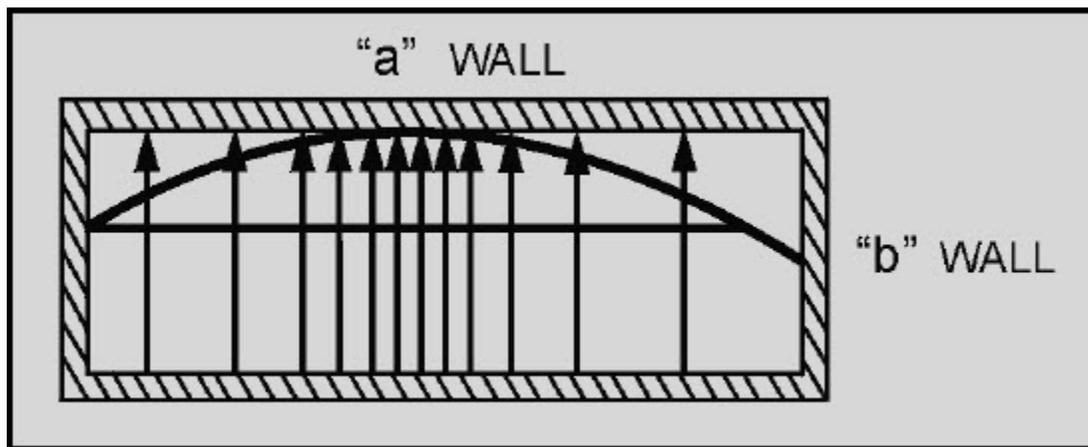


Figure 1-29 Half-sine E field distribution

The half-sine field is only one of many field configurations, or modes, that can exist in a rectangular waveguide. A full-sine field can also exist in a rectangular waveguide because, as shown in figure 1-30, the field is zero at the "b" walls.

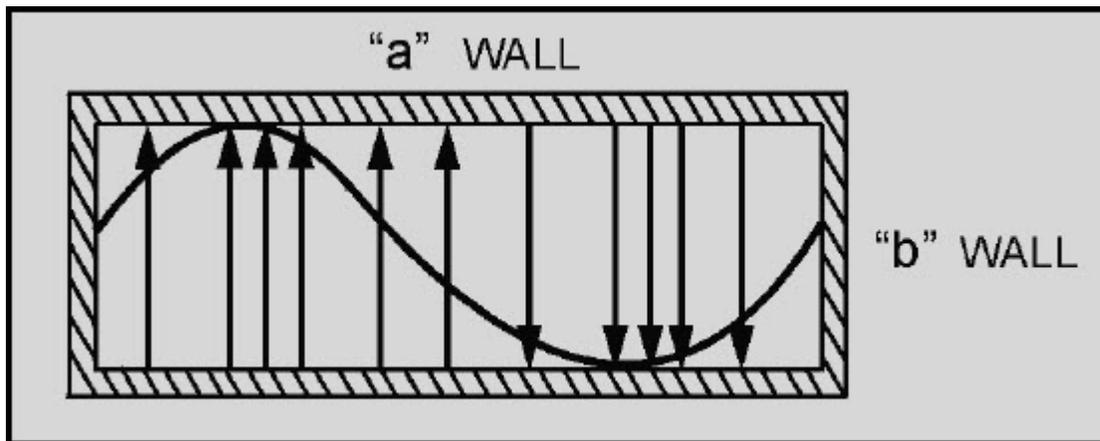


Figure 1-30 Full-sine E field distribution

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Similarly, a $1\frac{1}{2}$ sine-wave field can exist in a rectangular waveguide because this field also meets the boundary conditions. As shown in figure 1-31, the field is perpendicular to any conducting surface it touches and is zero along the "b" walls.

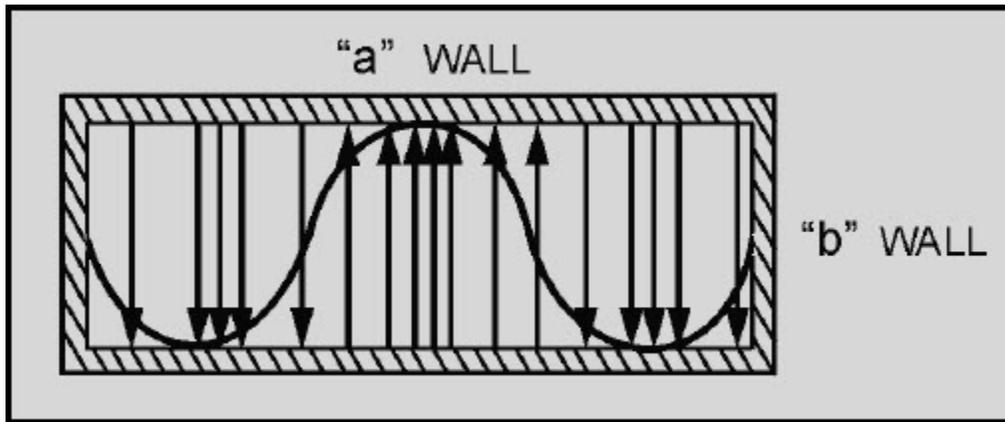


Figure 1-31 One and one-half sine E field distribution

The magnetic field in a rectangular waveguide is in the form of closed loops parallel to the surface of the conductors. The strength of the magnetic field is proportional to the electric field. Figure 1-32 illustrates the magnetic field pattern associated with a half-sine electric field distribution. The magnitude of the magnetic field varies in a sine-wave pattern down the center of the waveguide in "time phase" with the electric field. TIME PHASE means that the peak H lines and peak E lines occur at the same instant in time, although not necessarily at the same point along the length of the waveguide.

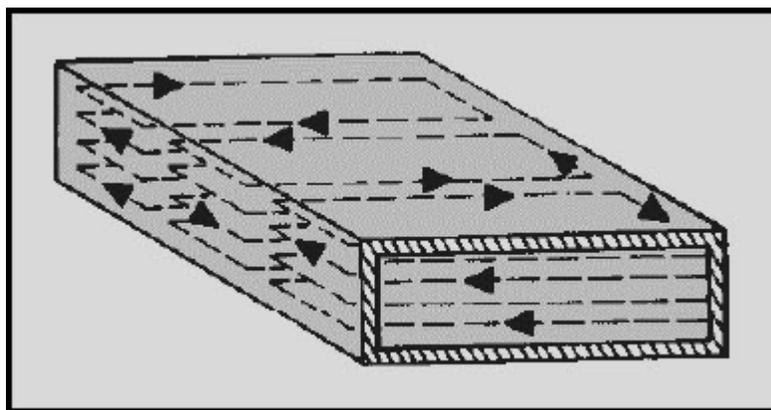


Figure 1-32 Magnetic field caused by a half-sine E field

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An electric field in a sine-wave pattern also exists down the center of a waveguide. In figure 1-33, view (A), consider the two wavefronts, C and D. Assume that they are positive at point 1 and negative at point 2. When the wavefronts cross at points 1 and 2, each field is at its maximum strength. At these points, the fields combine, further increasing their strength. This action is continuous because each wave is always followed by a replacement wave. Figure 1-33, view (B), illustrates the resultant sine configuration of the electric field at the center of the waveguide. This configuration is only one of the many field patterns that can exist in a waveguide. Each configuration forms a separate mode of operation. The easiest mode to produce is called the DOMINANT MODE. Other modes with different field configurations may occur accidentally or may be caused deliberately.

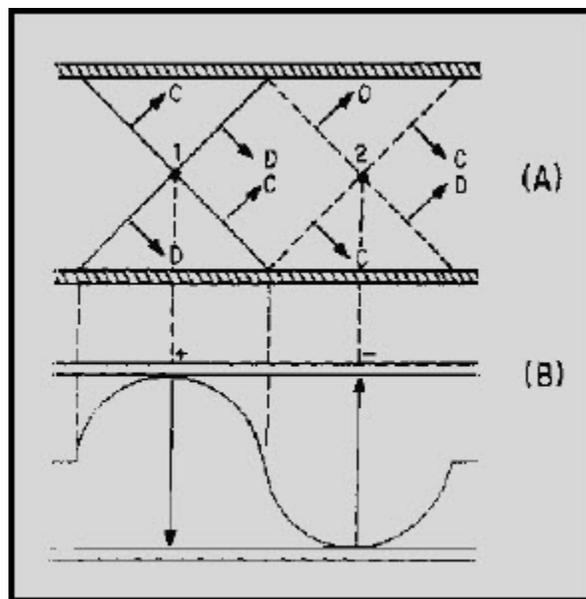


Figure 1-33 Crisscrossing wavefronts and the resultant E field

The dominant mode is the most efficient mode. Waveguides are normally designed so that only the dominant mode will be used. To operate in the dominant mode, a waveguide must have an "a" (wide) dimension of at least one half-wavelength of the frequency to be propagated. The "a" dimension of the waveguide must be kept near the minimum allowable value to ensure that only the dominant mode will exist. In practice, this dimension is usually 0.7 wavelength.

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Of the possible modes of operation available for a given waveguide, the dominant mode has the lowest cutoff frequency. The high-frequency limit of a rectangular waveguide is a frequency at which its "a" dimension becomes large enough to allow operation in a mode higher than that for which the waveguide has been designed.

Waveguides may be designed to operate in a mode other than the dominant mode. An example of a full-sine configuration mode is shown in figures 1-34A and 1-34B. The "a" dimension of the waveguide in this figure is one wavelength long. You may assume that the two-wire line is $1/4\lambda$ from one of the "b" walls, as shown in figure 1-34A. The remaining distance to the other "b" wall is $3/4\lambda$. The three-quarter wavelength section has the same high impedance as the quarter-wave section; therefore, the two-wire line is properly insulated. The field configuration shows a complete sine-wave pattern across the "a" dimension, as illustrated in figure 1-34B.

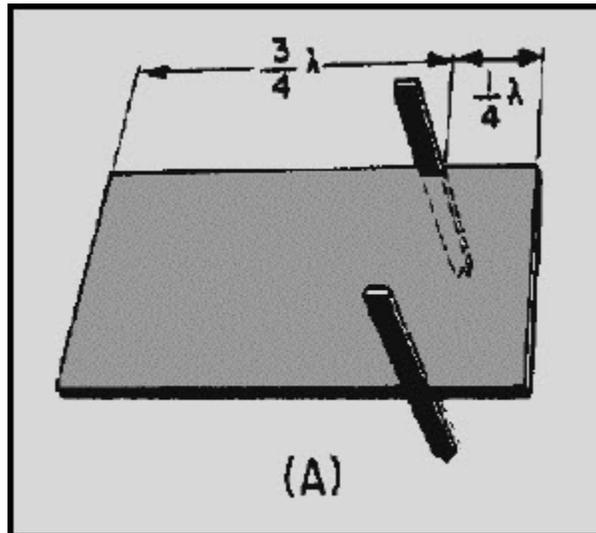


Figure 1-34A Waveguide operation in other than dominant mode

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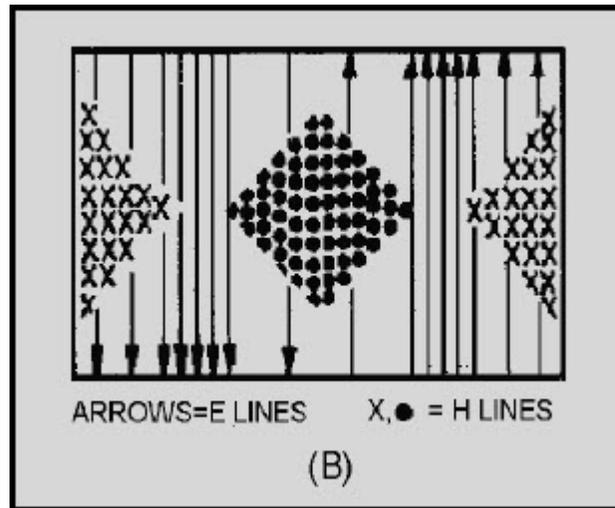


Figure 1-34B Waveguide operation in other than dominant mode

Circular waveguides are used in specific areas of radar and communications systems, such as rotating joints used at the mechanical point where the antennas rotate. Figure 1-35 illustrates the dominant mode of a circular waveguide. The cutoff wavelength of a circular guide is 1.71 times the diameter of the waveguide. Since the "a" dimension of a rectangular waveguide is approximately one half-wavelength at the cutoff frequency, the diameter of an equivalent circular waveguide must be $2 \div 1.71$, or approximately 1.17 times the "a" dimension of a rectangular waveguide.

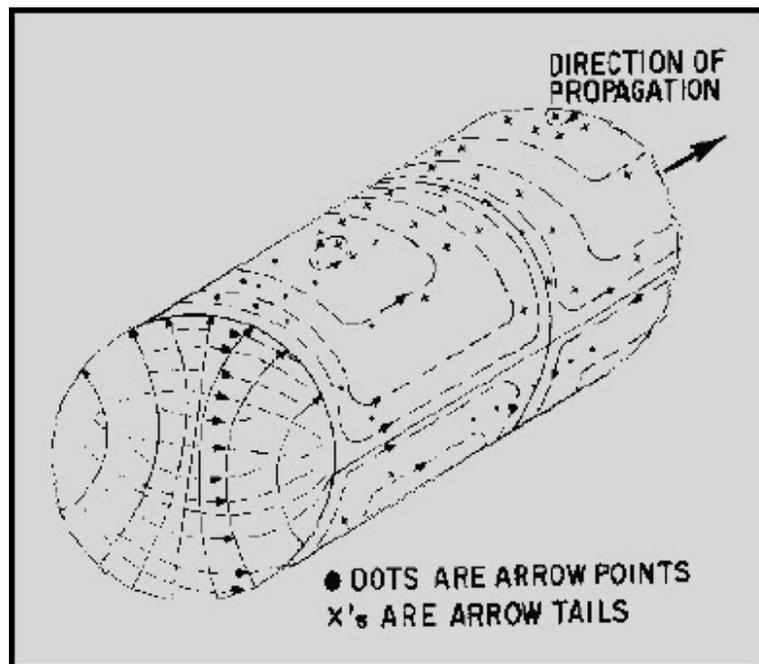


Figure 1-35 Dominant mode in a circular waveguide

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MODE NUMBERING SYSTEMS - So far, only the most basic types of E and H field arrangements have been shown. More complicated arrangements are often necessary to make possible coupling, isolation, or other types of operation. The field arrangements of the various modes of operation are divided into two categories: TRANSVERSE ELECTRIC (TE) and TRANSVERSE MAGNETIC (TM).

In the transverse electric (TE) mode, the entire electric field is in the transverse plane, which is perpendicular to the length of the waveguide (direction of energy travel). Part of the magnetic field is parallel to the length axis.

In the transverse magnetic (TM) mode, the entire magnetic field is in the transverse plane and has no portion parallel to the length axis.

Since there are several TE and TM modes, subscripts are used to complete the description of the field pattern. In rectangular waveguides, the first subscript indicates the number of half-wave patterns in the "a" dimension, and the second subscript indicates the number of half-wave patterns in the "b" dimension.

The dominant mode for rectangular waveguides is shown in figure 1-36. It is designated as the TE mode because the E fields are perpendicular to the "a" walls. The first subscript is 1 since there is only one half-wave pattern across the "a" dimension. There are no E-field patterns across the "b" dimension, so the second subscript is 0. The complete mode description of the dominant mode in rectangular waveguides is $TE_{1,0}$. Subsequent descriptions of waveguide operation in this text will assume the dominant ($TE_{1,0}$) mode unless otherwise noted.

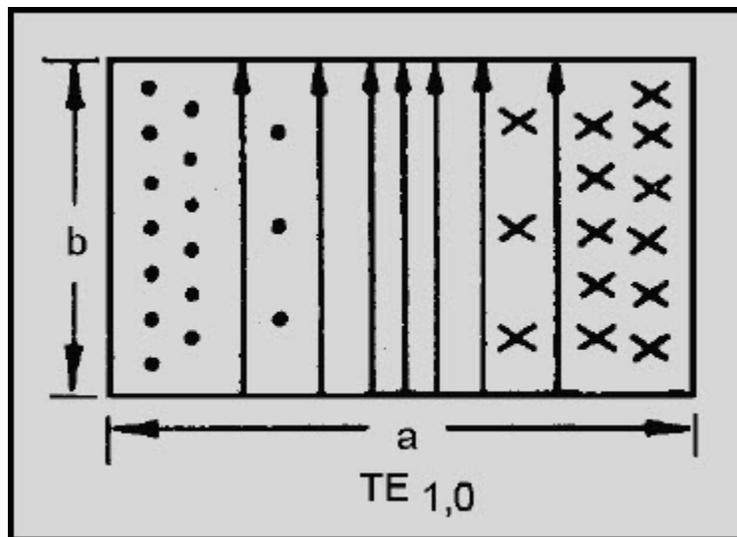


Figure 1-36 Dominant mode in a rectangular waveguide

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A similar system is used to identify the modes of circular waveguides. The general classification of TE and TM is true for both circular and rectangular waveguides. In circular waveguides the subscripts have a different meaning. The first subscript indicates the number of full-wave patterns around the circumference of the waveguide. The second subscript indicates the number of half-wave patterns across the diameter.

In the circular waveguide in figure 1-37, the E field is perpendicular to the length of the waveguide with no E lines parallel to the direction of propagation. Thus, it must be classified as operating in the TE mode. If you follow the E line pattern in a counterclockwise direction starting at the top, the E lines go from zero, through maximum positive (tail of arrows), back to zero, through maximum negative (head of arrows), and then back to zero again. This is one full wave, so the first subscript is 1. Along the diameter, the E lines go from zero through maximum and back to zero, making a half-wave variation.

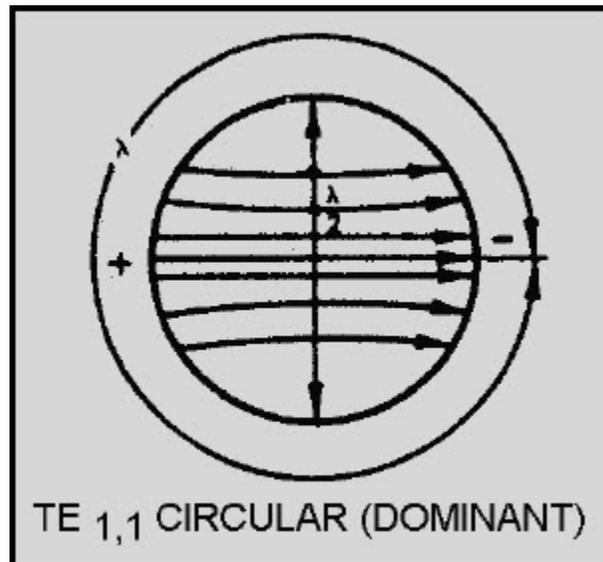


Figure 1-37 Counting wavelengths in a circular waveguide

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The second subscript, therefore, is also 1. $TE_{1,1}$ is the complete mode description of the dominant mode in circular waveguides. Several modes are possible in both circular and rectangular waveguides. Figure 1-38 illustrates several different modes that can be used to verify the mode numbering system.

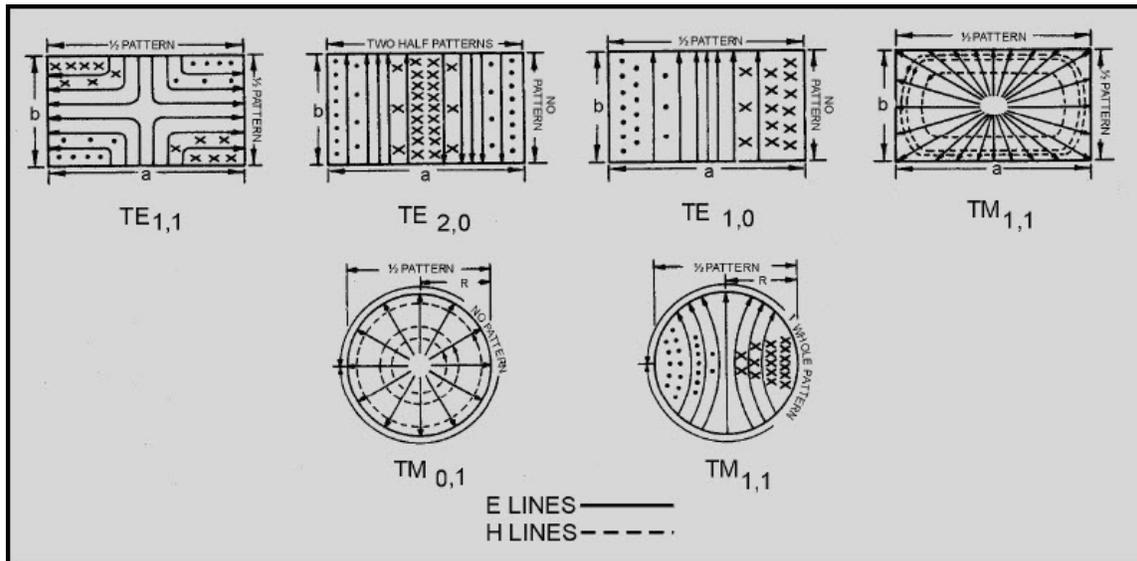


Figure 1-38 Various modes of operation for rectangular and circular waveguides

1.2.6 Waveguide Input/Output Methods

A waveguide, as explained earlier in this chapter, operates differently from an ordinary transmission line. Therefore, special devices must be used to put energy into a waveguide at one end and remove it from the other end.

The three devices used to inject or remove energy from waveguides are PROBES, LOOPS, and SLOTS. Slots may also be called APERTURES or WINDOWS.

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As previously discussed, when a small probe is inserted into a waveguide and supplied with microwave energy, it acts as a quarter-wave antenna. Current flows in the probe and sets up an E field such as the one shown in figure 1-39A. The E lines detach themselves from the probe. When the probe is located at the point of highest efficiency, the E lines set up an E field of considerable intensity.

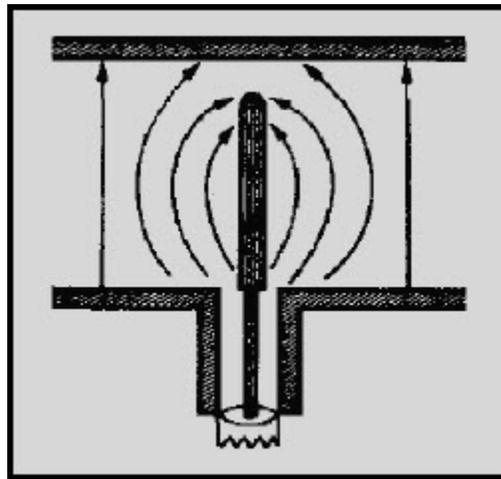


Figure 1-39A Probe coupling in a rectangular waveguide

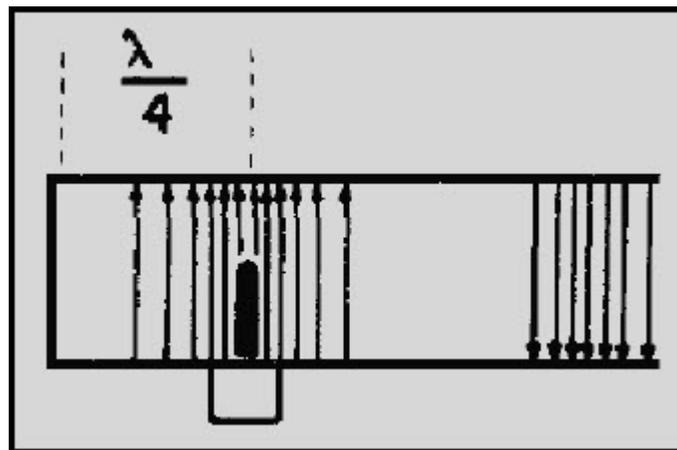


Figure 1-39B Probe coupling in a rectangular waveguide

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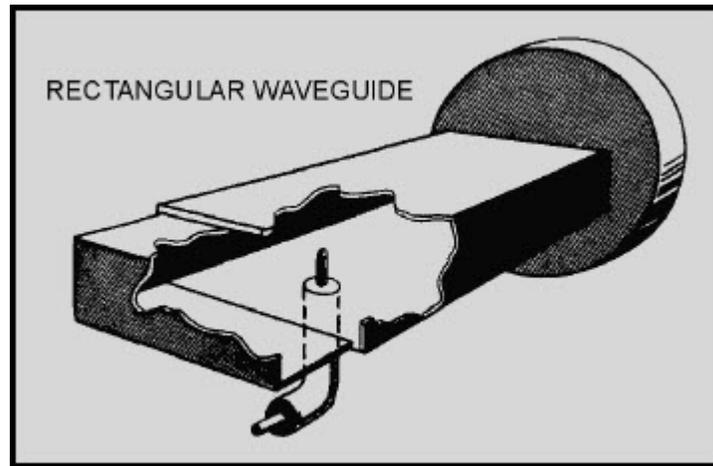


Figure 1-39C Probe coupling in a rectangular waveguide

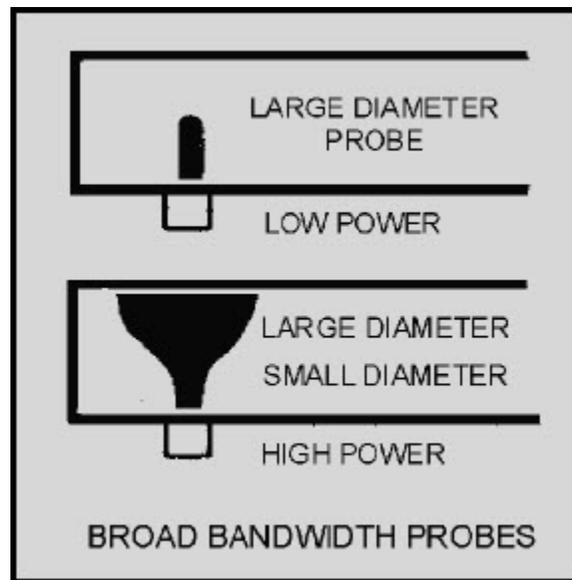


Figure 1-39D Probe coupling in a rectangular waveguide

The most efficient place to locate the probe is in the center of the "a" wall, parallel to the "b" wall, and one quarter-wavelength from the shorted end of the waveguide, as shown in figure 1-39B, and figure 1-39C. This is the point at which the E field is maximum in the dominant mode. Therefore, energy transfer (coupling) is maximum at this point. Note that the quarter-wavelength spacing is at the frequency required to propagate the dominant mode.

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In many applications a lesser degree of energy transfer, called loose coupling, is desirable. The amount of energy transfer can be reduced by decreasing the length of the probe, by moving it out of the center of the E field, or by shielding it. Where the degree of coupling must be varied frequently, the probe is made retractable so the length can be easily changed.

The size and shape of the probe determines its frequency, bandwidth, and power-handling capability. As the diameter of a probe increases, the bandwidth increases. A probe similar in shape to a door knob is capable of handling much higher power and a larger bandwidth than a conventional probe. The greater power-handling capability is directly related to the increased surface area. Two examples of broad-bandwidth probes are illustrated in figure 1-39D. Removal of energy from a waveguide is simply a reversal of the injection process using the same type of probe.

Another way of injecting energy into a waveguide is by setting up an H field in the waveguide. This can be accomplished by inserting a small loop which carries a high current into the waveguide, as shown in figure 1-40A. A magnetic field builds up around the loop and expands to fit the waveguide, as shown in figure 1-40B. If the frequency of the current in the loop is within the bandwidth of the waveguide, energy will be transferred to the waveguide.

For the most efficient coupling to the waveguide, the loop is inserted at one of several points where the magnetic field will be of greatest strength. Four of those points are shown in figure 1-40C.

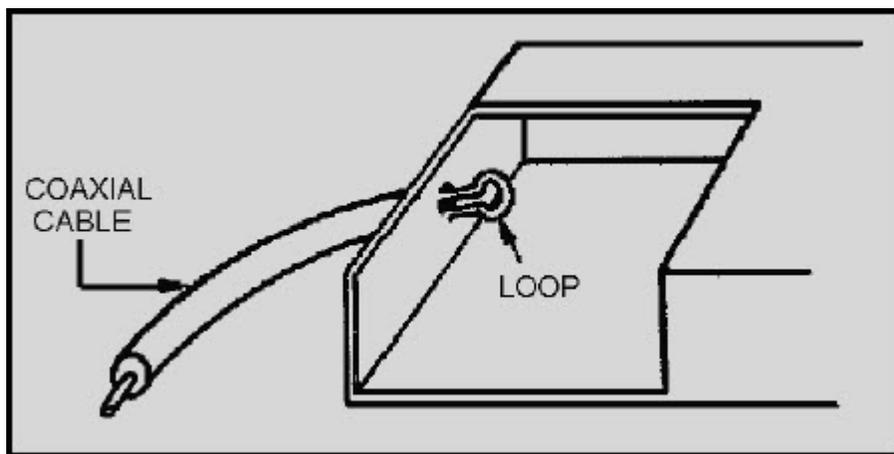


Figure 1-40A Loop coupling in a rectangular waveguide

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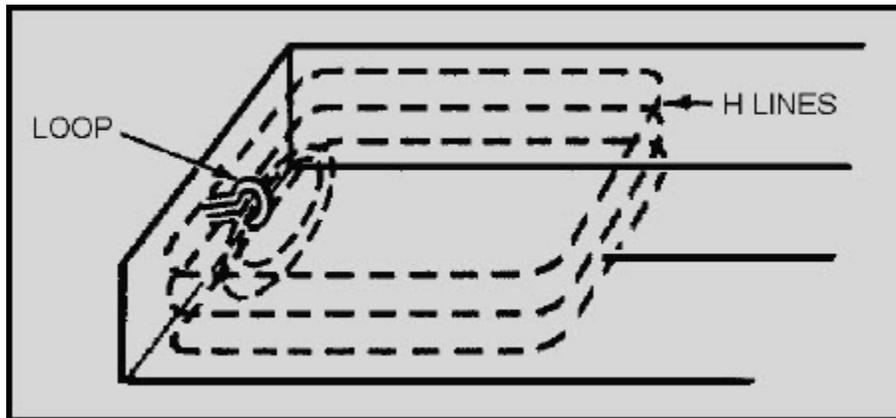


Figure 1-40B Loop coupling in a rectangular waveguide

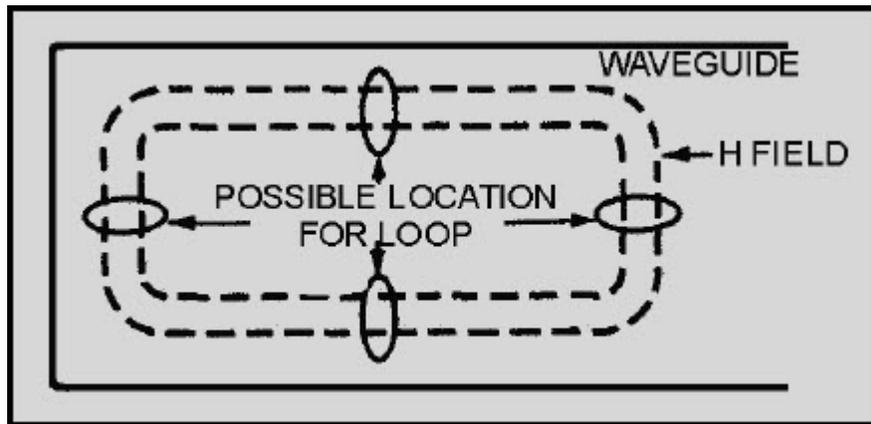


Figure 1-40C Loop coupling in a rectangular waveguide

When less efficient coupling is desired, you can rotate or move the loop until it encircles a smaller number of H lines. When the diameter of the loop is increased, its power-handling capability also increases. The bandwidth can be increased by increasing the size of the wire used to make the loop.

When a loop is introduced into a waveguide in which an H field is present, a current is induced in the loop. When this condition exists, energy is removed from the waveguide.

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Slots or apertures are sometimes used when very loose (inefficient) coupling is desired, as shown in figure 1-41. In this method energy enters through a small slot in the waveguide and the E field expands into the waveguide. The E lines expand first across the slot and then across the interior of the waveguide.

Minimum reflections occur when energy is injected or removed if the size of the slot is properly proportioned to the frequency of the energy.

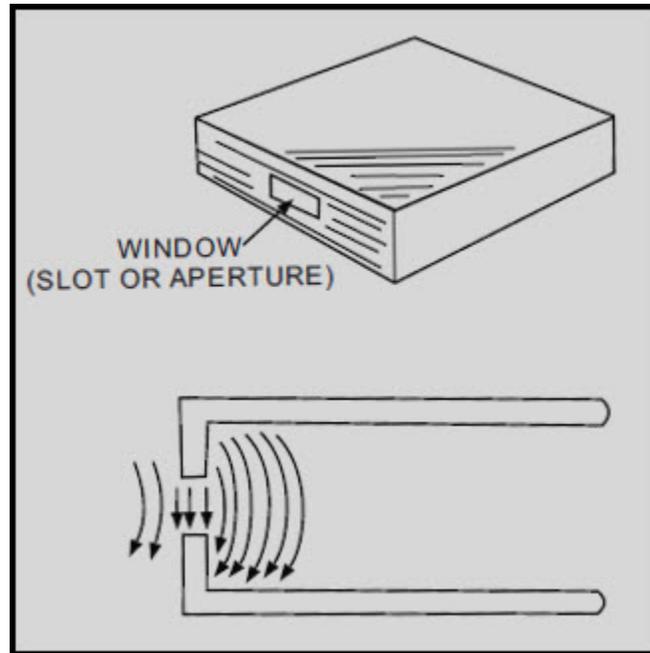


Figure 1-41 Slot coupling in a waveguide

After learning how energy is coupled into and out of a waveguide with slots, you might think that leaving the end open is the most simple way of injecting or removing energy in a waveguide. This is not the case, however, because when energy leaves a waveguide, fields form around the end of the waveguide. These fields cause an impedance mismatch which, in turn, causes the development of standing waves and a drastic loss in efficiency. Various methods of impedance matching and terminating waveguides will be covered in the next section.

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Q-24. What term is used to identify each of the many field configurations that can exist in waveguides?

Q-25. What field configuration is easiest to produce in a given waveguide?

Q-26. How is the cutoff wavelength of a circular waveguide figured?

Q-27. The field arrangements in waveguides are divided into what two categories to describe the various modes of operation?

Q-28. The electric field is perpendicular to the "a" dimension of a waveguide in what mode?

Q-29. The number of half-wave patterns in the "b" dimension of rectangular waveguides is indicated by which of the two descriptive subscripts?

Q-30. Which subscript, in circular waveguide classification, indicates the number of full-wave patterns around the circumference?

Q-31. What determines the frequency, bandwidth, and power-handling capability of a waveguide probe?

Q-32. Loose or inefficient coupling of energy into or out of a waveguide can be accomplished by the use of what method?

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1.2.7 Waveguide Impedance Matching

Waveguide transmission systems are not always perfectly impedance matched to their load devices. The standing waves that result from a mismatch cause a power loss, a reduction in power-handling capability, and an increase in frequency sensitivity. Impedance-changing devices are therefore placed in the waveguide to match the waveguide to the load. These devices are placed near the source of the standing waves.

Figure 1-42 illustrates three devices, called irises that are used to introduce inductance or capacitance into a waveguide. An iris is nothing more than a metal plate that contains an opening through which the waves may pass. The iris is located in the transverse plane.

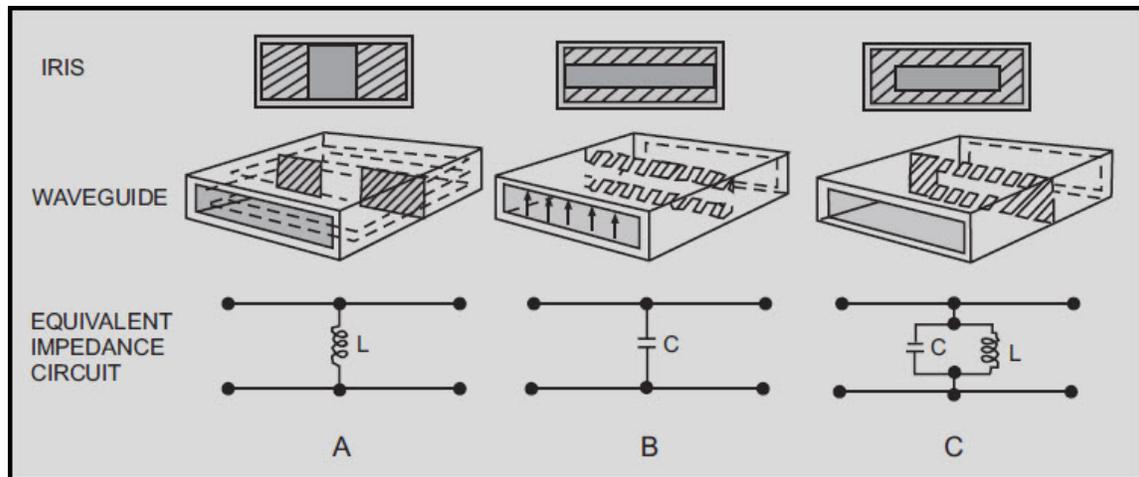


Figure 1-42 Waveguide irises

An inductive iris and its equivalent circuit are illustrated in figure 1-42, view (A). The iris places a shunt inductive reactance across the waveguide that is directly proportional to the size of the opening. Notice that the edges of the inductive iris are perpendicular to the magnetic plane. The shunt capacitive reactance, illustrated in view (B), basically acts the same way. Again, the reactance is directly proportional to the size of the opening, but the edges of the iris are perpendicular to the electric plane. The iris, illustrated in view (C), has portions across both the magnetic and electric planes and forms an equivalent parallel-LC circuit across the waveguide. At the resonant frequency, the iris acts as a high shunt resistance. Above or below resonance, the iris acts as a capacitive or inductive reactance.

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POSTS and SCREWS made from conductive material can be used for impedance-changing devices in waveguides. Figure 1-43A and 1-43B, illustrate two basic methods of using posts and screws. A post or screw which only partially penetrates into the waveguide acts as a shunt capacitive reactance. When the post or screw extends completely through the waveguide, making contact with the top and bottom walls, it acts as an inductive reactance. Note that when screws are used the amount of reactance can be varied.

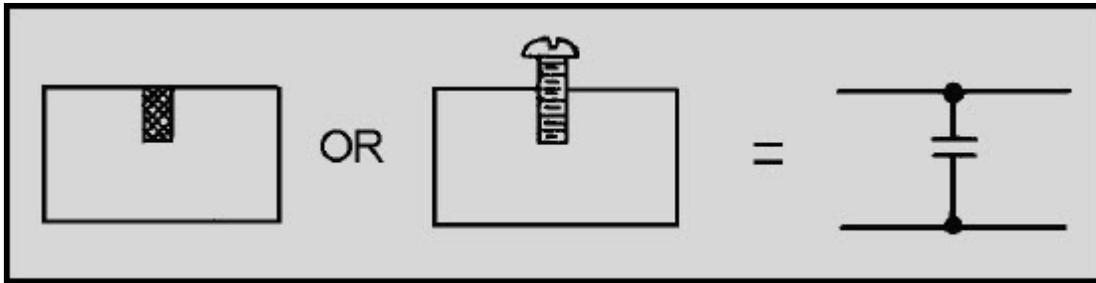


Figure 1-43A Conducting posts and screws. PENETRATING

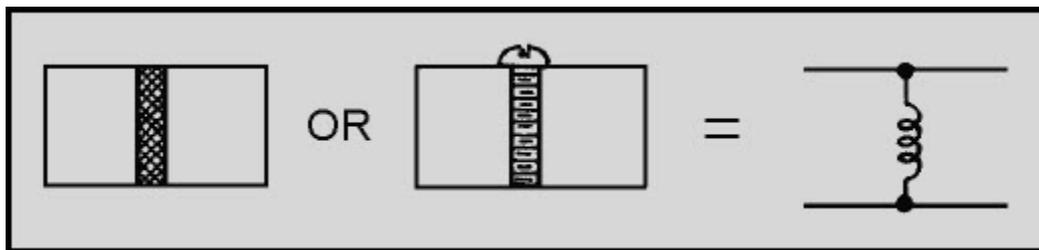


Figure 1-43B Conducting posts and screws. EXTENDING THROUGH

Q-33. What is the result of an impedance mismatch in a waveguide?

Q-34. What is used to construct irises?

Q-35. An iris placed along the "b" dimension wall produces what kind of reactance?

Q-36. How will an iris that has portions along both the "a" and "b" dimension walls act at the resonant frequency?

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1.2.8 Waveguide Terminations

Electromagnetic energy is often passed through a waveguide to transfer the energy from a source into space. As previously mentioned, the impedance of a waveguide does not match the impedance of space, and without proper impedance matching, standing waves cause a large decrease in the efficiency of the waveguide.

Any abrupt change in impedance causes standing waves, but when the change in impedance at the end of a waveguide is gradual, almost no standing waves are formed. Gradual changes in impedance can be obtained by terminating the waveguide with a funnel-shaped HORN, such as the three types illustrated in figures 1-44A, 1-44B, and 1-44C. The type of horn used depends upon the frequency and the desired radiation pattern.



Figure 1-44A Waveguide horns. E PLANE SECTORAL HORN



Figure 1-44B Waveguide horns. H PLANE SECTORAL HORN

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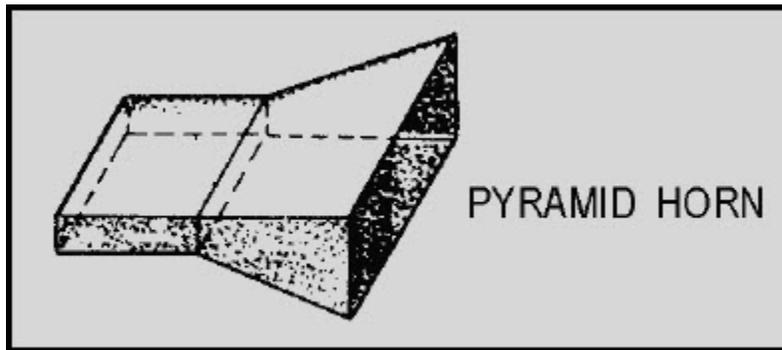


Figure 1-44C Waveguide horns. PYRAMID HORN

As you may have noticed, horns are really simple antennas. They have several advantages over other impedance-matching devices, such as their large bandwidth and simple construction. The use of horns as antennas will be discussed further in chapter 3.

A waveguide may also be terminated in a resistive load that is matched to the characteristic impedance of the waveguide. The resistive load is most often called a DUMMY LOAD, because its only purpose is to absorb all the energy in a waveguide without causing standing waves.

There is no place on a waveguide to connect a fixed termination resistor; therefore, several special arrangements are used to terminate waveguides. One method is to fill the end of the waveguide with a graphite and sand mixture, as illustrated in figure 1-45A. When the fields enter the mixture, they induce a current flow in the mixture which dissipates the energy as heat. Another method figure 1-45B is to use a high-resistance rod placed at the center of the E field. The E field causes current to flow in the rod, and the high resistance of the rod dissipates the energy as a power loss, again in the form of heat.

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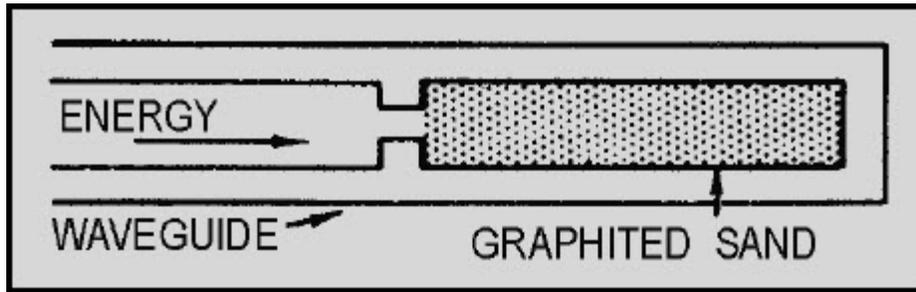


Figure 1-45A Terminating waveguides

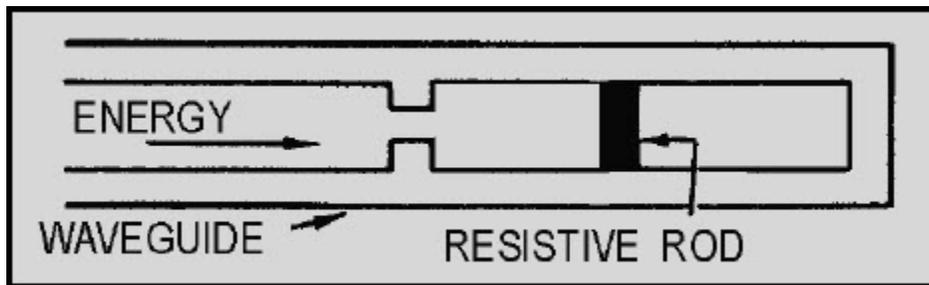


Figure 1-45B Terminating waveguides

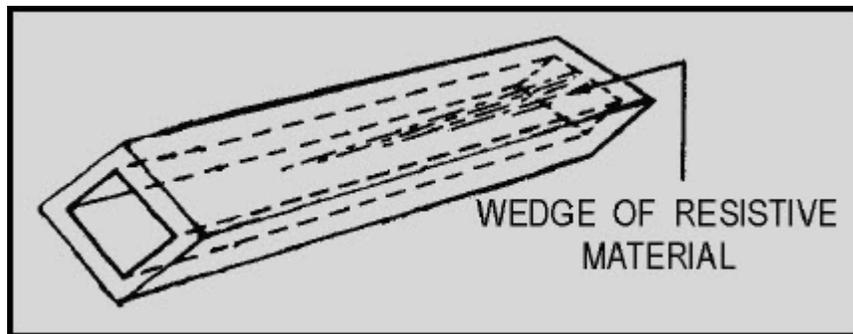


Figure 1-45C Terminating waveguides

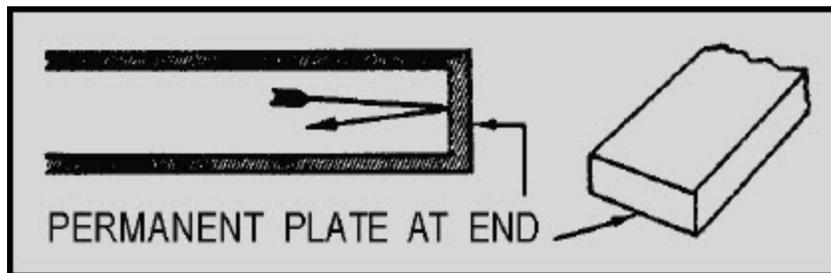


Figure 1-45D Terminating waveguides

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Still another method for terminating a waveguide is the use of a wedge of highly resistive material, as shown in of figure 1-45C. The plane of the wedge is placed perpendicular to the magnetic lines of force. When the H lines cut through the wedge, current flows in the wedge and causes a power loss. As with the other methods, this loss is in the form of heat. Since very little energy reaches the end of the waveguide, reflections are minimum.

All of the terminations discussed so far are designed to radiate or absorb the energy without reflections. In many instances, however, all of the energy must be reflected from the end of the waveguide. The best way to accomplish this is to permanently weld a metal plate at the end of the waveguide, as shown in figure 1-45D.

Q-37. What device is used to produce a gradual change in impedance at the end of a waveguide?

Q-38. When a waveguide is terminated in a resistive load, the load must be matched to what property of the waveguide?

Q-39. What is the primary purpose of a dummy load?

Q-40. The energy dissipated by a resistive load is most often in what form?

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1.2.9 Waveguide Plumbing

Since waveguides are really only hollow metal pipes, the installation and the physical handling of waveguides have many similarities to ordinary plumbing. In light of this fact, the bending, twisting, joining, and installation of waveguides is commonly called waveguide plumbing. Naturally, waveguides are different in design from pipes that are designed to carry liquids or other substances. The design of a waveguide is determined by the frequency and power level of the electromagnetic energy it will carry. The following paragraphs explain the physical factors involved in the design of waveguides.

WAVEGUIDE BENDS - The size, shape, and dielectric material of a waveguide must be constant throughout its length for energy to move from one end to the other without reflections. Any abrupt change in its size or shape can cause reflections and a loss in overall efficiency. When such a change is necessary, the bends, twists, and joints of the waveguides must meet certain conditions to prevent reflections.

Waveguides may be bent in several ways that do not cause reflections. One way is the gradual bend shown in figure 1-46. This gradual bend is known as an E bend because it distorts the E fields. The E bend must have a radius greater than two wavelengths to prevent reflections.

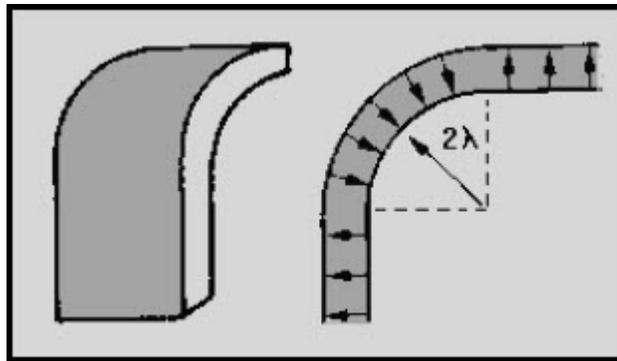


Figure 1-46 Gradual E bend

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Another common bend is the gradual H bend (figure 1-47). It is called an H bend because the H fields are distorted when a waveguide is bent in this manner. Again, the radius of the bend must be greater than two wavelengths to prevent reflections. Neither the E bend in the "a" dimension nor the H bend in the "b" dimension changes the normal mode of operation.

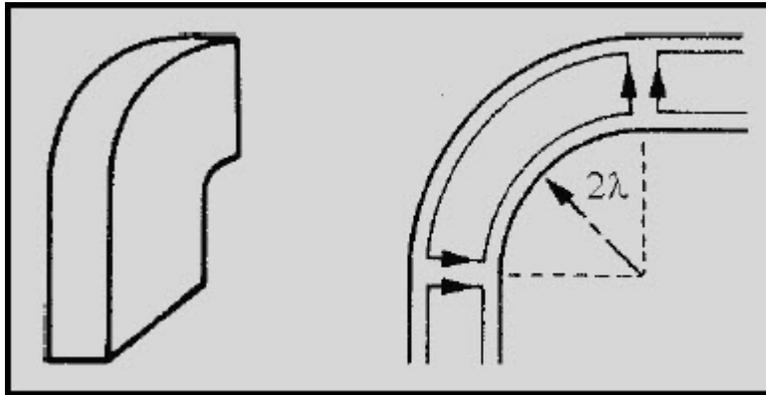


Figure 1-47 Gradual H bend

A sharp bend in either dimension may be used if it meets certain requirements. Notice the two 45-degree bends in figure 1-48; the bends are $1/4\lambda$ apart. The reflections that occur at the 45-degree bends cancel each other, leaving the fields as though no reflections have occurred.

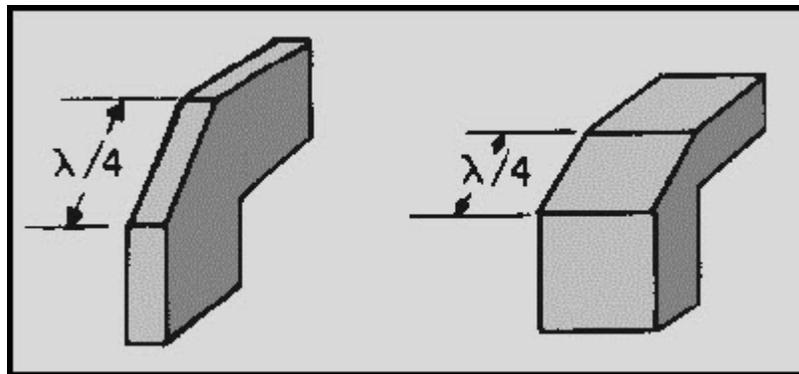


Figure 1-48 Sharp bends

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Sometimes the electromagnetic fields must be rotated so that they are in the proper phase to match the phase of the load. This may be accomplished by twisting the waveguide as shown in figure 1-49. The twist must be gradual and greater than 2λ .

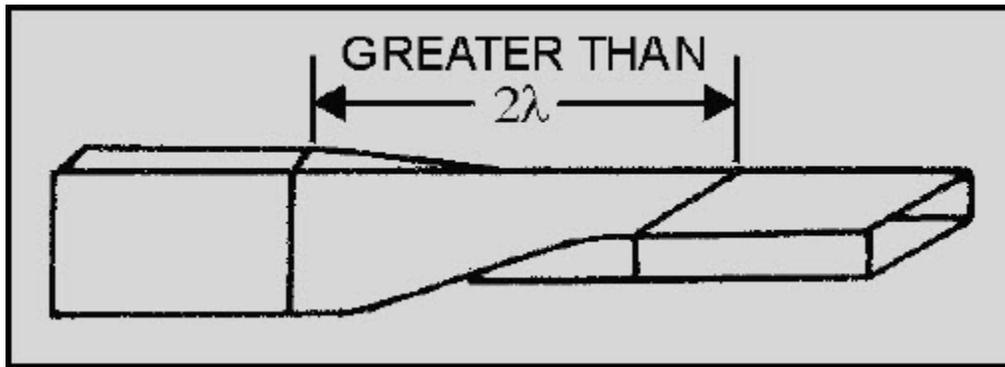


Figure 1-49 Waveguide twist

The flexible waveguide (figure 1-50) allows special bends which some equipment applications might require. It consists of a specially wound ribbon of conductive material, most commonly brass, with the inner surface plated with chromium. Power losses are greater in the flexible waveguide because the inner surfaces are not perfectly smooth. Therefore, it is only used in short sections where no other reasonable solution is available.

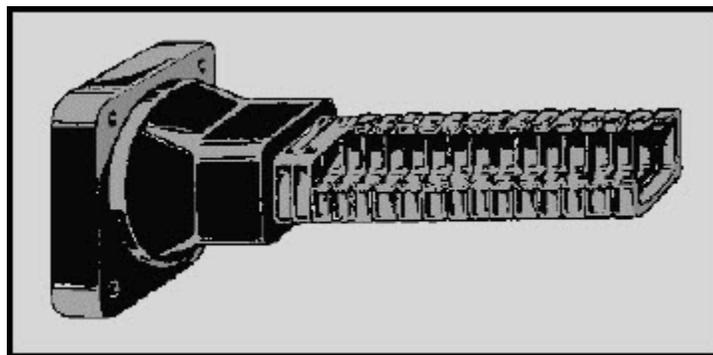


Figure 1-50 Flexible waveguide

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WAVEGUIDE JOINTS - Since an entire waveguide system cannot possibly be molded into one piece, the waveguide must be constructed in sections and the sections connected with joints. The three basic types of waveguide joints are the PERMANENT, the SEMIPERMANENT, and the ROTATING JOINTS. Since the permanent joint is a factory-welded joint that requires no maintenance, only the semi-permanent and rotating joints will be discussed.

Sections of waveguide must be taken apart for maintenance and repair. A semipermanent joint, called a CHOKE JOINT, is most commonly used for this purpose. The choke joint provides good electromagnetic continuity between sections of waveguide with very little power loss.

A cross-sectional view of a choke joint is shown in figures 1-51A and 1-51B. The pressure gasket shown between the two metal surfaces forms an airtight seal. Notice in figure 1-51B that the slot is exactly $1/4\lambda$ from the "a" wall of the waveguide. The slot is also $1/4\lambda$ deep, as shown in figure 1-51A, and because it is shorted at point (1), a high impedance results at point (2). Point (3) is $1/4\lambda$ from point (2). The high impedance at point (2) results in a low impedance, or short, at point (3). This effect creates a good electrical connection between the two sections that permits energy to pass with very little reflection or loss.

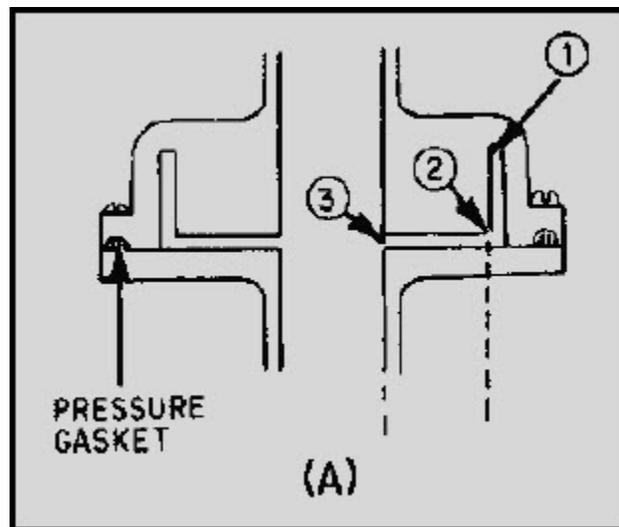


Figure 1-51A Choke joint

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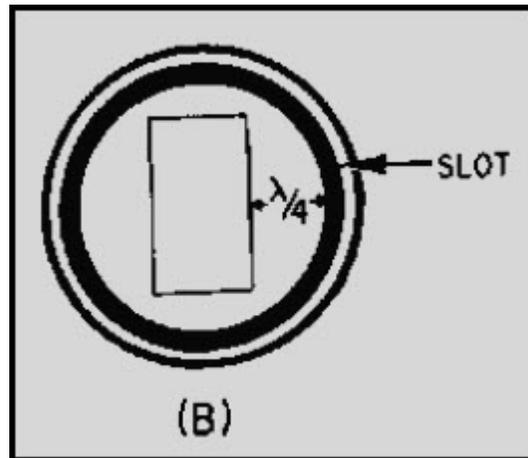


Figure 1-51B Choke joint

Whenever a stationary rectangular waveguide is to be connected to a rotating antenna, a rotating joint must be used. A circular waveguide is normally used in a rotating joint. Rotating a rectangular waveguide would cause field pattern distortion. The rotating section of the joint, illustrated in figure 1-52, uses a choke joint to complete the electrical connection with the stationary section. The circular waveguide is designed so that it will operate in the $TM_{0,1}$ mode. The rectangular sections are attached as shown in the illustration to prevent the circular waveguide from operating in the wrong mode.

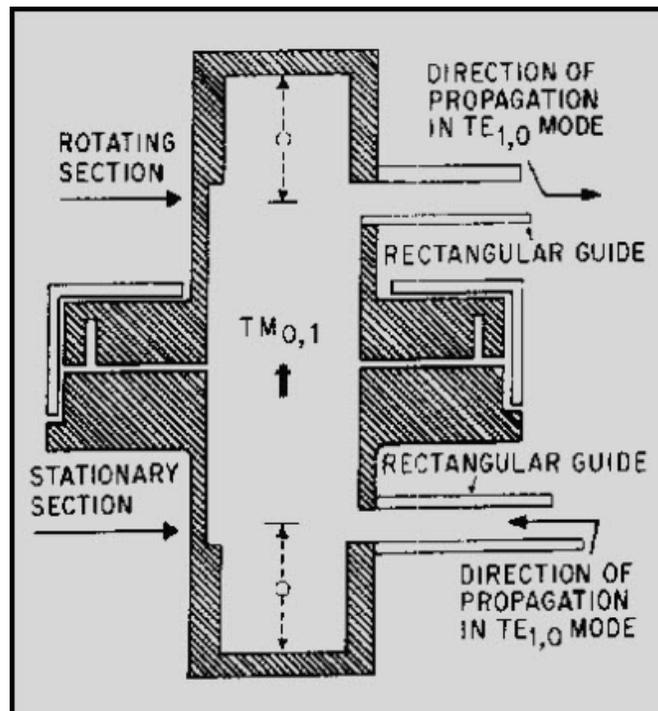


Figure 1-52 Rotating joint

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Distance "O" is $3/4\lambda$ so that a high impedance will be presented to any unwanted modes. This is the most common design used for rotating joints, but other types may be used in specific applications.

WAVEGUIDE MAINTENANCE - The installation of a waveguide system presents problems that are not normally encountered when dealing with other types of transmission lines. These problems often fall within the technician's area of responsibility. A brief discussion of waveguide handling, installation, and maintenance will help prepare you for this maintenance responsibility. Detailed information concerning waveguide maintenance in a particular system may be found in the technical manuals for the system.

Since a waveguide naturally has a low loss ratio, most losses in a waveguide system are caused by other factors. Improperly connected joints or damaged inner surfaces can decrease the efficiency of a system to the point that it will not work at all. Therefore, you must take great care when working with waveguides to prevent physical damage. Since waveguides are made from a soft, conductive material, such as copper or aluminum, they are very easy to dent or deform. Even the slightest damage to the inner surface of a waveguide will cause standing waves and, often, internal arcing. Internal arcing causes further damage to the waveguide in an action that is often self-sustaining until the waveguide is damaged beyond use. Part of your job as a technician will be to inspect the waveguide system for physical damage. The previously mentioned dents are only one type of physical damage that can decrease the efficiency of the system. Another problem occurs because waveguides are made from a conductive material such as copper while the structures of most ships are made from steel. When two dissimilar metals, such as copper and steel, are in direct contact, an electrical action called ELECTROLYSIS takes place that causes very rapid corrosion of the metals. Waveguides can be completely destroyed by electrolytic corrosion in a relatively short period of time if they are not isolated from direct contact with other metals. Any inspection of a waveguide system should include a detailed inspection of all support points to ensure that electrolytic corrosion is not taking place. Any waveguide that is exposed to the weather should be painted and all joints sealed. Proper painting prevents natural corrosion, and sealing the joints prevents moisture from entering the waveguide.

Moisture can be one of the worst enemies of a waveguide system. As previously discussed, the dielectric in waveguides is air, which is an excellent dielectric as long as it is free of moisture. Wet air, however, is a very poor dielectric and can cause serious internal arcing in a waveguide system. For this reason care is taken to ensure that waveguide systems are pressurized with air that is dry. Checking the pressure and moisture content of the waveguide air may be one of your daily system maintenance duties.

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More detailed waveguide installation and maintenance information can be found in the technical manuals that apply to your particular system. Another good source is the Electronics Installation and Maintenance Handbooks (EIMB) published by Naval Sea Systems Command. Installation Standards Handbook EIMB, NAVSEA 0967-LP-000-0110, is the volume that deals with waveguide installation and maintenance.

Q-41. What is the result of an abrupt change in the size, shape, or dielectric of a waveguide?

Q-42. A waveguide bend must have what minimum radius?

Q-43. What is the most common type of waveguide joint?

Q-44. What is the most likely cause of losses in waveguide systems?

1.3 WAVEGUIDE DEVICES

The discussion of waveguides, up to this point, has been concerned only with the transfer of energy from one point to another. Many waveguide devices have been developed, however, that modify the energy in some fashion during transit. Some devices do nothing more than change the direction of the energy. Others have been designed to change the basic characteristics or power level of the electromagnetic energy.

This section will explain the basic operating principles of some of the more common waveguide devices, such as DIRECTIONAL COUPLERS, CAVITY RESONATORS, and HYBRID JUNCTIONS

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1.3.1 Directional Couplers

The directional coupler is a device that provides a method of sampling energy from within a waveguide for measurement or use in another circuit. Most couplers sample energy traveling in one direction only. However, directional couplers can be constructed that sample energy in both directions. These are called BIDIRECTIONAL couplers and are widely used in radar and communications systems.

Directional couplers may be constructed in many ways. The coupler illustrated in figure 1-53 is constructed from an enclosed waveguide section of the same dimensions as the waveguide in which the energy is to be sampled. The "b" wall of this enclosed section is mounted to the "b" wall of the waveguide from which the sample will be taken. There are two holes in the "b" wall between the sections of the coupler. These two holes are $1/4\lambda$ apart. The upper section of the directional coupler has a wedge of energy-absorbing material at one end and a pickup probe connected to an output jack at the other end. The absorbent material absorbs the energy not directed at the probe and a portion of the overall energy that enters the section.

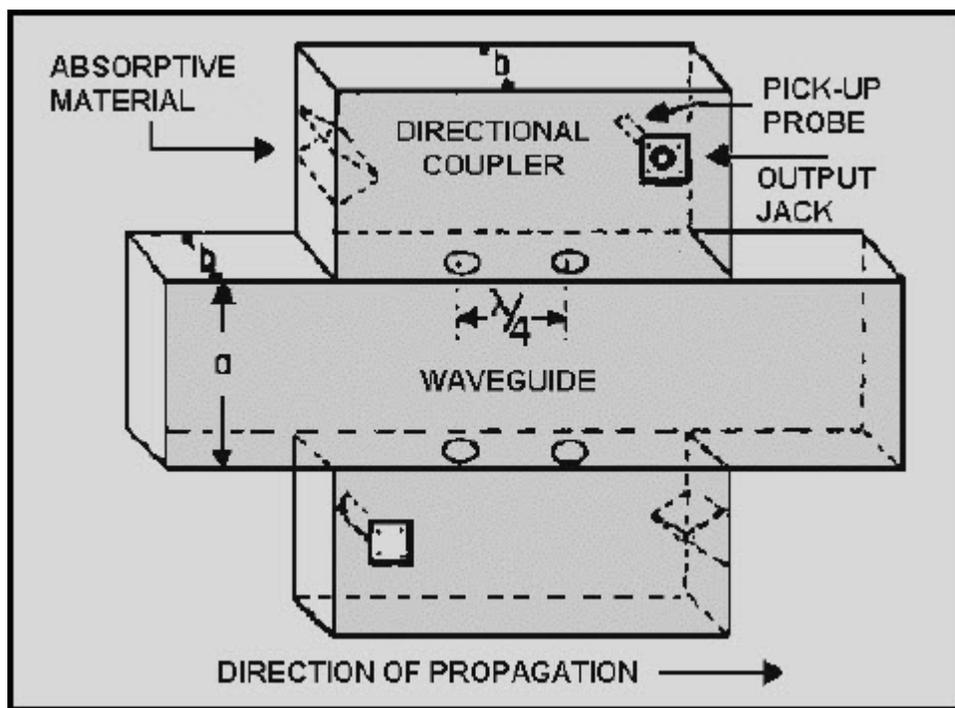


Figure 1-53 Directional coupler

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Figure 1-54 illustrates two portions of the incident wavefront in a waveguide. The waves travel down the waveguide in the direction indicated and enter the coupler section through both holes. Since both portions of the wave travel the same distance, they are in phase when they arrive at the pickup probe. Because the waves are in phase, they add together and provide a sample of the energy traveling down the waveguide. The sample taken is only a small portion of the energy that is traveling down the waveguide. The magnitude of the sample, however, is proportional to the magnitude of the energy in the waveguide. The absorbent material is designed to ensure that the ratio between the sample energy and the energy in the waveguide is constant. Otherwise the sample would contain no useful information.

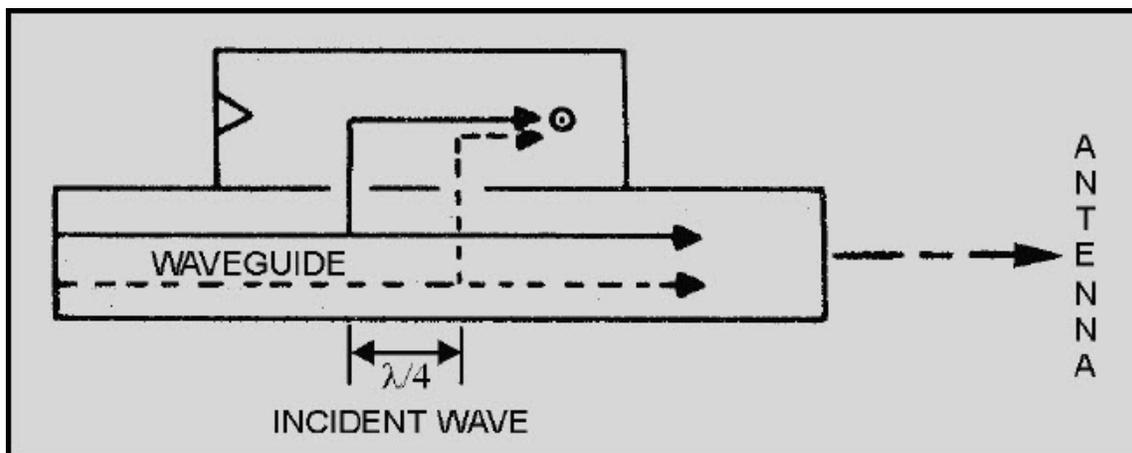


Figure 1-54 Incident wave in a directional coupler designed to sample incident waves

The ratio is usually stamped on the coupler in the form of an attenuation factor.

The effect of a directional coupler on any reflected energy is illustrated in figure 1-55. Note that these two waves do not travel the same distance to the pickup probe. The wave represented by the dotted line travels $1/2\lambda$ further and arrives at the probe 180 degrees out of phase with the wave represented by the solid line. Because the waves are 180 degrees out of phase at the probe, they cancel each other and no energy is induced in the pickup probe. When the reflected energy arrives at the absorbent material, it adds and is absorbed by the material.

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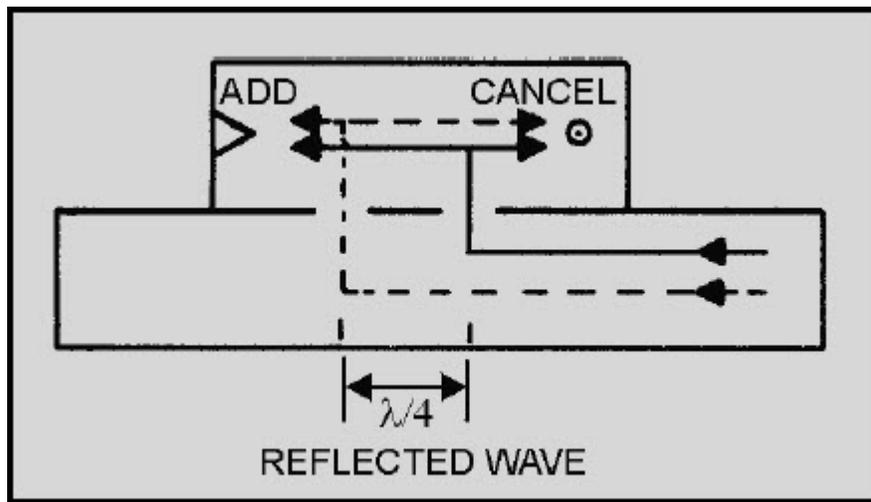


Figure 1-55 Reflected wave in a directional coupler

A directional coupler designed to sample reflected energy is shown in figure 1-56. The absorbent material and the probe are in opposite positions from the directional coupler designed to sample the incident energy. This positioning causes the two portions of the reflected energy to arrive at the probe in phase, providing a sample of the reflected energy. The sampled transmitted energy, however, is absorbed by the absorbent material.

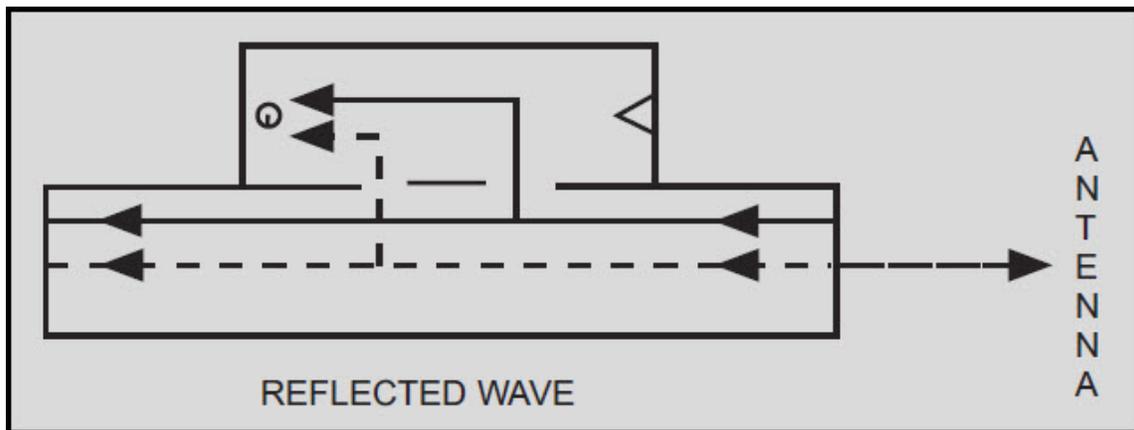


Figure 1-56 Directional coupler designed to sample reflected energy

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A simple bidirectional coupler for sampling both transmitted and reflected energy can be constructed by mounting two directional couplers on opposite sides of a waveguide, as shown in figure 1-57.

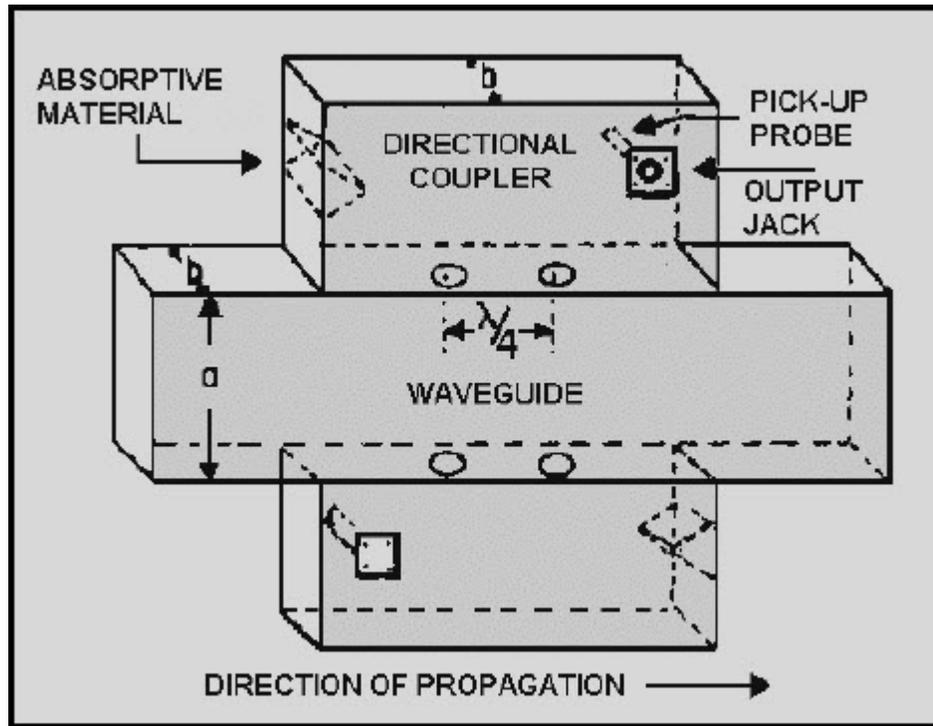


Figure 1-57 Bidirectional coupler

Q-45. What is the primary purpose of a directional coupler?

Q-46. How far apart are the two holes in a simple directional coupler?

Q-47. What is the purpose of the absorbent material in a directional coupler?

Q-48. In a directional coupler that is designed to sample the incident energy, what happens to the two portions of the wavefront when they arrive at the pickup probe?

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Q-49. What happens to reflected energy that enters a directional coupler that is designed to sample incident energy?

1.3.2 Cavity Resonators

In ordinary electronic equipment a resonant circuit consists of a coil and a capacitor that are connected either in series or in parallel. The resonant frequency of the circuit is increased by reducing the capacitance, the inductance, or both. A point is eventually reached where the inductance and the capacitance can be reduced no further. This is the highest frequency at which a conventional circuit can oscillate.

The upper limit for a conventional resonant circuit is between 2000 and 3000 megahertz. At these frequencies, the inductance may consist of a coil of one-half turn, and the capacitance may simply be the stray capacitance of the coil. Tuning a one-half turn coil is very difficult and tuning stray capacitance is even more difficult. In addition, such a circuit will handle only very small amounts of current.

NEETS, Module 10, *Introduction to Wave Propagation* explained that a $1/4\lambda$ section of transmission line can act as a resonant circuit. The same is true of a $1/4\lambda$ section of waveguide. Since a waveguide is hollow, it can also be considered as a RESONANT CAVITY.

By definition, a resonant cavity is any space completely enclosed by conducting walls that can contain oscillating electromagnetic fields and possess resonant properties. The cavity has many advantages and uses at microwave frequencies. Resonant cavities have a very high Q and can be built to handle relatively large amounts of power. Cavities with a Q value in excess of 30,000 are not uncommon. The high Q gives these devices a narrow bandpass and allows very accurate tuning. Simple, rugged construction is an additional advantage.

Although cavity resonators, built for different frequency ranges and applications, have a variety of shapes, the basic principles of operation are the same for all.

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One example of a cavity resonator is the rectangular box shown in figure 1-58A. It may be thought of as a section of rectangular waveguide closed at both ends by conducting plates. The frequency at which the resonant mode occurs is $1/2\lambda$ of the distance between the end plates. The magnetic and electric field patterns in the rectangular cavity are shown in figure 1-58B.

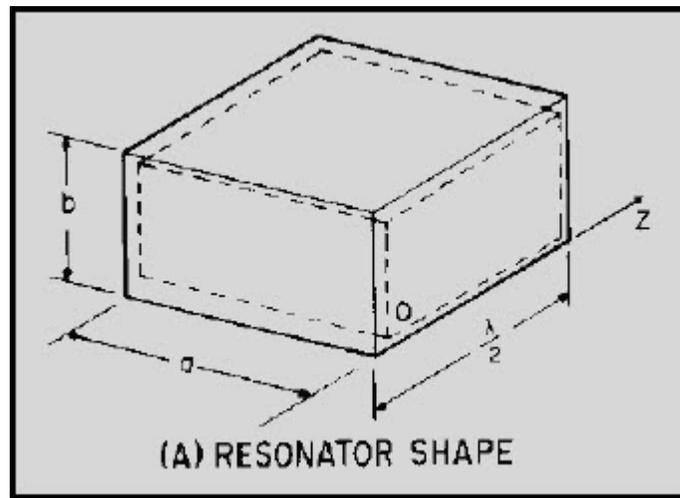


Figure 1-58A Rectangular waveguide cavity resonator.
RESONATOR SHAPE

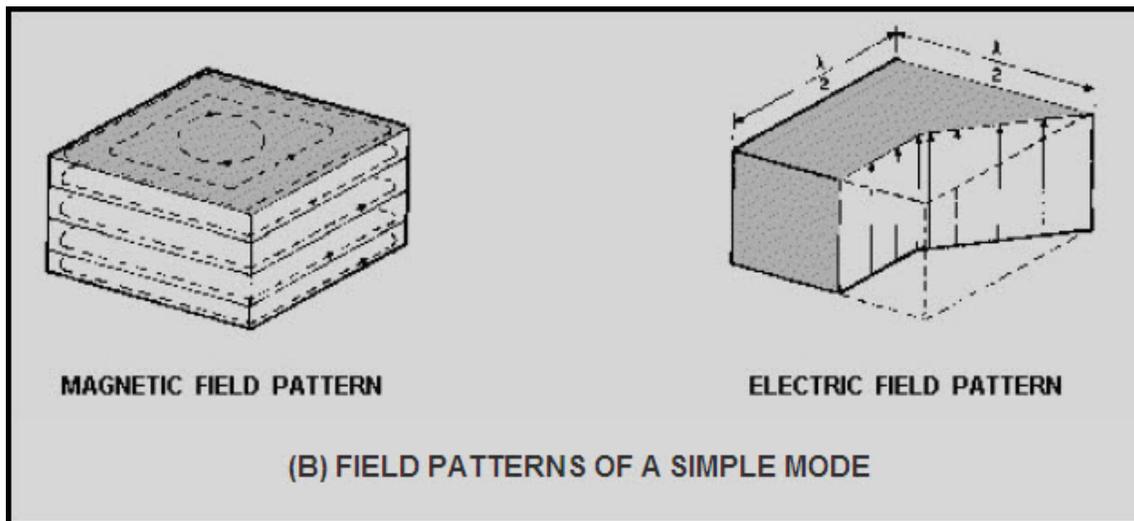


Figure 1-58B Rectangular waveguide cavity resonator. FIELD PATTERNS OF A SIMPLE
MODE

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The rectangular cavity is only one of many cavity devices that are useful as high-frequency resonators. Figures 1-59A, 1-59B, 1-59C, and 1-59D show the development of a cylindrical resonant cavity from an infinite number of quarter-wave sections of transmission line. In figure 1-59A the $1/4\lambda$ section is shown to be equivalent to a resonant circuit with a very small amount of inductance and capacitance. Three $1/4\lambda$ sections are joined in parallel in figure 1-59B. Note that although the current-carrying ability of several $1/4\lambda$ sections is greater than that of any one section, the resonant frequency is unchanged. This occurs because the addition of inductance in parallel lowers the total inductance, but the addition of capacitance in parallel increases the total capacitance by the same proportion. Thus, the resonant frequency remains the same as it was for one section. The increase in the number of current paths also decreases the total resistance and increases the Q of the resonant circuit. Figure 1-59C shows an intermediate step in the development of the cavity. Figure 1-59D shows a completed cylindrical resonant cavity with a diameter of $1/2\lambda$ at the resonant frequency.

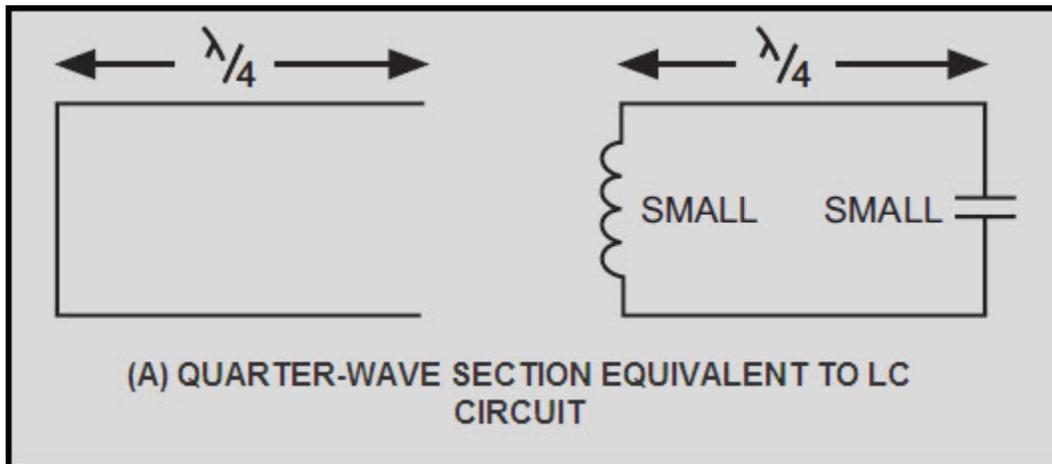


Figure 1-59A Development of a cylindrical resonant cavity. QUARTER-WAVE SECTION EQUIVALENT TO LC CIRCUIT

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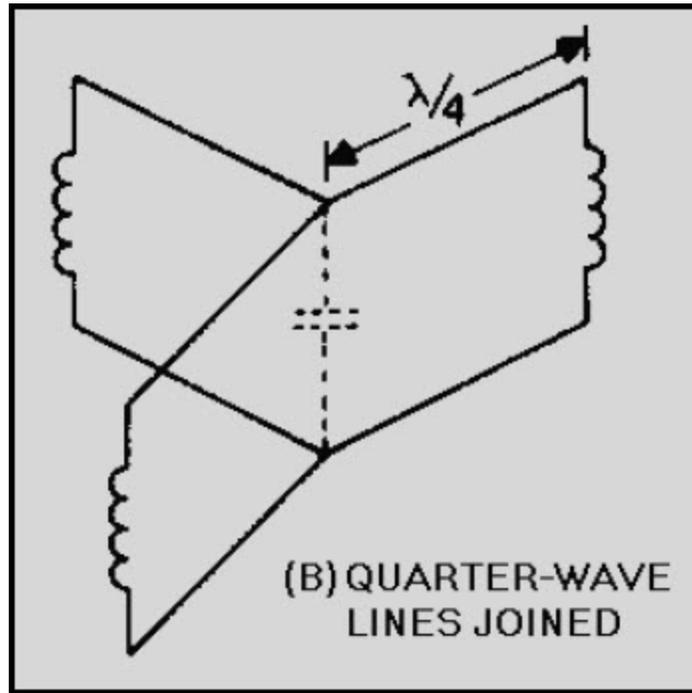
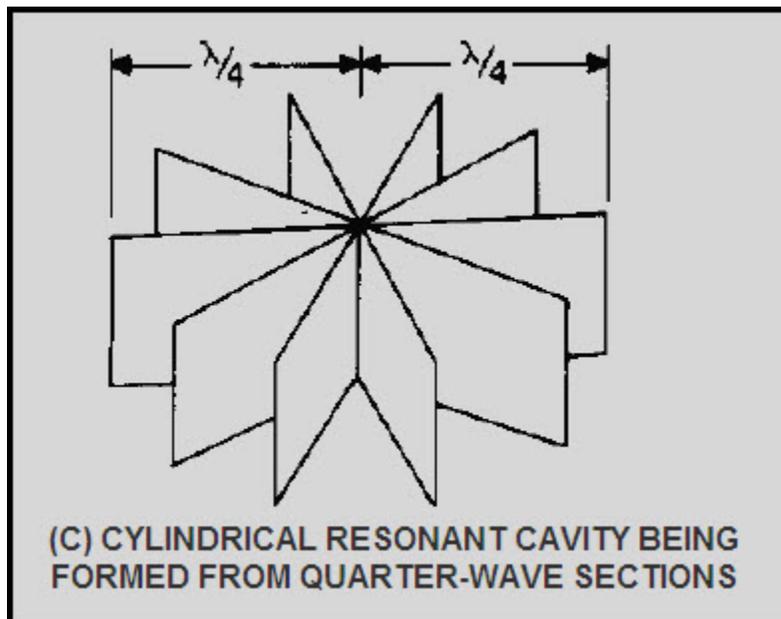


Figure 1-59B Development of a cylindrical resonant cavity. QUARTER-WAVE LINES JOINED



**Figure 1-59C Development of a cylindrical resonant cavity.
CYLINDRICAL RESONANT CAVITY BEING
FORMED FROM QUARTER-WAVE SECTIONS**

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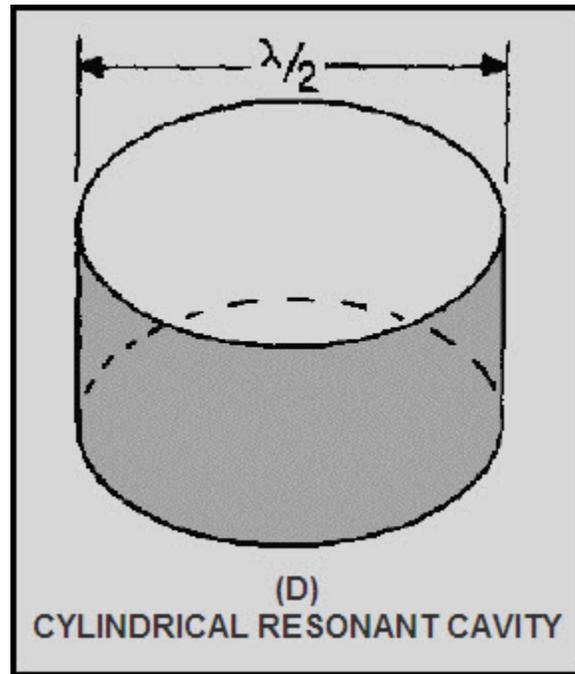


Figure 1-59D Development of a cylindrical resonant cavity. CYLINDRICAL RESONANT CAVITY

There are two variables that determine the primary frequency of any resonant cavity. The first variable is **PHYSICAL SIZE**. In general, the smaller the cavity, the higher its resonant frequency. The second controlling factor is the **SHAPE** of the cavity.

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Figure 1-60 illustrates several cavity shapes that are commonly used. Remember from the previously stated definition of a resonant cavity that any completely enclosed conductive surface, regardless of its shape, can act as a cavity resonator.

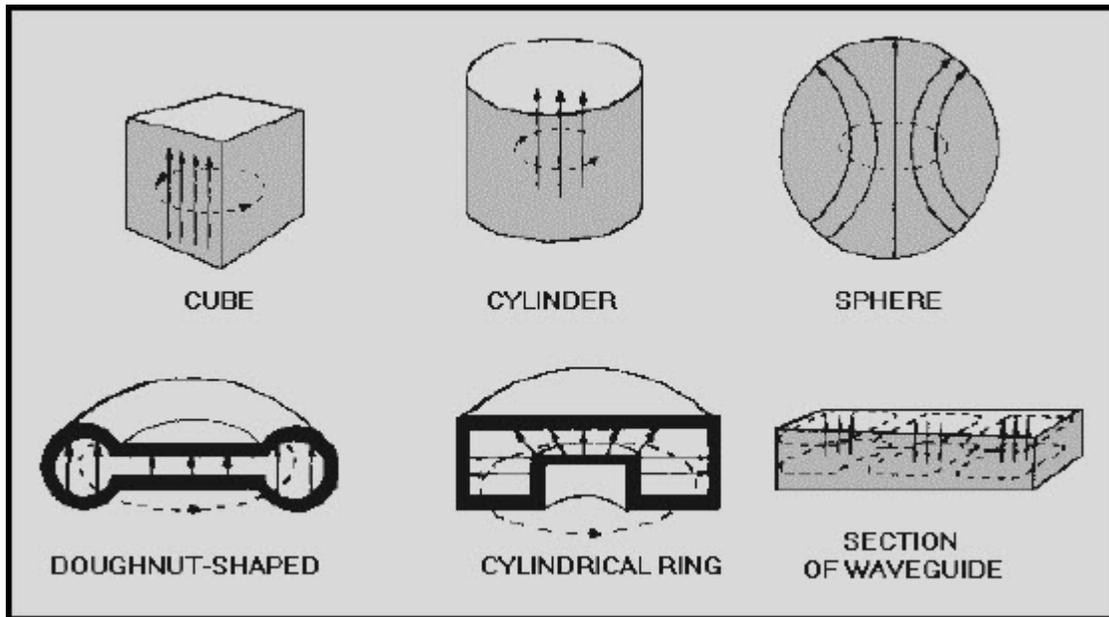


Figure 1-60 Several types of cavities

Cavity resonators are energized in basically the same manner as waveguides and have a similar field distribution. If the cavity shown in figure 1-61 were energized in the TE mode, the electromagnetic wave would reflect back and forth along the Z axis and form standing waves. These standing waves would form a field configuration within the cavity that would have to satisfy the same boundary conditions as those in a waveguide. Modes of operation in the cavity are described in terms of the fields that exist in the X, Y, and Z dimensions. Three subscripts are used; the first subscript indicates the number of $1/2\lambda$ along the X axis; the second subscript indicates the number of $1/2\lambda$ along the Y axis; and the third subscript indicates the number of $1/2\lambda$ along the Z axis.

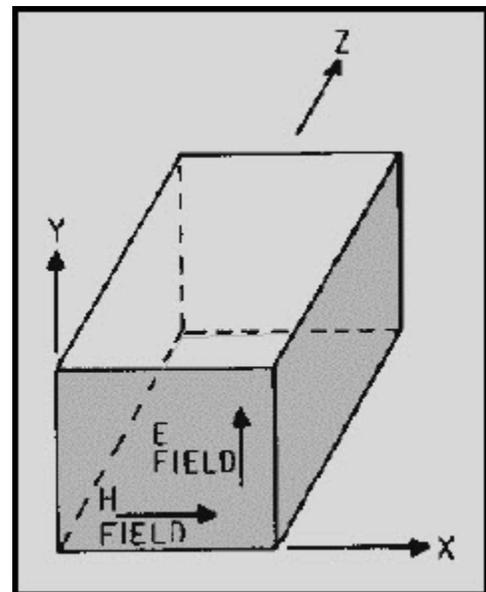


Figure 1-61 Rectangular cavity resonator

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Energy can be inserted or removed from a cavity by the same methods that are used to couple energy into and out of waveguides. The operating principles of probes, loops, and slots are the same whether used in a cavity or a waveguide. Therefore, any of the three methods can be used with cavities to inject or remove energy.

The resonant frequency of a cavity can be varied by changing any of three parameters: cavity volume, cavity capacitance, or cavity inductance. Changing the frequencies of a cavity is known as TUNING. The mechanical methods of tuning a cavity may vary with the application, but all methods use the same electrical principles.

A mechanical method of tuning a cavity by changing the volume (VOLUME TUNING) is illustrated in figure 1-62. Varying the distance d will result in a new resonant frequency because the inductance and the capacitance of the cavity are changed by different amounts. If the volume is decreased, the resonant frequency will be higher. The resonant frequency will be lower if the volume of the cavity is made larger.

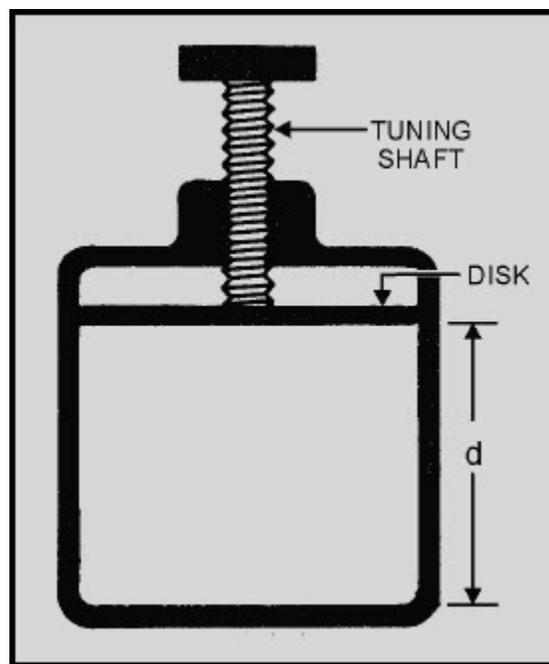


Figure 1-62 Cavity tuning by volume

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CAPACITIVE TUNING of a cavity is shown in figure 1-63A. An adjustable slug or screw is placed in the area of maximum E lines. The distance d represents the distance between two capacitor plates. As the slug is moved in, the distance between the two plates becomes smaller and the capacitance increases. The increase in capacitance causes a decrease in the resonant frequency. As the slug is moved out, the resonant frequency of the cavity increases.

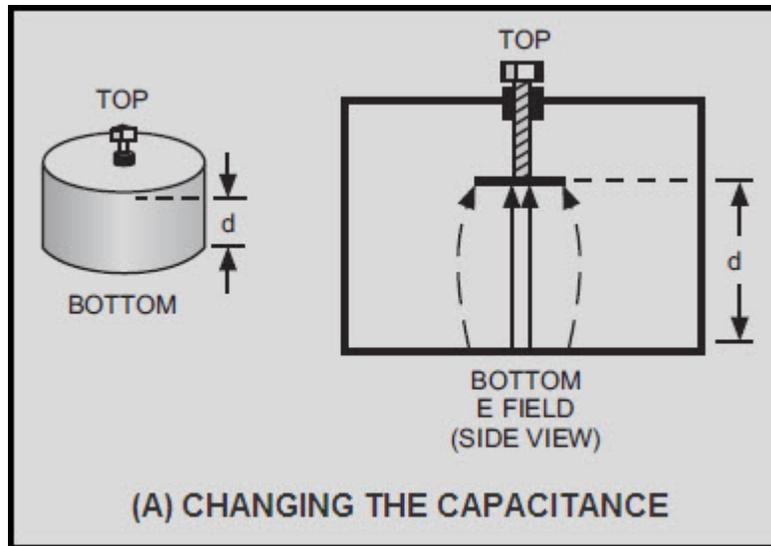


Figure 1-63A Methods of changing the resonant frequency of a cavity. CHANGING THE CAPACITANCE

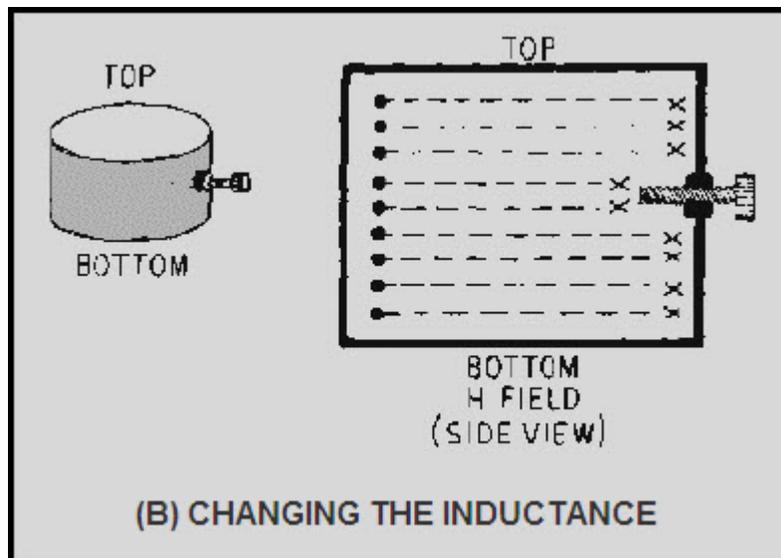


Figure 1-63B Methods of changing the resonant frequency of a cavity. CHANGING THE INDUCTANCE

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INDUCTIVE TUNING is accomplished by placing a nonmagnetic slug in the area of maximum H lines, as shown in figure 1-63B. The changing H lines induce a current in the slug that sets up an opposing H field. The opposing field reduces the total H field in the cavity, and therefore reduces the total inductance. Reducing the inductance, by moving the slug in, raises the resonant frequency. Increasing the inductance, by moving the slug out, lowers the resonant frequency.

Resonant cavities are widely used in the microwave range, and many of the applications will be studied in chapter 2. For example, most microwave tubes and transmitting devices use cavities in some form to generate microwave energy. Cavities are also used to determine the frequency of the energy traveling in a waveguide, since conventional measurement devices do not work well at microwave frequencies.

Q-50. What two variables determine the primary frequency of a resonant cavity?

Q-51. Energy can be inserted or removed from a cavity by what three methods?

Q-52. Inductive tuning of a resonant cavity is accomplished by placing a nonmagnetic slug in what area?

1.3.3 Waveguide Junctions

You may have assumed that when energy traveling down a waveguide reaches a junction, it simply divides and follows the junction. This is not strictly true. Different types of junctions affect the energy in different ways. Since waveguide junctions are used extensively in most systems, you need to understand the basic operating principles of those most commonly used.

The T JUNCTION is the most simple of the commonly used waveguide junctions. T junctions are divided into two basic types, the E-TYPE and the H-TYPE. HYBRID JUNCTIONS are more complicated developments of the basic T junctions. The MAGIC-T and the HYBRID RING are the two most commonly used hybrid junctions.

E-TYPE T JUNCTION - An E-type T junction is illustrated in figure 1-64, view (A). It is called an E-type T junction because the junction arm extends from the main waveguide in the same direction as the E field in the waveguide.

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Figure 1-64, view (B), illustrates cross-sectional views of the E-type T junction with inputs fed into the various arms. For simplicity, the magnetic lines that are always present with an electric field have been omitted. In view (K), the input is fed into arm b and the outputs are taken from the a and c arms. When the E field arrives between points 1 and 2, point 1 becomes positive and point 2 becomes negative. The positive charge at point 1 then induces a negative charge on the wall at point 3. The negative charge at point 2 induces a positive charge at point 4. These charges cause the fields to form 180 degrees out of phase in the main waveguide; therefore, the outputs will be 180 degrees out of phase with each other. In view (L), two in-phase inputs of equal amplitude are fed into the a and c arms. The signals at points 1 and 2 have the same phase and amplitude. No difference of potential exists across the entrance to the b arm, and no energy will be coupled out. However, when the two signals fed into the a and c arms are 180 degrees out of phase, as shown in view (M), points 1 and 2 have a difference of potential. This difference of potential induces an E field from point 1 to point 2 in the b arm, and energy is coupled out of this arm. Views (N) and (P) illustrate two methods of obtaining two outputs with only one input.

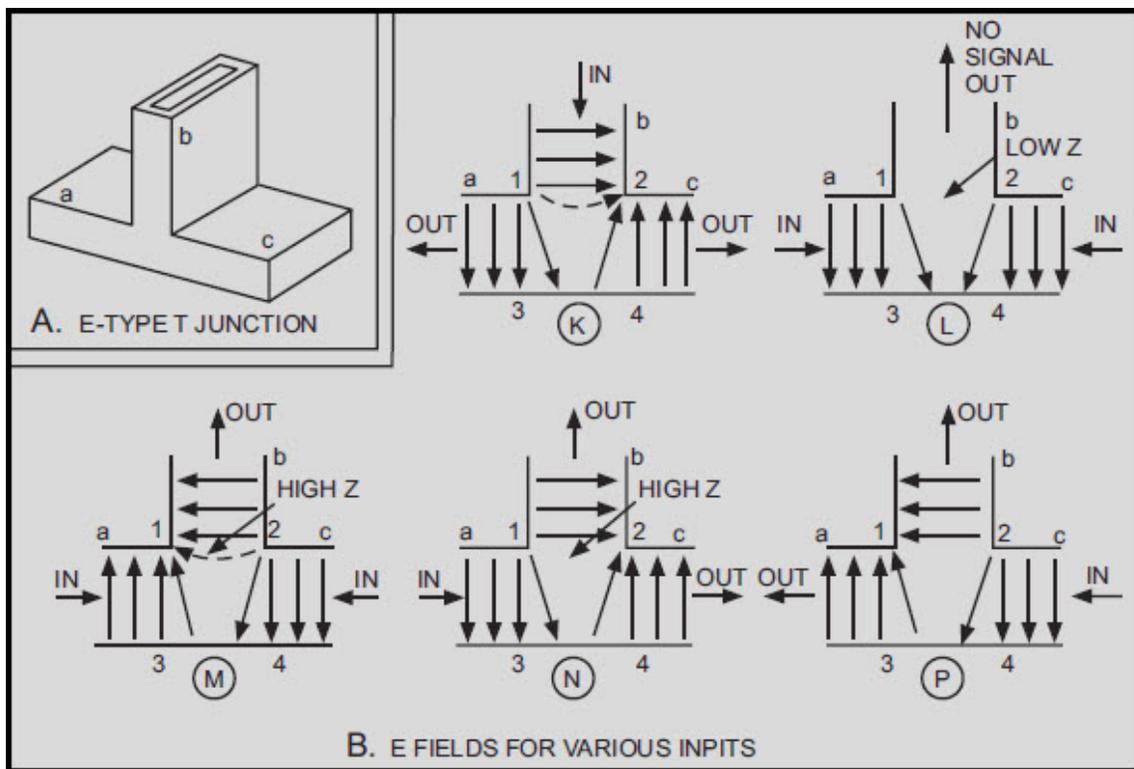


Figure 1-64 E fields in an E-type T junction

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H-TYPE T JUNCTION..An H-type T junction is illustrated in figure 1-65A. It is called an H-type T junction because the long axis of the "b" arm is parallel to the plane of the magnetic lines of force in the waveguide. Again, for simplicity, only the E lines are shown in this figure. Each X indicates an E line moving away from the observer. Each dot indicates an E line is moving toward the observer.

In view (1) of figure 1-65B, the signal is fed into arm b and in-phase outputs are obtained from the a and c arms. In view (2), in-phase signals are fed into arms a and c and the output signal is obtained from the b arm because the fields add at the junction and induce E lines into the b arm. If 180-degree-out-of-phase signals are fed into arms a and c, as shown in view (3), no output is obtained from the b arm because the opposing fields cancel at the junction. If a signal is fed into the a arm, as shown in view (4), outputs will be obtained from the b and c arms. The reverse is also true. If a signal is fed into the c arm, outputs will be obtained from the a and b arms.

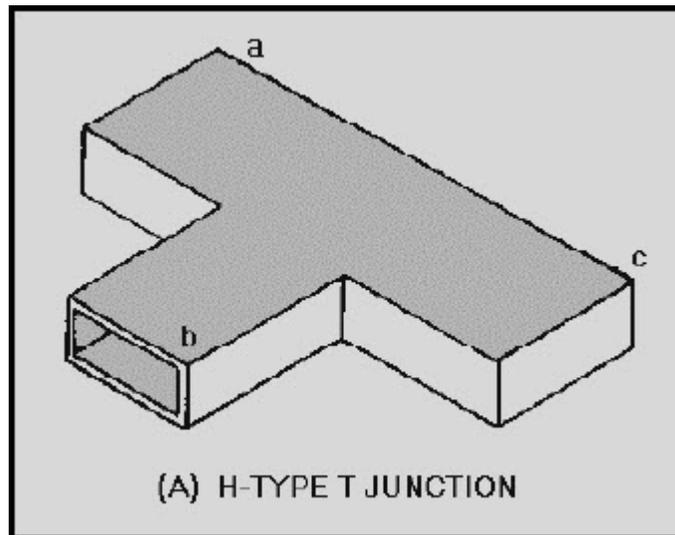


Figure 1-65A E fields in an H-type junction. H-TYPE T JUNCTION

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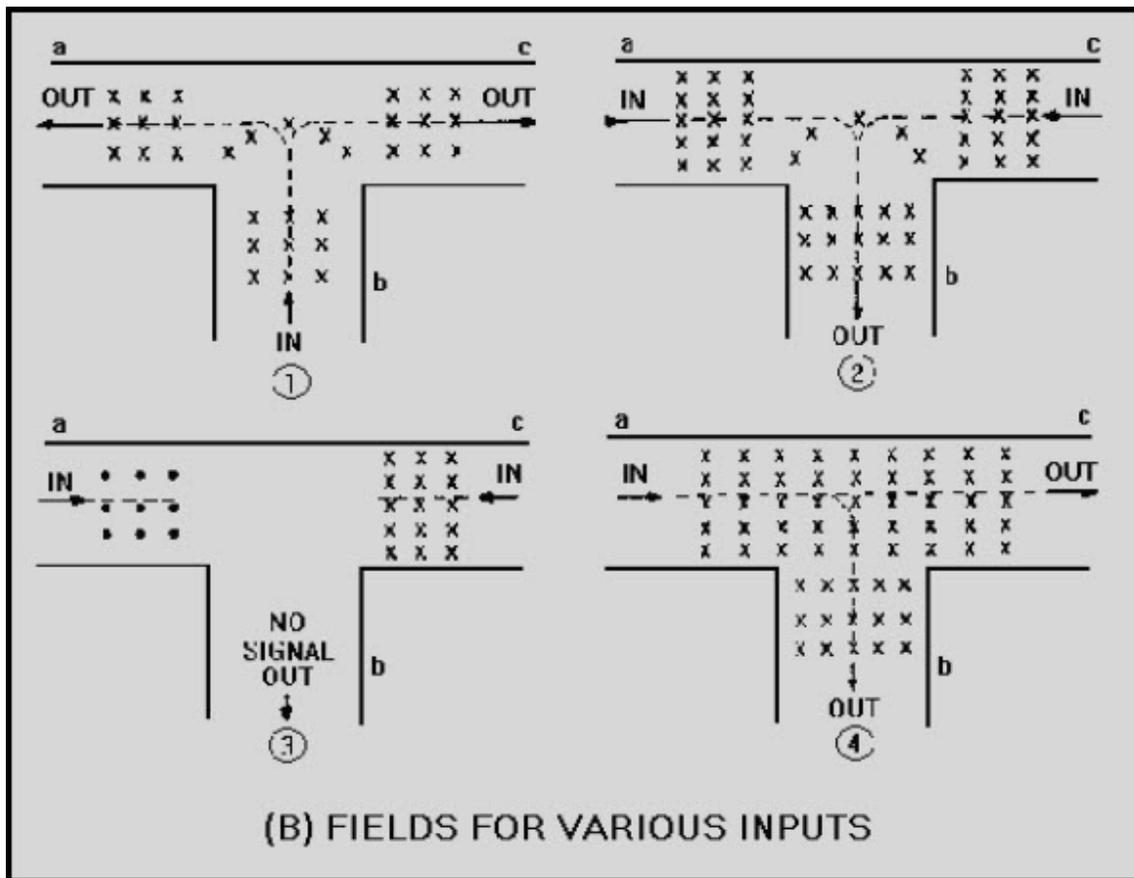


Figure 1-65B E fields in an H-type junction. FIELDS FOR VARIOUS INPUTS

MAGIC-T HYBRID

JUNCTION - A simplified version of the magic-T hybrid junction is shown in figure 1-66. The magic-T is a combination of the H-type and E-type T junctions. The most common application of this type of junction is as the mixer section for microwave radar receivers. Its operation as a mixer will be discussed in later NEETS modules. At present, only the fields within the magic-T junction will be discussed.

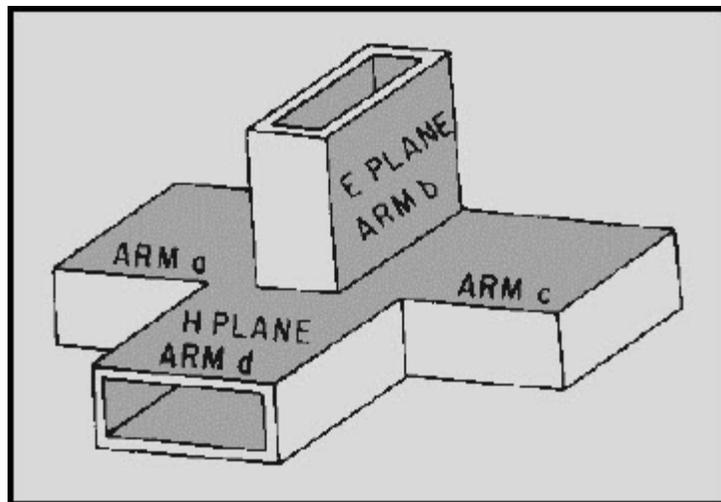


Figure 1-66 Magic-T hybrid junction

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If a signal is fed into the b arm of the magic-T, it will divide into two out-of-phase components. As shown in figure 1-67A, these two components will move into the a and c arms. The signal entering the b arm will not enter the d arm because of the zero potential existing at the entrance of the d arm. The potential must be zero at this point to satisfy the boundary conditions of the b arm. This absence of potential is illustrated in figures 1-67B and 1-67C where the magnitude of the E field in the b arm is indicated by the length of the arrows. Since the E lines are at maximum in the center of the b arm and minimum at the edge where the d arm entrance is located, no potential difference exists across the mouth of the d arm.

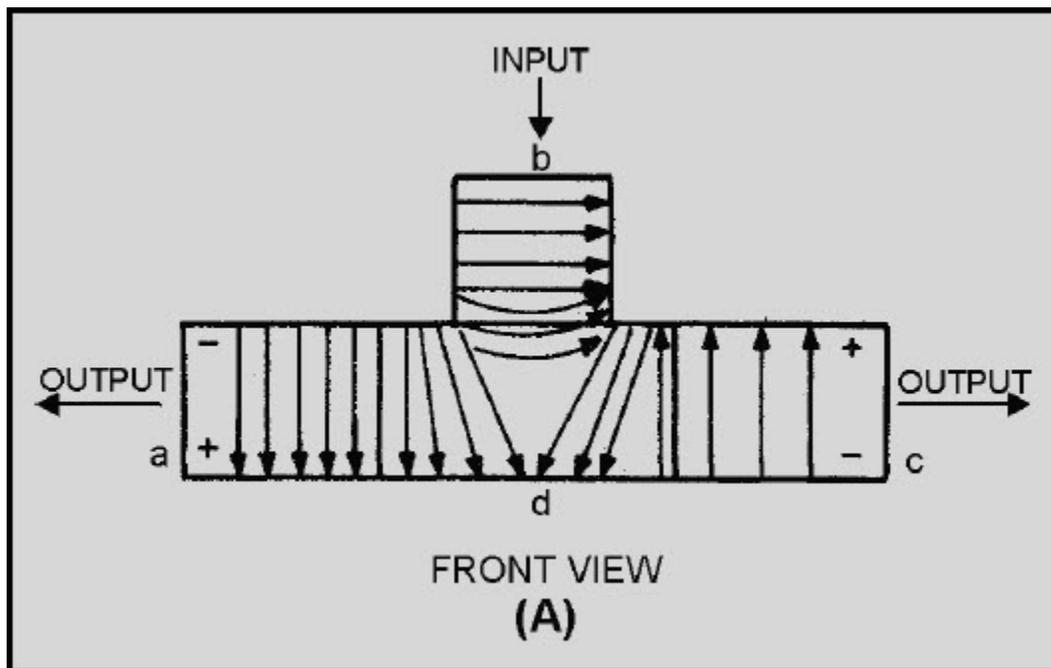


Figure 1-67A Magic-T with input to arm b

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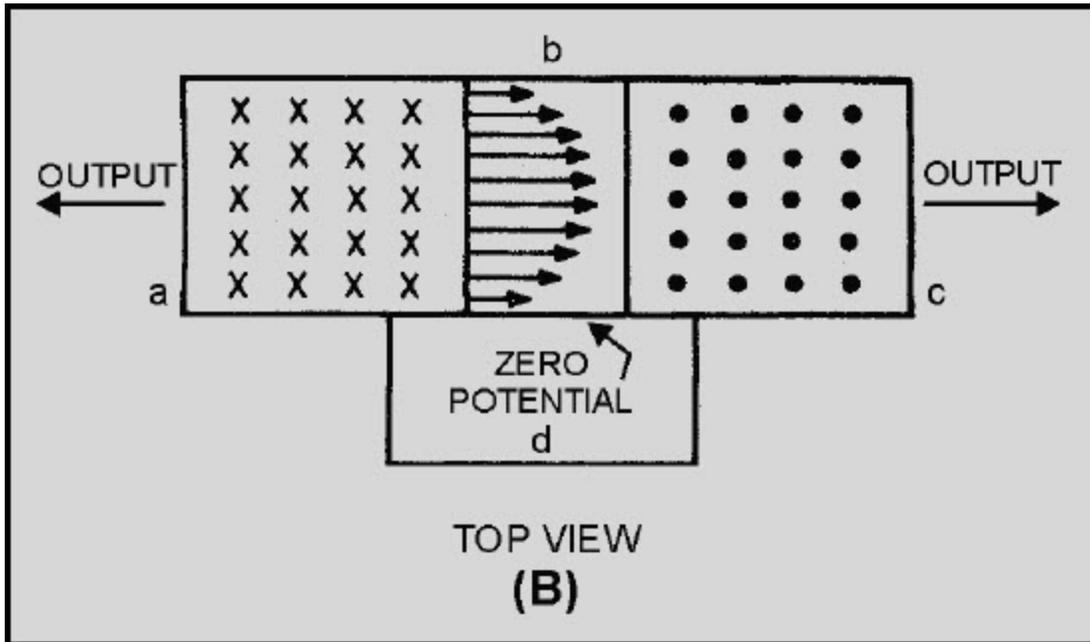


Figure 1-67B Magic-T with input to arm b

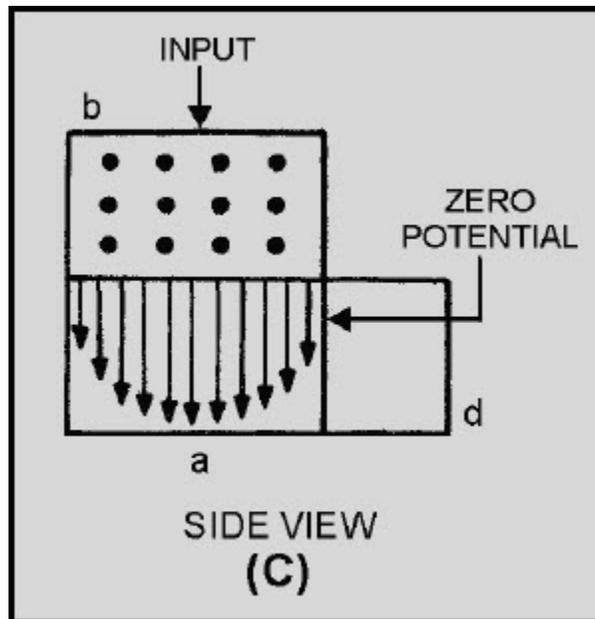


Figure 1-67C Magic-T with input to arm b

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In summary, when an input is applied to arm b of the magic-T hybrid junction, the output signals from arms a and c are 180 degrees out of phase with each other, and no output occurs at the d arm.

The action that occurs when a signal is fed into the d arm of the magic-T is illustrated in figure 1-68. As with the H-type T junction, the signal entering the d arm divides and moves down the a and c arms as outputs which are in phase with each other and with the input. The shape of the E fields in motion is shown by the numbered curved slices. As the E field moves down the d arm, points 2 and 3 are at an equal potential. The energy divides equally into arms a and c, and the E fields in both arms become identical in shape. Since the potentials on both sides of the b arm are equal, no potential difference exists at the entrance to the b arm, resulting in no output.

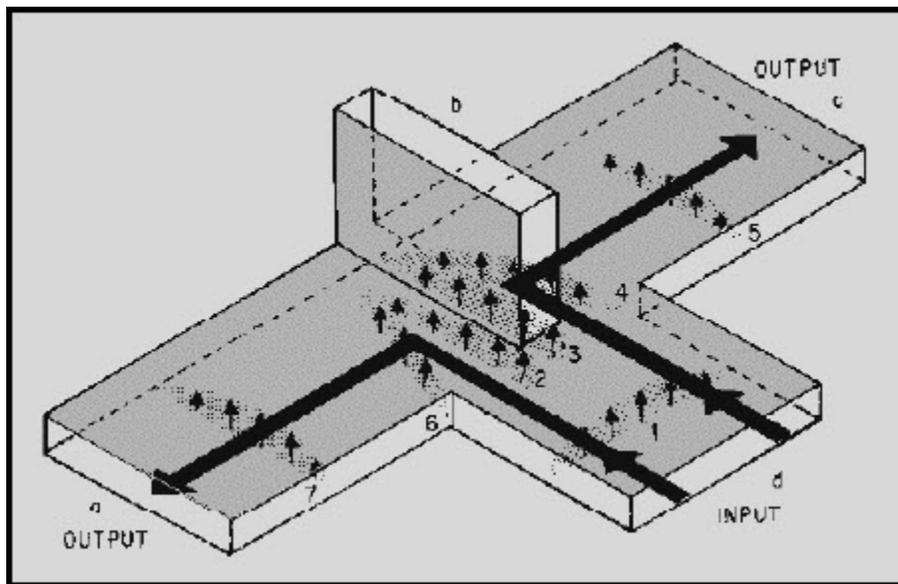


Figure 1-68 Magic-T with input to arm d

When an input signal is fed into the a arm as shown in figure 1-69, a portion of the energy is coupled into the b arm as it would be in an E-type T junction. An equal portion of the signal is coupled through the d arm because of the action of the H-type junction. The c arm has two fields across it that are out of phase with each other. Therefore the fields cancel, resulting in no output at the c arm. The reverse of this action takes place if a signal is fed into the c arm, resulting in outputs at the b and d arms and no output at the a arm.

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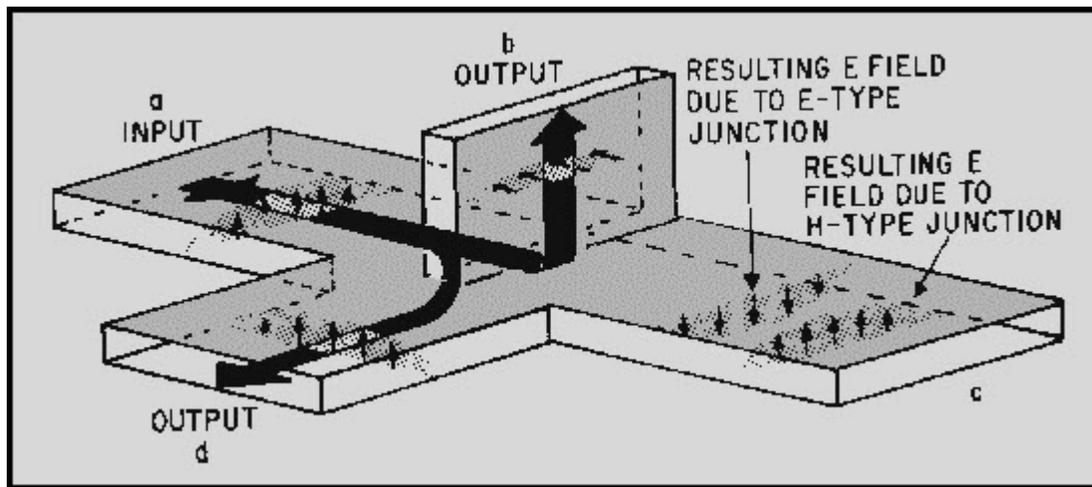


Figure 1-69 Magic-T with input to arm a

Unfortunately, when a signal is applied to any arm of a magic-T, the flow of energy in the output arms is affected by reflections. Reflections are caused by impedance mismatching at the junctions. These reflections are the cause of the two major disadvantages of the magic-T. First, the reflections represent a power loss since all the energy fed into the junction does not reach the load which the arms feed. Second, the reflections produce standing waves that can result in internal arcing. Thus the maximum power a magic-T can handle is greatly reduced.

Reflections can be reduced by using some means of impedance matching that does not destroy the shape of the junctions. One method is shown in figure 1-70. A post is used to match the H plane, and an iris is used to match the E plane. Even though this method reduces reflections, it lowers the power-handling capability even further.

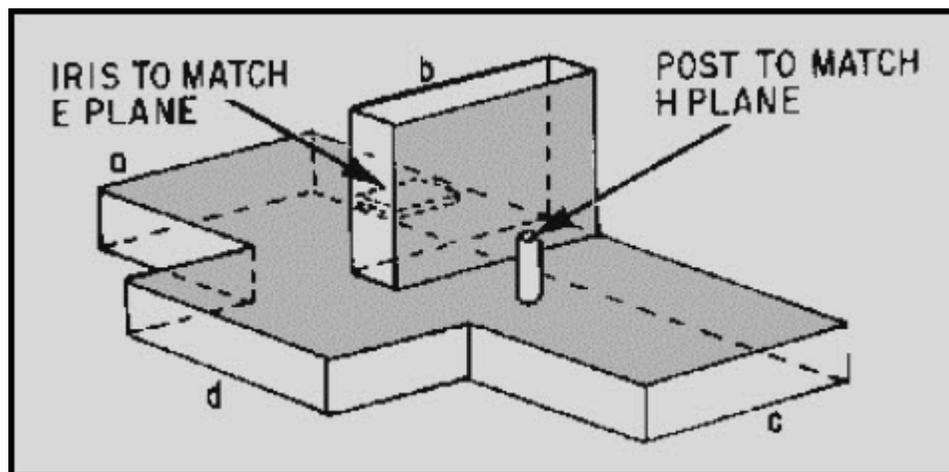


Figure 1-70 Magic-T impedance matching

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HYBRID RING - A type of hybrid junction that overcomes the power limitation of the magic-T is the hybrid ring, also called a RAT RACE. The hybrid ring, illustrated in figure 1-71A, is actually a modification of the magic-T. It is constructed of rectangular waveguides molded into a circular pattern. The arms are joined to the circular waveguide to form E-type T junctions. Figure 1-71B shows, in wavelengths, the dimensions required for a hybrid ring to operate properly.

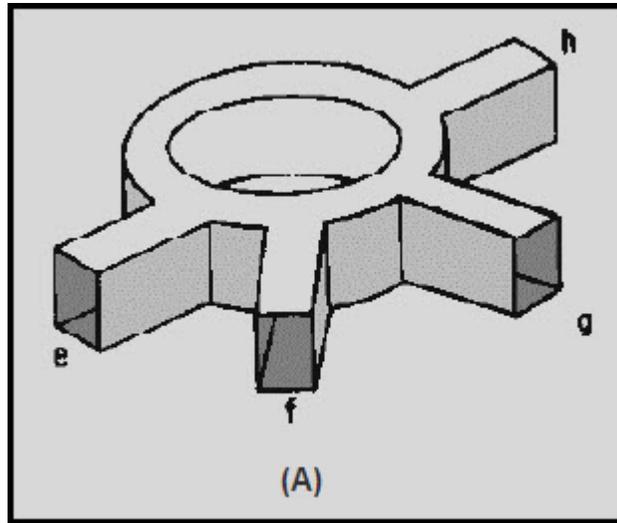


Figure 1-71 A Hybrid ring with wavelength measurements

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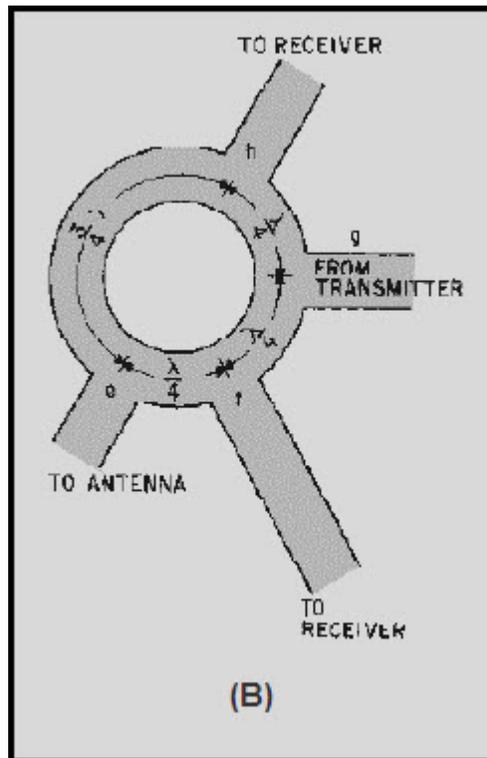


Figure 1-71B Hybrid ring with wavelength measurements

The hybrid ring is used primarily in high-powered radar and communications systems to perform two functions. During the transmit period, the hybrid ring couples microwave energy from the transmitter to the antenna and allows no energy to reach the receiver. During the receive cycle, the hybrid ring couples energy from the antenna to the receiver and allows no energy to reach the transmitter. Any device that performs both of these functions is called a DUPLEXER. A duplexer permits a system to use the same antenna for both transmitting and receiving. Since the only common application of the hybrid ring is as a duplexer, the details of hybrid ring operation will be explained in later *NEETS* modules concerning duplexers.

Q-53. What are the two basic types of T junctions?

Q-54. Why is the H-type T junction so named?

Q-55. The magic-T is composed of what two basic types of T junctions?

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Q-56. What are the primary disadvantages of the magic-T?

Q-57. What type of junctions are formed where the arms of a hybrid ring meet the main ring?

Q-58. Hybrid rings are used primarily for what purpose?

1.3.4 Ferrite Devices

A FERRITE is a device that is composed of material that causes it to have useful magnetic properties and, at the same time, high resistance to current flow. The primary material used in the construction of ferrites is normally a compound of iron oxide with impurities of other oxides added. The compound of iron oxide retains the properties of the ferromagnetic atoms, and the impurities of the other oxides increase the resistance to current flow. This combination of properties is not found in conventional magnetic materials. Iron, for example, has good magnetic properties but a relatively low resistance to current flow. The low resistance causes eddy currents and significant power losses at high frequencies (You may want to review NEETS, Module 2, Introduction to Alternating Current and Transformers, Chapter 5). Ferrites, on the other hand, have sufficient resistance to be classified as semiconductors.

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The compounds used in the composition of ferrites can be compared to the more familiar compounds used in transistors. As in the construction of transistors, a wide range of magnetic and electrical properties can be produced by the proper choice of atoms in the right proportions. An example of a ferrite device is shown in figure 1-72.

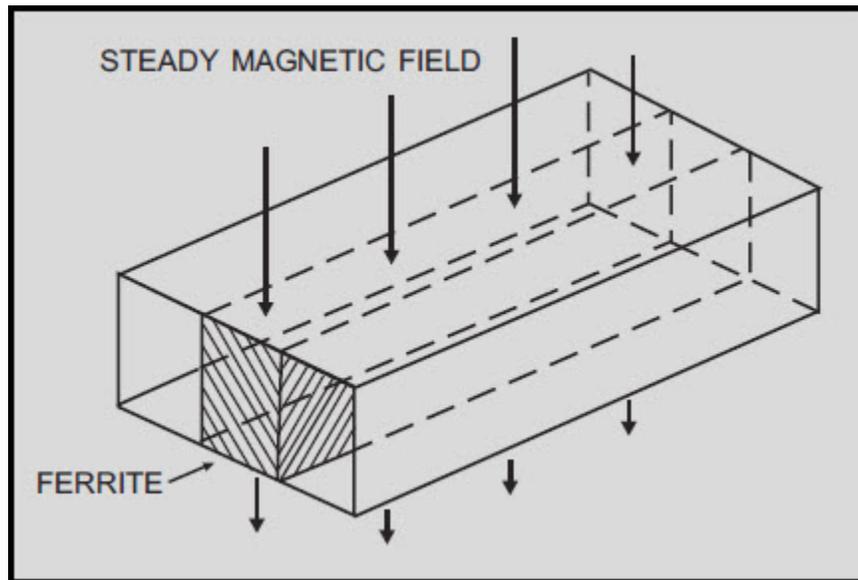


Figure 1-72 Ferrite attenuator

Ferrites have long been used at conventional frequencies in computers, television, and magnetic recording systems. The use of ferrites at microwave frequencies is a relatively new development and has had considerable influence on the design of microwave systems. In the past, the microwave equipment was made to conform to the frequency of the system and the design possibilities were limited. The unique properties of ferrites provide a variable reactance by which microwave energy can be manipulated to conform to the microwave system. At present, ferrites are used as LOAD ISOLATORS, PHASE SHIFTERS, VARIABLE ATTENUATORS, MODULATORS, and SWITCHES in microwave systems. The operation of ferrites as isolators, attenuators, and phase shifters will be explained in the following paragraphs. The operation of ferrites in other applications will be explained in later *NEETS* modules. Ferromagnetism is a continuation of the conventional domain theory of magnetism that was explained in *NEETS*, Module 1, *Matter, Energy, and Direct Current*. A review of the section on magnetism might be helpful to you at this point.

The magnetic property of any material is a result of electron movement within the atoms of the material. Electrons have two basic types of motion. The most familiar is the ORBITAL movement of the electron about the nucleus of the atom. Less familiar, but even more important, is the movement of the electron about its own axis, called ELECTRON SPIN.

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You will recall that magnetic fields are generated by current flow. Since current is the movement of electrons, the movement of the electrons within an atom create magnetic fields. The magnetic fields caused by the movement of the electrons about the nucleus have little effect on the magnetic properties of a material. The magnetic fields caused by electron spin combine to give a material magnetic properties. The different types of electron movement are illustrated in figure 1-73. In most materials the spin axes of the electrons are so randomly arranged that the magnetic fields largely cancel out and the material displays no significant magnetic properties. The electron spin axes within some materials, such as iron and nickel, can be caused to align by applying an external magnetic field. The alignment of the electrons within a material causes the magnetic fields to add, and the material then has magnetic properties.

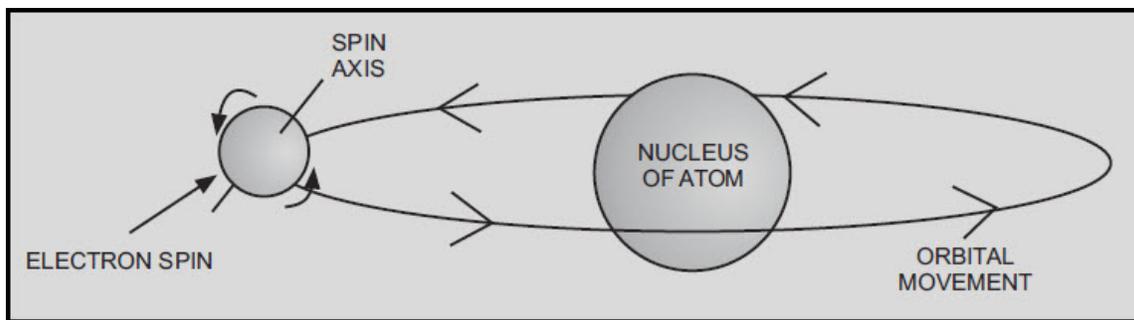


Figure 1-73 Two types of electron movement

In the absence of an external force, the axis of any spinning object tends to remain pointed in one direction. Spinning electrons behave the same way. Therefore, once the electrons are aligned, they tend to remain aligned even when the external field is removed. Electron alignment in a ferrite is caused by the orbital motion of the electrons about the nucleus and the force that holds the atom together. When a static magnetic field is applied, the electrons try to align their spin axes with the new force. The attempt of the electrons to balance between the interaction of the new force and the binding force causes the electrons to wobble on their axes, as shown in figure 1-74. The wobble of the electrons has a natural resonant **WOBBLE FREQUENCY** that varies with the strength of the applied field. Ferrite action is based on this behavior of the electrons under the influence of an external field and the resulting wobble frequency.

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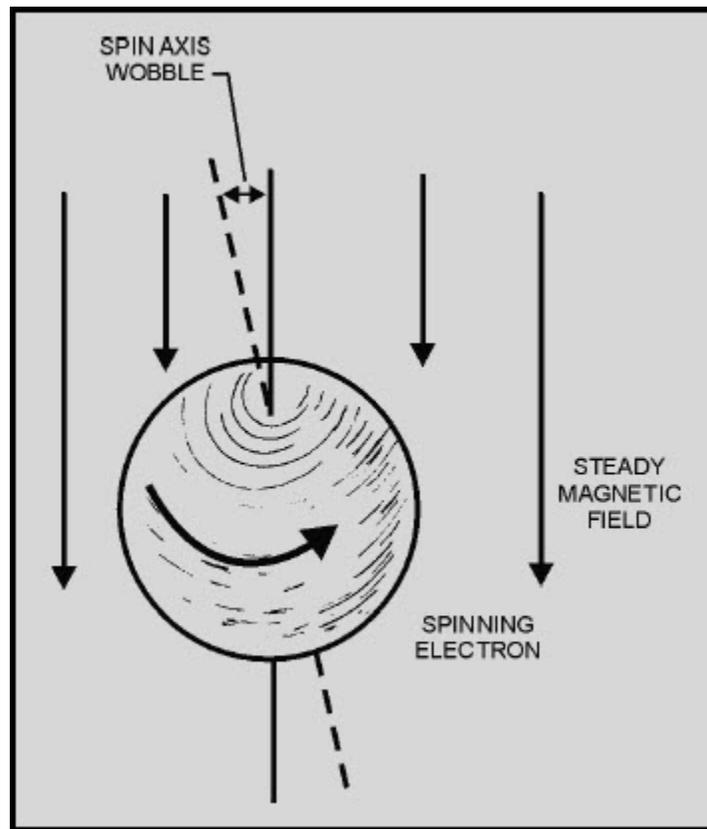


Figure 1-74 Electron wobble in a magnetic field

FERRITE ATTENUATORS - A ferrite attenuator can be constructed that will attenuate a particular microwave frequency and allow all others to pass unaffected. This can be done by placing a ferrite in the center of a waveguide, as shown in figure 1-72. The ferrite must be positioned so that the magnetic fields caused by its electrons are perpendicular to the energy in the waveguide. A steady external field causes the electrons to wobble at the same frequency as the energy that is to be attenuated.

Since the wobble frequency is the same as the energy frequency, the energy in the waveguide always adds to the wobble of the electrons. The spin axis of the electron changes direction during the wobble motion and energy is used. The force causing the increase in wobble is the energy in the waveguide. Thus, the energy in the waveguide is attenuated by the ferrite and is given off as heat. Energy in the waveguide that is a different frequency from the wobble frequency of the ferrite is largely unaffected because it does not increase the amount of electron wobble. The resonant frequency of electron wobble can be varied over a limited range by changing the strength of the applied magnetic field.

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FERRITE ISOLATORS - An isolator is a ferrite device that can be constructed so that it allows microwave energy to pass in one direction but blocks energy in the other direction in a waveguide. This isolator is constructed by placing a piece of ferrite off-center in a waveguide, as shown in figure 1-75. A magnetic field is applied by the magnet and adjusted to make the electron wobble frequency of the ferrite equal to the frequency of the energy traveling down the waveguide. Energy traveling down the waveguide from left to right will set up a rotating magnetic field that rotates through the ferrite material in the same direction as the natural wobble of the electrons. The aiding magnetic field increases the wobble of the ferrite electrons so much that almost all of the energy in the waveguide is absorbed and dissipated as heat. The magnetic fields caused by energy traveling from right to left rotate in the opposite direction through the ferrite and have very little effect on the amount of electron wobble. In this case the fields attempt to push the electrons in the direction opposite the natural wobble and no large movements occur. Since no overall energy exchange takes place, energy traveling from right to left is affected very little.

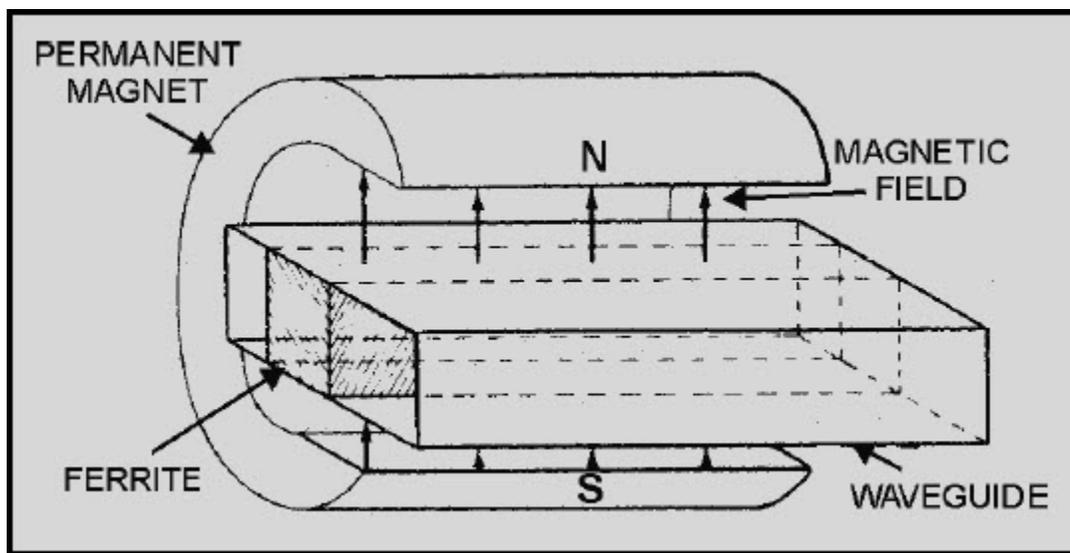


Figure 1-75 One-way isolator

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FERRITE PHASE SHIFTER - When microwave energy is passed through a piece of ferrite in a magnetic field, another effect occurs. If the frequency of the microwave energy is much greater than the electron wobble frequency, the plane of polarization of the wavefront is rotated. This is known as the FARADAY ROTATION EFFECT and is illustrated in figure 1-76. A ferrite rod is placed along the axis of the waveguide, and a magnetic field is set up along the axis by a coil. As a wavefront enters the section containing the ferrite, it sets up a limited motion in the electrons. The magnetic fields of the wavefront and the wobbling electrons interact, and the polarization of the wavefront is rotated. The amount of rotation depends upon the length of the ferrite rod. The direction of rotation depends upon the direction of the external magnetic field and can be reversed by reversing the field. The direction of rotation will remain constant, no matter what direction the energy in the waveguide travels, as long as the external field is not changed.

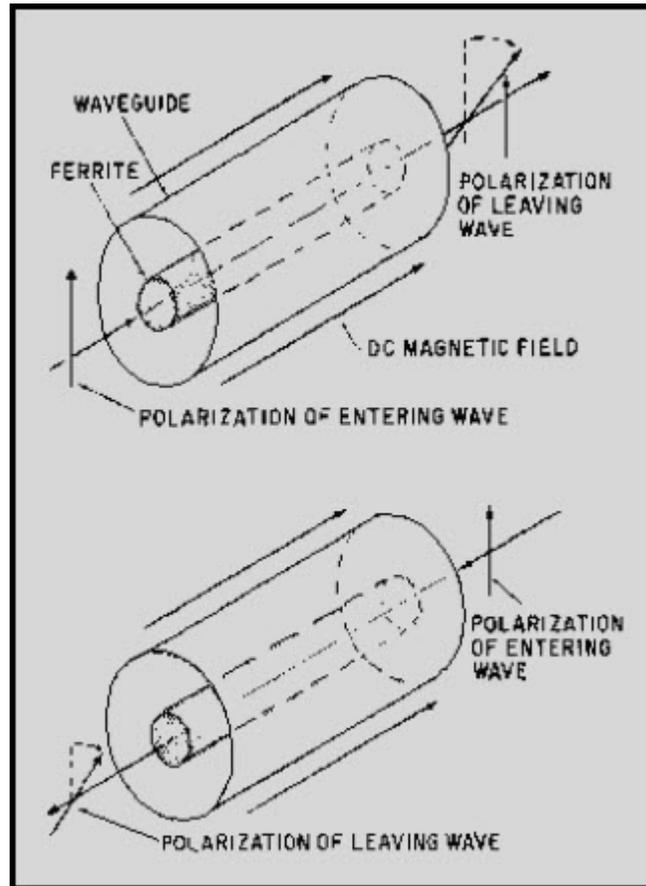


Figure 1-76 Faraday rotation

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Q-59. Ferrite devices are useful in microwave applications because they possess what properties?

Q-60. Which of the two types of electron motion (orbital movement and electron spin) is more important in the explanation of magnetism?

Q-61. The interaction between an external field and the binding force of an atom causes electrons to do what?

Q-62. The resonant frequency of electron wobble can be changed by variation of what force?

Q-63. Rotating the plane of polarization of a wavefront by passing it through a ferrite device is called what?

1.4 SUMMARY

This chapter has presented information on waveguide theory and application. The information that follows summarizes the important points of this chapter.

WAVEGUIDES are the primary methods of transporting microwave energy. Waveguides have fewer losses and greater power-handling capability than transmission lines. The physical size of the waveguides becomes too great for use at frequencies less than 1000 megahertz. Waveguides are made in three basic shapes, as shown in the first illustration. The wide, or "a," dimension determines the frequency range of the waveguide, and the narrow, or "b," dimension determines power-handling capability as shown in the second illustration. Waveguides handle a small range of frequencies both above and below the primary operating frequency. Energy is transported through waveguides by the interaction of electric and magnetic fields, abbreviated E FIELD and H FIELD, respectively. The density of the E field varies at the same rate as the applied voltage. If energy is to travel through a waveguide, two BOUNDARY CONDITIONS must be met: (1) An electric field, to exist at the surface of a conductor, must be perpendicular to the conductor, and (2) a varying magnetic field must exist in closed loops parallel to the conductors and perpendicular to the electric field.

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WAVEFRONTS travel down a waveguide by reflecting from the side walls in a zigzag pattern, as shown in the figure. The striking angle, or angle of incidence (θ), is the same as the angle of reflection (θ), causing the reflected wavefront to have the same shape as the incident wavefront. The velocity of wavefronts traveling down a waveguide is called the **GROUP VELOCITY** because of the zigzag path of these wavefronts. The group velocity is slower than the velocity of wavefronts through space.

The **MODES** in waveguides are divided into two categories: (1) the **TRANSVERSE ELECTRIC (TE)** mode and (2) the **TRANSVERSE MAGNETIC (TM)** mode. Subscripts are used to complete the description of the various TE and TM modes.

WAVEGUIDE INPUT/OUTPUT METHODS are divided into three basic categories: **PROBES**, **LOOPS**, and **SLOTS**. Size, shape, and placement in the waveguide are critical factors in the efficiency of all three input/output methods.

WAVEGUIDE/IMPEDANCE MATCHING is often necessary to reduce reflections caused by a **MISMATCH** between the waveguide and the load. Matching devices called **IRISES**, shown in the illustration, are used to introduce either capacitance or inductance (or a combination of both) into a waveguide. **Conductive POSTS** and **SCREWS** can also be used for impedance matching in waveguides.

WAVEGUIDE TERMINATIONS prevent standing waves at the end of a waveguide system. They are usually specially constructed **HORNS** or absorptive loads called **DUMMY LOADS**.

WAVEGUIDE PLUMBING refers to the bends, twists, and joints necessary to install waveguides. E bends, H bends, and twists must have a radius greater than two wavelengths. The **CHOKE JOINT**, shown in the figure, is most often used to connect two pieces of waveguide. The **ROTATING JOINT** is used when a waveguide must be connected to a rotating load such as an antenna.

DIRECTIONAL COUPLERS are devices that permit the sampling of the energy in a waveguide. Directional couplers may be constructed to sample energy in one direction only or in both directions. The energy removed by the directional coupler is a small sample that is proportional to the magnitude of the energy in the waveguide. An example of a directional coupler is shown in the illustration.

A **RESONANT CAVITY** is any space completely enclosed by conductive walls that can contain oscillating electromagnetic fields and can possess resonant properties. Several cavity shapes are shown in the illustration

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WAVEGUIDE JUNCTIONS are of several basic types. The T-JUNCTION may be either of the E-TYPE or the H-TYPE. The effect on the input energy depends upon which arm is used as the input. The MAGIC-T HYBRID JUNCTION, shown at the right, is a combination of the E- and H-type T junctions.

FERRITE DEVICES combine magnetic properties with a high resistance to current flow. Ferrites are constructed from compounds of ferrous metal oxides to achieve the desired characteristics. The fact that the spin axes of electrons will wobble at a natural resonant frequency when subjected to an external magnetic field is the basic principle of operation of ferrite devices.

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ANSWERS TO QUESTIONS Q1. THROUGH Q63.

- A-1. *Microwave region.*
- A-2. *Electromagnetic field theory.*
- A-3. *The electromagnetic fields are completely confined.*
- A-4. *Conductive material.*
- A-5. *Copper loss.*
- A-6. *Skin effect.*
- A-7. *Air.*
- A-8. *Physical size.*
- A-9. *The characteristics of the dielectric of a capacitor.*
- A-10. *A shorted quarter-wave section called a metallic insulator.*
- A-11. *The "a" dimension.*
- A-12. *The bus bar becomes wider.*
- A-13. *Energy will no longer pass through the waveguide.*
- A-14. *The interaction of the electric and magnetic fields.*
- A-15. *The relative strength of the field.*
- A-16. *Magnetic lines of force must form a continuous closed loop.*
- A-17. *The H lines cancel.*
- A-18. *The field must be perpendicular to the conductors.*
- A-19. *Decrease to zero.*
- A-20. *The angles are equal.*
- A-21. *Cutoff frequency.*
- A-22. *Slower.*
- A-23. *Group velocity.*
- A-24. *Mode of operation.*
- A-25. *Dominant mode.*
- A-26. *1.71 times the diameter.*
- A-27. *Transverse electric (TE) and transverse magnetic (TM).*
- A-28. *TE.*
- A-29. *Second.*
- A-30. *First.*
- A-31. *Size and shape.*
- A-32. *Slots and apertures.*
- A-33. *Standing waves that cause power losses, a reduction in power-handling capability, and an increase in frequency and sensitivity.*
- A-34. *Metal plates.*
- A-35. *Inductive.*
- A-36. *As a shunt resistance.*
- A-37. *Horn.*
- A-38. *Characteristic impedance.*
- A-39. *Absorb all energy without producing standing waves.*
- A-40. *Heat.*

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- A-41. Reflections.
- A-42. Greater than 2 wavelengths.
- A-43. Choke joint.
- A-44. Improperly connected joints or damaged inner surface.
- A-45. Sampling energy within a waveguide.
- A-46. 1/4 wavelength.
- A-47. Absorb the energy not directed at the pick-up probe and a portion of the overall energy.
- A-48. The wavefront portions add.
- A-49. The reflected energy adds at the absorbent material and is absorbed.
- A-50. Size and shape of the cavity.
- A-51. Probes, loops, and slots.
- A-52. The area of maximum H lines.
- A-53. E-type and H-type.
- A-54. The junction arm extends in a direction parallel to the H lines in the main waveguide.
- A-55. E-type and H-type.
- A-56. Low power-handling capability and power losses.
- A-57. Basic E-type junctions.
- A-58. High-power duplexes.
- A-59. Magnetic properties and high resistance.
- A-60. Electron spin.
- A-61. Wobble at a natural resonant frequency.
- A-62. The applied magnetic field.
- A-63. Faraday rotation.

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2 MICROWAVE COMPONENTS AND CIRCUITS

LEARNING OBJECTIVES

Upon completion of this chapter the student will be able to:

1. Explain the basic principles of microwave tubes and describe the limitations of conventional tubes.
2. Describe the basic principles of velocity modulation.
3. Outline the development of microwave tubes.
4. Describe the basic theory of operation of klystrons including multi-cavity and reflex klystrons.
5. Explain the basic theory of operation of traveling-wave tubes and backward-wave oscillators.
6. Describe the construction, basic theories of operation, and typical applications of magnetrons and amplitrons.
7. Describe the basic theory of operation of tunnel diodes when used in oscillator-, amplifier-, and frequency-converter circuits.
8. Explain the operation of varactors when used in parametric amplifiers and frequency converters.
9. State the basic principles of operation of bulk-effect diodes and the gunn oscillator.
10. Explain the basic operation of passive microwave diodes in terms of theory and application.
11. Explain the basic operation of microwave transistors in terms of theory and application.

2.1 MICROWAVE COMPONENTS

The waveguides discussed in chapter 1 serve to transport microwave energy from one place to another. Energy is transported after it has been generated or amplified in a previous stage of the circuit. In this chapter you will be introduced to the special components used in those circuits.

Microwave energy is used in both radar and communications applications. The fact that the frequencies are very high and the wavelengths very short presents special problems in circuit design. Components that were previously satisfactory for signal generation and amplification use are no longer useful in the microwave region. The theory of operation for these components is discussed in this chapter. Because the theory of operation is sometimes difficult to understand, you need to pay particular attention to detail as you study this chapter. It is written in the simplest manner possible while retaining the necessary technical complexity.

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2.2 MICROWAVE TUBE PRINCIPLES

The efficiency of conventional tubes is largely independent of frequency up to a certain limit. When frequency increases beyond that limit, several factors combine to rapidly decrease tube efficiency. Tubes that are efficient in the microwave range usually operate on the theory of VELOCITY MODULATION, a concept that avoids the problems encountered in conventional tubes. Velocity modulation is more easily understood if the factors that limit the frequency range of a conventional tube are thoroughly understood. Therefore, the frequency limitations of conventional tubes will be discussed before the concepts and applications of velocity modulation are explained. You may want to review NEETS, Module 6, Introduction to Electronic Emission, Tubes, and Power Supplies, Chapters 1 and 2, for a refresher on vacuum tubes before proceeding.

2.2.1 Frequency Limitations of Conventional Tubes

Three characteristics of ordinary vacuum tubes become increasingly important as frequency rises. These characteristics are interelectrode capacitance, lead inductance, and electron transit time.

The INTERELECTRODE CAPACITANCES in a vacuum tube, at low or medium radio frequencies, produce capacitive reactances that are so large that no serious effects upon tube operation are noticeable. However, as the frequency increases, the reactances become small enough to materially affect the performance of a circuit. For example, in figure 2-1A, a 1-picofarad capacitor has a reactance of 159,000 ohms at 1 megahertz. If this capacitor was the interelectrode capacitance between the grid and plate of a tube, and the rf voltage between these electrodes was 500 volts, then 3.15 milliamperes of current would flow through the interelectrode capacitance. Current flow in this small amount would not seriously affect circuit performance. On the other hand, at a frequency of 100 megahertz the reactance would decrease to approximately 1,590 ohms and, with the same voltage applied, current would increase to 315 milliamperes (figure 2-1B). Current in this amount would definitely affect circuit performance.

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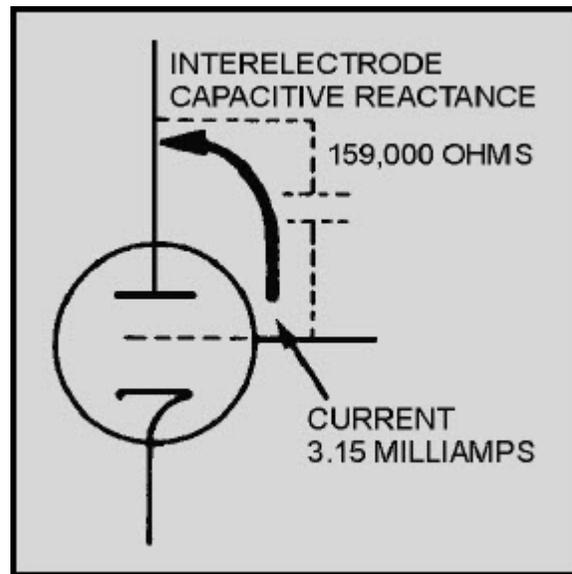


Figure 2-1A Interelectrode capacitance in a vacuum tube. 1 MEGAHERTZ

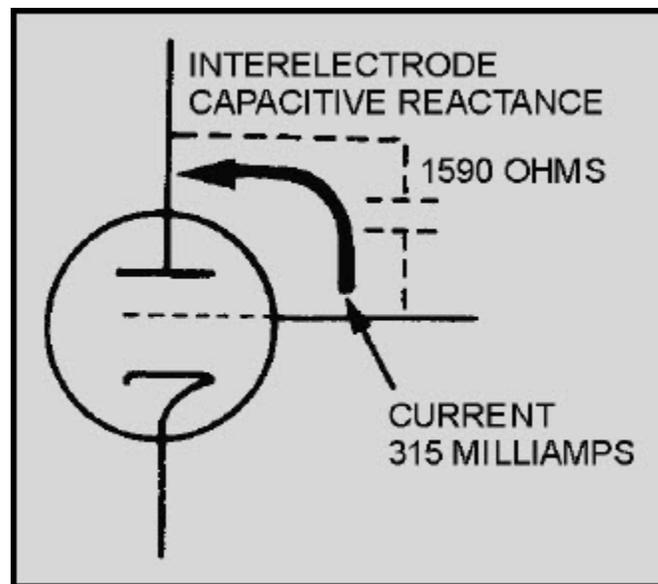


Figure 2-1B Interelectrode capacitance in a vacuum tube. 100 MEGAHERTZ

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A good point to remember is that the higher the frequency, or the larger the interelectrode capacitance, the higher will be the current through this capacitance. The circuit in figure 2-1C, shows the interelectrode capacitance between the grid and the cathode (C_{gk}) in parallel with the signal source. As the frequency of the input signal increases, the effective grid-to-cathode impedance of the tube decreases because of a decrease in the reactance of the interelectrode capacitance. If the signal frequency is 100 megahertz or greater, the reactance of the grid-to-cathode capacitance is so small that much of the signal is short-circuited within the tube. Since the interelectrode capacitances are effectively in parallel with the tuned circuits, as shown in figures 2-1A, B, and C, they will also affect the frequency at which the tuned circuits resonate.

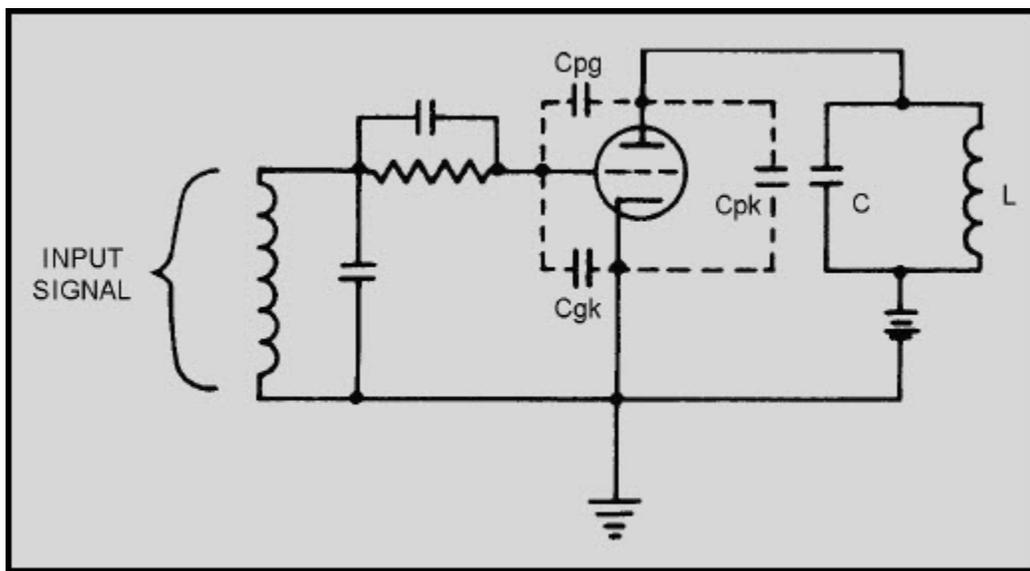


Figure 2-1C Interelectrode capacitance in a vacuum tube. INTERELECTRODE CAPACITANCE IN A TUNED-PLATE TUNED-GRID OSCILLATOR

Another frequency-limiting factor is the LEAD INDUCTANCE of the tube elements. Since the lead inductances within a tube are effectively in parallel with the interelectrode capacitance, the net effect is to raise the frequency limit. However, the inductance of the cathode lead is common to both the grid and plate circuits. This provides a path for degenerative feedback which reduces overall circuit efficiency.

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A third limitation caused by tube construction is TRANSIT TIME. Transit time is the time required for electrons to travel from the cathode to the plate. While some small amount of transit time is required for electrons to travel from the cathode to the plate, the time is insignificant at low frequencies. In fact, the transit time is so insignificant at low frequencies that it is generally not considered to be a hindering factor. However, at high frequencies, transit time becomes an appreciable portion of a signal cycle and begins to hinder efficiency. For example, a transit time of 1 nanosecond, which is not unusual, is only 0.001 cycle at a frequency of 1 megahertz. The same transit time becomes equal to the time required for an entire cycle at 1,000 megahertz. Transit time depends on electrode spacing and existing voltage potentials. Transit times in excess of 0.1 cycle cause a significant decrease in tube efficiency. This decrease in efficiency is caused, in part, by a phase shift between plate current and grid voltage.

If the tube is to operate efficiently, the plate current must be in phase with the grid-signal voltage and 180 degrees out of phase with the plate voltage. When transit time approaches 1/4 cycle, this phase relationship between the elements does not hold true. A positive swing of a high-frequency grid signal causes electrons to leave the cathode and flow to the plate. Initially this current is in phase with the grid voltage. However, since transit time is an appreciable part of a cycle, the current arriving at the plate now lags the grid-signal voltage. As a result, the power output of the tube decreases and the plate power dissipation increases. Another loss of power occurs because of ELECTROSTATIC INDUCTION.

The electrons forming the plate current also electrostatically induce potentials in the grid as they move past it. This electrostatic induction in the grid causes currents of positive charges to move back and forth in the grid structure. This back and forth action is similar to the action of hole current in semiconductor devices. When transit-time effect is not a factor (as in low frequencies), the current induced in one side of the grid by the approaching electrons is equal to the current induced on the other side by the receding electrons. The net effect is zero since the currents are in opposite directions and cancel each other. However, when transit time is an appreciable part of a cycle, the number of electrons approaching the grid is not always equal to the number going away. As a result, the induced currents do not cancel. This uncancelled current produces a power loss in the grid that is considered resistive in nature. In other words, the tube acts as if a resistor were connected between the grid and the cathode. The resistance of this imaginary resistor decreases rapidly as the frequency increases. The resistance may become so low that the grid is essentially short-circuited to the cathode, preventing proper operation of the tube.

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Several methods are available to reduce the limitations of conventional tubes, but none work well when frequency increases beyond 1,000 megahertz. Interelectrode capacitance can be reduced by moving the electrodes further apart or by reducing the size of the tube and its electrodes. Moving the electrodes apart increases the problems associated with transit time, and reducing the size of the tube lowers the power-handling capability. You can see that efforts to reduce certain limitations in conventional tubes are compromises that are often in direct opposition to each other. The net effect is an upper limit of approximately 1,000 megahertz, beyond which conventional tubes are not practical.

Q-1. What happens to the impedance of interelectrode capacitance as frequency increases?

Q-2. What undesirable effect is caused by the inductance of the cathode lead?

Q-3. How does transit time affect the relationship of the grid voltage and the plate current at high frequencies?

Q-4. Moving tube electrodes apart to decrease interelectrode capacitance causes an increase in the effect of what property?

2.2.2 Velocity Modulation

The microwave tube was developed when the use of the frequency spectrum went beyond 1,000 megahertz and into the microwave range. The microwave tube uses transit time in the conversion of dc power to radio-frequency (rf) power. The interchange of power is accomplished by using the principle of electron VELOCITY MODULATION and low-loss resonant cavities in the microwave tube.

A clear understanding of microwave tubes must start with an understanding of how electrons and electric fields interact. An electron has mass and thus exhibits kinetic energy when in motion. The amount of kinetic energy in an electron is directly proportional to its velocity; that is, the higher the velocity, the higher the energy level. The basic concept of the electron energy level being directly related to electron velocity is the key principle of energy transfer and amplification in microwave tubes.

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An electron can be accelerated or decelerated by an electrostatic field. Figure 2-2 shows an electron moving in an electrostatic field. The direction of travel (shown by the heavy arrow) is against the electrostatic lines of force which are from positive to negative. The negatively charged electron will be attracted to the positively charged body and will increase in velocity. As its velocity increases, the energy level of the electron will also increase. Where does the electron acquire its additional energy? The only logical source is from the electrostatic field. Thus, the conclusion is clear. An electron traveling in a direction opposite to electrostatic lines of force will absorb energy and increase in velocity (accelerate).

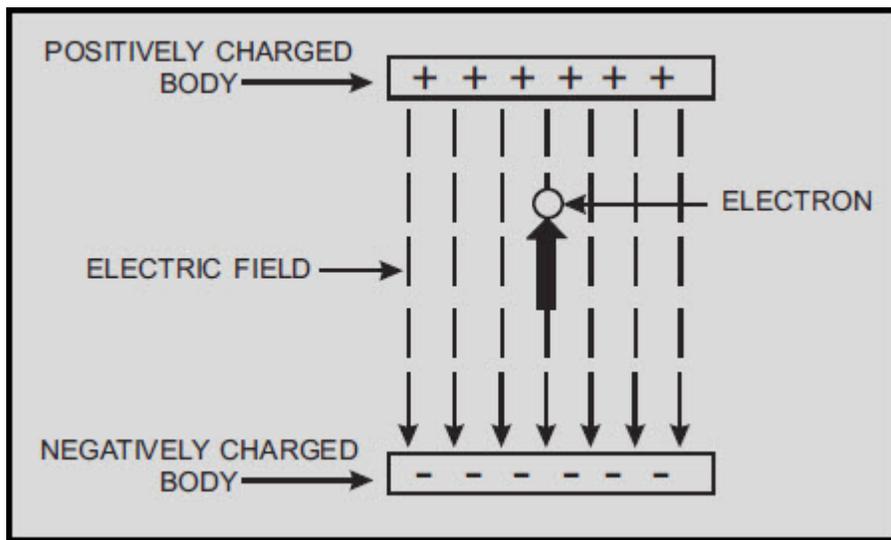


Figure 2-2 Moving electron gaining velocity and energy

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As figure 2-3 illustrates, the opposite condition is also true. An electron traveling in the same direction as the electrostatic lines of force will decelerate by giving up energy to the field. The negatively charged body will repel the electron and cause it to decrease in velocity. When the velocity is reduced, the energy level is also reduced. The energy lost by the electron is gained by the electrostatic field.

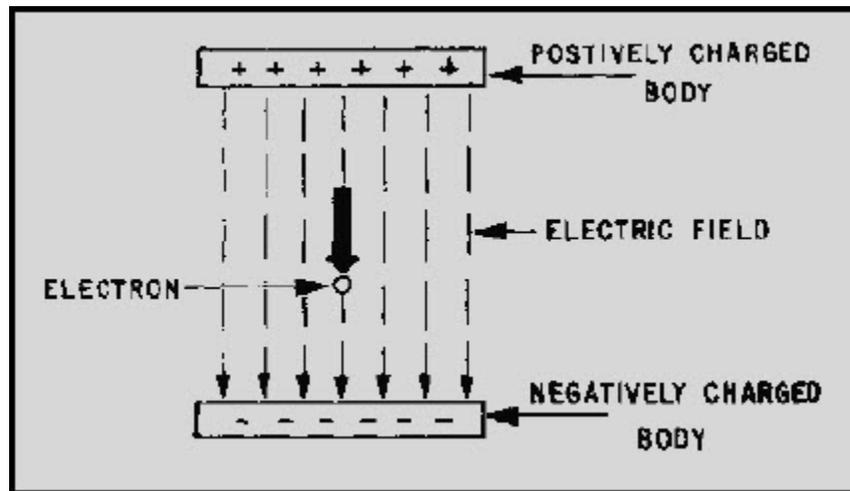


Figure 2-3 Moving electron losing energy and velocity

The operation of a velocity-modulated tube depends on a change in the velocity of the electrons passing through its electrostatic field. A change in electron velocity causes the tube to produce BUNCHES of electrons. These bunches are separated by spaces in which there are relatively few electrons. Velocity modulation is then defined as that variation in the velocity of a beam of electrons caused by the alternate speeding up and slowing down of the electrons in the beam. This variation is usually caused by a voltage signal applied between the grids through which the beam must pass.

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The first requirement in obtaining velocity modulation is to produce a stream of electrons which are all traveling at the same speed. The electron stream is produced by an electron gun. A simplified version of an electron gun is shown in figure 2-4A. Electrons emitted from the cathode are attracted toward the positive accelerator grid and all but a few of the electrons pass through the grid and form a beam. The electron beam then passes through a pair of closely spaced grids, called BUNCHER GRIDS. Each grid is connected to one side of a tuned circuit. The parallel-resonant tuned circuit (figure 2-4A) in the illustration represents the doughnut-shaped resonant cavity surrounding the electron stream (figure 2-4B). The buncher grids are the dashed lines at the center of the cavity and are at the same dc potential as the accelerator grid. The alternating voltage which exists across the resonant circuit causes the velocity of the electrons leaving the buncher grids to differ from the velocity of the electrons arriving at the buncher grids. The amount of difference depends on the strength and direction of the electrostatic field within the resonant cavity as the electrons pass through the grids.

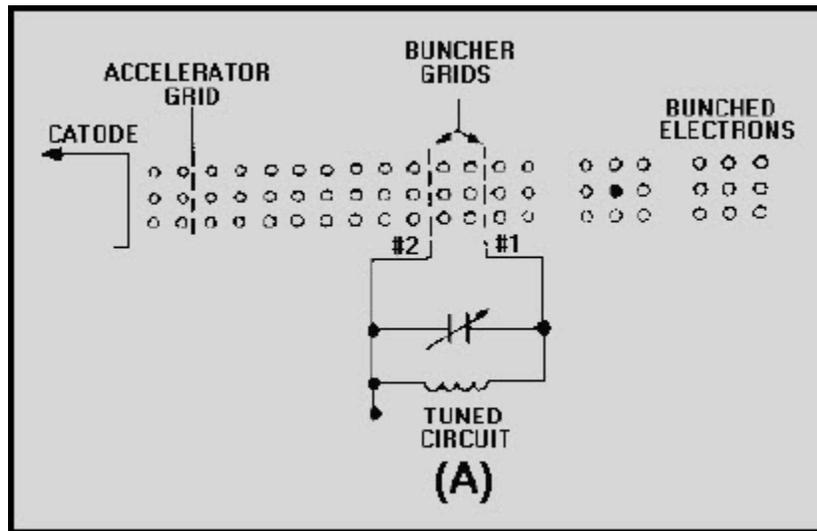


Figure 2-4A Electron gun with buncher grids

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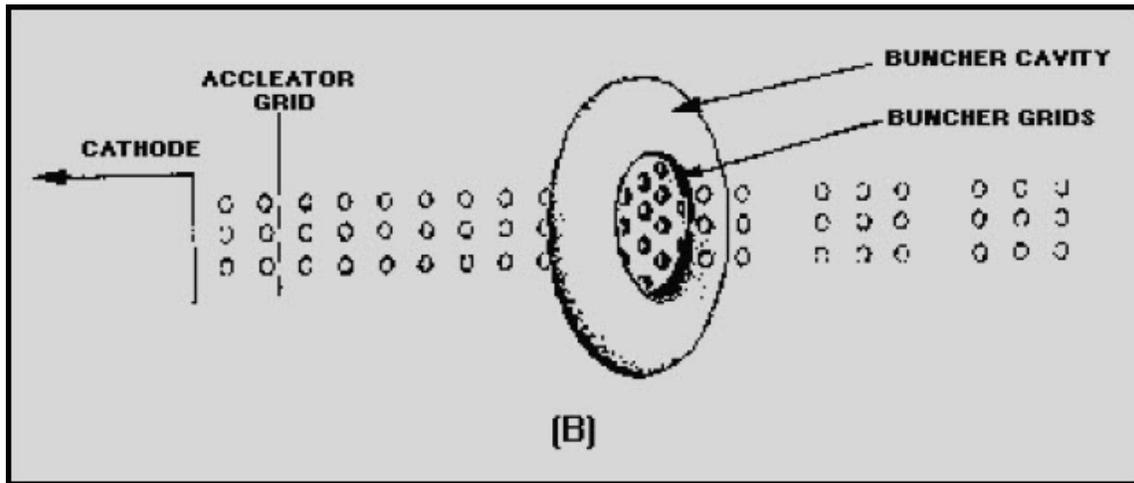


Figure 2-4B Electron gun with buncher grids

The manner in which the buncher produces bunches of electrons is better understood by considering the motions of individual electrons, as illustrated in figure 2-5A.

When the voltage across the grids is negative, as shown in figure 2-5B, electron 1 crossing the gap at that time is slowed. Figure 2-5C shows the potential across the gap at 0 volts; electron 2 is not affected. Electron 3 enters the gap (figure 2-5D) when the voltage across the gap is positive and its velocity is increased. The combined effect is shown in figure 2-5E. All of the electrons in the group have been bunched closer together.

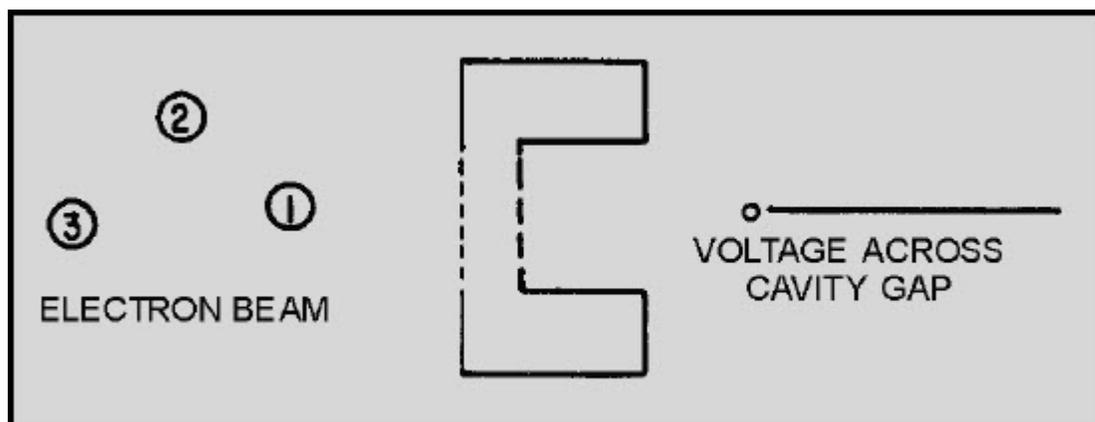


Figure 2-5A Buncher cavity action. BUNCHER CAVITY

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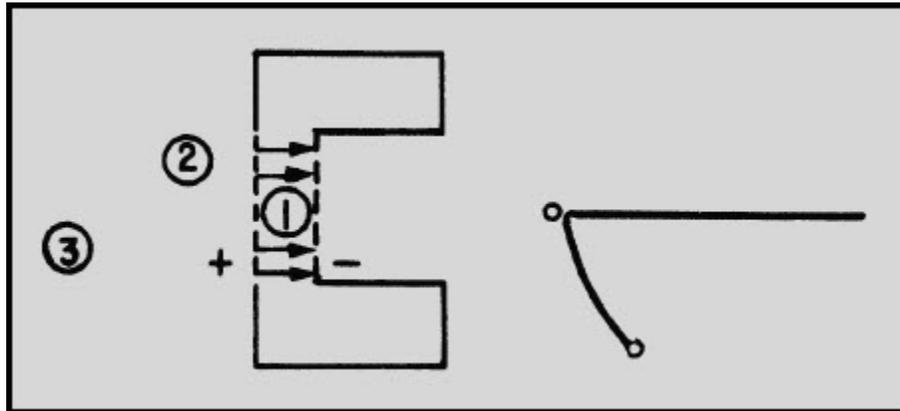


Figure 2-5B Buncher cavity action. ELECTRON #1 DECELERATED

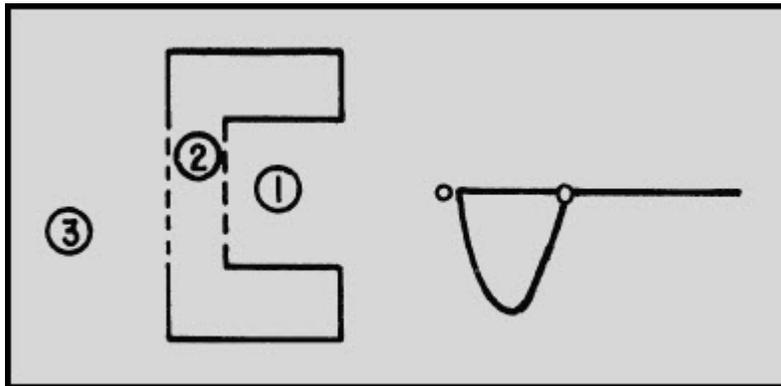


Figure 2-5C Buncher cavity action. ELECTRON #2 VELOCITY

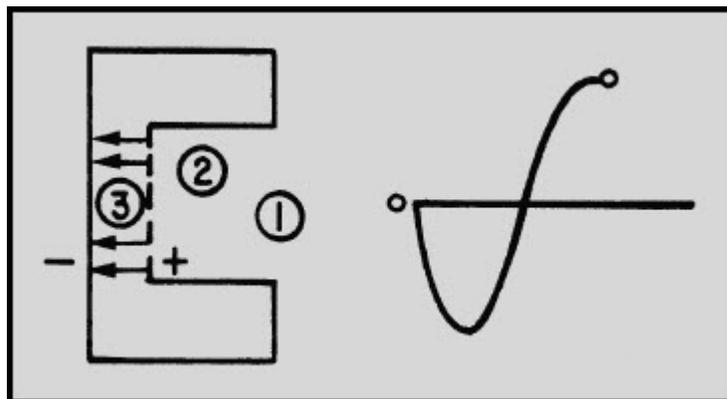
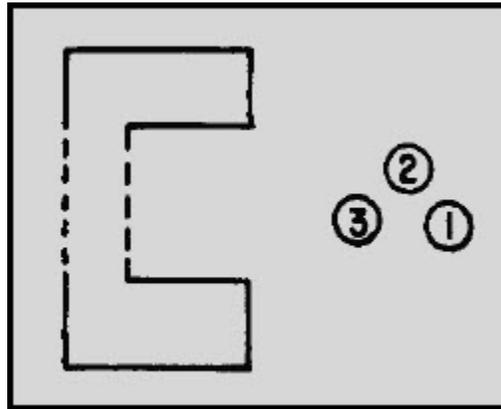


Figure 2-5D Buncher cavity action. ELECTRON #3
ACCELERATED

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**Figure 2-5E Buncher cavity action.
ELECTRONS BEGINNING TO
BUNCH, DUE TO VELOCITY
DIFFERENCES**

The velocity modulation of the beam is merely a means to an end. No useful power has been produced at this point. The energy gained by the accelerated electrons is balanced by the energy lost by the decelerated electrons. However, a new and useful beam distribution will be formed if the velocity-modulated electrons are allowed to drift into an area that has no electrostatic field.

As the electrons drift into the field-free area beyond the buncher cavity, bunches continue to form because of the new velocity relationships between the electrons. Unless the beam is acted upon by some other force, these bunches will tend to form and disperse until the original beam distribution is eventually reformed. The net effect of velocity modulation is to form a current-density modulated beam that varies at the same rate as the grid-signal frequency. The next step is to take useful power from the beam.

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The current-modulated (bunched) electron beam in figure 2-6A and B is shown in various stages of formation and dispersion. A second cavity, called a CATCHER CAVITY, must be placed at a point of maximum bunching to take useful energy from the beam (shown in figure 2-6B). The physical position of the catcher cavity is determined by the frequency of the buncher-grid signal because this signal determines the transit time of the electron bunches. Note also that both cavities are resonant at the buncher-grid frequency. The electron bunches will induce an rf voltage in the grid gap of the second cavity causing it to oscillate. Proper placement of the second cavity will cause the induced grid-gap voltage to decelerate the electron bunches as they arrive at the gap. Since the largest concentration of electrons is in the bunches, slowing the bunches causes a transfer of energy to the output cavity. The balance of energy has been disturbed because the placement of the catcher cavity is such that bunches are slowed down when they arrive at the cavity. The areas between bunches arrive at the cavity at just the right time. At this time the voltage is of the correct polarity to increase the velocity of the electrons and the beam absorbs energy. The areas between the bunches have very few electrons, so the energy removed from the beam is much greater than the energy required to speed up the electrons between the bunches. Therefore, if the second cavity is properly positioned, useful energy can be removed from a velocity-modulated electron beam.

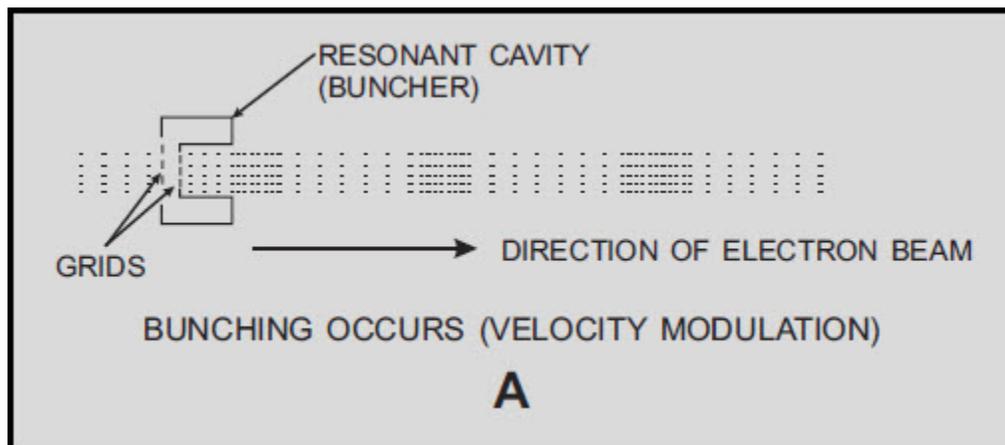


Figure 2-6A Removing energy from a velocity-modulated beam

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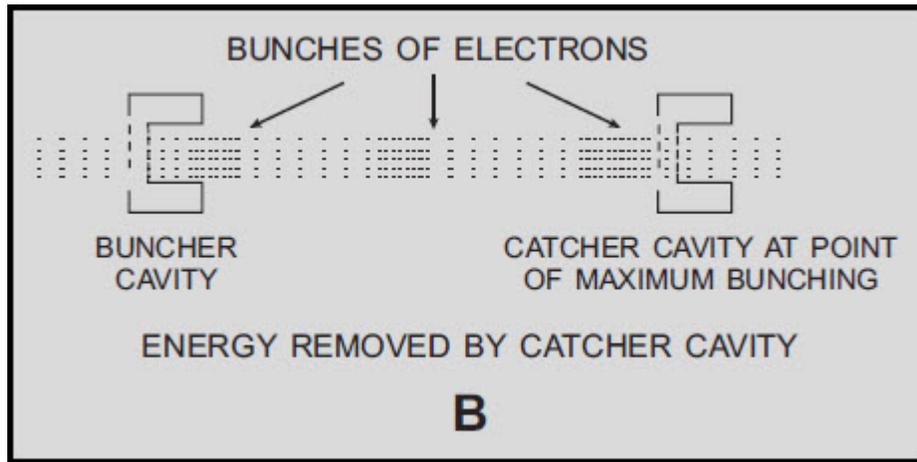


Figure 2-6B Removing energy from a velocity-modulated beam

Q-5. The kinetic energy of an electron is directly proportional to what property?

Q-6. What will be the effect upon an electron traveling in the opposite direction to the lines of force in an electrostatic field?

Q-7. How is a beam of electrons velocity-modulated?

Q-8. What portion of an electron gun causes the electrons to accelerate or decelerate?

Q-9. What is the effect upon an electron that enters the buncher gap when the potential across the grids is at 0 volts?

Q-10. What determines the placement of the catcher cavity?

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2.3 MICROWAVE TUBES

Microwave tubes perform the same functions of generation and amplification in the microwave portion of the frequency spectrum that vacuum tubes perform at lower frequencies. This section will explain the basic operation of the most widely used microwave tubes, including klystrons, traveling-wave tubes, backward-wave oscillators, magnetrons, and crossed-field amplifiers. The variations of these tubes for use in specific applications are so numerous that all of them cannot be discussed in this module. However, general principles of operation are similar in all of the variations so the explanations will be restricted to the general principles of operation.

2.3.1 The Basic Two-Cavity Klystron

Klystrons are velocity-modulated tubes that are used in radar and communications equipment as oscillators and amplifiers. Klystrons make use of the transit-time effect by varying the velocity of an electron beam in much the same manner as the previously discussed velocity-modulation process. Strong electrostatic fields are necessary in the klystron for efficient operation. This is necessary because the interaction of the signal and the electron beam takes place in a very short distance.

The construction and essential components of a TWO-CAVITY KLYSTRON are shown in figure 2-7A. Figure 2-7B is a schematic representation of the same tube. When the tube is energized, the cathode emits electrons which are focused into a beam by a low positive voltage on the control grid. The beam is then accelerated by a very high positive dc potential that is applied in equal amplitude to both the accelerator grid and the buncher grids. The buncher grids are connected to a cavity resonator that superimposes an ac potential on the dc voltage. Ac potentials are produced by oscillations within the cavity that begin spontaneously when the tube is energized. The initial oscillations are caused by random fields and circuit imbalances that are present when the circuit is energized. The oscillations within the cavity produce an oscillating electrostatic field between the buncher grids that is at the same frequency as the natural frequency of the cavity. The direction of the field changes with the frequency of the cavity. These changes alternately accelerate and decelerate the electrons of the beam passing through the grids. The area beyond the buncher grids is called the DRIFT SPACE. The electrons form bunches in this area when the accelerated electrons overtake the decelerated electrons.

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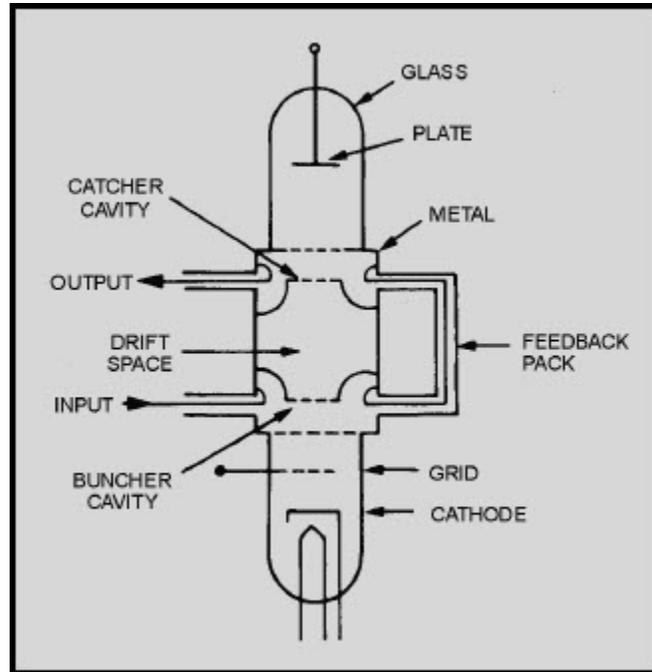


Figure 2-7A Functional and schematic diagram of a two-cavity klystron

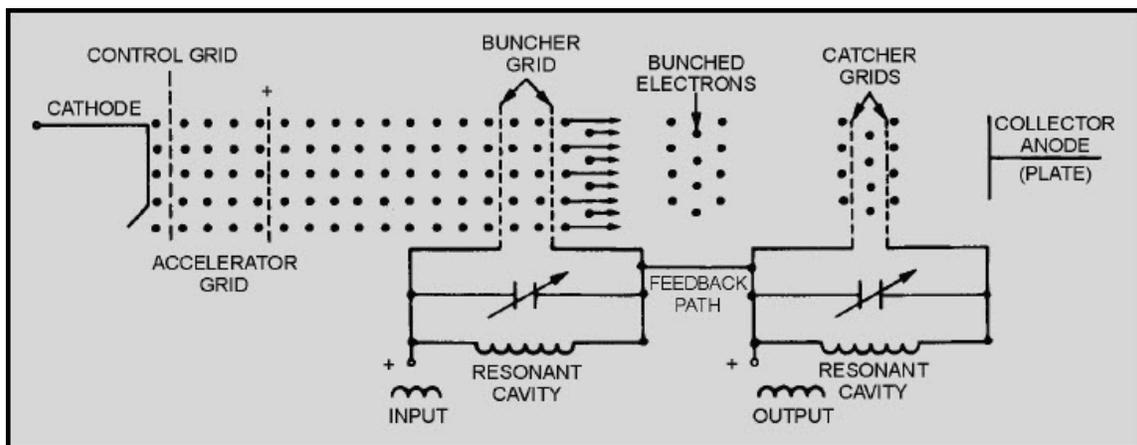


Figure 2-7B Functional and schematic diagram of a two-cavity klystron

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The function of the CATCHER GRIDS is to absorb energy from the electron beam. The catcher grids are placed along the beam at a point where the bunches are fully formed. The location is determined by the transit time of the bunches at the natural resonant frequency of the cavities (the resonant frequency of the catcher cavity is the same as the buncher cavity). The location is chosen because maximum energy transfer to the output (catcher) cavity occurs when the electrostatic field is of the correct polarity to slow down the electron bunches.

The two-cavity klystron in figure 2-7A and B may be used either as an oscillator or an amplifier. The configuration shown in the figure is correct for oscillator operation. The feedback path provides energy of the proper delay and phase relationship to sustain oscillations. A signal applied at the buncher grids will be amplified if the feedback path is removed.

Q-11. What is the basic principle of operation of a klystron?

Q-12. The electrons in the beam of a klystron are speeded up by a high dc potential applied to what elements?

Q-13. The two-cavity klystron uses what cavity as an output cavity?

Q-14. A two-cavity klystron without a feedback path will operate as what type of circuit?

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2.3.2 The Multicavity Power Klystron

Klystron amplification, power output, and efficiency can be greatly improved by the addition of intermediate cavities between the input and output cavities of the basic klystron. Additional cavities serve to velocity-modulate the electron beam and produce an increase in the energy available at the output. Since all intermediate cavities in a multicavity klystron operate in the same manner, a representative THREE-CAVITY KLYSTRON will be discussed.

A three-cavity klystron is illustrated in figure 2-8. The entire DRIFT-TUBE ASSEMBLY, the three CAVITIES, and the COLLECTOR PLATE of the three-cavity klystron are operated at ground potential for reasons of safety. The electron beam is formed and accelerated toward the drift tube by a large negative pulse applied to the cathode. MAGNETIC FOCUS COILS are placed around the drift tube to keep the electrons in a tight beam and away from the side walls of the tube. The focus of the beam is also aided by the concave shape of the cathode in high-powered klystrons.

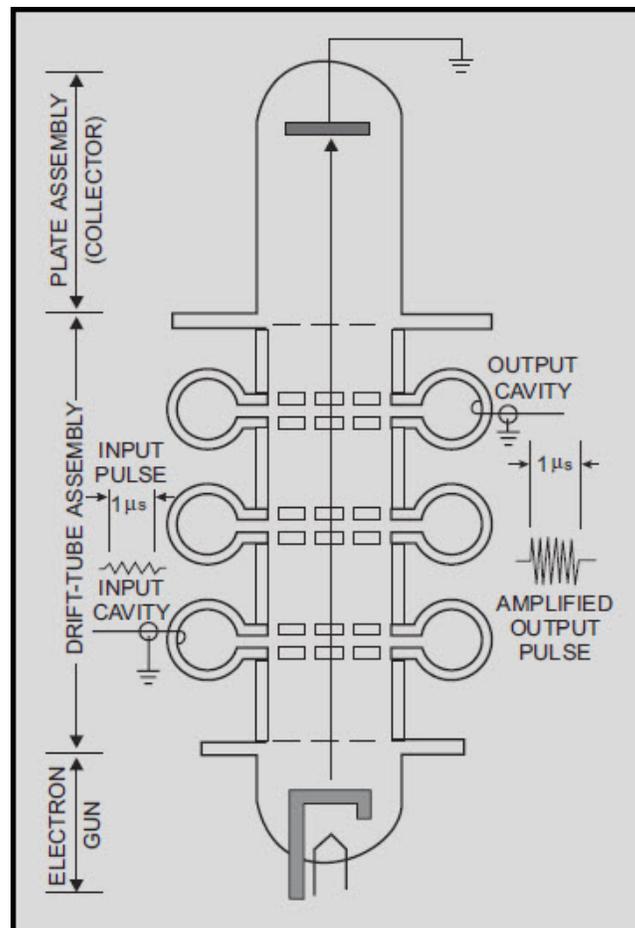


Figure 2-8 Three-cavity klystron

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The output of any klystron (regardless of the number of cavities used) is developed by velocity modulation of the electron beam. The electrons that are accelerated by the cathode pulse are acted upon by rf fields developed across the input and middle cavities. Some electrons are accelerated, some are decelerated, and some are unaffected. Electron reaction depends on the amplitude and polarity of the fields across the cavities when the electrons pass the cavity gaps. During the time the electrons are traveling through the drift space between the cavities, the accelerated electrons overtake the decelerated electrons to form bunches. As a result, bunches of electrons arrive at the output cavity at the proper instant during each cycle of the rf field and deliver energy to the output cavity.

Only a small degree of bunching takes place within the electron beam during the interval of travel from the input cavity to the middle cavity. The amount of bunching is sufficient, however, to cause oscillations within the middle cavity and to maintain a large oscillating voltage across the input gap. Most of the velocity modulation produced in the three-cavity klystron is caused by the voltage across the input gap of the middle cavity. The high voltage across the gap causes the bunching process to proceed rapidly in the drift space between the middle cavity and the output cavity. The electron bunches cross the gap of the output cavity when the gap voltage is at maximum negative. Maximum energy transfer from the electron beam to the output cavity occurs under these conditions. The energy given up by the electrons is the kinetic energy that was originally absorbed from the cathode pulse.

Klystron amplifiers have been built with as many as five intermediate cavities in addition to the input and output cavities. The effect of the intermediate cavities is to improve the electron bunching process which improves amplifier gain. The overall efficiency of the tube is also improved to a lesser extent. Adding more cavities is roughly the same as adding more stages to a conventional amplifier. The overall amplifier gain is increased and the overall bandwidth is reduced if all the stages are tuned to the same frequency. The same effect occurs with multicavity klystron tuning. A klystron amplifier tube will deliver high gain and a narrow bandwidth if all the cavities are tuned to the same frequency. This method of tuning is called **SYNCHRONOUS TUNING**. If the cavities are tuned to slightly different frequencies, the gain of the amplifier will be reduced but the bandwidth will be appreciably increased. This method of tuning is called **STAGGERED TUNING**.

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Q-15. What can be added to the basic two-cavity klystron to increase the amount of velocity modulation and the power output?

Q-16. How is the electron beam of a three-cavity klystron accelerated toward the drift tube?

Q-17. Which cavity of a three-cavity klystron causes most of the velocity modulation?

Q-18. In a multicavity klystron, tuning all the cavities to the same frequency has what effect on the bandwidth of the tube?

Q-19. The cavities of a multicavity klystron are tuned to slightly different frequencies in what method of tuning?

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2.3.3 The Reflex Klystron

Another tube based on velocity modulation, and used to generate microwave energy, is the REFLEX KLYSTRON (figure 2-9). The reflex klystron contains a REFLECTOR PLATE, referred to as the REPELLER, instead of the output cavity used in other types of klystrons. The electron beam is modulated as it was in the other types of klystrons by passing it through an oscillating resonant cavity, but here the similarity ends. The feedback required to maintain oscillations within the cavity is obtained by reversing the beam and sending it back through the cavity. The electrons in the beam are velocity-modulated before the beam passes through the cavity the second time and will give up the energy required to maintain oscillations. The electron beam is turned around by a negatively charged electrode that repels the beam. This negative element is the repeller mentioned earlier. This type of klystron oscillator is called a reflex klystron because of the reflex action of the electron beam.

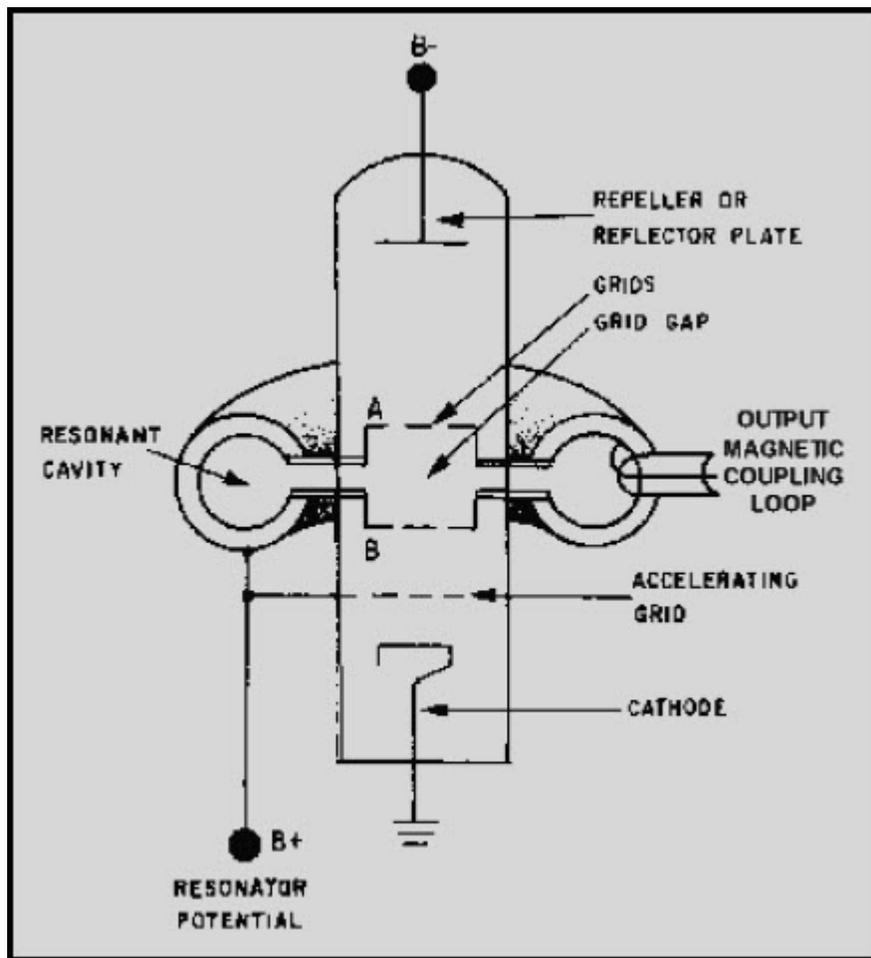


Figure 2-9 Functional diagram of a reflex klystron

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Three power sources are required for reflex klystron operation: (1) filament power, (2) positive resonator voltage (often referred to as beam voltage) used to accelerate the electrons through the grid gap of the resonant cavity, and (3) negative repeller voltage used to turn the electron beam around. The electrons are focused into a beam by the electrostatic fields set up by the resonator potential (B+) in the body of the tube. Note in figure 2-9 that the resonator potential is common to the resonator cavity, the accelerating grid, and the entire body of the tube.

The resonator potential also causes the resonant cavity to begin oscillating at its natural frequency when the tube is energized. These oscillations cause an electrostatic field across the grid gap of the cavity that changes direction at the frequency of the cavity. The changing electrostatic field affects the electrons in the beam as they pass through the grid gap. Some are accelerated and some are decelerated, depending upon the polarity of the electrostatic field as they pass through the gap. Figure 2-10, view (A), illustrates the three possible ways an electron can be affected as it passes through the gap (velocity increasing, decreasing, or remaining constant). Since the resonant cavity is oscillating, the grid potential is an alternating voltage that causes the electrostatic field between the grids to follow a sine-wave curve as shown in figure 2-10, view (B). As a result, the velocity of the electrons passing through the gap is affected uniformly as a function of that sine wave. The amount of velocity change is dependent on the strength and polarity of the grid voltage.

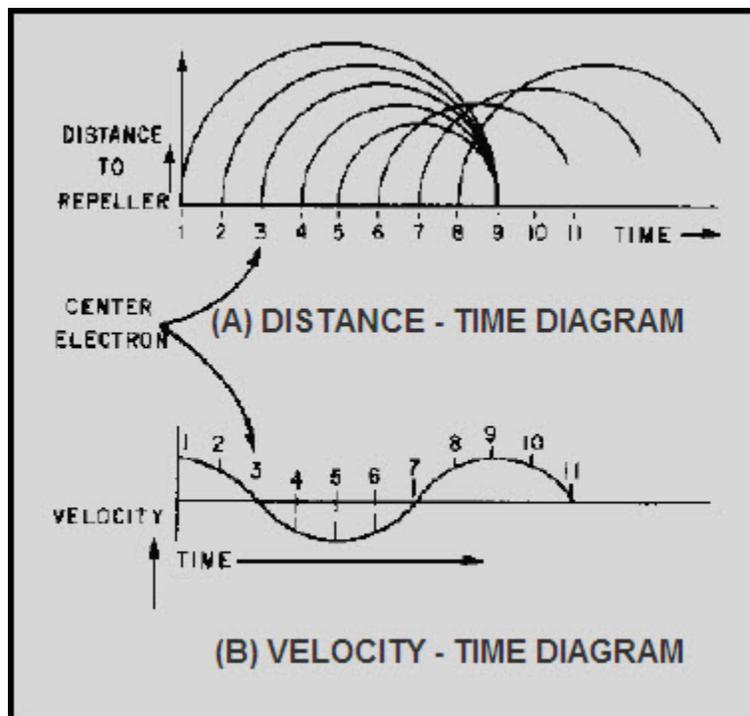


Figure 2-10 Electron bunching diagram

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The variation in grid voltage causes the electrons to enter the space between the grid and the repeller at various velocities. For example, in figure 2-10, views (A) and (B), the electrons at times 1 and 2 are speeded up as they pass through the grid. At time 3, the field is passing through zero and the electron is unaffected. At times 4 and 5, the grid field is reversed; the electrons give up energy because their velocity is reduced as they pass through the grids.

The distance the electrons travel in the space separating the grid and the repeller depends upon their velocity. Those moving at slower velocities, such as the electron at time 4, move only a short distance from the grid before being affected by the repeller voltage. When this happens, the electron is forced by the repeller voltage to stop, reverse direction, and return toward the grid. The electrons moving at higher velocities travel further beyond the grid before reversing direction because they have greater momentum. If the repeller voltage is set at the correct value, the electrons will form a bunch around the constant-speed electrons. The electrons will then return to the grid gap at the instant the electrostatic field is at the correct polarity to cause maximum deceleration of the bunch. This action is also illustrated in figure 2-10, view (A). When the grid field provides maximum deceleration, the returning electrons release maximum energy to the grid field which is in phase with cavity current. Thus, the returning electrons supply the regenerative feedback required to maintain cavity oscillations.

The constant-speed electrons must remain in the reflecting field space for a minimum time of $\frac{3}{4}$ cycle of the grid field for maximum energy transfer. The period of time the electrons remain in the repeller field is determined by the amount of negative repeller voltage. The reflex klystron will continue to oscillate if the electrons remain in the repeller field longer than $\frac{3}{4}$ cycle (as long as the electrons return to the grid gap when the field is of the proper polarity to decelerate the electrons). Figure 2-11 shows the effect of the repeller field on the electron bunch for $\frac{3}{4}$ cycle and for $1\frac{3}{4}$ cycles. Although not shown in the figure, the constant-velocity electrons may remain in the repeller field for any number of cycles over the minimum $\frac{3}{4}$ cycle. If the electrons remain in the field for longer than $\frac{3}{4}$ cycle, the difference in electron transit time causes the tube performance characteristics to change. The differences in operating characteristics are identified by MODES OF OPERATION.

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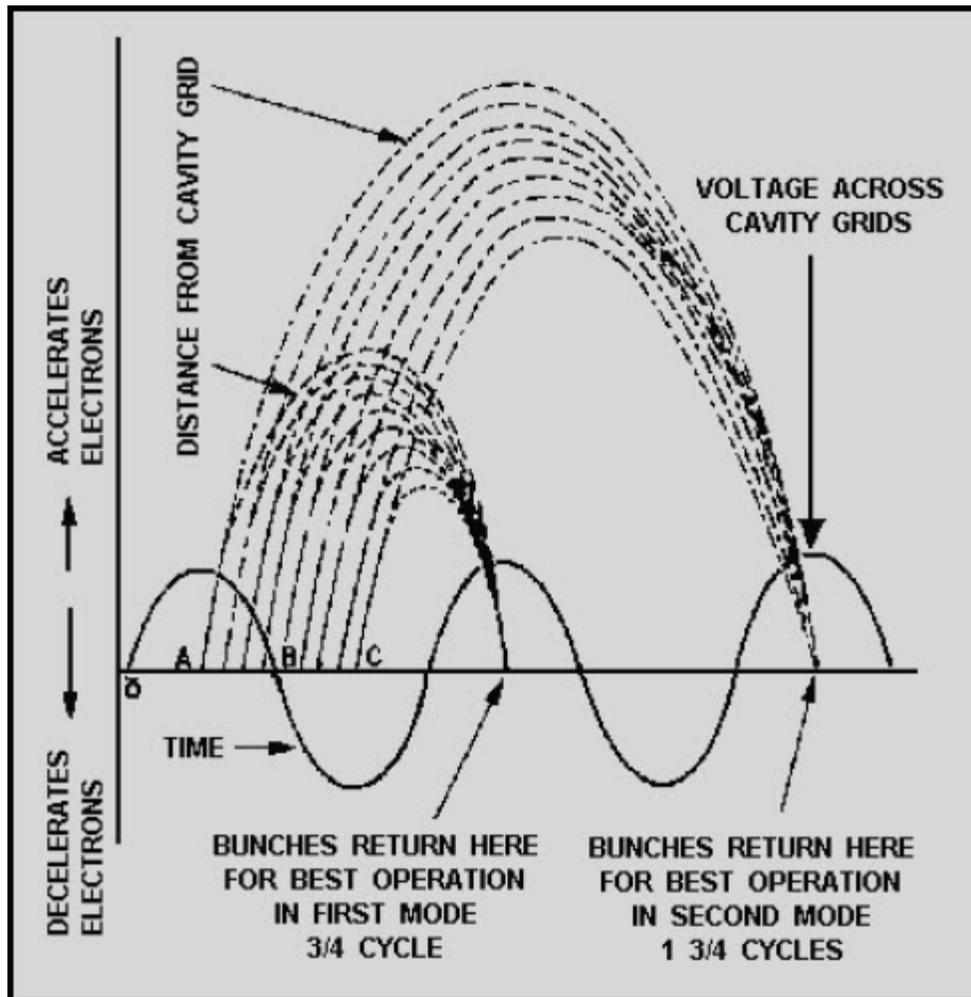


Figure 2-11 Bunching action of a reflex klystron

The reflex klystron operates in a different mode for each additional cycle that the electrons remain in the repeller field. Mode 1 is obtained when the repeller voltage produces an electron transit time of $\frac{3}{4}$ cycle. Additional modes follow in sequence. Mode 2 has an electron transit time of $1 \frac{3}{4}$ cycles; mode 3 has an electron transit time of $2 \frac{3}{4}$ cycles; etc. The physical design of the tube limits the number of modes possible in practical applications. A range of four modes of operation are normally available. The actual mode used ($1 \frac{3}{4}$ cycles through $4 \frac{3}{4}$ cycles, $2 \frac{3}{4}$ cycles through $6 \frac{3}{4}$ cycles, etc.) depends upon the application. The choice of mode is determined by the difference in power available from each mode and the band of frequencies over which the circuit can be tuned.

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OUTPUT POWER - The variation in output power for different modes of operation can be explained by examining the factors which limit the amplitude of oscillations. Power and amplitude limitations are caused by the **DEBUNCHING** process of the electrons in the repeller field space. Debunching is simply the spreading out of the electron bunches before they reach electrostatic fields across the cavity grid. The lower concentration of electrons in the returning bunches provides less power for delivery to the oscillating cavity. This reduced power from the bunches, in turn, reduces the amplitude of the cavity oscillations and causes a decrease in output power. In higher modes of operation the electron bunches are formed more slowly. They are more likely to be affected by debunching because of mutual repulsion between the negatively charged electrons. The long drift time in the higher modes allows more time for this electron interaction and, as a result, the effects of debunching are more severe. The mutual repulsion changes the relative velocity between the electrons in the bunches and causes the bunches to spread out.

Figure 2-12 illustrates the **ELECTRONIC TUNING** (tuning by altering the repeller voltage) range and output power of a reflex klystron. Each mode has a center frequency of 3,000 megahertz which is predetermined by the physical size of the cavity. The output power increases as the repeller voltage is made more negative. This is because the transit time of the electron bunches is decreased.

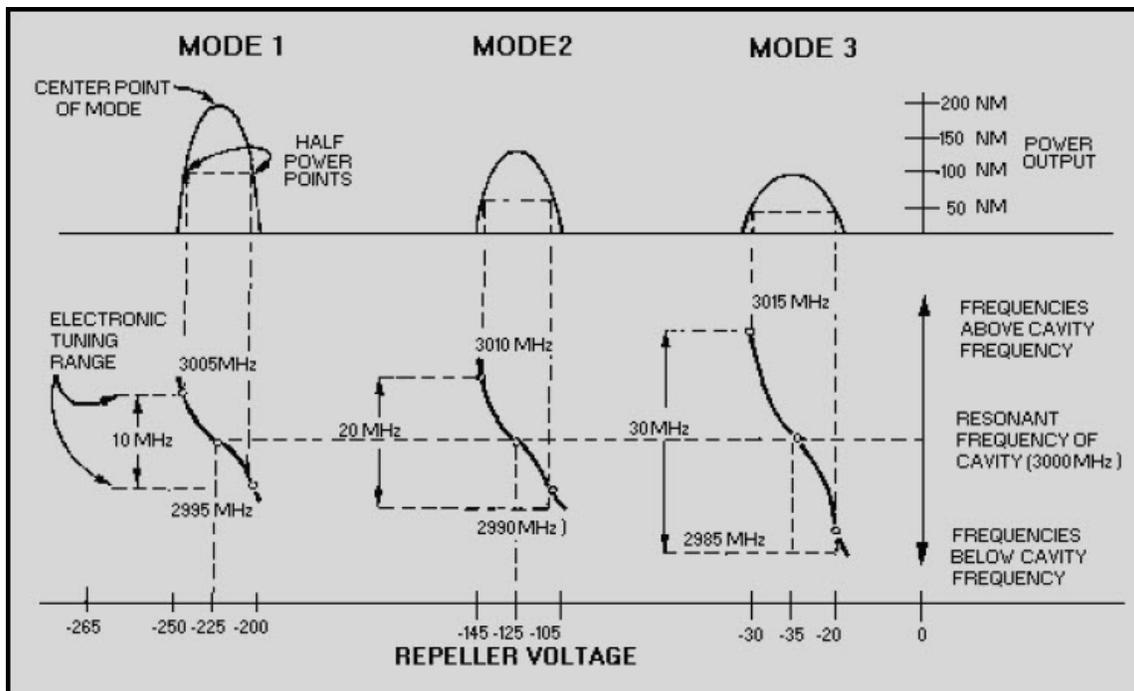


Figure 2-12 Electronic tuning and output power of a reflex klystron

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Electronic tuning does not change the center frequency of the cavity, but does vary the frequency within the mode of operation. The amount the frequency can be varied above or below the center frequency is limited by the half-power points of the mode, as shown in figure 2-12. The center frequency can be changed by one of two methods. One method, GRID-GAP TUNING, varies the cavity frequency by altering the distance between the grids to change the physical size of the cavity. This method varies the capacitance of the cavity by using a tuning screw to change the distance between the grids mechanically. The cavity can also be tuned by PADDLES or SLUGS that change the inductance of the cavity.

Q-20. What element of the reflex klystron replaces the output cavity of a normal klystron?

Q-21. When the repeller potential is constant, what property of the electron determines how long it will remain in the drift space of the reflex klystron?

Q-22. The constant-speed electrons of an electron bunch in a reflex klystron must remain in the repeller field for what minimum time?

Q-23. If the constant-speed electrons in a reflex klystron remain in the repeller field for $1\frac{3}{4}$ cycles, what is the mode of operation?

Q-24. Debunching of the electron bunches in the higher modes of a reflex klystron has what effect on output power?

Q-25. What limits the tuning range around the center frequency of a reflex klystron in a particular mode of operation?

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2.3.4 The Decibel Measurement System

Because of the use of the decibel measurement system in the following paragraphs, you will be introduced to it at this point. Technicians who deal with communications and radar equipment most often speak of the gain of an amplifier or a system in terms of units called DECIBELS (dB). Throughout your Navy career you will use decibels as an indicator of equipment performance; therefore, you need to have a basic understanding of the decibel system of measurement. Because the actual calculation of decibel measurements is seldom required in practical applications, the explanation given in this module is somewhat simplified. Most modern test equipment is designed to measure and indicate decibels directly which eliminates the need for complicated mathematical calculations. Nevertheless, a basic explanation of the decibel measurement system is necessary for you to understand the significance of dB readings and equipment gain ratings which are expressed in decibels.

The basic unit of measurement in the system is not the decibel, but the bel, named in honor of the American inventor, Alexander Graham Bell. The bel is a unit that expresses the logarithmic ratio between the input and output of any given component, circuit, or system and may be expressed in terms of voltage, current, or power. Most often it is used to show the ratio between input and output power. The formula is as follows:

$$N = \log_{10} \frac{P_1}{P_2} \text{ bel}$$

The gain of an amplifier can be expressed in bels by dividing the output (P₁) by the input (P₂) and taking the base 10 logarithm of the resulting quotient. Thus, if an amplifier doubles the power, the quotient will be 2. If you consult a logarithm table, you will find that the base 10 logarithm of 2 is 0.3; so the power gain of the amplifier is 0.3 bel. Experience has taught that because the bel is a rather large unit, it is difficult to apply. A more practical unit that can be applied more easily is the decibel (1/10 bel). Any figure expressed in bels can easily be converted to decibels by multiplying the figure by 10 or simply by moving the decimal one place to the right. The previously found ratio of 0.3 is therefore equal to 3 decibels.

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The reason for using the decibel system when expressing signal strength may be seen in the power ratios in table 2-1. For example, to say that a reference signal has increased 50 dB is much easier than to say the output has increased 100,000 times. The amount of increase or decrease from a chosen reference level is the basis of the decibel measurement system, not the reference level itself. Whether the input power is increased from 1 watt to 100 watts or from 1,000 watts to 100,000 watts, the amount of increase is still 20 decibels.

Table 2-1 Decibel Power Ratios

Source Level (dB)	=	Power Ratio
1	=	1.3
3	=	2.0
5	=	3.2
6	=	4.0
7	=	5.0
10	=	10 = 10 ₁
20	=	100 = 10 ₂
30	=	1000 = 10 ₃
40	=	10,000 = 10 ₄
50	=	100,000 = 10 ₅
60	=	1,000,000 = 10 ₆
70	=	10,000,000 = 10 ₇
100	=	10 ₁₀
110	=	10 ₁₁
140	=	10 ₁₄

Examine table 2-1 again, and take particular note of the power ratios for source levels of 3 dB and 6 dB. As the table illustrates, an increase of 3 dB represents a doubling of power. The reverse is also true. If a signal decreases by 3 dB, half the power is lost. For example, a 1,000 watt signal decreased by 3 dB will equal 500 watts while a 1,000 watt signal increased by 3 dB equals 2,000 watts.

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The attenuator is a widely used piece of test equipment that can be used to demonstrate the importance of the decibel as a unit of measurement. Attenuators are used to reduce a signal to a smaller level for use or measurement. Most attenuators are rated by the number of decibels the signal is reduced. The technician's job is to know the relationship between the dB rating and the power reduction it represents. This is so important, in fact, that every student of electronics should memorize the relationships in table 2-1 through the 60 dB range. The technician will have to apply this knowledge to prevent damage to valuable equipment. A helpful hint is to note that the first digit of the source level (on the chart) is the same number as the corresponding power of 10 exponent; i.e., $40 \text{ dB} = 1 \times 10^4$ or 10,000. A 20 dB attenuator, for example, will reduce an input signal by a factor of 100. In other words, a 100-milliwatt signal will be reduced to 1 milliwatt. A 30 dB attenuator will reduce the same 100-milliwatt signal by a factor of 1,000 and produce an output of 0.1 milliwatt. When an attenuator of the required size is not available, attenuators of several smaller sizes may be added directly together to reach the desired amount of attenuation. A 10 dB attenuator and a 20 dB attenuator add directly to equal 30 dB of attenuation. The same relationship exists with amplifier stages as well. If an amplifier has two stages rated at 10 dB each, the total amplifier gain will be 20 dB.

When you speak of the dB level of a signal, you are really speaking of a logarithmic comparison between the input and output signals. The input signal is normally used as the reference level. However, the application sometimes requires the use of a standard reference signal. The most widely used reference level is a 1-milliwatt signal. The standard decibel abbreviation of dB is changed to dBm to indicate the use of the 1-milliwatt standard reference. Thus, a signal level of +3 dBm is 3 dB above 1 milliwatt, and a signal level of -3 dBm is 3 dB below 1 milliwatt. Whether using dB or dBm, a plus (+) sign (or no sign at all) indicates the output signal is larger than the reference; a minus (-) sign indicates the output signal is less than the reference.

The Navy student of electronics will encounter the dBm system of measurement most often as a figure indicating the receiver sensitivity of radar or communications equipment. Typically, a radar receiver will be rated at approximately -107 dBm, which means the receiver will detect a signal 107 dB below 1 milliwatt.

The importance of understanding the decibel system of measurement can easily be seen in the case of receiver-sensitivity measurements. At first glance a loss of 3 dBm from a number as large as -107 dBm seems insignificant; however, it becomes extremely important when the number indicates receiver sensitivity in the decibel system. When the sensitivity falls to -104 dBm, the receiver will only detect a signal that is twice as large as a signal at -107 dBm.

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2.3.5 The Traveling-Wave Tube

The TRAVELING-WAVE TUBE (twt) is a high-gain, low-noise, wide-bandwidth microwave amplifier. It is capable of gains greater than 40 dB with bandwidths exceeding an octave. (A bandwidth of 1 octave is one in which the upper frequency is twice the lower frequency.) Traveling-wave tubes have been designed for frequencies as low as 300 megahertz and as high as 50 gigahertz. The twt is primarily a voltage amplifier. The wide-bandwidth and low-noise characteristics make the twt ideal for use as an rf amplifier in microwave equipment.

The physical construction of a typical twt is shown in figure 2-13. The twt contains an electron gun which produces and then accelerates an electron beam along the axis of the tube. The surrounding magnet provides a magnetic field along the axis of the tube to focus the electrons into a tight beam. The HELIX, at the center of the tube, is a coiled wire that provides a low-impedance transmission line for the rf energy within the tube. The rf input and output are coupled onto and removed from the helix by directional couplers that have no physical connection to the helix. If the rf energy is transported on coaxial cables, the coaxial couplers are wound in a helical manner similar to that shown in figure 2-13. If the rf energy is transported in waveguides, waveguide directional couplers are used. The attenuator prevents any reflected waves from traveling back down the helix.

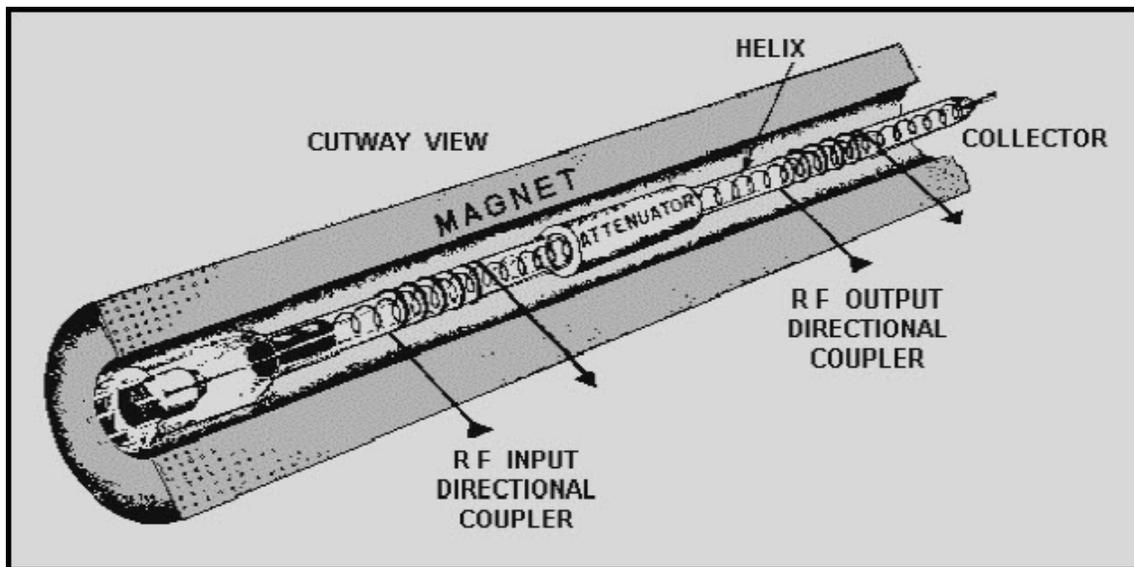


Figure 2-13 Physical construction of a twt

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A simplified version of twt operation is shown in figure 2-14. In the figure, an electron beam is passing along a nonresonant transmission line represented by a straight wire. The input to the transmission line is an rf wave which travels on the line from input to output. The line will transport a wide range of rf frequencies if it is terminated in the characteristic impedance of the line. The electromagnetic waves traveling down the line produce electric fields that interact with the electrons of the beam.

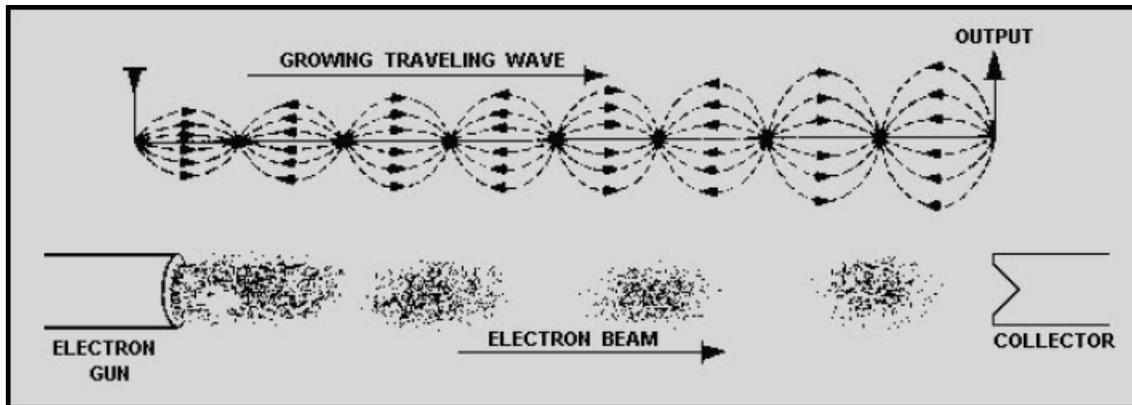


Figure 2-14 Simplified twt

If the electrons of the beam were accelerated to travel faster than the waves traveling on the wire, bunching would occur through the effect of velocity modulation. Velocity modulation would be caused by the interaction between the traveling-wave fields and the electron beam. Bunching would cause the electrons to give up energy to the traveling wave if the fields were of the correct polarity to slow down the bunches. The energy from the bunches would increase the amplitude of the traveling wave in a progressive action that would take place all along the length of the twt, as shown in figure 2-14.

However, because the waves travel along the wire at the speed of light, the simple twt shown in figure 2-14 will not work. At present no way is known to accelerate an electron beam to the speed of light. Since the electron beam cannot travel faster than the wave on the wire, bunching will not take place and the tube will not work. The twt is therefore designed with a delay structure to slow the traveling wave down to or below the speed of the electrons in the beam. A common twt delay structure is a wire, wound in the form of a long coil or helix, as shown in figure 2-15, view (A). The shape of the helix slows the effective velocity of the wave along the common axis of the helix and the tube to about one-tenth the speed of light. The wave still travels down the helix wire at the speed of light, but the coiled shape causes the wave to travel a much greater total distance than the electron beam. The speed at which the wave travels down the tube can be varied by changing the number of turns or the diameter of the turns in the helix wire. The helical delay structure works well because it has the added advantage of causing a large proportion of electric fields that are parallel to the electron beam. The parallel fields provide maximum interaction between the fields and the electron beam.

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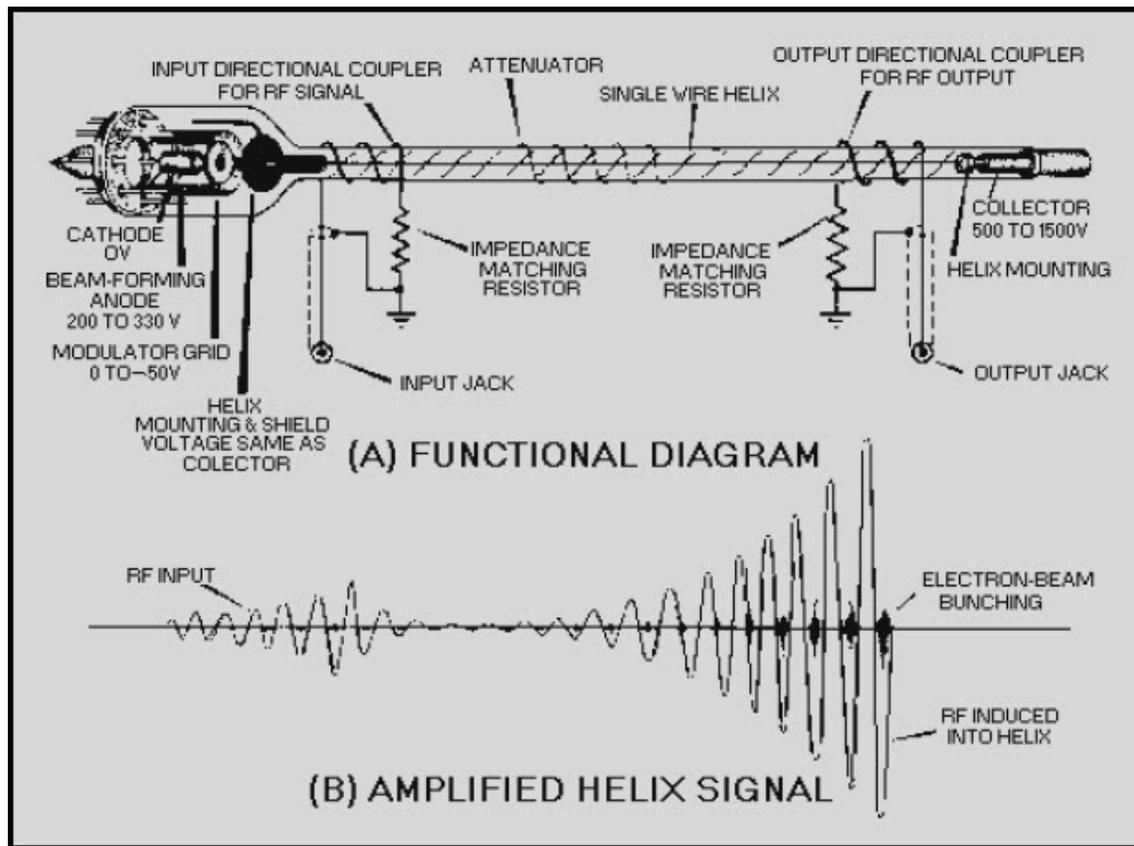


Figure 2-15 Functional diagram of a twt

In a typical twt, the electron beam is directed down the center of the helix while, at the same time, an rf signal is coupled onto the helix. The electrons of the beam are velocity-modulated by the electric fields produced by the rf signal.

Amplification begins as the electron bunches form and release energy to the signal on the helix. The slightly amplified signal causes a denser electron bunch which, in turn, amplifies the signal even more. The amplification process is continuous as the rf wave and the electron beam travel down the length of the tube.

Any portion of the twt output signal that reflects back to the input will cause oscillations within the tube which results in a decrease in amplification. Attenuators are placed along the length of the helix to prevent reflections from reaching the input. The attenuator causes a loss in amplitude, as can be seen in figure 2-15, view (B), but it can be placed so as to minimize losses while still isolating the input from the output.

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The relatively low efficiency of the twt partially offsets the advantages of high gain and wide bandwidth. The internal attenuator reduces the gain of the tube, and the power required to energize the focusing magnet is an operational loss that cannot be recovered. The twt also produces heat which must be dissipated by either air-conditioning or liquid-cooling systems. All of these factors reduce the overall efficiency of the twt, but the advantages of high gain and wide bandwidth are usually enough to overcome the disadvantages.

2.3.6 The Backward-Wave Oscillator

The BACKWARD-WAVE OSCILLATOR (bwo) is a microwave-frequency, velocity-modulated tube that operates on the same principle as the twt. However, a traveling wave that moves from the electron gun end of the tube toward the collector is not used in the bwo. Instead, the bwo extracts energy from the electron beam by using a backward wave that travels from the collector toward the electron gun (cathode). Otherwise, the electron bunching action and energy extraction from the electron beam is very similar to the actions in a twt.

The typical bwo is constructed from a folded transmission line or waveguide that winds back and forth across the path of the electron beam, as shown in figure 2-16. The folded waveguide in the illustration serves the same purpose as the helix in a twt. The fixed spacing of the folded waveguide limits the bandwidth of the bwo. Since the frequency of a given waveguide is constant, the frequency of the bwo is controlled by the transit time of the electron beam. The transit time is controlled by the collector potential. Thus, the output frequency can be changed by varying the collector voltage, which is a definite advantage. As in the twt, the electron beam in the bwo is focused by a magnet placed around the body of the tube.

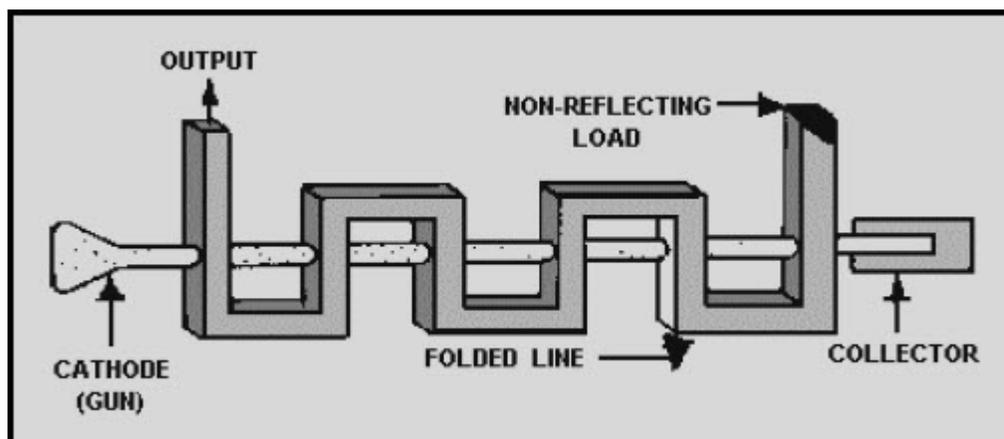


Figure 2-16 Typical bwo

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Q-26. What is the primary use of the twt?

Q-27. The magnet surrounding the body of a twt serves what purpose?

Q-28. How are the input and output directional couplers in a twt connected to the helix?

Q-29. What relationship must exist between the electron beam and the traveling wave for bunching to occur in the electron beam of a twt?

Q-30. What structure in the twt delays the forward progress of the traveling wave?

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2.3.7 The Magnetron

The MAGNETRON, shown in figure 2-17A, is a self-contained microwave oscillator that operates differently from the linear-beam tubes, such as the twt and the klystron. Figure 2-17B is a simplified drawing of the magnetron. **CROSSED-ELECTRON** and **MAGNETIC** fields are used in the magnetron to produce the high-power output required in radar and communications equipment.

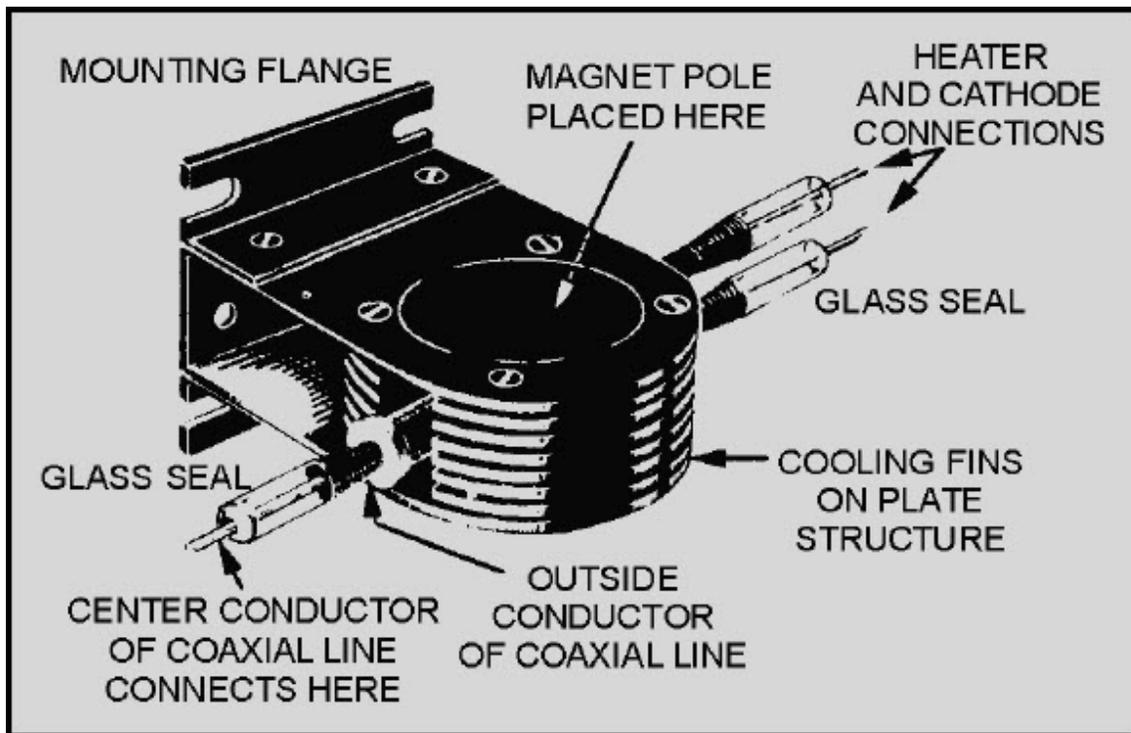


Figure 2-17A Magnetron

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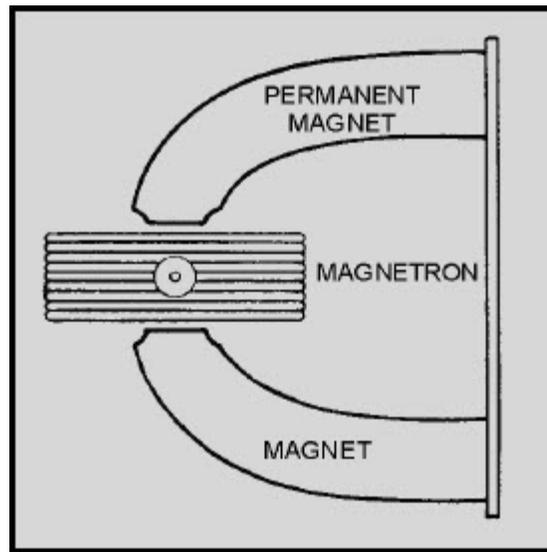


Figure 2-17B Magnetron

The magnetron is classed as a diode because it has no grid. A magnetic field located in the space between the plate (anode) and the cathode serves as a grid. The plate of a magnetron does not have the same physical appearance as the plate of an ordinary electron tube. Since conventional inductive-capacitive (LC) networks become impractical at microwave frequencies, the plate is fabricated into a cylindrical copper block containing resonant cavities which serve as tuned circuits. The magnetron base differs considerably from the conventional tube base. The magnetron base is short in length and has large diameter leads that are carefully sealed into the tube and shielded.

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The cathode and filament are at the center of the tube and are supported by the filament leads. The filament leads are large and rigid enough to keep the cathode and filament structure fixed in position. The output lead is usually a probe or loop extending into one of the tuned cavities and coupled into a waveguide or coaxial line. The plate structure, shown in figure 2-18, is a solid block of copper. The cylindrical holes around its circumference are resonant cavities. A narrow slot runs from each cavity into the central portion of the tube dividing the inner structure into as many segments as there are cavities. Alternate segments are strapped together to put the cavities in parallel with regard to the output. The cavities control the output frequency. The straps are circular, metal bands that are placed across the top of the block at the entrance slots to the cavities. Since the cathode must operate at high power, it must be fairly large and must also be able to withstand high operating temperatures. It must also have good emission characteristics, particularly under return bombardment by the electrons. This is because most of the output power is provided by the large number of electrons that are emitted when high-velocity electrons return to strike the cathode. The cathode is indirectly heated and is constructed of a high-emission material. The open space between the plate and the cathode is called the INTERACTION SPACE. In this space the electric and magnetic fields interact to exert force upon the electrons.

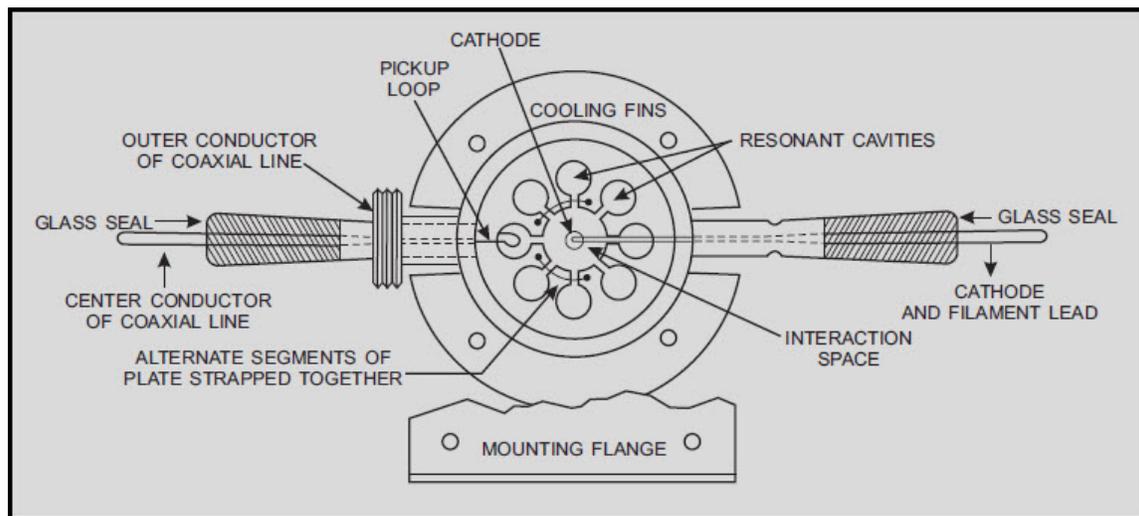


Figure 2-18 Cutaway view of a magnetron

The magnetic field is usually provided by a strong, permanent magnet mounted around the magnetron so that the magnetic field is parallel with the axis of the cathode. The cathode is mounted in the center of the interaction space.

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BASIC MAGNETRON OPERATION - Magnetron theory of operation is based on the motion of electrons under the influence of combined electric and magnetic fields. The following information presents the laws governing this motion.

The direction of an electric field is from the positive electrode to the negative electrode. The law governing the motion of an electron in an electric field (E field) states:

The force exerted by an electric field on an electron is proportional to the strength of the field. Electrons tend to move from a point of negative potential toward a positive potential.

This is shown in figure 2-19. In other words, electrons tend to move against the E field. When an electron is being accelerated by an E field, as shown in figure 2-19, energy is taken from the field by the electron.

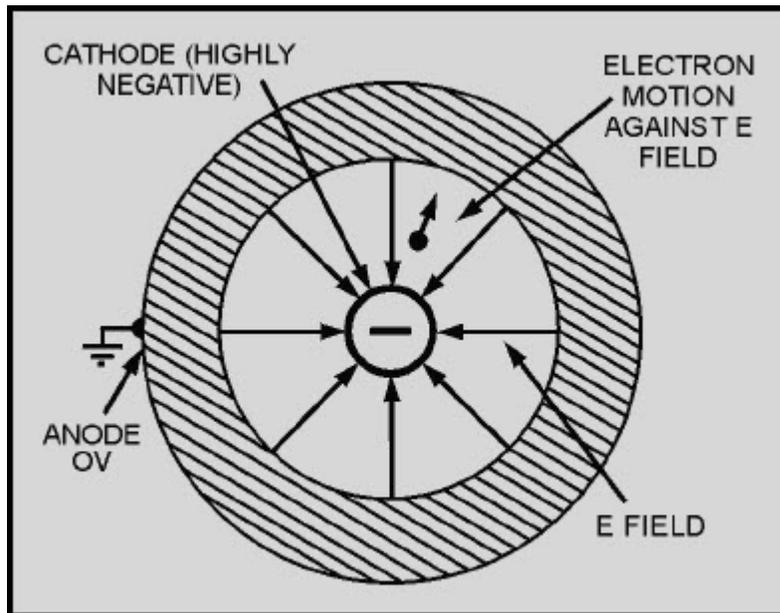


Figure 2-19 Electron motion in an electric field

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The law of motion of an electron in a magnetic field (H field) states:

The force exerted on an electron in a magnetic field is at right angles to both the field and the path of the electron. The direction of the force is such that the electron trajectories are clockwise when viewed in the direction of the magnetic field.

This is shown in figure 2-20.

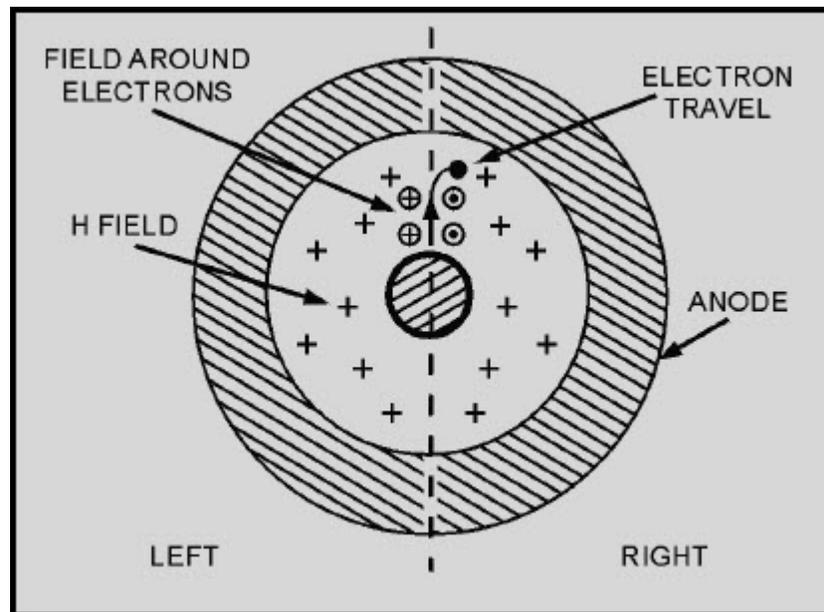


Figure 2-20 Electron motion in a magnetic field

In figure 2-20, assume that a south pole is below the figure and a north pole is above the figure so that the magnetic field is going into the paper. When an electron is moving through space, a magnetic field builds around the electron just as it would around a wire when electrons are flowing through a wire. In figure 2-20 the magnetic field around the moving electron adds to the permanent magnetic field on the left side of the electron's path and subtracts from the permanent magnetic field on the right side. This action weakens the field on the right side; therefore, the electron path bends to the right (clockwise). If the strength of the magnetic field is increased, the path of the electron will have a sharper bend. Likewise, if the velocity of the electron increases, the field around it increases and the path will bend more sharply.

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A schematic diagram of a basic magnetron is shown in figure 2-21A. The tube consists of a cylindrical plate with a cathode placed along the center axis of the plate. The tuned circuit is made up of cavities in which oscillations take place and are physically located in the plate.

When no magnetic field exists, heating the cathode results in a uniform and direct movement of the field from the cathode to the plate, as illustrated in figure 2-21B.

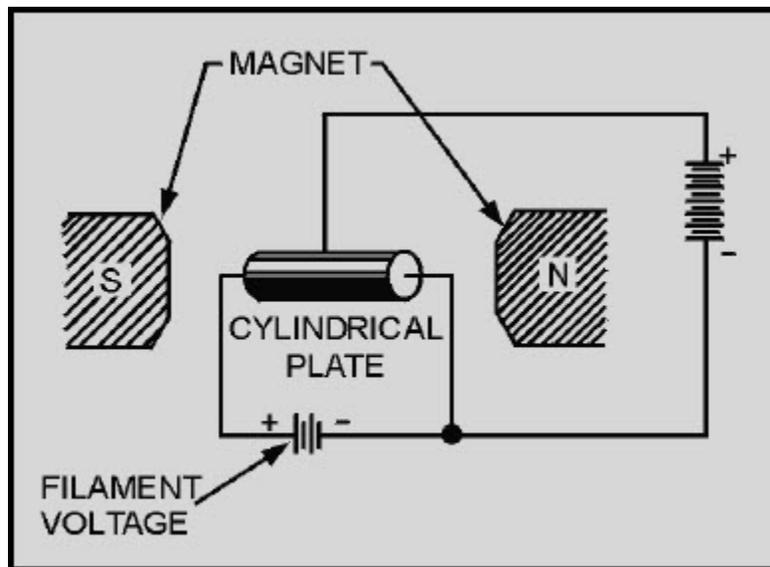


Figure 2-21A Basic magnetron. SIDE VIEW

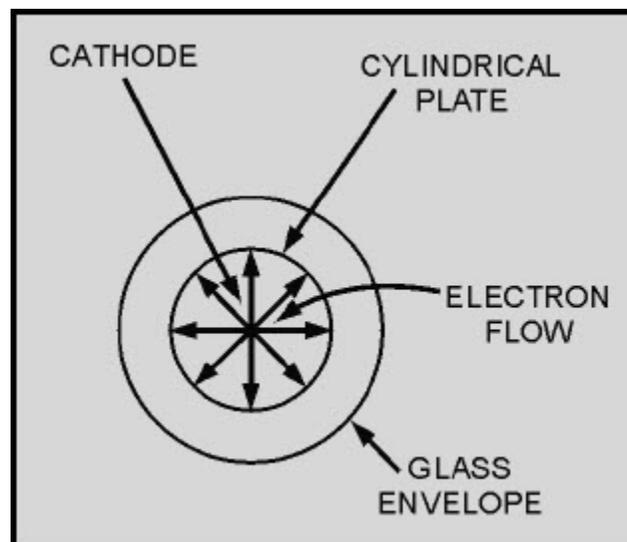


Figure 2-21B Basic magnetron. END VIEW
OMITTING MAGNETS

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However, as the magnetic field surrounding the tube is increased, a single electron is affected, as shown in figure 2-22. In figure 2-22, view (A), the magnetic field has been increased to a point where the electron proceeds to the plate in a curve rather than a direct path.

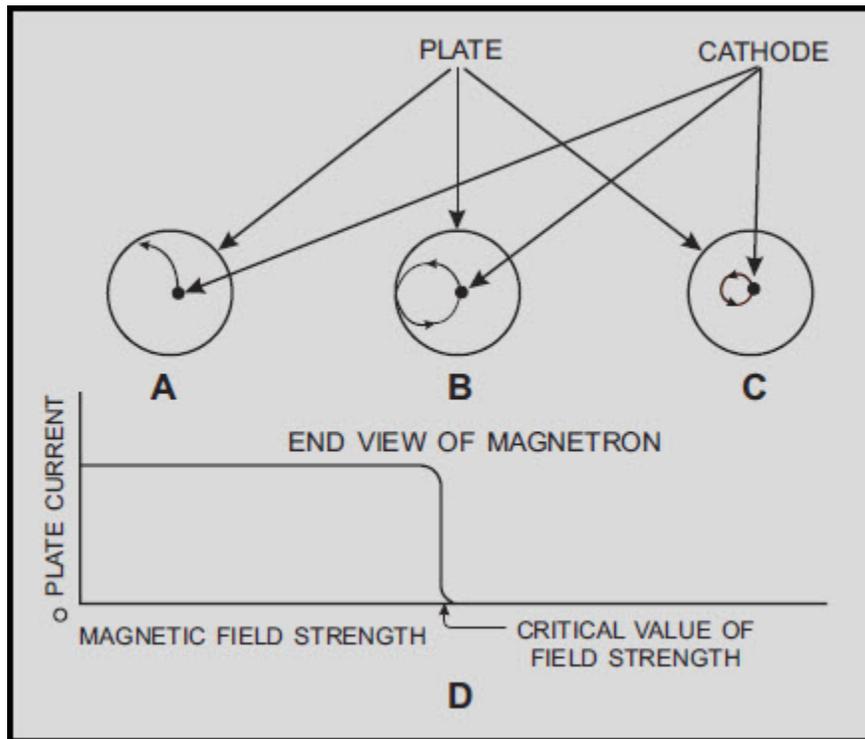


Figure 2-22 Effect of a magnetic field on a single electron

In view (B) of figure 2-22, the magnetic field has reached a value great enough to cause the electron to just miss the plate and return to the filament in a circular orbit. This value is the **CRITICAL VALUE** of field strength. In view (C), the value of the field strength has been increased to a point beyond the critical value; the electron is made to travel to the cathode in a circular path of smaller diameter.

View (D) of figure 2-22. shows how the magnetron plate current varies under the influence of the varying magnetic field. In view (A), the electron flow reaches the plate, so a large amount of plate current is flowing. However, when the critical field value is reached, as shown in view (B), the electrons are deflected away from the plate and the plate current then drops quickly to a very small value. When the field strength is made still greater, as shown in view (C), the plate current drops to zero.

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When the magnetron is adjusted to the cutoff, or critical value of the plate current, and the electrons just fail to reach the plate in their circular motion, it can produce oscillations at microwave frequencies. These oscillations are caused by the currents induced electrostatically by the moving electrons. The frequency is determined by the time it takes the electrons to travel from the cathode toward the plate and back again. A transfer of microwave energy to a load is made possible by connecting an external circuit between the cathode and the plate of the magnetron. Magnetron oscillators are divided into two classes: NEGATIVE-RESISTANCE and ELECTRON-RESONANCE MAGNETRON OSCILLATORS.

A negative-resistance magnetron oscillator is operated by a static negative resistance between its electrodes. This oscillator has a frequency equal to the frequency of the tuned circuit connected to the tube.

An electron-resonance magnetron oscillator is operated by the electron transit time required for electrons to travel from cathode to plate. This oscillator is capable of generating very large peak power outputs at frequencies in the thousands of megahertz. Although its average power output over a period of time is low, it can provide very high-powered oscillations in short bursts of pulses.

Q-31. The folded waveguide in a bwo serves the same purpose as what component in a twt?

Q-32. What serves as a grid in a magnetron?

Q-33. A cylindrical copper block with resonant cavities around the circumference is used as what component of a magnetron?

Q-34. What controls the output frequency of a magnetron?

Q-35. What element in the magnetron causes the curved path of electron flow?

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Q-36. What is the term used to identify the amount of field strength required to cause the electrons to just miss the plate and return to the filament in a circular orbit?

Q-37. A magnetron will produce oscillations when the electrons follow what type of path?

NEGATIVE-RESISTANCE MAGNETRON - The split-anode, negative-resistance magnetron is a variation of the basic magnetron which operates at a higher frequency. The negative-resistance magnetron is capable of greater power output than the basic magnetron. Its general construction is similar to the basic magnetron except that it has a split plate, as shown in figure 2-23A and B. These half plates are operated at different potentials to provide an electron motion, as shown in figure 2-24. The electron leaving the cathode and progressing toward the high-potential plate is deflected by the magnetic field and follows the path shown in figure 2-24. After passing the split between the two plates, the electron enters the electrostatic field set up by the lower-potential plate.

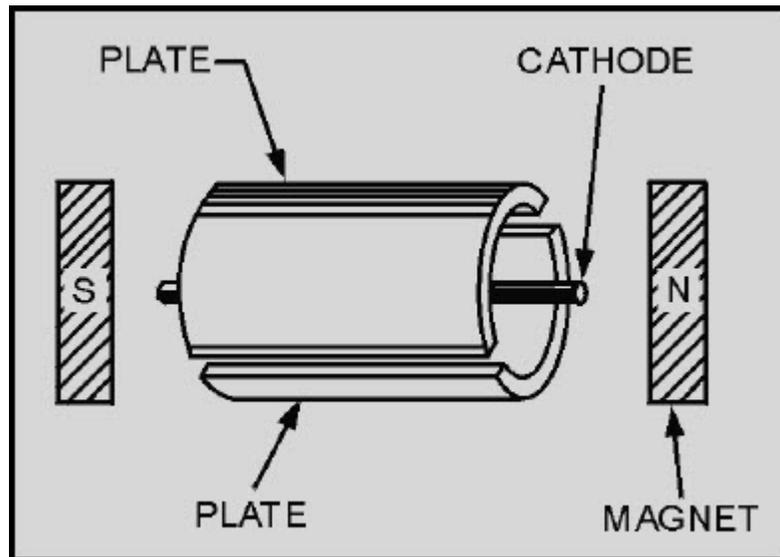


Figure 2-23A Split-anode magnetron

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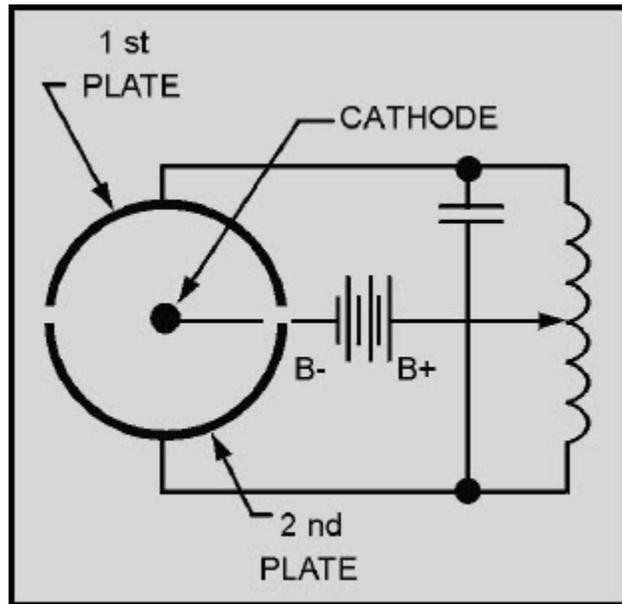


Figure 2-23B Split-anode magnetron

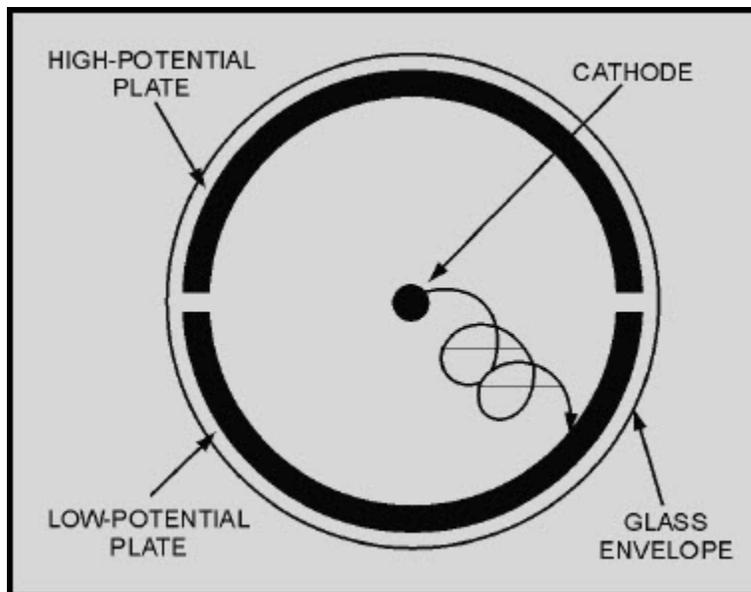


Figure 2-24 Movement of an electron in a split-anode magnetron

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Here the magnetic field has more effect on the electron and deflects it into a tighter curve. The electron then continues to make a series of loops through the magnetic field and the electric field until it finally arrives at the low-potential plate.

Oscillations are started by applying the proper magnetic field to the tube. The field value required is slightly higher than the critical value. In the split-anode tube, the critical value is the field value required to cause all the electrons to miss the plate when its halves are operating at the same potential. The alternating voltages impressed on the plates by the oscillations generated in the tank circuit will cause electron motion, such as that shown in figure 2-24, and current will flow. Since a very concentrated magnetic field is required for the negative-resistance magnetron oscillator, the length of the tube plate is limited to a few centimeters to keep the magnet at reasonable dimensions. In addition, a small diameter tube is required to make the magnetron operate efficiently at microwave frequencies. A heavy-walled plate is used to increase the radiating properties of the tube. Artificial cooling methods, such as forced-air or water-cooled systems, are used to obtain still greater dissipation in these high-output tubes.

The output of a magnetron is reduced by the bombardment of the filament by electrons which travel in loops, shown in figure 2-22, views (B) and (C). This action causes an increase of filament temperature under conditions of a strong magnetic field and high plate voltage and sometimes results in unstable operation of the tube. The effects of filament bombardment can be reduced by operating the filament at a reduced voltage. In some cases, the plate voltage and field strength are also reduced to prevent destructive filament bombardment.

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ELECTRON-RESONANCE MAGNETRON - In the electron-resonance magnetron, the plate is constructed to resonate and function as a tank circuit. Thus, the magnetron has no external tuned circuits. Power is delivered directly from the tube through transmission lines, as shown in figure 2-25. The constants and operating conditions of the tube are such that the electron paths are somewhat different from those in figure 2-24. Instead of closed spirals or loops, the path is a curve having a series of sharp points, as illustrated in figure 2-26. Ordinarily, this type of magnetron has more than two segments in the plate. For example, figure 2-26 illustrates an eight-segment plate.

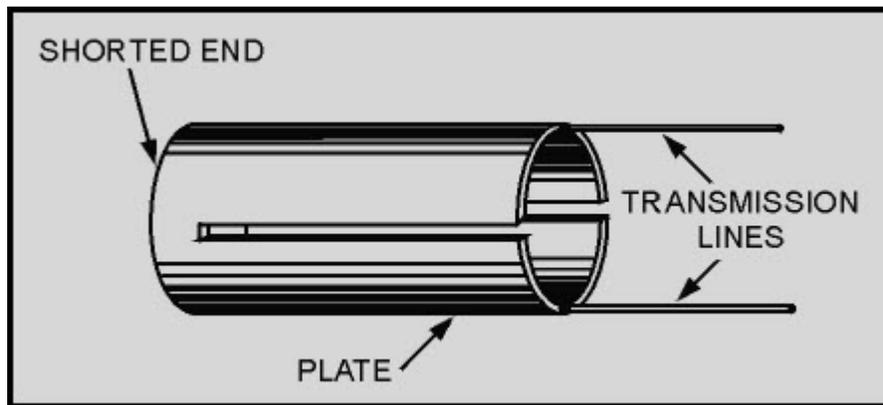


Figure 2-25 Plate tank circuit of a magnetron

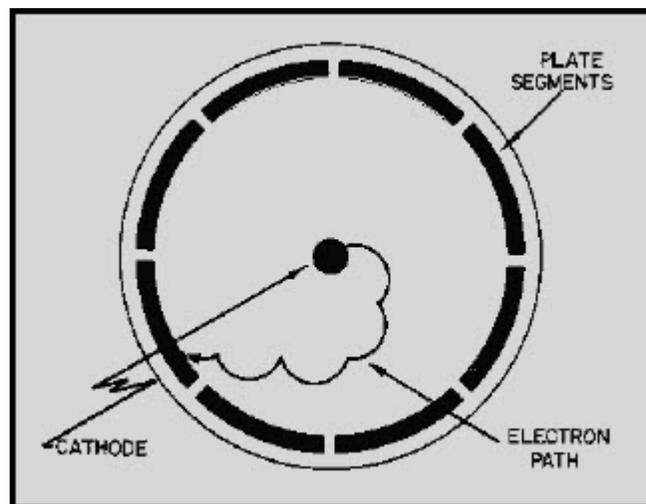


Figure 2-26 Electron path in an electron-resonance magnetron

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The electron-resonance magnetron is the most widely used for microwave frequencies because it has reasonably high efficiency and relatively high output. The average power of the electron-resonance magnetron is limited by the amount of cathode emission, and the peak power is limited by the maximum voltage rating of the tube components. Three common types of anode blocks used in electron-resonance magnetrons are shown in figure 2-27.

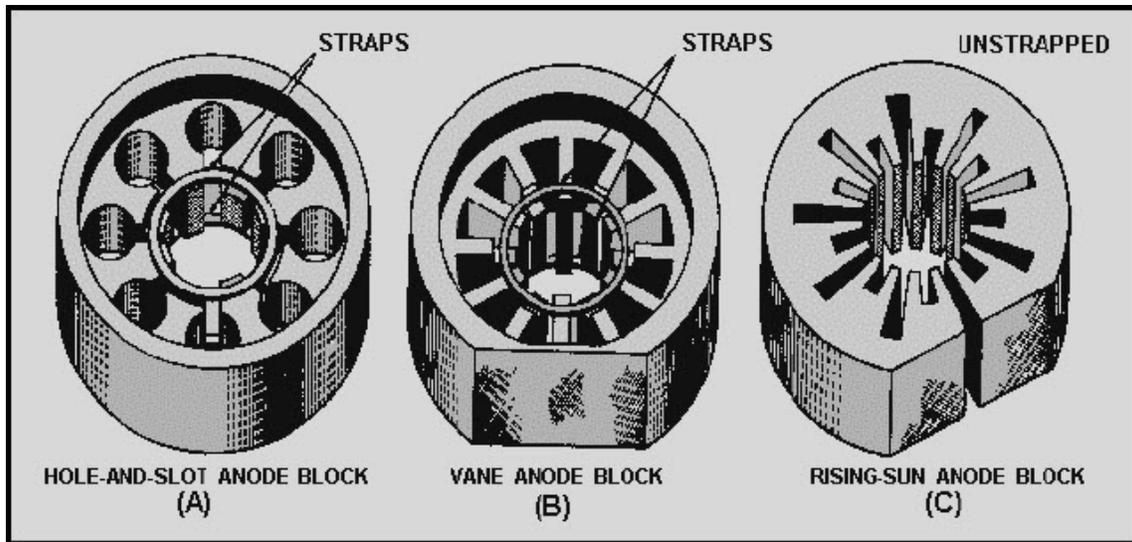


Figure 2-27 Common types of anode blocks

The anode block shown in figure 2-27, view (A), has cylindrical cavities and is called a HOLEAND-SLOT ANODE. The anode block in view (B) is called the VANE ANODE which has trapezoidal cavities. The first two anode blocks operate in such a way that alternate segments must be connected, or strapped, so that each segment is opposite in polarity to the segment on either side, as shown in figure 2-28. This also requires an even number of cavities.

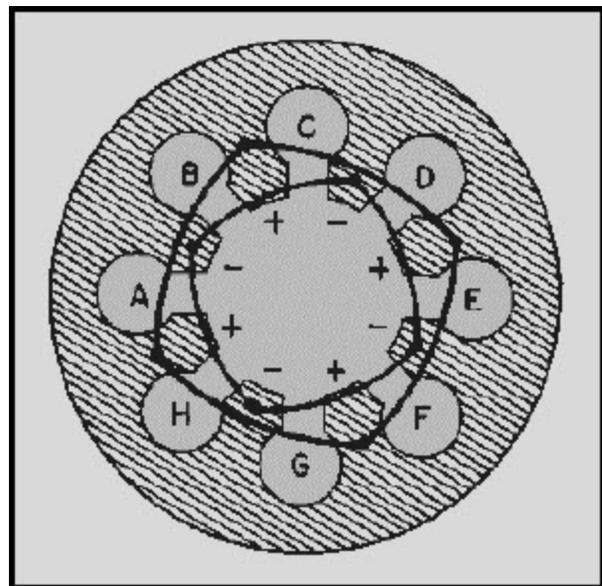


Figure 2-28 Strapping alternate segments

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The anode block illustrated in figure 2-27, view (C), is called a RISING-SUN BLOCK. The alternate large and small trapezoidal cavities in this block result in a stable frequency between the resonant frequencies of the large and small cavities.

Figure 2-29A, shows the physical relationships of the resonant cavities contained in the hole-and-slot anode (figure 2-27, view (A)). This will be used when analyzing the operation of the electron-resonance magnetron.

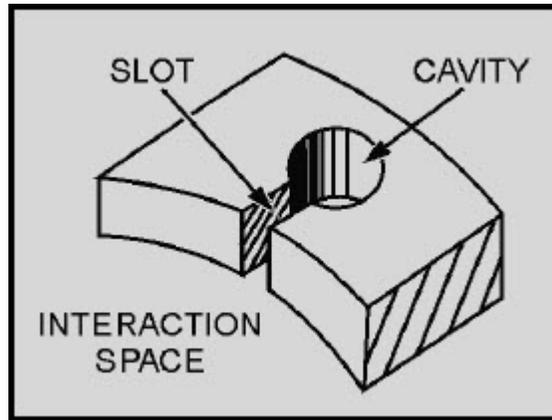


Figure 2-29A Equivalent circuit of a hole-and-slot cavity

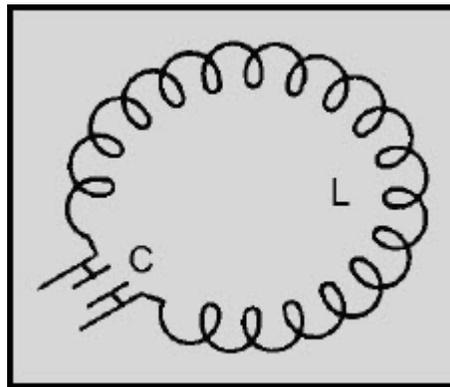


Figure 2-29B Equivalent circuit of a hole-and-slot cavity

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Electrical Equivalent - Notice in figure 2-29A, that the cavity consists of a cylindrical hole in the copper anode and a slot which connects the cavity to the interaction space.

The equivalent electrical circuit of the hole and slot is shown in figure 2-29B. The parallel sides of the slot form the plates of a capacitor while the walls of the hole act as an inductor. The hole and slot thus form a high-Q, resonant LC circuit. As shown in figure 2-27, the anode of a magnetron has a number of these cavities.

An analysis of the anodes in the hole-and-slot block reveals that the LC tanks of each cavity are in series (assuming the straps have been removed), as shown in figure 2-30. However, an analysis of the anode block after alternate segments have been strapped reveals that the cavities are connected in parallel because of the strapping. Figure 2-31 shows the equivalent circuit of a strapped anode.

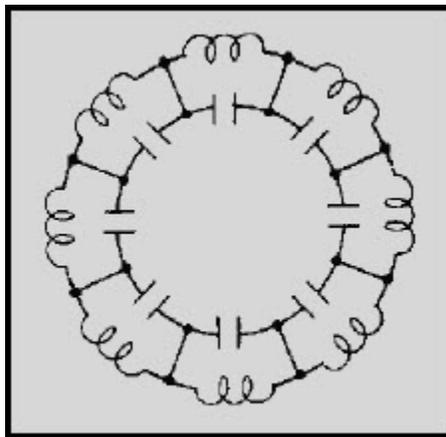


Figure 2-30 Cavities connected in series

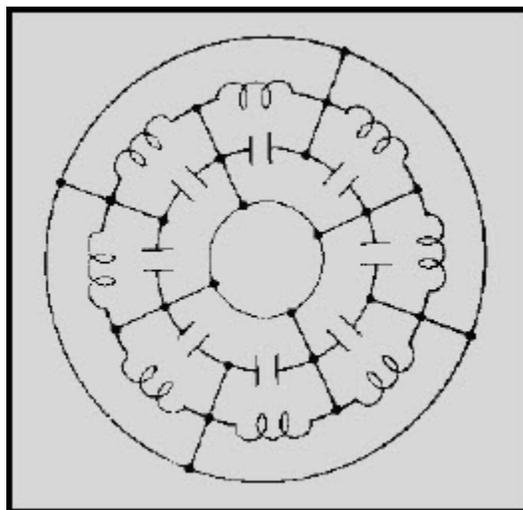


Figure 2-31 Cavities in parallel because of strapping

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Electric Field - The electric field in the electron-resonance oscillator is a product of ac and dc fields. The dc field extends radially from adjacent anode segments to the cathode, as shown in figure 2-32. The ac fields, extending between adjacent segments, are shown at an instant of maximum magnitude of one alternation of the rf oscillations occurring in the cavities.

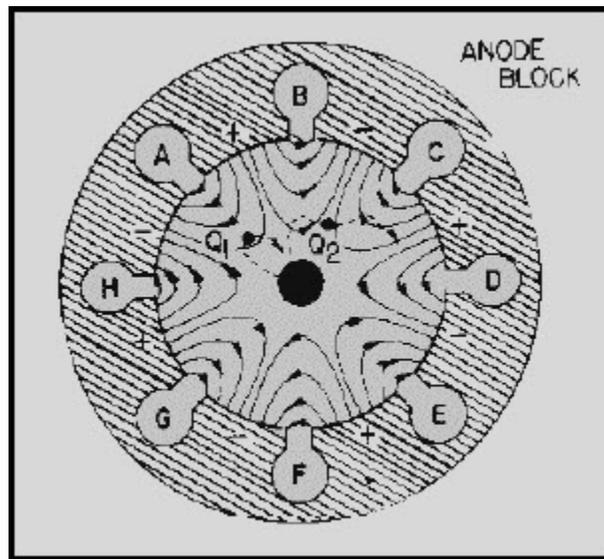


Figure 2-32 Probable electron paths in an electron-resonance magnetron oscillator

The strong dc field going from anode to cathode is created by a large, negative dc voltage pulse applied to the cathode. This strong dc field causes electrons to accelerate toward the plate after they have been emitted from the cathode. Recall that an electron moving against an E field is accelerated by the field and takes energy from the field. Also, an electron gives up energy to a field and slows down if it is moving in the same direction as the field (positive to negative). Oscillations are sustained in a magnetron because as electrons pass through the ac and dc fields, they gain energy from the dc field and give up energy to the ac field. The electrons that give up energy to the ac field are called **WORKING ELECTRONS**. However, not all of the electrons give up energy to the ac field. Some electrons take energy from the ac field, which is an undesirable action.

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In figure 2-32, consider electron Q1, which is shown entering the field around the slot entrance to cavity A. The clockwise rotation of the electron path is caused by the interaction of the magnetic field around the moving electron with the permanent magnetic field. The permanent magnetic field is assumed to be going into the paper in figure 2-32 (the action of an electron moving in an H field was explained earlier). Notice that electron Q1 is moving against the ac field around cavity A. The electron takes energy from the ac field and then accelerates, turning more sharply when its velocity increases. Thus, electron Q1 turns back toward the cathode. When it strikes the cathode, it gives up the energy it received from the ac field. This bombardment also forces more electrons to leave the cathode and accelerate toward the anode. Electron Q2 is slowed down by the field around cavity B and gives up some of its energy to the ac field. Since electron Q2 loses velocity, the deflective force exerted by the H field is reduced. The electron path then deviates to the left in the direction of the anode, rather than returning to the cathode as did electron Q1.

The cathode to anode potential and the magnetic field strength determine the amount of time for electron Q2 to travel from a position in front of cavity B to a position in front of cavity C. Cavity C is equal to approximately 1/2 cycle of the rf oscillations of the cavities. When electron Q2 reaches a position in front of cavity C, the ac field of cavity C is reversed from that shown. Therefore, electron Q2 gives up energy to the ac field of cavity C and slows down even more. Electron Q2 actually gives up energy to each cavity as it passes and eventually reaches the anode when its energy is expended. Thus, electron Q2 has helped sustain oscillations because it has taken energy from the dc field and given it to the ac field. Electron Q1, which took energy from the ac field around cavity A, did little harm because it immediately returned to the cathode.

The cumulative action of many electrons returning to the cathode while others are moving toward the anode forms a pattern resembling the moving spokes of a wheel known as a SPACE-CHARGE WHEEL, as indicated in figure 2-33. Electrons in the spokes of the wheel are the working electrons.

The space-charge wheel rotates about the cathode at an angular velocity of 2 poles (anode segments) per cycle of the ac field. This phase relationship enables the concentration of electrons to continuously deliver energy to sustain the rf oscillations. Electrons emitted from the area of the cathode between the spokes are quickly returned to the cathode.

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In figure 2-33 the alternate segments between cavities are assumed to be at the same potential at the same instant. An ac field is assumed to exist across each individual cavity. This mode of operation is called the PI MODE, since adjacent segments of the anode have a phase difference of 180 degrees, or one-pi radian. Several other modes of oscillation are possible, but a magnetron operating in the pi mode has greater power and output and is the most commonly used.

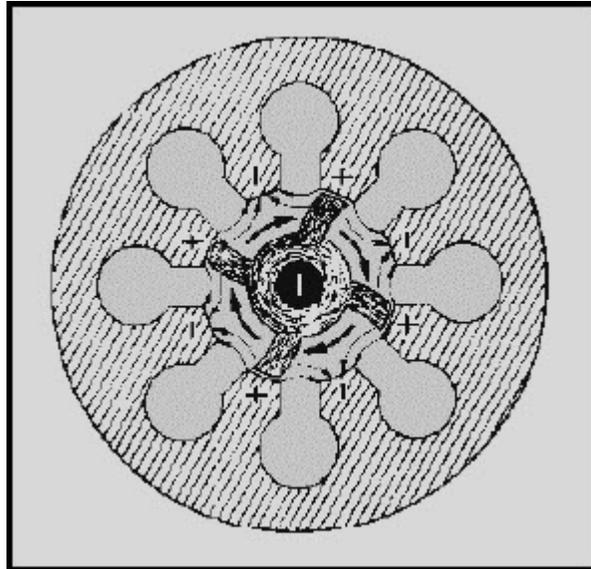


Figure 2-33 Rotating space-charge wheel in an eight-cavity magnetron

An even number of cavities, usually six or eight, are used and alternate segments are strapped to ensure that they have identical polarities. The frequency of the pi mode is separated from the frequency of the other modes by strapping.

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For the pi mode, all parts of each strapping ring are at the same potential; but the two rings have alternately opposing potentials, as shown in figure 2-34. Stray capacitance between the rings adds capacitive loading to the resonant mode. For other modes, however, a phase difference exists between the successive segments connected to a given strapping ring which causes current to flow in the straps.

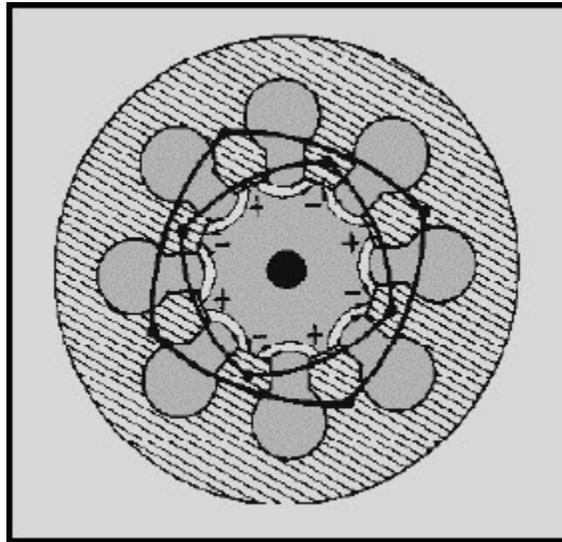


Figure 2-34 Alternate segments connected by strapping rings

The straps contain inductance, and an inductive shunt is placed in parallel with the equivalent circuit. This lowers the inductance and increases the frequency at modes other than the pi mode.

Q-38. What is the primary difference in construction between the basic magnetron and the negative-resistance magnetron?

Q-39. What starts the oscillations in a negative-resistance magnetron?

Q-40. Why is the negative-resistance magnetron often operated with reduced filament voltage?

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Q-41. What type of electron-resonance anode block does not require strapping?

Q-42. Without strapping, the resonant cavities of a hole-and-slot anode are connected in what manner?

Q-43. What are the electrons called that give up energy to the ac field in a magnetron?

COUPLING METHODS - Energy (rf) can be removed from a magnetron by means of a **COUPLING LOOP**. At frequencies lower than 10,000 megahertz, the coupling loop is made by bending the inner conductor of a coaxial cable into a loop. The loop is then soldered to the end of the outer conductor so that it projects into the cavity, as shown in figure 2-35A. Locating the loop at the end of the cavity, as shown in figure 2-35B, causes the magnetron to obtain sufficient pickup at higher frequencies.

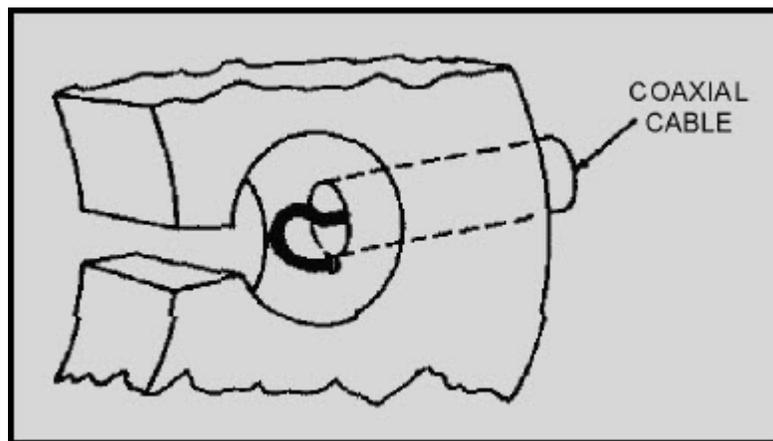


Figure 2-35A Magnetron coupling methods

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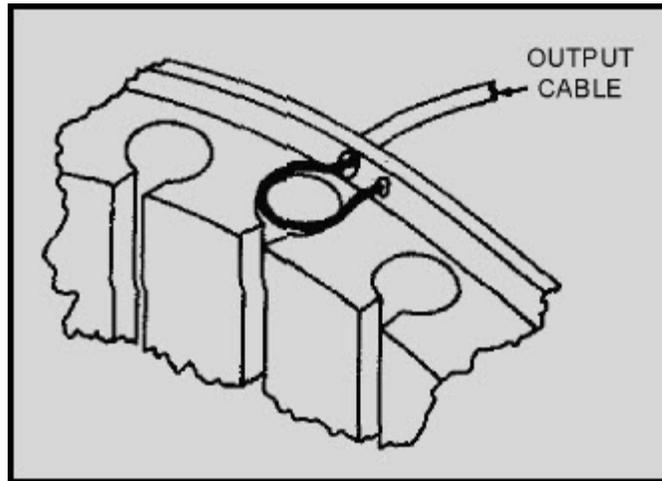


Figure 2-35B Magnetron coupling methods

The SEGMENT-FED LOOP METHOD is shown in figure 2-35C. The loop intercepts the magnetic lines passing between cavities. The STRAP-FED LOOP METHOD (figure 2-35D), intercepts the energy between the strap and the segment. On the output side, the coaxial line feeds another coaxial line directly or feeds a waveguide through a choke joint. The vacuum seal at the inner conductor helps to support the line. APERTURE, OR SLOT, COUPLING is illustrated in figure 2-35E. Energy is coupled directly to a waveguide through an iris.

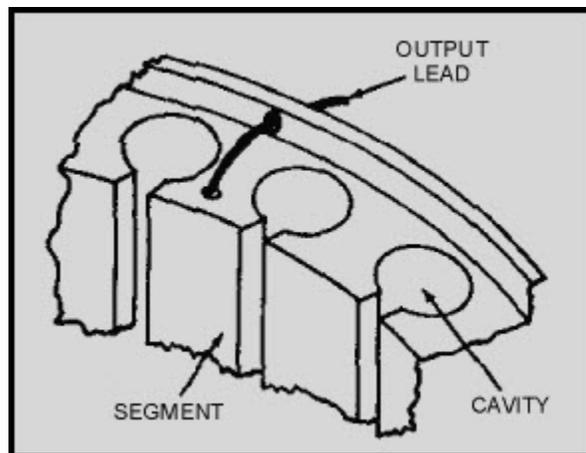


Figure 2-35C Magnetron coupling methods

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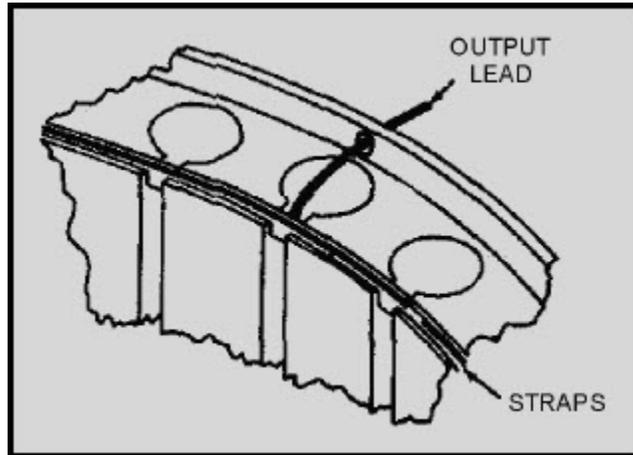


Figure 2-35D Magnetron coupling methods

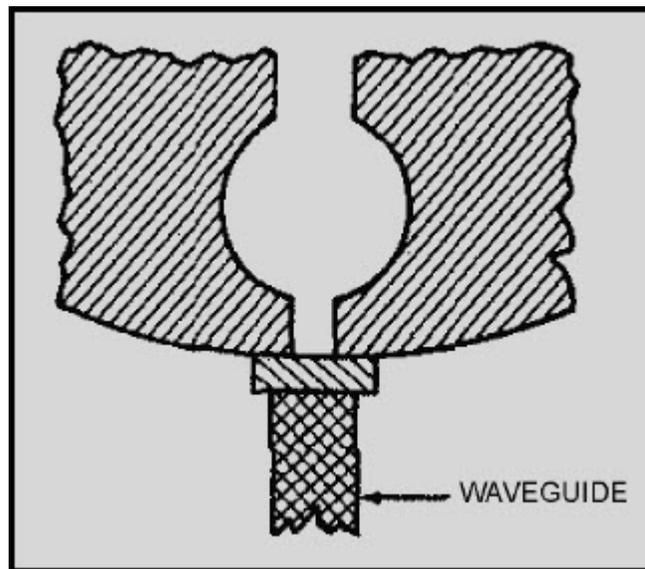


Figure 2-35E Magnetron coupling methods

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MAGNETRON TUNING - A tunable magnetron permits the system to be operated at a precise frequency anywhere within a band of frequencies, as determined by magnetron characteristics.

The resonant frequency of a magnetron may be changed by varying the inductance or capacitance of the resonant cavities. In figure 2-36, an inductive tuning element is inserted into the hole portion of the hole-and-slot cavities. It changes the inductance of the resonant circuits by altering the ratio of surface area to cavity volume in a high-current region. The type of tuner illustrated in figure 2-36 is called a SPROCKET TUNER or CROWN-OF-THORNS TUNER. All of its tuning elements are attached to a frame which is positioned by a flexible bellows arrangement. The insertion of the tuning elements into each anode hole decreases the inductance of the cavity and therefore increases the resonant frequency. One of the limitations of inductive tuning is that it lowers the unloaded Q of the cavities and therefore reduces the efficiency of the tube.

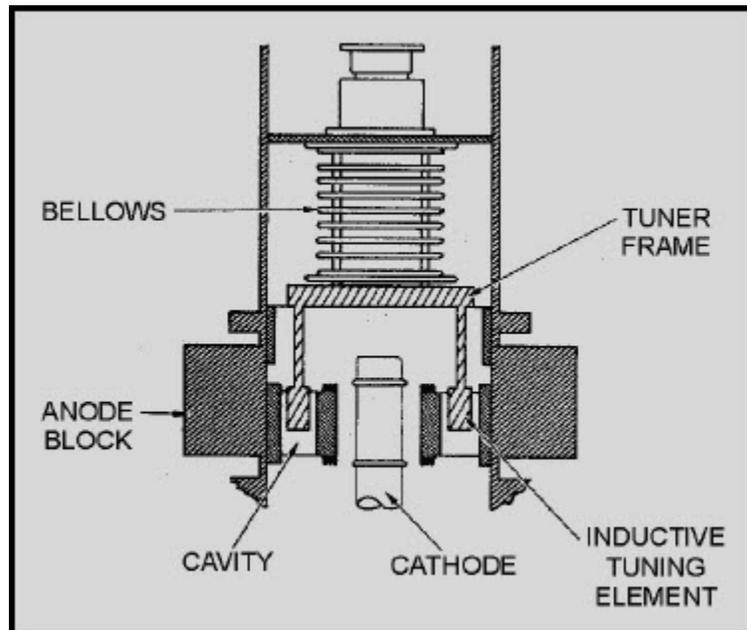


Figure 2-36 Inductive magnetron tuning

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The insertion of an element (ring) into the cavity slot, as shown in figure 2-37, increases the slot capacitance and decreases the resonant frequency. Because the gap is narrowed in width, the breakdown voltage is lowered. Therefore, capacitively tuned magnetrons must be operated with low voltages and at low-power outputs. The type of capacitive tuner illustrated in figure 2-37 is called a COOKIE-CUTTER TUNER. It consists of a metal ring inserted between the two rings of a double-strapped magnetron, which serves to increase the strap capacitance. Because of the mechanical and voltage breakdown problems associated with the cookie-cutter tuner, it is more suitable for use at longer wavelengths. Both the capacitance and inductance tuners described are symmetrical; that is, each cavity is affected in the same manner, and the pi mode is preserved.

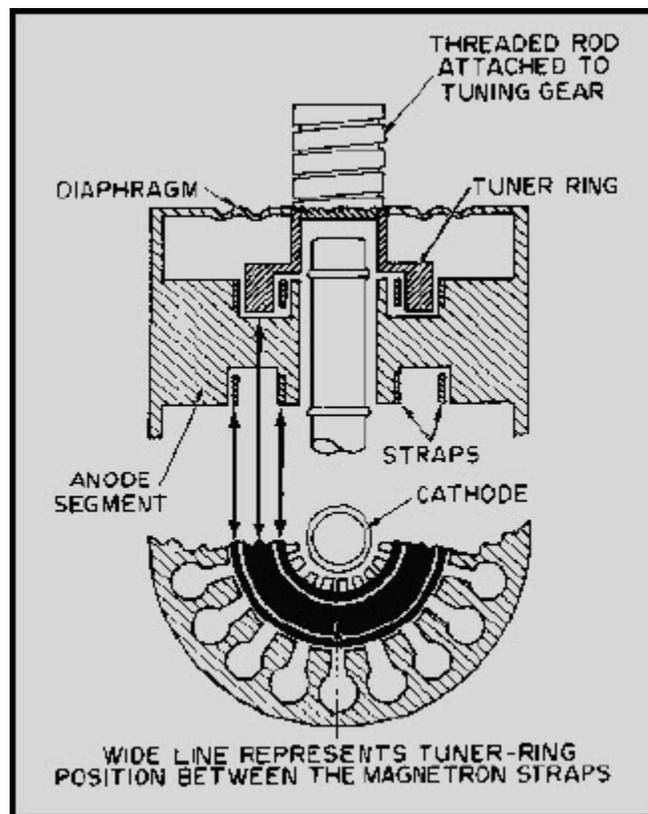


Figure 2-37 Capacitive magnetron tuning

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A 10-percent frequency range may be obtained with either of the two tuning methods described above. Also, the two tuning methods may be used in combination to cover a larger tuning range than is possible with either one alone.

ARCING IN MAGNETRONS - During initial operation a high-powered magnetron arcs from cathode to plate and must be properly **BROKEN IN** or **BAKED IN**. Actually, arcing in magnetrons is very common. It occurs with a new tube or following long periods of idleness.

One of the prime causes of arcing is the release of gas from tube elements during idle periods. Arcing may also be caused by the presence of sharp surfaces within the tube, mode shifting, and by drawing excessive current. While the cathode can withstand considerable arcing for short periods of time, continued arcing will shorten the life of the magnetron and may destroy it entirely. Therefore, each time excessive arcing occurs, the tube must be baked in again until the arcing ceases and the tube is stabilized.

The baking-in procedure is relatively simple. Magnetron voltage is raised from a low value until arcing occurs several times a second. The voltage is left at that value until arcing dies out. Then the voltage is raised further until arcing again occurs and is left at that value until the arcing again ceases. Whenever the arcing becomes very violent and resembles a continuous arc, the applied voltage is excessive and should be reduced to permit the magnetron to recover. When normal rated voltage is reached and the magnetron remains stable at the rated current, the baking-in is complete. A good maintenance practice is to bake-in magnetrons left idle in the equipment or those used as spares when long periods of non-operating time have accumulated.

The preceding information is general in nature. The recommended times and procedures in the technical manuals for the equipment should be followed when baking-in a specific type magnetron.

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2.3.8 The Crossed-Field Amplifier (Amplitron)

The CROSSED-FIELD AMPLIFIER (cfa), commonly known as an AMPLITRON and sometimes referred to as a PLATINOTRON, is a broadband microwave amplifier that can also be used as an oscillator. The cfa is similar in operation to the magnetron and is capable of providing relatively large amounts of power with high efficiency. The bandwidth of the cfa, at any given instant, is approximately plus or minus 5 percent of the rated center frequency. Any incoming signals within this bandwidth are amplified. Peak power levels of many megawatts and average power levels of tens of kilowatts average are, with efficiency ratings in excess of 70 percent, possible with crossed-field amplifiers.

Because of the desirable characteristics of wide bandwidth, high efficiency, and the ability to handle large amounts of power, the cfa is used in many applications in microwave electronic systems. This high efficiency has made the cfa useful for space-telemetry applications, and the high power and stability have made it useful in high-energy, linear atomic accelerators. When used as the intermediate or final stage in high-power radar systems, all of the advantages of the cfa are used.

Since the cfa operates in a manner so similar to the magnetron, the detailed theory is not presented in this module. Detailed information of cfa operation is available in NAVSHIPS 0967-443-2230, *Handling*, Installation and Operation of Crossed-Field Amplifiers. As mentioned earlier, crossed-field amplifiers are commonly called Amplitrons. You should note, however, that Amplitron is a trademark of the Raytheon Manufacturing Company for the Raytheon line of crossed-field amplifiers. An illustration of a crossed-field amplifier is shown in figure 2-38.

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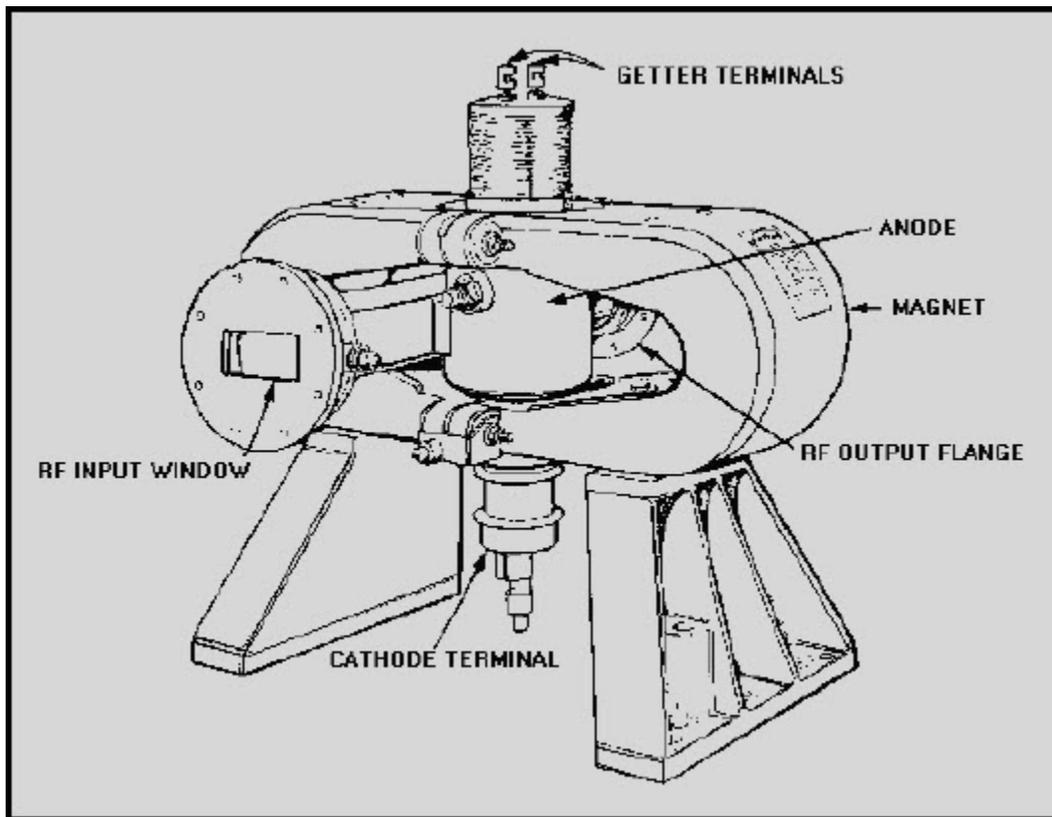


Figure 2-38 Crossed-field amplifier (Amplitron)

Q-44. Why is the pi mode the most commonly used magnetron mode of operation?

Q-45. What two methods are used to couple energy into and out of magnetrons?

Q-46. Magnetron tuning by altering the surface-to-volume ratio of the hole portion of a hole-and-slot cavity is what type of tuning?

Q-47. Capacitive tuning by inserting a ring into the cavity slot of a magnetron is accomplished by what type of tuning mechanism?

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2.4 SOLID-STATE MICROWAVE DEVICES

As with vacuum tubes, the special electronics effects encountered at microwave frequencies severely limit the usefulness of transistors in most circuit applications. The need for small-sized microwave devices has caused extensive research in this area. This research has produced solid-state devices with higher and higher frequency ranges. The new solid-state microwave devices are predominantly active, two-terminal diodes, such as tunnel diodes, varactors, transferred-electron devices, and avalanche transit-time diodes. This section will describe the basic theory of operation and some of the applications of these relatively new solid-state devices.

2.4.1 Tunnel Diode Devices

The TUNNEL DIODE is a pn junction with a very high concentration of impurities in both the p and n regions. The high concentration of impurities causes it to exhibit the properties of a negative-resistance element over part of its range of operation, as shown in the characteristic curve in figure 2-39. In other words, the resistance to current flow through the tunnel diode increases as the applied voltage increases over a portion of its region of operation. Outside the negative-resistance region, the tunnel diode functions essentially the same as a normal diode. However, the very high impurity density causes a junction depletion region so narrow that both holes and electrons can transfer across the pn junction by a quantum mechanical action called TUNNELING. Tunneling causes the negative-resistance action and is so fast that no transit-time effects occur even at microwave frequencies. The lack of a transit-time effect permits the use of tunnel diodes in a wide variety of microwave circuits, such as amplifiers, oscillators, and switching devices. The detailed theory of tunnel-diode operation and the negative-resistance property exhibited by the tunnel diode was discussed in *NEETS, Module 7, Introduction to Solid-State Devices and Power Supplies*, Chapter 3.

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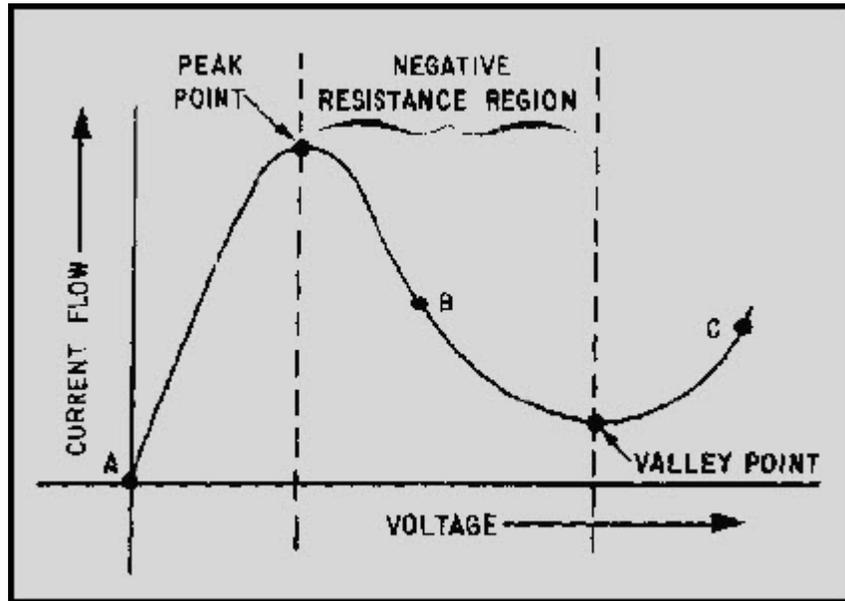


Figure 2-39 Tunnel-diode characteristic curve

TUNNEL-DIODE OSCILLATORS - A tunnel diode, biased at the center point of the negative-resistance range (point B in figure 2-39) and coupled to a tuned circuit or cavity, produces a very stable oscillator. The oscillation frequency is the same as the tuned circuit or cavity frequency.

Microwave tunnel-diode oscillators are useful in applications that require microwatts or, at most, a few milliwatts of power, such as local oscillators for microwave superheterodyne receivers. Tunnel-diode oscillators can be mechanically or electronically tuned over frequency ranges of about one octave and have a top-end frequency limit of approximately 10 gigahertz.

Tunnel-diode oscillators that are designed to operate at microwave frequencies generally use some form of transmission line as a tuned circuit. Suitable tuned circuits can be built from coaxial lines, transmission lines, and waveguides.

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An example of a highly stable tunnel-diode oscillator is shown in figure 2-40. A tunnel-diode is loosely coupled to a high-Q tunable cavity. Loose coupling is achieved by using a short, antenna feed probe placed off-center in the cavity. Loose coupling is used to increase the stability of the oscillations and the output power over a wider bandwidth.

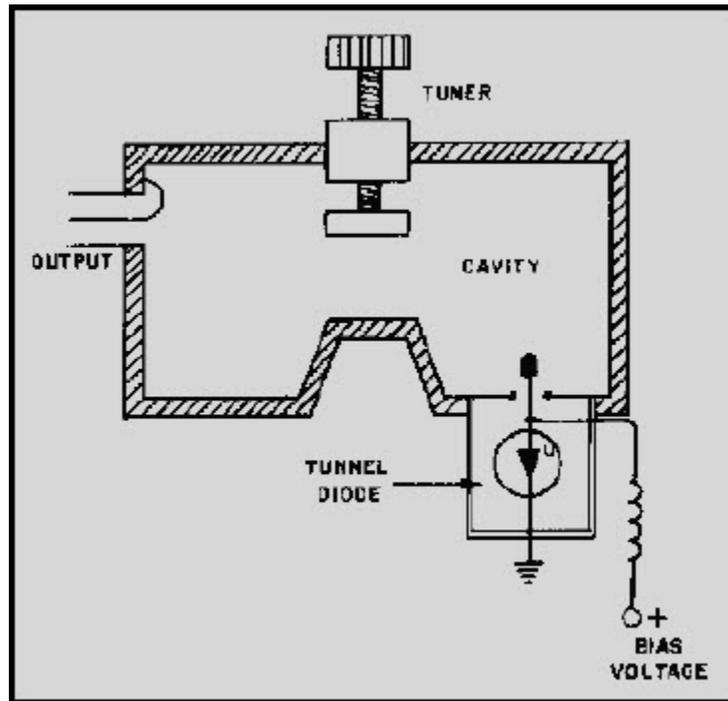


Figure 2-40 Tunnel-diode oscillator

The output power produced is in the range of a few hundred microwatts, sufficient for many microwave applications. The frequency at which the oscillator operates is determined by the physical positioning of the tuner screw in the cavity. Changing the output frequency by this method is called MECHANICAL TUNING. In addition to mechanical tuning, tunnel-diode oscillators may be tuned electronically. One method is called BIAS TUNING and involves nothing more than changing the bias voltage to change the bias point on the characteristic curve of the tunnel-diode. Another method is called VARACTOR TUNING and requires the addition of a varactor to the basic circuit. Varactors were discussed in NEETS, Module 7, Introduction to Solid-State Devices, and Power Supplies, Chapter 3. Tuning is achieved by changing the voltage applied across the varactor which alters the capacitance of the tuned circuit.

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TUNNEL-DIODE AMPLIFIERS - Low-noise, tunnel-diode amplifiers represent an important microwave application of tunnel diodes. Tunnel-diode amplifiers with frequencies up to 85 gigahertz have been built in waveguides, coaxial lines, and transmission lines. The low-noise generation, gain ratios of up to 30 dB, high reliability, and light weight make these amplifiers ideal for use as the first stage of amplification in communications and radar receivers.

Most microwave tunnel-diode amplifiers are REFLECTION-TYPE, CIRCULATOR-COUPLED AMPLIFIERS. As in oscillators, the tunnel diode is biased to the center point of its negative-resistance region, but a CIRCULATOR replaces the tuned cavity.

A circulator is a waveguide device that allows energy to travel in one direction only, as shown in figure 2-41. The tunnel diode in figure 2-41 is connected across a tuned-input circuit. This arrangement normally produces feedback that causes oscillations if the feedback is allowed to reflect back to the tuned-input circuit. The feedback is prevented because the circulator carries all excess energy to the absorptive load (R_L). In this configuration the tunnel diode cannot oscillate, but will amplify.

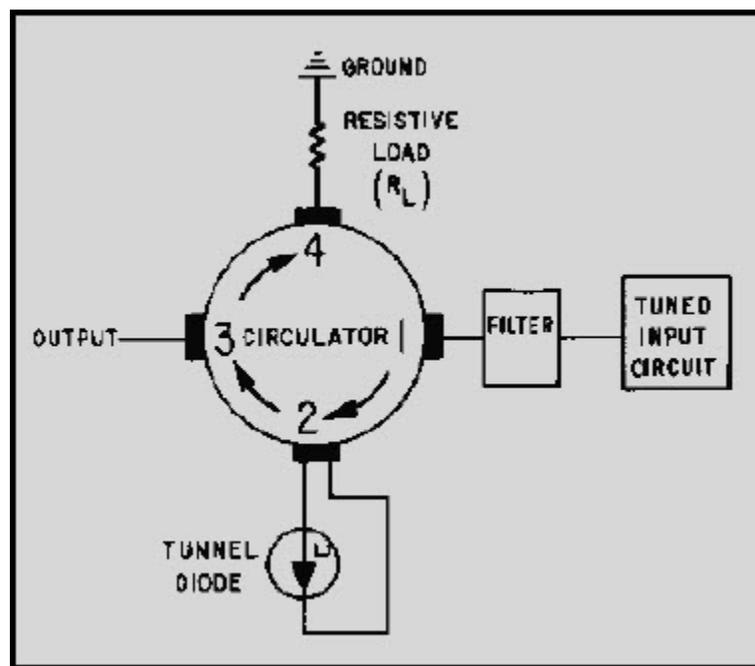


Figure 2-41 Tunnel-diode amplifier

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The desired frequency input signal is fed to port 1 of the circulator through a bandpass filter. The filter serves a dual purpose as a bandwidth selector and an impedance-matching device that improves the gain of the amplifiers. The input energy enters port 2 of the circulator and is amplified by the tunnel diode. The amplified energy is fed from port 2 to port 3 and on to the mixer. If any energy is reflected from port 3, it is passed to port 4, where it is absorbed by the matched load resistance.

TUNNEL-DIODE FREQUENCY CONVERTERS AND MIXERS - Tunnel diodes make excellent mixers and frequency converters because their current-voltage characteristics are highly nonlinear. While other types of frequency converters usually have a conversion power loss, tunnel-diode converters can actually have a conversion power gain. A single tunnel diode can also be designed to act as both the nonlinear element in a converter and as the negative-resistance element in a local oscillator at the same time.

Practical tunnel-diode frequency converters usually have either a unity conversion gain or a small conversion loss. Conversion gains as high as 20 dB are possible if the tunnel diode is biased near or into the negative-resistance region. Although high gain is useful in some applications, it presents problems in stability. For example, the greatly increased sensitivity to variations in input impedance can cause high-gain converters to be unstable unless they are protected by isolation circuitry.

As with tunnel-diode amplifiers, low-noise generation is one of the more attractive characteristics of tunnel-diode frequency converters. Low-noise generation is a primary concern in the design of today's extremely sensitive communications and radar receivers. This is one reason tunnel-diode circuits are finding increasingly wide application in these fields.

Q-48. Name the procedure used to reduce excessive arcing in a magnetron?

Q-49. What causes the negative-resistance property of tunnel diodes?

Q-50. What determines the frequency of a tunnel-diode oscillator?

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Q-51. Why is the tunnel diode loosely coupled to the cavity in a tunnel-diode oscillator?

Q-52. What is the purpose of the circulator in a tunnel-diode amplifier?

2.4.2 Varactor Devices

The VARACTOR is another of the active two-terminal diodes that operates in the microwave range. Since the basic theory of varactor operation was presented in *NEETS, Module 7, Introduction to Solid-State Devices and Power Supplies*, Chapter 3, only a brief review of the basic principles is presented here.

The varactor is a semiconductor diode with the properties of a voltage-dependent capacitor. Specifically, it is a variable-capacitance, pn-junction diode that makes good use of the voltage dependency of the depletion-area capacitance of the diode.

In figure 2-42A, two materials are brought together to form a pn-junction diode. The different energy levels in the two materials cause a diffusion of the holes and electrons through both materials which tends to balance their energy levels. When this diffusion process stops, the diode is left with a small area on either side of the junction, called the depletion area, which contains no free electrons or holes. The movement of electrons through the materials creates an electric field across the depletion area that is described as a barrier potential and has the electrical characteristics of a charged capacitor.

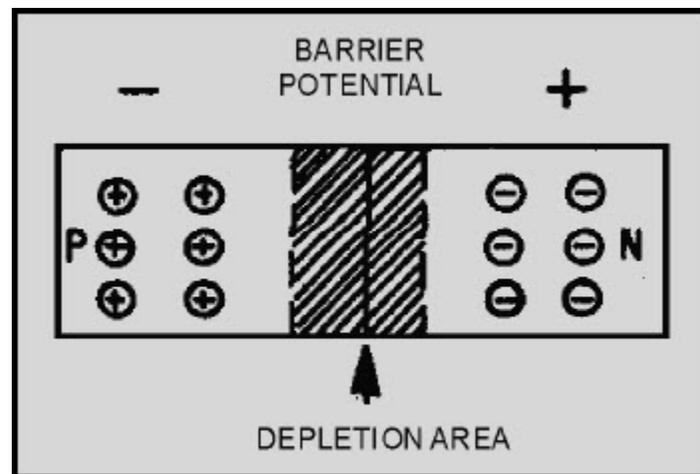


Figure 2-42A Pn-junction diode as a variable capacitor

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External bias, applied in either the forward or reverse direction, as shown in figure 2-42B and C, affects the magnitude, barrier potential, and width of the depletion area. Enough forward or reverse bias will overcome the barrier potential and cause current to flow through the diode. The width of the depletion region can be controlled by keeping the bias voltage at levels that do not allow current flow. Since the depletion area acts as a capacitor, the diode will perform as a variable capacitor that changes with the applied bias voltage. The capacitance of a typical varactor can vary from 2 to 50 picofarads for a bias variation of just 2 volts.

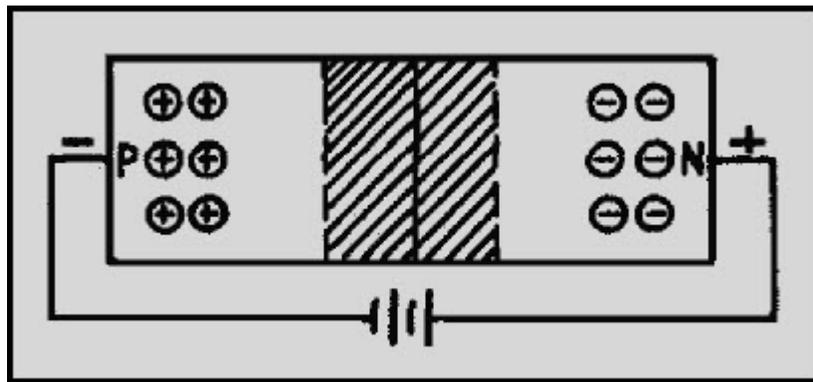


Figure 2-42B Pn-junction diode as a variable capacitor

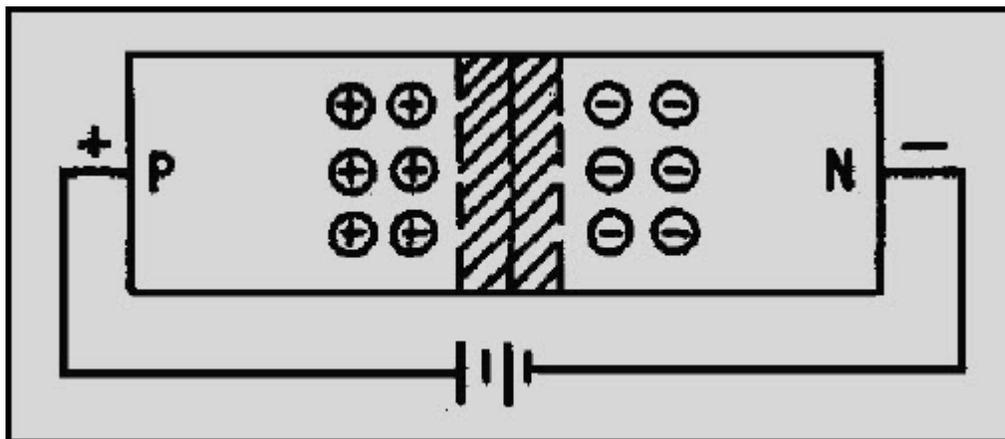


Figure 2-42C Pn-junction diode as a variable capacitor

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The variable capacitance property of the varactor allows it to be used in circuit applications, such as amplifiers, that produce much lower internal noise levels than circuits that depend upon resistance properties. Since noise is of primary concern in receivers, circuits using varactors are an important development in the field of low-noise amplification. The most significant use of varactors to date has been as the basic component in parametric amplifiers.

PARAMETRIC AMPLIFIERS - The parametric amplifier is named for the time-varying parameter, or value of capacitance, associated with the operation. Since the underlying principle of operation is based on reactance, the parametric amplifier is sometimes called a REACTANCE AMPLIFIER.

The conventional amplifier is essentially a variable resistance that uses energy from a dc source to increase ac energy. The parametric amplifier uses a nonlinear variable reactance to supply energy from an ac source to a load. Since reactance does not add thermal noise to a circuit, parametric amplifiers produce much less noise than most conventional amplifiers.

Because the most important feature of the parametric amplifier is the low-noise characteristic, the nature of ELECTRONIC NOISE and the effect of this type of noise on receiver operation must first be discussed. Electronic noise is the primary limitation on receiver sensitivity and is the name given to very small randomly fluctuating voltages that are always present in electronic circuits. The sensitivity limit of the receiver is reached when the incoming signal falls below the level of the noise generated by the receiver circuits. At this point the incoming signal is hidden by the noise, and further amplification has no effect because the noise is amplified at the same rate as the signal. The effects of noise can be reduced by careful circuit design and control of operating conditions, but it cannot be entirely eliminated. Therefore, circuits such as the parametric amplifier are important developments in the fields of communication and radar.

The basic theory of parametric amplification centers around a capacitance that varies with time. Consider the simple series circuit shown in figure 2-43. When the switch is closed, the capacitor charges to value (Q). If the switch is opened, the isolated capacitor has a voltage across the plates determined by the charge Q divided by the capacitance C.

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

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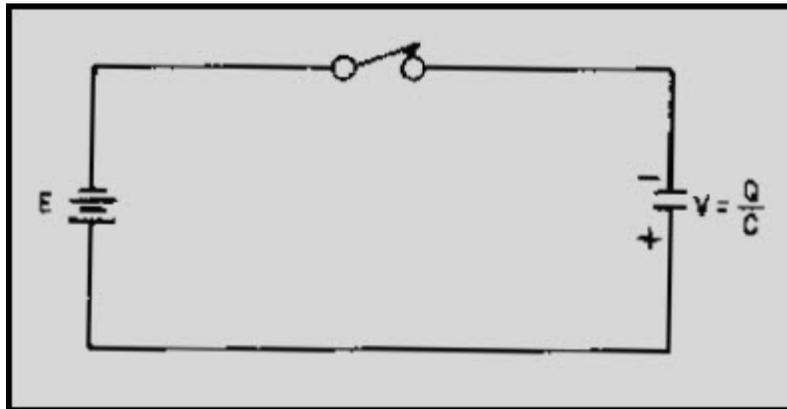


Figure 2-43 Voltage amplification from a varying capacitor

An increase in the charge Q or a decrease in the capacitance C causes an increase in the voltage across the plates. Thus, a voltage increase, or amplification, can be obtained by mechanically or electronically varying the amount of capacitance in the circuit. In practice a voltage-variable capacitance, such as a varactor, is used. The energy required to vary the capacitance is obtained from an electrical source called a PUMP.

Figure 2-44, view (A), shows a circuit application using a voltage-variable capacitor and a pump circuit. The pump circuit decreases the capacitance each time the input signal (E) across the capacitor reaches maximum. The decreased capacitance causes a voltage buildup as shown by the dotted line in view (B). Therefore, each time the pump decreases capacitance (view (C)), energy transfers from the pump circuit to the input signal. The step-by-step buildup of the input-signal energy level is shown in view (D).

Proper phasing between the pump and the input signal is crucial in this circuit. The electrical pump action is simply a sine-wave voltage applied to a varactor located in a resonant cavity. For proper operation, the capacitance must be decreased when the input voltage is maximum and increased when the input voltage is minimum. In other words, the pump signal frequency must be exactly double the frequency of the input signal. This relationship can be seen when you compare views (B) and (C). A parametric amplifier of the type shown in figure 2-44 is quite phase-sensitive. The input signal and the capacitor variation are often in the wrong phase for long periods of time.

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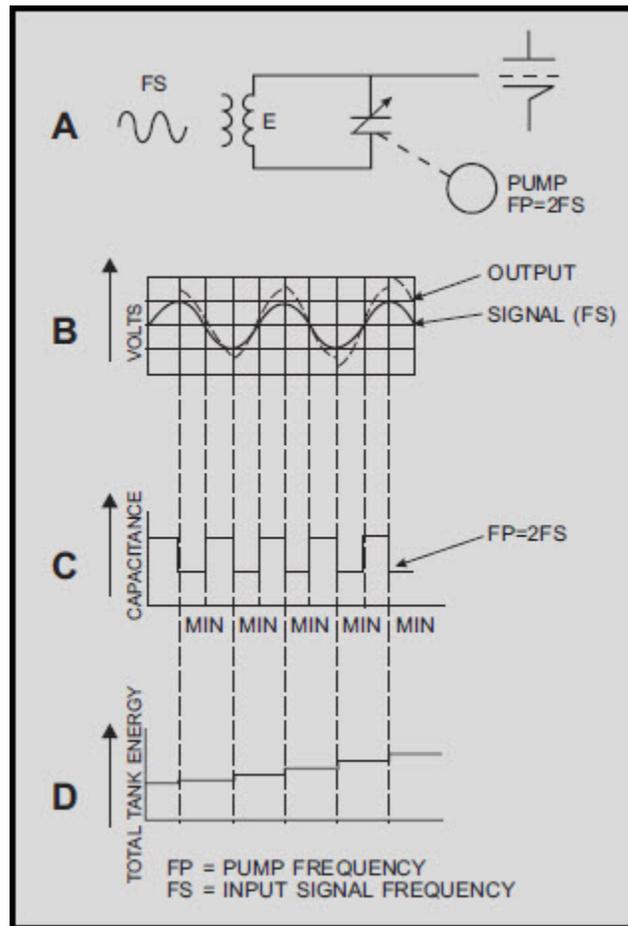


Figure 2-44 Energy transfer from pump signal to input signal

A parametric amplifier that is not phase-sensitive, referred to as a **NONDEGENERATIVE PARAMETRIC AMPLIFIER**, uses a pump circuit with a frequency higher than twice the input signal. The higher-frequency pump signal mixes with the input signal and produces additional frequencies that represent both the sum and difference of the input signal and pump frequencies.

Figure 2-45A, is a diagram of a typical nondegenerative parametric amplifier with the equivalent circuit shown in figure 2-45B. The pump signal (f_p) is applied to the varactor. The cavity on the left is resonant at the input frequency (f_s), and the cavity on the right is resonant at the difference frequency ($f_p - f_s$). The difference frequency is called the **IDLER-** or **LOWER-SIDEBAND** frequency. The varactor is located at the high-voltage points of the two cavities and is reverse biased by a small battery. The pump signal varies the bias above and below the fixed-bias level.

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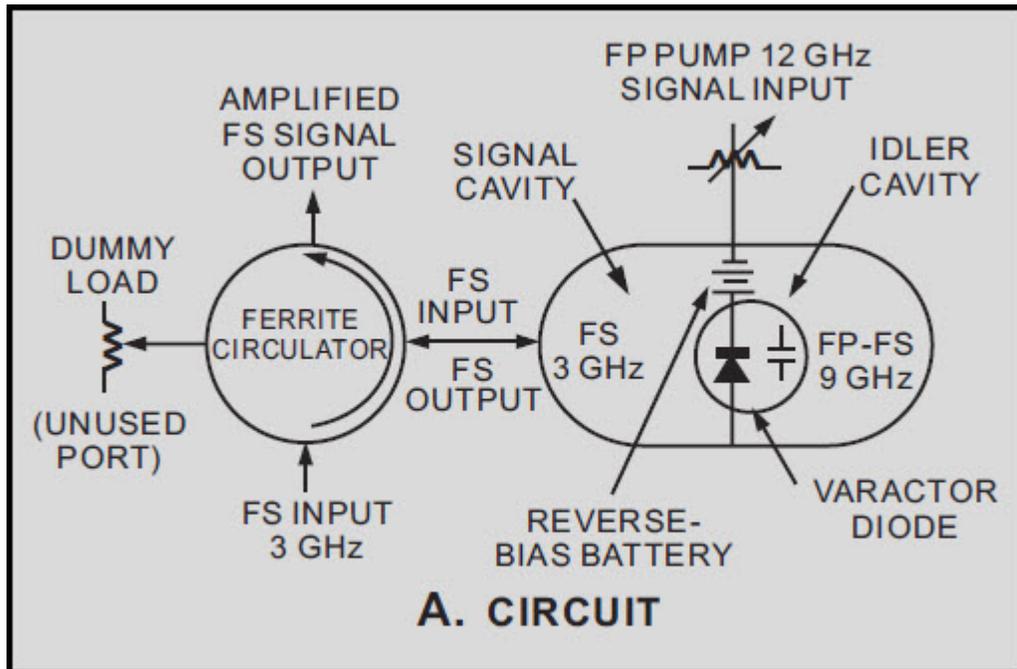


Figure 2-45A Nondegenerative parametric amplifier. CIRCUIT

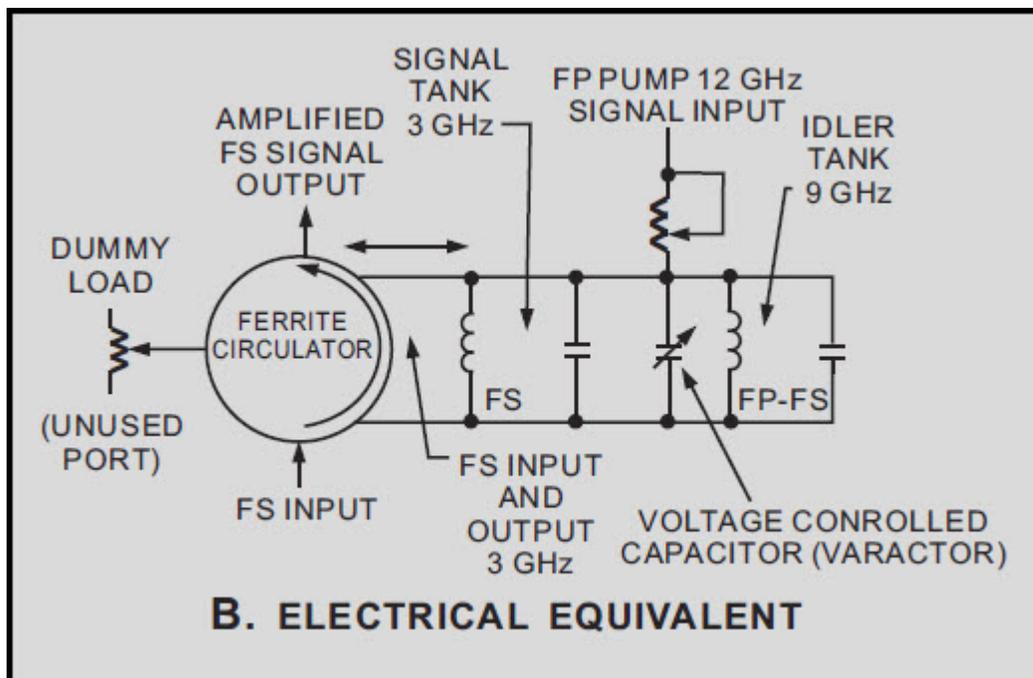


Figure 2-45B Nondegenerative parametric amplifier. ELECTRICAL EQUIVALENT

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The pump signal causes the capacitor in figure 2-45A to vary at a 12-gigahertz rate. The 3-gigahertz input signal enters via a four-port ferrite circulator, is developed in the signal cavity, and applied across the varactor. The nonlinear action of the varactor produces a 9-gigahertz difference frequency ($f_p - f_s$) with an energy-level higher than the original input signal.

The difference (idler) frequency is reapplied to the varactor to increase the gain and to produce an output signal of the correct frequency. The 9-gigahertz idler frequency recombines with the 12-gigahertz pump signal and produces a 3-gigahertz difference signal that has a much larger amplitude than the original 3-gigahertz input signal. The amplified signal is sent to the ferrite circulator for transfer to the next stage.

As with tunnel-diode amplifiers, the circulator improves stability by preventing reflection of the signal back into the amplifier. Reflections would be amplified and cause uncontrollable oscillations. The ferrite circulator also serves as an isolator to prevent source and load impedance changes from affecting gain.

Typically, the gain of a parametric amplifier is about 20 dB. The gain can be controlled with a variable attenuator that changes the amount of pump power applied to the varactor.

Parametric amplifiers are relatively simple in construction. The only component is a varactor diode placed in an arrangement of cavities and waveguides. The most elaborate feature of the amplifier is the mechanical tuning mechanism. Figure 2-46 illustrates an actual parametric amplifier.

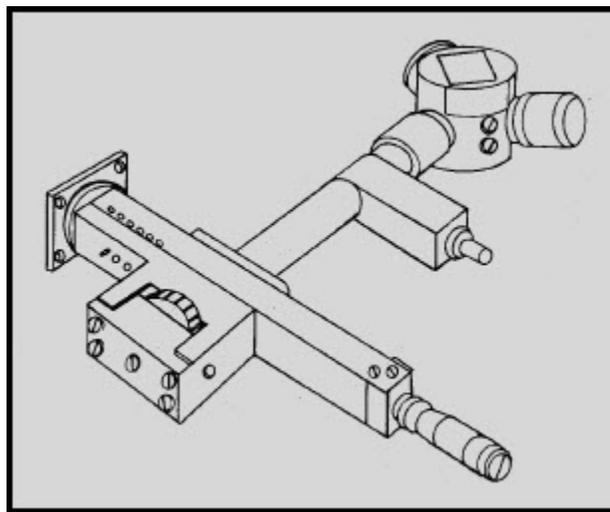


Figure 2-46 Parametric amplifier

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PARAMETRIC FREQUENCY CONVERTERS - Parametric frequency converters, using varactors, are of three basic types. The UPPER-SIDEBAND PARAMETRIC UP-CONVERTER produces an output frequency that is the SUM of the input frequency and the pump frequency. The LOWER-SIDEBAND PARAMETRIC DOWN-CONVERTER produces an output frequency that is the DIFFERENCE between the pump frequency and the input frequency. The DOUBLE-SIDEBAND PARAMETRIC UP-CONVERTER produces an output in which both the SUM and the DIFFERENCE of the pump and input frequencies are available.

Parametric frequency converters are very similar to parametric amplifiers in both construction and operation. Figure 2-47 is a functional diagram of a parametric down-converter.

The parametric frequency converter operates in the same manner as the parametric amplifier except that the sideband frequencies are not reapplied to the varactor. Therefore, the output is one or both of the sideband frequencies and is not the same as the input frequency. The output frequency is determined by the cavity used as an output. For example, the idler cavity in figure 2-47 could be replaced by a cavity that is resonant at the upper-sideband frequency (22 gigahertz) to produce an upper-sideband parametric up-converter. Since input and output signals are at different frequencies, the parametric frequency converter does not require a ferrite circulator. However, a ferrite isolator is used to isolate the converter from changes in source impedance.

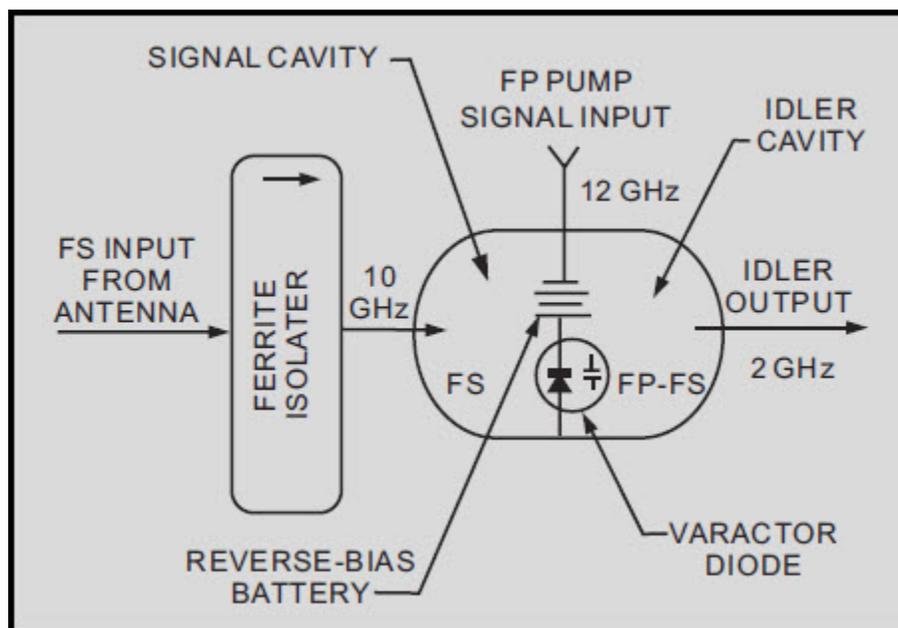


Figure 2-47 Lower-sideband parametric down-converter

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Q-53. What limits the usefulness of high-gain, tunnel-diode frequency converters?

Q-54. The varactor is a pn junction that acts as what type of electronic device?

Q-55. The underlying principle of operation of the parametric amplifier is based on what property?

Q-56. What is the most important feature of the parametric amplifier?

Q-57. How is amplification achieved in the circuit shown in figure 2-43?

Q-58. What is the purpose of the pump in a parametric amplifier?

Q-59. The pump signal frequency must be of what value when compared to the input signal of a simple parametric amplifier?

Q-60. What is the primary difference between the pump signal of a simple parametric amplifier and the pump signal of a nondegenerative parametric amplifier?

Q-61. In a nondegenerative parametric amplifier the difference between the input frequency and the pump frequency is called what?

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2.4.3 Bulk-Effect Semiconductors

BULK-EFFECT SEMICONDUCTORS are unlike normal pn-junction diodes in both construction and operation. Some types have no junctions and the processes necessary for operation occur in a solid block of semiconductor material. Other types have more than one junction but still use bulk-effect action. Bulk-effect devices are among the latest of developments in the field of microwave semiconductors and new applications are being developed rapidly. They seem destined to revolutionize the field of high-power, solid-state microwave generation because they can produce much larger microwave power outputs than any currently available pn-junction semiconductors.

Bulk-effect semiconductors are of two basic types: the transferred-electron devices and the avalanche transit-time devices.

TRANSFERRED-ELECTRON SEMICONDUCTORS - The discovery that microwaves could be generated by applying a steady voltage across a chip of n-type gallium-arsenide (GaAs) crystal was made in 1963 by J.B. Gunn. The device is operated by raising electrons in the crystal to conduction-band energy levels that are higher than the level they normally occupy. The overall effect is called the transferred-electron effect.

In a gallium-arsenide semiconductor, empty electron conduction bands exist that are at a higher energy level than the conduction bands occupied by most of the electrons. Any electrons that do occupy the higher conduction band essentially have no mobility. If an electric field of sufficient intensity is applied to the semiconductor electrons, they will move from the low-energy conduction band to the high-energy conduction band and become essentially immobile. The immobile electrons no longer contribute to the current flow and the applied voltage progressively increases the rate at which the electrons move from the low band to the high band. As the curve in figure 2-48 shows, the maximum current rate is reached and begins to decrease even though the applied voltage continues to increase. The point at which the current on the curve begins to decrease is called the THRESHOLD. This point is the beginning of the negative-resistance region. Negative resistance is caused by electrons moving to the higher conduction band and becoming immobile.

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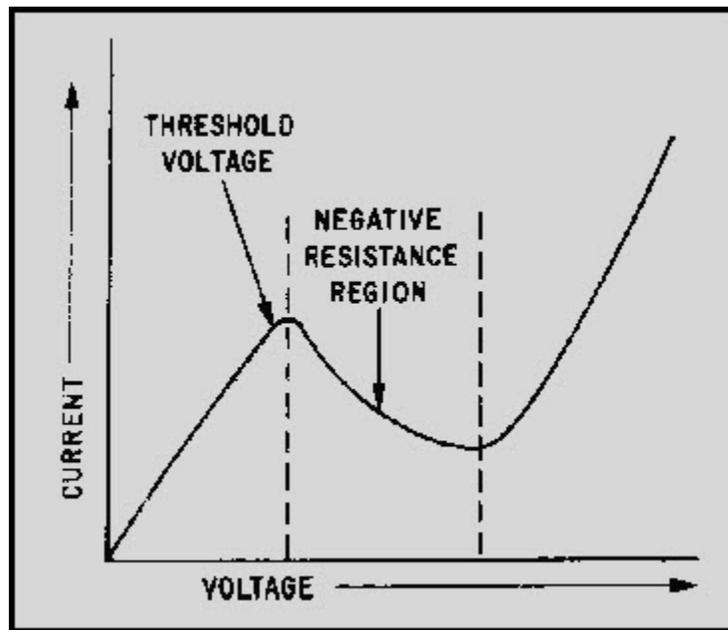


Figure 2-48 Characteristic curve for a bulk-effect semiconductor

If an increase in voltage is applied to a gallium-arsenide semiconductor, which is biased to operate in the negative-resistance region, it divides into regions of varying electric fields. A tiny region, known as a DOMAIN, forms that has an electric field of much greater intensity than the fields in the rest of the semiconductor. The applied voltage causes the domain to travel across the semiconductor chip from the cathode to the anode. The high field intensity of the domain is caused by the interaction of the slow electrons in the high-energy band and the faster electrons in the low-energy band. The electrons in the low-energy band travel faster than the moving domain and continually catch up during the transit from cathode to anode. When the fast electrons catch up to the domain, the high field intensity forces them into the higher band where they lose most of their mobility. This also causes them to fall behind the moving domain. Random scattering causes the electrons to lose some energy and drop back into the lower, faster, energy band and race again after the moving domain. The movement from the low-energy band to the high-energy band causes the electrons to bunch up at the back of the domain and to provide the electron-transfer energy that creates the high field intensity in the domain.

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The domains form at or near the cathode and move across the semiconductor to the anode, as shown in figure 2-49. As the domain disappears at the anode, a new domain forms near the cathode and repeats the process.

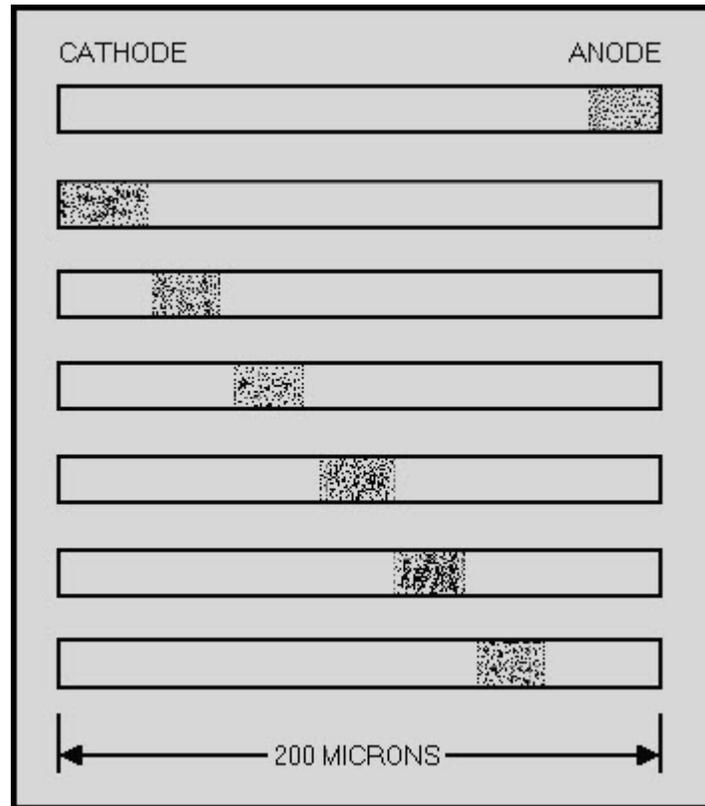


Figure 2-49 Gallium-arsenide semiconductor domain movement

The GUNN OSCILLATOR is a source of microwave energy that uses the bulk-effect, gallium-arsenide semiconductor. The basic frequency of a gunn oscillator is inversely proportional to the transit time of a domain across the semiconductor. The transit time is proportional to the length of semiconductor material, and to some extent, the voltage applied. Each domain causes a pulse of current at the output; thus, the output is a frequency determined by the physical length of the semiconductor chip.

The gunn oscillator can deliver continuous power up to about 65 milliwatts and pulsed outputs of up to about 200 watts peak. The power output of a solid chip is limited by the difficulty of removing heat from the small chip. Much higher power outputs have been achieved using wafers of gallium-arsenide as a single source.

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AVALANCHE TRANSIT-TIME DIODES - Avalanche transit-time diodes, also called IMPATT (Impact Avalanche and Transit-Time) diodes, are multilayer diodes of several different types used to generate microwave power. The earliest of the avalanche transit-time diodes consists of four layers in a pnin arrangement. The intrinsic (i) layer has neither p nor n properties.

The pn junction for the pnin diode, shown in figure 2-50, is strongly reverse biased to cause an avalanche in its depletion layer when the positive half cycle of a microwave signal is applied. The avalanche effect causes the electrons in the n region, which is very thin, to cross over to the intrinsic layer. The intrinsic layer is constructed so that the drift transit time causes the current to lag the signal voltage by more than 90 degrees at the desired frequency. Such a lag represents a negative resistance at the desired frequency. The pnin avalanche transit-time diode, when inserted in a microwave cavity with the proper dc bias, amplifies microwave signals introduced to the cavity.

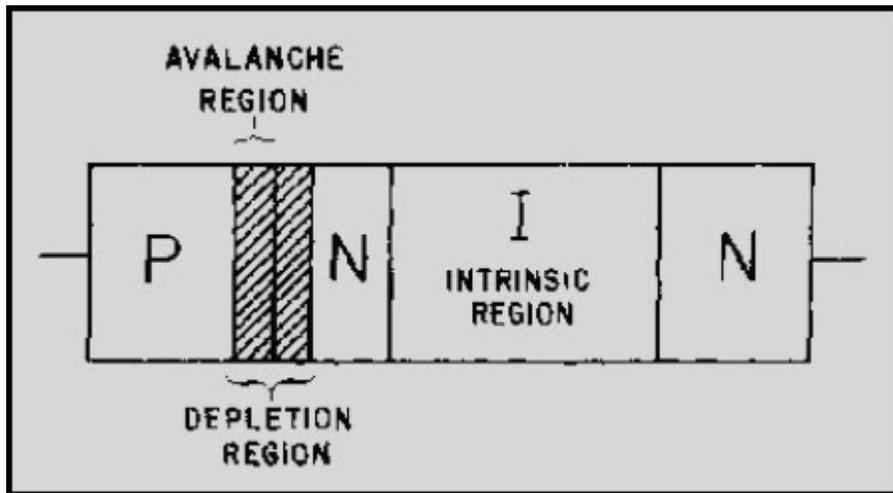


Figure 2-50 Avalanche transit time for a pnin diode

More recent research has shown that pin-junction diodes and simple pn-junction diodes can show negative resistance and amplification at microwave frequencies when they are reverse biased into an avalanche condition. The negative resistance in a simple pn-junction or pin diode is the result of a more complicated internal mechanism than in the pnin diode. The avalanche region and the drift region of the pnin diode are physically separate. Diodes of the pn and pin type must use the same physical region for both avalanche and drift-time control. In all types of avalanche transit-time diodes, the negative-resistance property causes dc bias energy to be absorbed by electrons in the avalanche process and given up to the applied microwave field.

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Q-62. What is the output frequency of an upper-sideband parametric-frequency converter?

Q-63. What is the primary advantage of bulk-effect devices over normal pn-junction semiconductors?

Q-64. What happens to the electrons of a gallium-arsenide semiconductor when they move from the normal low-energy conduction band to the high-energy conduction band?

Q-65. The point on the current curve of a gallium-arsenide semiconductor at which it begins to exhibit negative resistance is called what?

Q-66. The domain in a gallium-arsenide semiconductor has what type of electrical field when compared to the other regions across the body of a semiconductor?

Q-67. What characteristic of a gunn oscillator is inversely proportional to the transit time of the domain across the semiconductor?

Q-68. What is the junction arrangement of the original avalanche transit-time diode?

Q-69. What causes dc bias energy to be absorbed by avalanche electrons and given up to the microwave field applied to an avalanche transit-time diode?

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2.4.4 The Point-Contact Diode

POINT-CONTACT DIODES, commonly called CRYSTALS, are the oldest microwave semiconductor devices. They were developed during World War II for use in microwave receivers and are still in widespread use as receiver mixers and detectors.

Unlike the pn-junction diode, the point-contact diode depends on the pressure of contact between a point and a semiconductor crystal for its operation. Figure 2-51A and B, illustrate a point-contact diode. One section of the diode consists of a small rectangular crystal of n-type silicon. A fine beryllium-copper, bronze-phosphor, or tungsten wire called the CATWHISKER presses against the crystal and forms the other part of the diode. During the manufacture of the point contact diode, a relatively large current is passed from the catwhisker to the silicon crystal. The result of this large current is the formation of a small region of p-type material around the crystal in the vicinity of the point contact. Thus, a pn-junction is formed which behaves in the same way as a normal pn-junction.

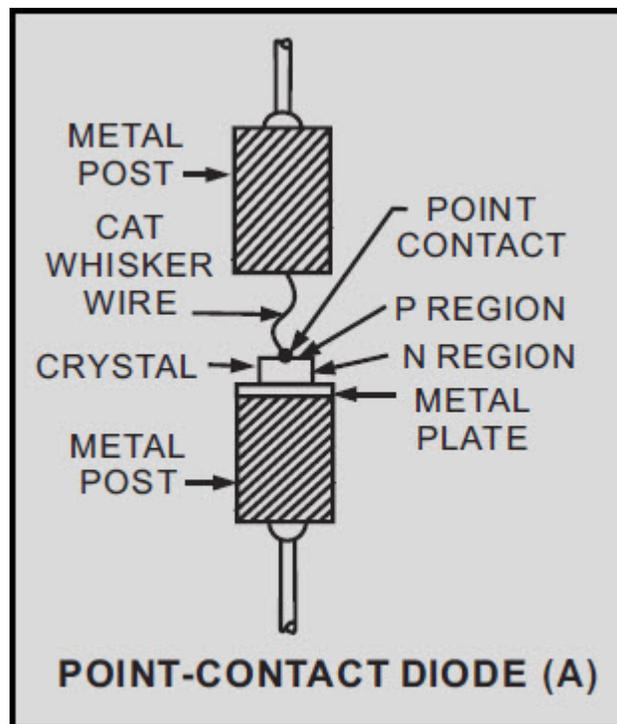


Figure 2-51A Point-contact diode. P REGION AROUND POINT

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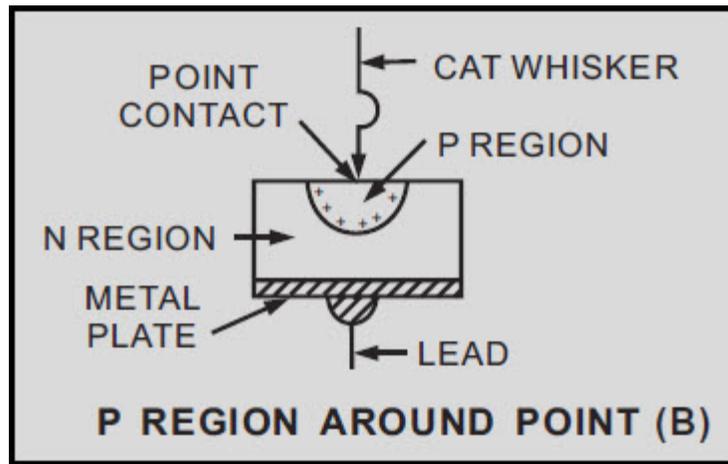


Figure 2-51B Point-contact diode. P REGION AROUND POINT

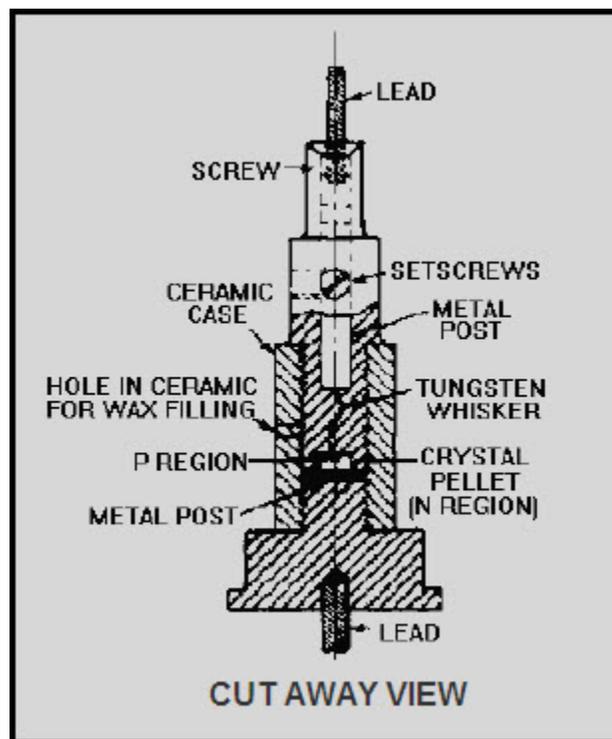
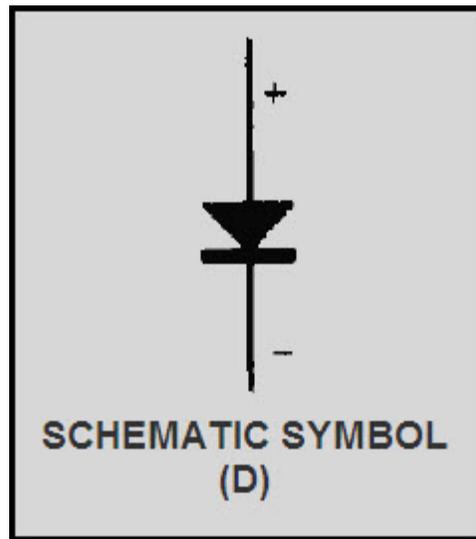


Figure 2-51C Point-contact diode. CUT AWAY VIEW

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**Figure 2-51D Point-contact diode.
SCHEMATIC SYMBOL**

The pointed wire is used instead of a flat metal plate to produce a high-intensity electric field at the point contact without using a large external source voltage. It is not possible to apply large voltages across the average semiconductor because of the excessive heating.

The end of the catwhisker is one of the terminals of the diode. It has a low-resistance contact to the external circuit. A flat metal plate on which the crystal is mounted forms the lower contact of the diode with the external circuit. Both contacts with the external circuit are low-resistance contacts.

The characteristics of the point-contact diode under forward and reverse bias are somewhat different from those of the junction diode.

With forward bias, the resistance of the point-contact diode is higher than that of the junction diode. With reverse bias, the current flow through a point-contact diode is not as independent of the voltage applied to the crystal as it is in the junction diode. The point-contact diode has an advantage over the junction diode because the capacitance between the catwhisker and the crystal is less than the capacitance between the two sides of the junction diode. As such, the capacitive reactance existing across the point-contact diode is higher and the capacitive current that will flow in the circuit at high frequencies is smaller. A cutaway view of the entire point-contact diode is shown in figure 2-51C. The schematic symbol of a point-contact diode is shown in figure 2-51D.

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2.4.5 Schottky Barrier Diode

The SCHOTTKY BARRIER DIODE is actually a variation of the point-contact diode in which the metal semiconductor junction is a surface rather than a point contact. The large contact area, or barrier, between the metal and the semiconductor in the Schottky barrier diode provides some advantages over the point-contact diode. Lower forward resistance and lower noise generation are the most important advantages of the Schottky barrier diode. The applications of the Schottky barrier diode are the same as those of the point-contact diode. The low noise level generated by Schottky diodes makes them especially suitable as microwave receiver detectors and mixers.

The Schottky barrier diode is sometimes called the HOT-ELECTRON or HOT-CARRIER DIODE because the electrons flowing from the semiconductor to the metal have a higher energy level than the electrons in the metal. The effect is the same as it would be if the metal were heated to a higher temperature than normal. Figure 2-52 is an illustration of the construction of a Schottky barrier diode.

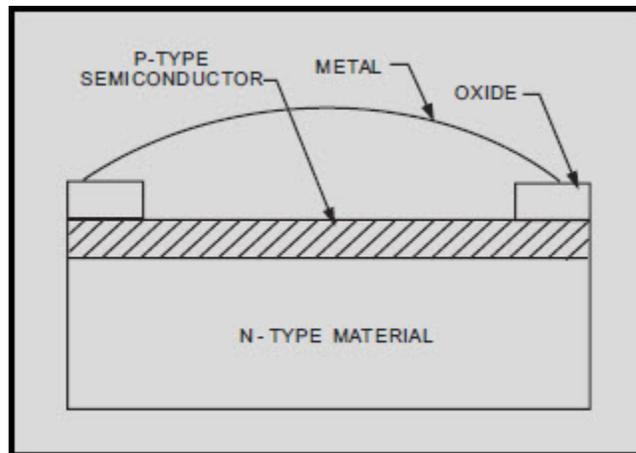


Figure 2-52 Schottky-barrier diode

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2.4.6 PIN Diodes

The pin diode consists of two narrow, but highly doped, semiconductor regions separated by a thicker, lightly-doped material called the intrinsic region. As suggested in the name, pin, one of the heavily doped regions is p-type material and the other is n-type. The same semiconductor material, usually silicon, is used for all three areas. Silicon is used most often for its power-handling capability and because it provides a highly resistive intrinsic (i) region. The pin diode acts as an ordinary diode at frequencies up to about 100 megahertz, but above this frequency the operational characteristics change.

The large intrinsic region increases the transit time of electrons crossing the region. Above 100 megahertz, electrons begin to accumulate in the intrinsic region. The carrier storage in the intrinsic region causes the diode to stop acting as a rectifier and begin acting as a variable resistance. The equivalent circuit of a pin diode at microwave frequencies is shown in figure 2-53A. A resistance versus voltage characteristic curve is shown in figure 2-53B.

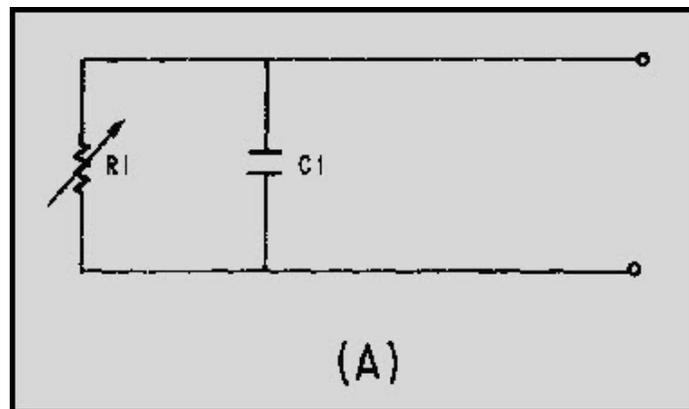


Figure 2-53A Diode equivalent circuit (pin)

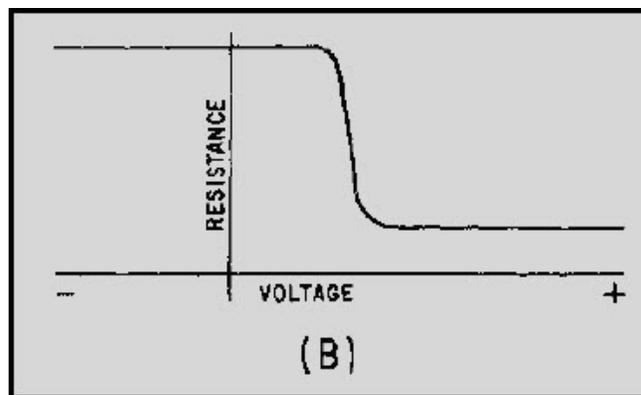


Figure 2-53B Diode equivalent circuit (pin)

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When the bias on a pin diode is varied, the microwave resistance changes from a typical value of 6 kilohms under negative bias to about 5 ohms when the bias is positive. Thus when the diode is mounted across a transmission line or waveguide, the loading effect is insignificant while the diode is reverse biased, and the diode presents no interference to power flow. When the diode is forward biased, the resistance drops to approximately 5 ohms and most power is reflected. In other words, the diode acts as a switch when mounted in parallel with a transmission line or waveguide. Several diodes in parallel can switch power in excess of 150 kilowatts peak. The upper power limit is determined by the ability of the diode to dissipate power. The upper frequency limit is determined by the shunt capacitance of the pn junction, shown as C1 in figure 2-53A. Pin diodes with upper limit frequencies in excess of 30 gigahertz are available.

Q-70. During the manufacture of a point-contact diode, what is the purpose of passing a relatively large current from the catwhisker to the silicon crystal?

Q-71. What is the capacitive reactance across a point-contact diode as compared to a normal junction diode?

Q-72. What are the most important advantages of the Schottky barrier diode?

Q-73. At frequencies above 100 megahertz, the intrinsic (i) region causes a pin diode to act as what?

Q-74. The pin diode is primarily used for what purpose?

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2.4.7 Microwave Transistors

Transistors, like vacuum tubes, have had a very limited application in the microwave range. Many of the same problems encountered with vacuum tubes, such as transit-time effects, also limit the upper frequency range of transistors. However, research in the area of microwave transistors, and especially MICROWAVE INTEGRATED CIRCUITS (ICs), is proceeding rapidly.

GALLIUM-ARSENIDE FET AMPLIFIERS have been developed which provide low-noise amplification up to about 30 dB in the 7- to 18-gigahertz range. The power output of many of these amplifiers is relatively low, approximately 20 to 200 milliwatts, but that is satisfactory for many microwave applications. Research has extended both the frequency range and the power output of gallium-arsenide FET amplifiers to frequencies as high as 26.5 gigahertz and power levels in excess of 1 watt in multistage amplifiers.

SILICON BIPOLAR-TRANSISTOR AMPLIFIERS in integrated circuit form have been developed that provide up to 40 watts peak power in the 1- to 1.5-gigahertz range. Other types of microwave transistor amplifiers combined into multistage modules are capable of providing power outputs approaching 100 watts.

Microwave transistor amplifiers, because of their stability, light weight, and long life, are rapidly replacing microwave tubes in the first stages of high-powered radar and communications transmitters. In the future new systems will be almost completely solid state.

2.5 SUMMARY

The information that follows summarizes the important points presented in this chapter.

The use of microwave frequencies forced the development of special tubes to offset the limitations caused by interelectrode capacitance, lead inductance, and electron transit-time effects in conventional tubes. Microwave tubes, such as the klystron and twt, take advantage of transit-time effects through the use of VELOCITY MODULATION to amplify and generate microwave energy.

The **KLYSTRON** is a velocity-modulated tube which may be used as an amplifier or oscillator. The klystron, when used as an amplifier, requires at least two resonant cavities, the buncher and the catcher.

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The **REFLEX KLYSTRON** is used only as an oscillator and uses only one cavity to bunch and collect the electrons. The frequency is determined by the size and shape of the cavity. The reflex klystron has several possible modes of operation which are determined by electron transit time. Electron transit time is controlled by the **REPELLER** voltage.

The **TWT** is a wide-bandwidth, velocity-modulated tube used primarily as an amplifier. The electron beam is bunched by a signal applied to the **HELIX**. The bunching causes an energy transfer from the electron beam to the traveling wave on the helix.

The **MAGNETRON** is a **DIODE OSCILLATOR** capable of delivering microwave energy at very high power levels. Three fields exist within a magnetron that influence operation: (1) the **DC ELECTRIC FIELD** between the anode and cathode; (2) the **AC ELECTRIC FIELD** produced by the oscillating resonant cavities and on the same plane as the dc field; and (3) the **MAGNETIC FIELD** produced by the permanent magnet which is perpendicular to the dc electric field.

Magnetrons are of two basic types, the **NEGATIVE-RESISTANCE MAGNETRON** and the **ELECTRON-RESONANCE MAGNETRON**.

SOLID-STATE MICROWAVE DEVICES are becoming increasingly widespread in microwave equipment with new developments almost daily. Most of the currently available solid-state devices are two-terminal diodes with the capability to generate or amplify microwave energy. Many of the solid-state devices, such as the **TUNNEL DIODE** and the **BULK-EFFECT DIODE**, apply the property of **NEGATIVE RESISTANCE** to amplify microwave signals or generate microwave energy.

The **VARACTOR** is a two-terminal diode that acts as a variable capacitance and is the active element of **PARAMETRIC AMPLIFIERS**. The parametric amplifier is a low-noise microwave amplifier that uses variable reactance to amplify microwave signals.

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ANSWERS TO QUESTIONS Q1. THROUGH Q74.

A-1. *Impedance decreases.*

A-2. *Degenerative feedback.*

A-3. *Transit time causes the grid voltage and plate current to be out of phase.*

A-4. *Transit time.*

A-5. *Velocity.*

A-6. *The electron will be accelerated.*

A-7. *By alternately speeding up or slowing down the electrons.*

A-8. *The buncher grids.*

A-9. *There is no effect.*

A-10. *The frequency period of the buncher grid signal.*

A-11. *Velocity modulation.*

A-12. *The accelerator grid and the buncher grids.*

A-13. *The catcher cavity.*

A-14. *Amplifier.*

A-15. *Intermediate cavities between the input and output cavities.*

A-16. *A large negative pulse is applied to the cathode.*

A-17. *The middle cavity.*

A-18. *The bandwidth decreases.*

A-19. *Stagger tuning.*

A-20. *The reflector or repeller.*

A-21. *Velocity.*

A-22. *Three-quarter cycle.*

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A-23. Mode 2.

A-24. Power is reduced.

A-25. The half-power points of the mode.

A-26. Voltage amplification.

A-27. Used to focus the electrons into a tight beam.

A-28. The directional couplers are not physically connected to the helix.

A-29. The traveling wave must have a forward velocity equal to or less than the speed of the electrons in the beam.

A-30. The helix.

A-31. Helix.

A-32. A magnetic field.

A-33. Anode or plate.

A-34. The resonant cavities.

A-35. The permanent magnet.

A-36. The critical value of field strength.

A-37. Circular.

A-38. The negative-resistance magnetron has a split plate.

A-39. The application of the proper magnetic field.

A-40. To reduce the effects of filament bombardment.

A-41. Rising-sun block.

A-42. Series.

A-43. Working electrons.

A-44. Greater power output.

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A-45. Loops and slots.

A-46. Inductive.

A-47. A cookie-cutter tuner.

A-48. Baking in.

A-49. The tunneling action.

A-50. The tuned circuit or cavity frequency.

A-51. To increase the stability.

A-52. Prevent feedback to the tuned input circuit.

A-53. Stability problems.

A-54. Variable capacitor.

A-55. Reactance.

A-56. The low-noise characteristic.

A-57. By varying the amount of capacitance in the circuit.

A-58. Supplies the electrical energy required to vary the capacitance.

A-59. Exactly double the input frequency.

A-60. The pump signal of a nondegenerative parametric amplifier is higher than twice the input signal.

A-61. Idler- or lower-sideband frequency.

A-62. The sum of the input frequency and the pump frequency.

A-63. Larger microwave power outputs.

A-64. The electrons become immobile.

A-65. Threshold.

A-66. A field of much greater intensity.

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A-67. The frequency.

A-68. P_{nin}.

A-69. The negative-resistance property.

A-70. To form a small region of p-type material.

A-71. Lower.

A-72. Lower forward resistance and low noise.

A-73. Variable resistance.

A-74. A switching device.

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3 MICROWAVE ANTENNAS

LEARNING OBJECTIVES

Upon completion of this chapter the student will be able to:

1. Explain the basic characteristics of coupling, directivity, reciprocity, and efficiency in microwave antennas.
2. Describe the construction and basic theory of operation of reflector antennas and horn radiators.
3. Explain construction and operation of microwave lens antennas.
4. Describe the construction and theory of operation of driven and parasitic antenna arrays.
5. Explain the basic operation and applications of frequency-sensitive antennas.

3.1 INTRODUCTION

In this chapter you will study the general characteristics of microwave antennas that are widely used in radar and communications applications. The basic principles of operation of microwave antennas are similar to those of antennas used at lower frequencies. You might want to review the principles presented in NEETS, Module 10, Introduction to Wave Propagation, Transmission Lines, and Antennas, at this time. Pay particular attention to basic antenna principles in chapter 4 for a review of microwave antennas. Antennas are devices used to radiate electromagnetic energy into space. The characteristics of transmitting and receiving antennas are similar, so a good transmitting antenna is often a good receiving antenna. A single antenna performs both functions in many modern applications.

3.2 ANTENNA CHARACTERISTICS

Since the operating principles of low-frequency and microwave antennas are essentially the same, the electrical characteristics are also very similar. You will need a fundamental knowledge of radar and communications antenna electrical theory in your shipboard antenna maintenance work. Antenna theory is primarily a design consideration of antenna size and shape requirements that depend on the frequency used. A brief description of antenna electrical characteristics is sufficient for the needs of most students of electronics.

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3.2.1 Antenna Efficiency

The effectiveness of an antenna depends upon its ability to couple or radiate energy into the air. An efficient antenna is one which wastes very little energy during the radiation process. The efficiency of an antenna is usually referred to as the **POWER GAIN** or **POWER RATIO** as compared to a standard reference antenna. The power gain of an antenna is a ratio of the radiated power to that of the reference antenna, which is usually a basic dipole. Both antennas must be fed rf energy in the same manner and must be in the same position when the energy is radiated. The power gain of a single dipole without a reflector is unity (one). An array of several dipoles in the same position as the single dipole, and fed with the same line, has a power gain of more than one.

The effectiveness of an entire transmitting/ receiving system depends largely on impedance matching between the elements of the system. Impedance matching is particularly critical at the antenna connection. If a good impedance match is maintained between the system and the antenna throughout the operating frequency band, power transfer to and from the antenna is always maximum. The transmission line or waveguide used to transport energy to and from the antenna should have a characteristic impedance equal to that of the antenna. A proper impedance match allows all available power to be absorbed and radiated by the antenna without reflections back down the line.

If you have a transmission line or waveguide with an impedance mismatch at the termination, standing waves are set up by the reflections. Standing waves cause losses in the form of unwanted radiations, heat losses in transmission lines, and arcing in waveguides.

The **STANDING-WAVE RATIO**, abbreviated **swr**, is a way to measure the degree of mismatch between the transmission line and its load. The **swr** can be expressed as a ratio of the maximum and minimum values of the current or voltage in the standing waves that are set up on the lines as follows:

$$v_{swr} = \frac{E_{max}}{E_{min}}$$

or

$$i_{swr} = \frac{I_{max}}{I_{min}}$$

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A transmission line or waveguide approaches a perfectly matched condition when the swr approaches a value of 1. A ratio that is a little higher than 1 is usually acceptable in practical applications.

Measurement of swr is the only practical method of detecting an impedance mismatch between a transmitting/receiving system and its antenna. As such, the system swr is an important indication of the overall efficiency of the system during operation.

The line impedance can usually be matched to the antenna at only one frequency. However, the swr will NOT become too high if the antenna is used over a small range of frequencies and the line is matched to the center frequency.

3.2.2 Antenna Directivity

You can divide antennas into two general classes based on directivity, omnidirectional and directional. OMNIDIRECTIONAL antennas radiate and receive energy from all directions at once (SPHERICAL WAVEFRONT). They are seldom used in modern radar systems as the primary antenna, but are commonly used in radio equipment and iff (identification friend or foe) receivers. DIRECTIONAL antennas radiate energy in LOBES (or BEAMS) that extend outward from the antenna in either one or two directions. The radiation pattern contains small minor lobes, but these lobes are weak and normally have little effect on the main radiation pattern. Directional antennas also receive energy efficiently from only one or two directions, depending upon whether it is unidirectional or bidirectional.

Directional antennas have two characteristics that are important to you in radar and communications systems. One is DIRECTIVITY and the other is POWER GAIN. The directivity of an antenna refers to the NARROWNESS of the radiated beam. If the beam is NARROW in either the horizontal or vertical plane, the antenna has a high degree of directivity in that plane. An antenna may be designed for high directivity in one plane only or in both planes, depending on the application. The power gain of an antenna increases as the degree of directivity increases because the power is concentrated into a narrow beam and less power is required to cover the same distance.

Since microwave antennas are predominantly unidirectional, the examples you will study in this chapter are all of the unidirectional type.

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3.2.3 Reciprocity

You read in this chapter that an antenna is able to both transmit and receive electromagnetic energy. This is known as RECIPROCITY. Antenna reciprocity is possible because antenna characteristics are essentially the same regardless of whether an antenna is transmitting or receiving electromagnetic energy. Reciprocity allows most radar and communications systems to operate with only one antenna. An automatic switch, called a DUPLEXER, connects either the transmitter or the receiver to the antenna at the proper time. Duplexer operation will be explained in later *NEETS* modules dealing with radar and communications systems. Because of the reciprocity of antennas, this chapter will discuss antennas from the viewpoint of the transmitting cycle. However, you should understand that the same principles apply on the receiving cycle.

3.2.4 Radar Fundamentals

Radio, television, radar, and the human eye have much in common because they all process the same type of electromagnetic energy. The major difference between the light processed by the human eye and the radio-frequency energy processed by radio and radar is frequency. For example, radio transmitters send out signals in all directions. These signals can be detected by receivers tuned to the same frequency. Radar works somewhat differently because it uses reflected energy (echo) instead of directly transmitted energy. The echo, as it relates to sound, is a familiar concept to most of us. An experienced person can estimate the distance and general direction of an object causing a sound echo. Radar uses microwave electromagnetic energy in much the same way.

Radar transmits microwave energy that reflects off an object and returns to the radar. The returned portion of the energy is called an ECHO, as it is in sound terminology. It is used to determine the direction and distance of the object causing the reflection. Determination of direction and distance to an object is the primary function of most radar systems.

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Telescopes and radars, in terms of locating objects in space, have many common problems. Both have a limited field of view and both require a geographic reference system to describe the position of an object (target). The position of an object viewed with a telescope is usually described by relating it to a familiar object with a known position. Radar uses a standard system of reference coordinates to describe the position of an object in relation to the position of the radar. Normally ANGULAR measurements are made from true north in an imaginary flat plane called the HORIZONTAL PLANE. All angles in the UP direction are measured in a second imaginary plane perpendicular to the horizontal plane called the VERTICAL PLANE. The center of the coordinate system is the radar location. As shown in figure 3-1, the target position with respect to the radar is defined as 60 degrees true, 10 degrees up, and 10 miles distant. The line directly from the radar to the target is called the LINE OF SIGHT. The distance from point 1 to point 2, measured along the line of sight, is called TARGET RANGE. The angle between the horizontal plane and the line of sight is known as the ELEVATION ANGLE. The angle measured in a clockwise direction in the horizontal plane between true north and the line of sight is known as BEARING (sometimes referred to as AZIMUTH). These three coordinates of range, bearing, and elevation determine the location of the target with respect to the radar.

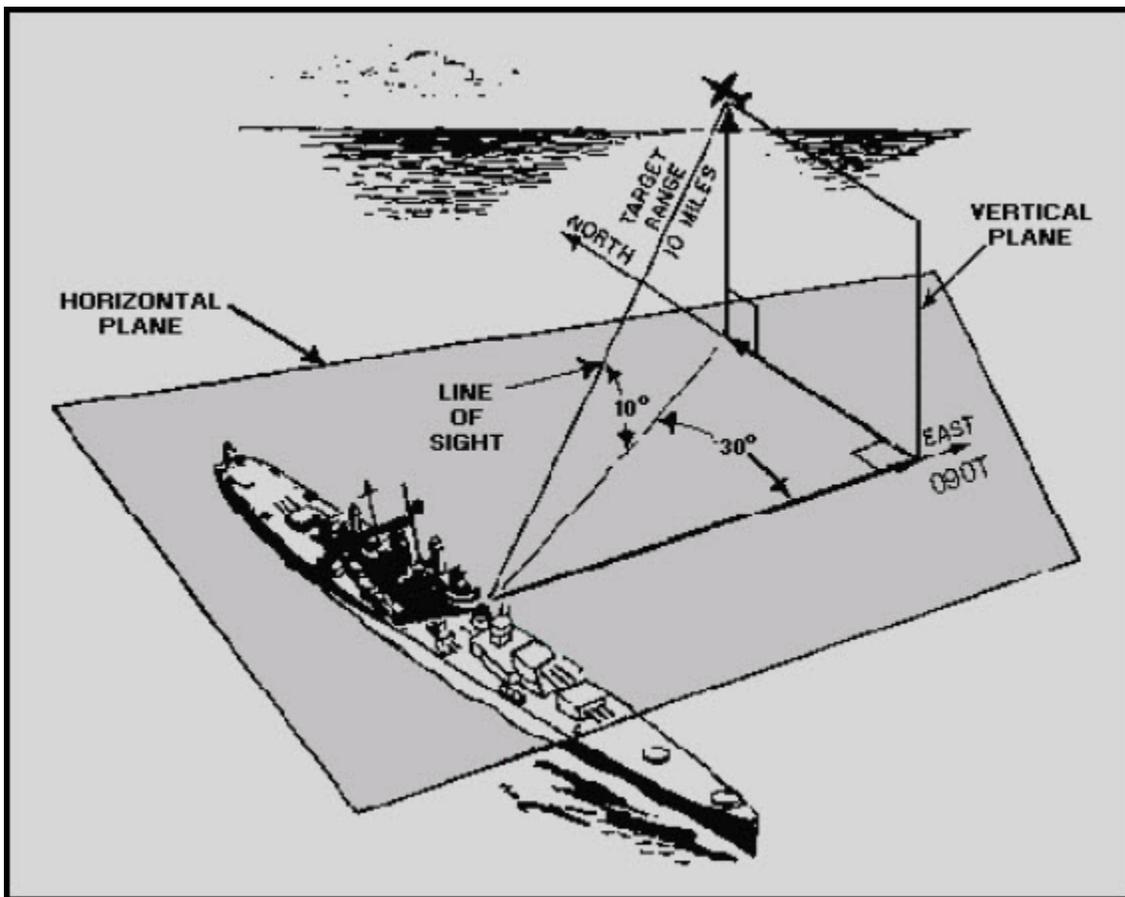


Figure 3-1 Radar target position

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Bearing and elevation angles are determined by measuring the angular position of the radar antenna (the transmitted beam) when it is pointing directly at the target. Range is more difficult to determine because it cannot be directly measured. The radar system is designed to measure range as a function of time. Since the speed of electromagnetic energy is the same as the speed of light, range is determined by measuring the time required for a pulse of energy to reach the target and return to the radar. Because the speed of the pulse is known, the two-way distance can be determined by multiplying the time by the speed of travel. The total must be divided by two to obtain the one-way range because the time value used initially is the time required for the pulse to travel to the target and return.

The discussion of microwave antennas in this chapter requires only the most basic understanding of radar concepts! Radar fundamentals will be discussed in more detail in a later *NEETS* module.

Q-1. Microwave antennas and low-frequency antennas are similar in what ways?

Q-2. What term is used to express the efficiency of an antenna?

Q-3. What term is used to express the measurement of the degree of mismatch between a line and its load?

Q-4. What type of antenna radiates in and receives energy from all directions at once?

Q-5. What is the term that is used to describe narrowness in the radiated beam of an antenna?

Q-6. What characteristic allows the same antenna to both transmit and receive?

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3.3 REFLECTOR ANTENNAS

A spherical wavefront (one in which the energy spreads out in all directions) spreads out as it travels away from the antenna and produces a pattern that is not very directional. A wavefront that exists in only one plane does not spread because all of the wavefront moves forward in the same direction. For an antenna to be highly directive, it must change the normally spherical wavefront into a plane wavefront. Many highly directive microwave antennas produce a plane wavefront by using a reflector to focus the radiated energy. The PARABOLIC REFLECTOR is most often used for high directivity.

Microwaves travel in straight lines as do light rays. They can also be focused and reflected just as light rays can, as illustrated by the antenna shown in figure 3-2. A microwave source is placed at focal point F. The field leaves this antenna as a spherical wavefront. As each part of the wavefront reaches the reflecting surface, it is phase-shifted 180 degrees. Each part is then sent outward at an angle that results in all parts of the field traveling in parallel paths. Because of the special shape of a parabolic surface, all paths from F to the reflector and back to line XY are the same length. Therefore, when the parts of the field are reflected from the parabolic surface, they travel to line XY in the same amount of time.

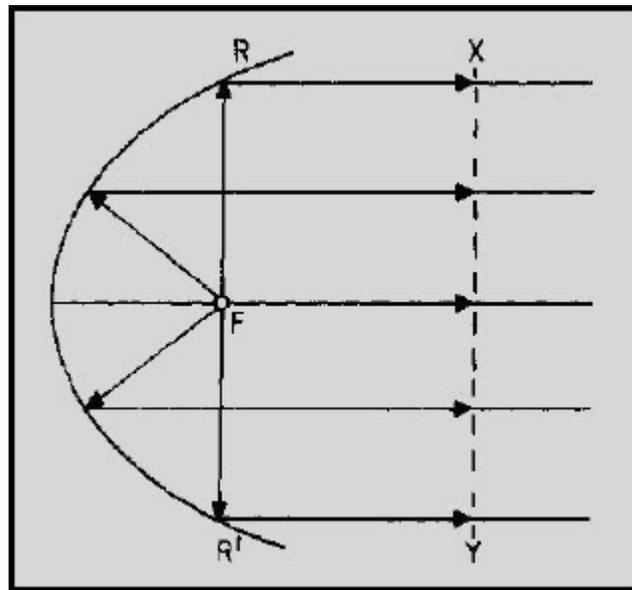


Figure 3-2 Parabolic reflector radiation

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If a dipole is used as the source of transmission, energy will be radiated from the antenna into space as well as toward the reflector. Energy which is not directed toward the paraboloid has a wide-beam characteristic which will destroy the narrow pattern of the parabolic reflector. However, a HEMISPHERICAL SHIELD (not shown) may be used to direct most of the radiation toward the parabolic surface and thus prevent the destruction of the narrow pattern. Direct radiation into space is eliminated, the beam is made sharper, and more power is concentrated in the beam. Without the shield, some of the radiated field would leave the radiator directly. Since this part of the field that would leave the radiator would not be reflected, it would not become a part of the main beam and could serve no useful purpose.

In figure 3-3 the radiation pattern of a paraboloid reflector contains a major lobe and several minor lobes. The major lobe is directed along the axis of revolution. Very narrow beams are possible with this type of reflector. Figure 3-4 illustrates the basic paraboloid reflector.

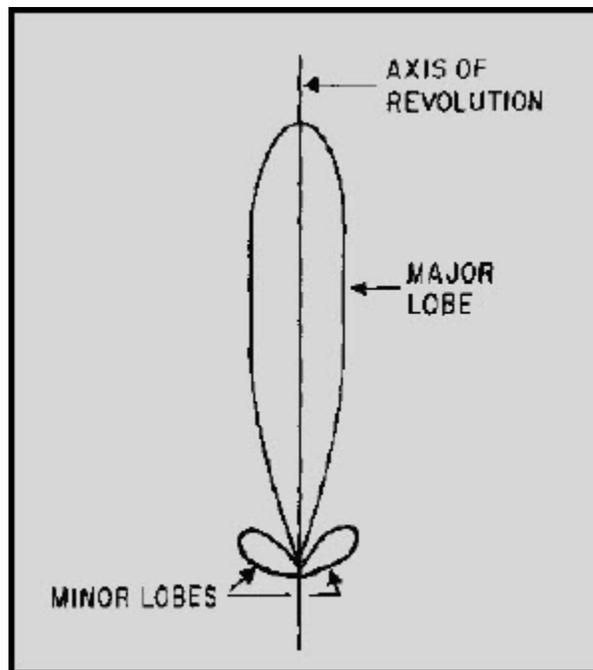


Figure 3-3 Parabolic radiation pattern

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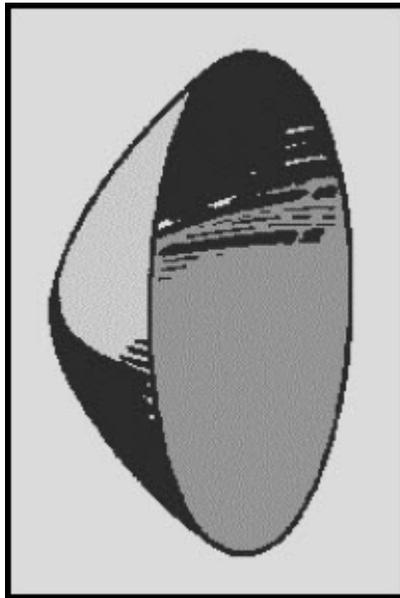


Figure 3-4 Basic paraboloid reflector

You may see several variations of the basic paraboloid reflector used to produce different beam shapes required by special applications. The basic characteristics of the most commonly used paraboloids are presented in the following paragraphs.

3.3.1 Truncated Paraboloid

Figure 3-5A, shows a TRUNCATED PARABOLOID. Since the reflector is parabolic in the horizontal plane, the energy is focused into a narrow beam. With the reflector TRUNCATED (cut) so that it is shortened vertically, the beam spreads out vertically instead of being focused. This fan-shaped beam is used in radar detection applications for the accurate determination of bearing. Since the beam is spread vertically, it will detect aircraft at different altitudes without changing the tilt of the antenna. The truncated paraboloid also works well for surface search radar applications to compensate for the pitch and roll of the ship.

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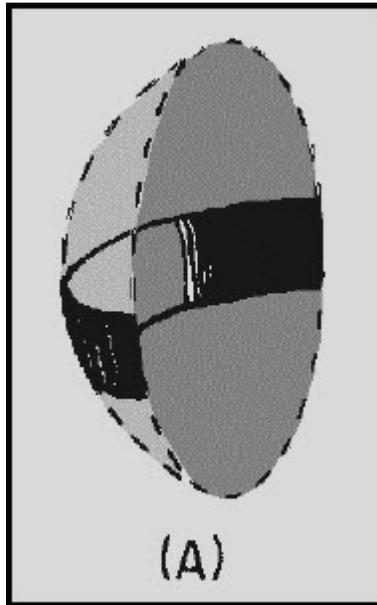


Figure 3-5A Truncated paraboloid

The truncated paraboloid may be used in target height-finding systems if the reflector is rotated 90 degrees, as shown in figure 3-5B. Since the reflector is now parabolic in the vertical plane, the energy is focused vertically into a narrow beam. If the reflector is truncated, or cut, so that it is shortened horizontally, the beam will spread out horizontally instead of being focused. Such a fan-shaped beam is used to accurately determine elevation.

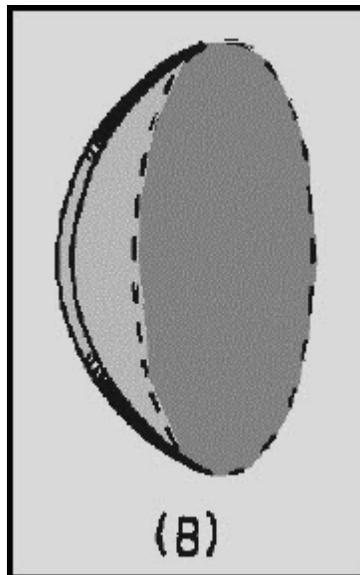


Figure 3-5B Truncated paraboloid.

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3.3.2 Orange-Peel Paraboloid

A section of a complete circular paraboloid, often called an ORANGE-PEEL REFLECTOR because of its orange-peel shape, is shown in figure 3-6. Since the reflector is narrow in the horizontal plane and wide in the vertical plane, it produces a beam that is wide in the horizontal plane and narrow in the vertical plane. In shape, the beam resembles a huge beaver tail. The microwave energy is sent into the parabolic reflector by a horn radiator (not shown) which is fed by a waveguide. The horn radiation pattern covers nearly the entire shape of the reflector, so almost all of the microwave energy strikes the reflector and very little escapes at the sides. Antenna systems which use orange-peel paraboloids are often used in height-finding equipment.

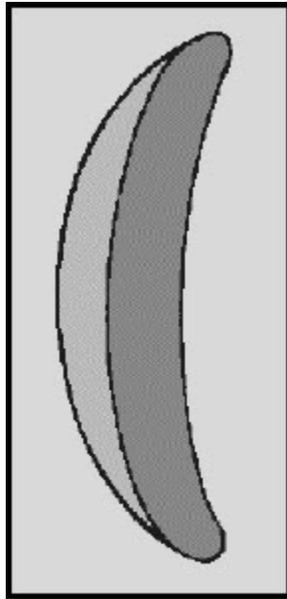


Figure 3-6 Orange-peel paraboloid

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3.3.3 Cylindrical Paraboloid

When a beam of radiated energy that is noticeably wider in one cross-sectional dimension than in another is desired, a cylindrical paraboloidal section which approximates a rectangle can be used. Figure 3-7 illustrates such an antenna. A PARABOLIC CYLINDER has a parabolic cross section in just one dimension which causes the reflector to be directive in one plane only. The cylindrical paraboloid reflector is fed either by a linear array of dipoles, a slit in the side of a waveguide, or by a thin waveguide radiator. It also has a series of focal points forming a straight line rather than a single focal point. Placing the radiator, or radiators, along this focal line produces a directed beam of energy. As the width of the parabolic section is changed, different beam shapes are obtained. You may see this type of antenna system used in search radar systems and in ground control approach (gca) radar systems.

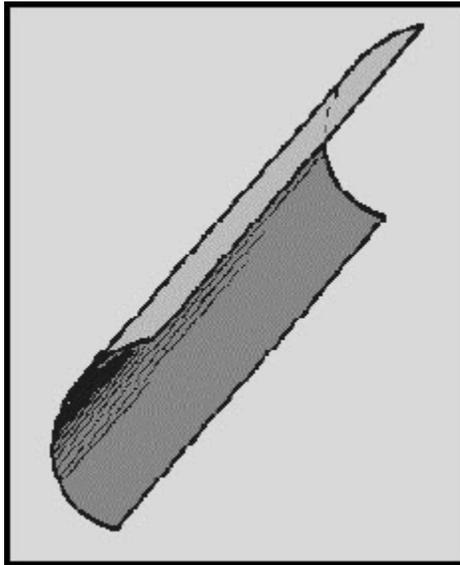


Figure 3-7 Cylindrical paraboloid

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3.3.4 Corner Reflector

The CORNER-REFLECTOR ANTENNA consists of two flat conducting sheets that meet at an angle to form a corner, as shown in figure 3-8. The corner reflector is normally driven by a HALF-WAVE RADIATOR located on a line which bisects the angle formed by the sheet reflectors.

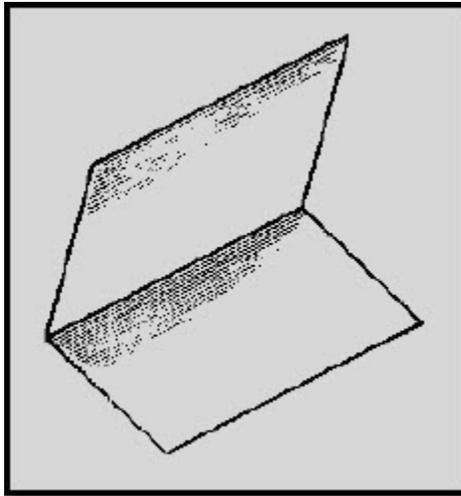


Figure 3-8 Corner reflector

Q-7. What type of reflector is most often used in directive antennas?

Q-8. Microwaves can be focused and reflected in the same way as what other type of waves?

Q-9. How many major lobes are radiated by a parabolic reflector?

Q-10. A horizontally truncated paraboloid antenna is used for what purpose?

Q-11. The beam from a horizontally positioned cylindrical paraboloid is narrow in what plane?

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3.4 HORN RADIATORS

Like parabolic reflectors, you can use HORN RADIATORS to obtain directive radiation at microwave frequencies. Because they do not use resonant elements, horns have the advantage of being useful over a wide frequency band.

The operation of a horn as an rf radiating device is similar to that of an automobile horn radiating sound waves. However, the throat of an automobile horn usually is sized much smaller than the sound wavelengths for which it is used. The throat of the rf radiating horn is sized to be comparable to the wavelength being used.

Horn radiators are used with waveguides because they serve both as an impedance-matching device and as a directional radiator. Horn radiators may be fed by coaxial and other types of lines.

Horn radiators are constructed in a variety of shapes, as illustrated in figure 3-9. The shape of the horn determines the shape of the field pattern. The ratio of the horn length to the size of its mouth determines the beam angle and directivity. In general, the larger the mouth of the horn, the more directive is the field pattern.

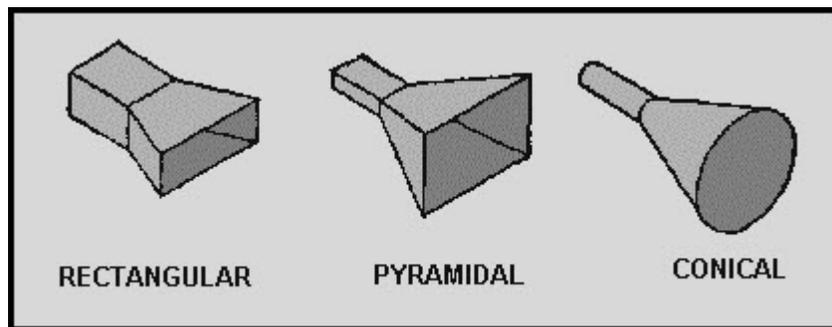


Figure 3-9 Horn radiators

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3.5 LENS ANTENNAS

With a LENS ANTENNA you can convert spherically radiated microwave energy into a plane wave (in a given direction) by using a point source (open end of the waveguide) with a COLLIMATING LENS. A collimating lens forces all radial segments of the spherical wavefront into parallel paths. The point source can be regarded as a gun which shoots the microwave energy toward the lens. The point source is often a horn radiator or a simple dipole antenna.

3.5.1 Waveguide Type

The WAVEGUIDE-TYPE LENS is sometimes referred to as a conducting-type. It consists of several parallel concave metallic strips which are placed parallel to the electric field of the radiated energy fed to the lens, as shown in figure 3-10A and 3-10B. These strips act as waveguides in parallel for the incident (radiated) wave. The strips are placed slightly more than a half wavelength apart.

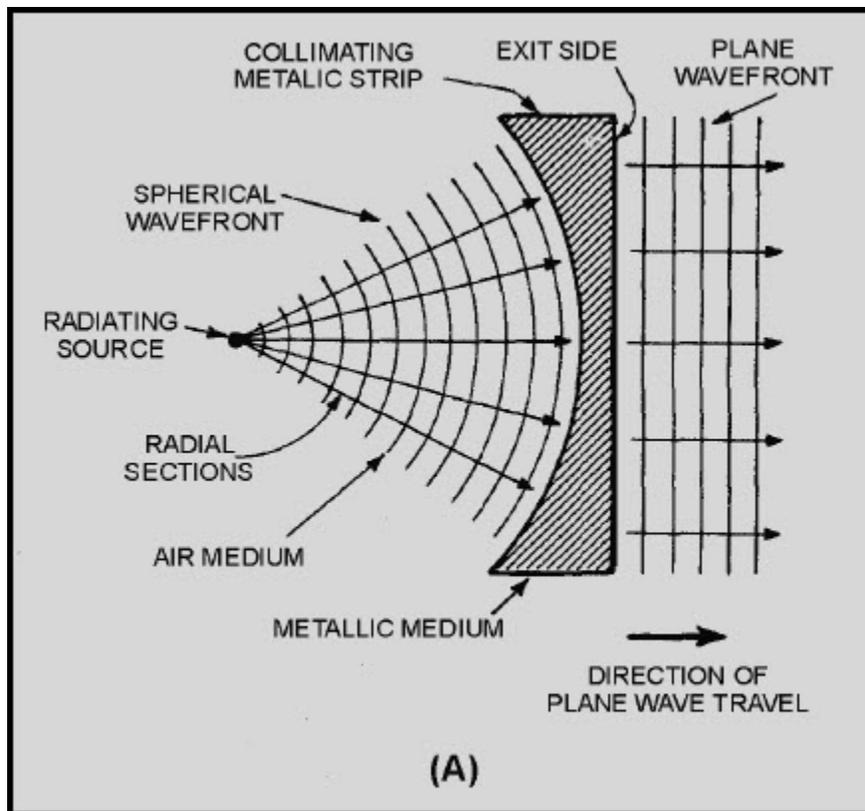


Figure 3-10A Waveguide lens

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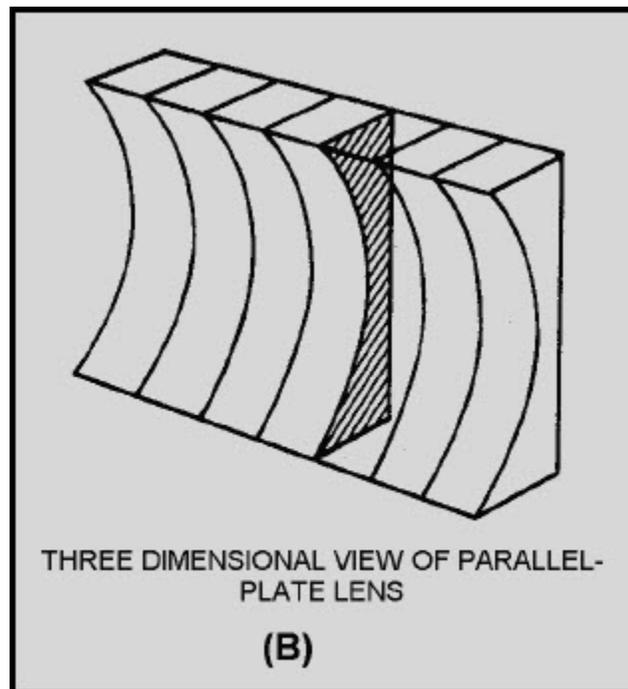


Figure 3-10B Waveguide lens

The radiated energy consists of an infinite number of RADIAL SECTIONS (RAYS). Each of the radial sections contains mutually perpendicular E and H lines and both are perpendicular to the direction of travel. Because each of the radial sections travels in a different direction, the point source, in itself, has poor directivity. The purpose of the lens is to convert the input spherical microwave segment (which consists of all of the radial sections) into parallel (collimated) lines in a given direction at the exit side of the lens. The focusing action of the lens is accomplished by the refracting qualities of the metallic strips. The collimating effect of the lens is possible because the velocity of electromagnetic energy propagation through metals is greater than its velocity through air. Because of the concave construction of the lens, wavefronts arriving near the ends of the lens travel farther in the same amount of time than do those at the center. Thus, the wavefront emerging from the exit side of the lens appears as a plane wave. It consists of an infinite number of parallel sections (with both the E field and H field components) mutually perpendicular to the direction of travel.

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3.5.2 Delay Lens

Another type of lens that you may see is the DIELECTRIC or METALLIC DELAY LENS shown in figure 3-11. The delay lens, as its name implies, slows down the phase propagation (velocity) as the wave passes through the lens. The delay lens is convex and is constructed of dielectric material. The delay in the phase of the wave passing through the lens is determined by the DIELECTRIC CONSTANT (REFRACTIVE INDEX) of the material. In most cases, artificial dielectrics, consisting of conducting rods or spheres that are small compared to the wavelength, are used. (Artificial dielectrics are of three-dimensional construction and act as a dielectric to electromagnetic waves.) In this case the inner portion of the transmitted wave is decelerated for a longer interval of time than the outer portions. The delay causes the radiated wave to be collimated.

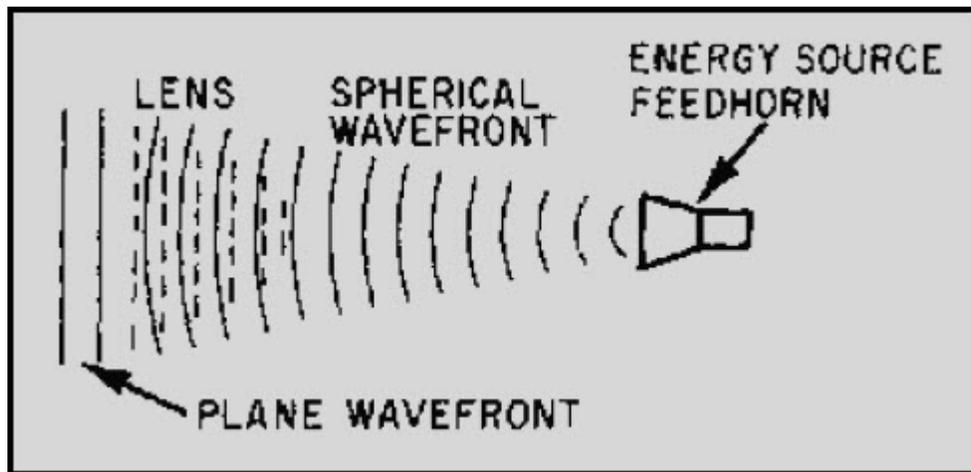


Figure 3-11 Delay-type lens

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3.5.3 Loaded Microwave Lens

The LOADED MICROWAVE LENS, shown in figure 3-12, is a multi-cellular array of thousands of cells. Each cell contains a slow-wave (delayed), serrated-metal, plastic-supported waveguide element which acts as a phase-controlling device. A loaded lens can focus microwave energy in much the same way as the waveguide type. The reason is that the speed of propagation is higher in the region between parallel plates than in free space. The parallel plates support the cells.

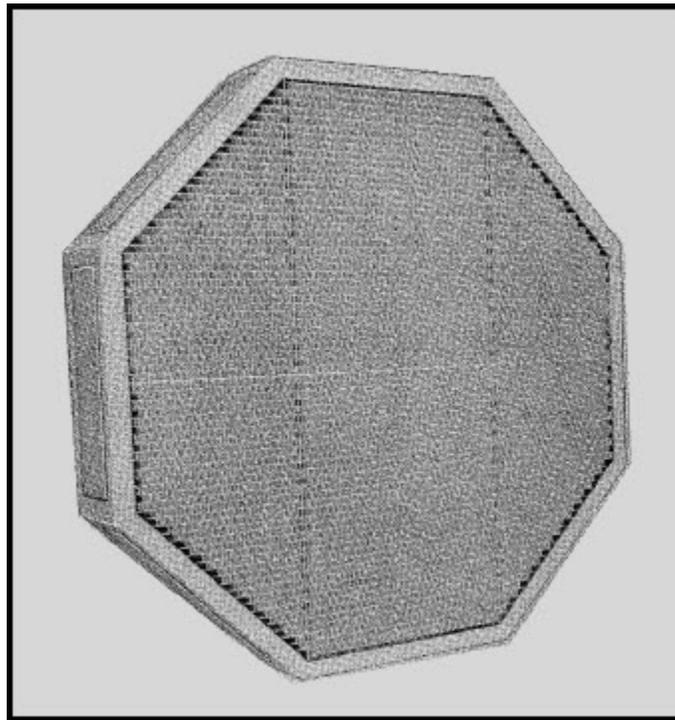


Figure 3-12 Loaded lens

The lens shown in figure 3-12 has an egg-crate appearance because it is really two lenses occupying the same volume. Vertical plates make up a lens that focuses a vertically polarized beam, and horizontal plates handle beams which are horizontally polarized. In other words, this type of construction can be used in multiple-beam applications where the polarization of the beams is different.

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Q-12. What is the purpose of a collimating lens?

Q-13. How does a waveguide-type lens focus spherical wavefront microwave energy?

Q-14. What type of lens decelerates a portion of a spherical wavefront?

3.6 ANTENNA ARRAYS

Sharply directive antennas can be constructed from two or more simple half-wave dipole elements. They must be positioned so that the fields from the elements add in some directions and cancel in others. Such a set of antenna elements is called an ANTENNA ARRAY. When a reflector is placed behind the dipole array, radiation occurs in one direction with a pattern similar to the one shown in figure 3-13.

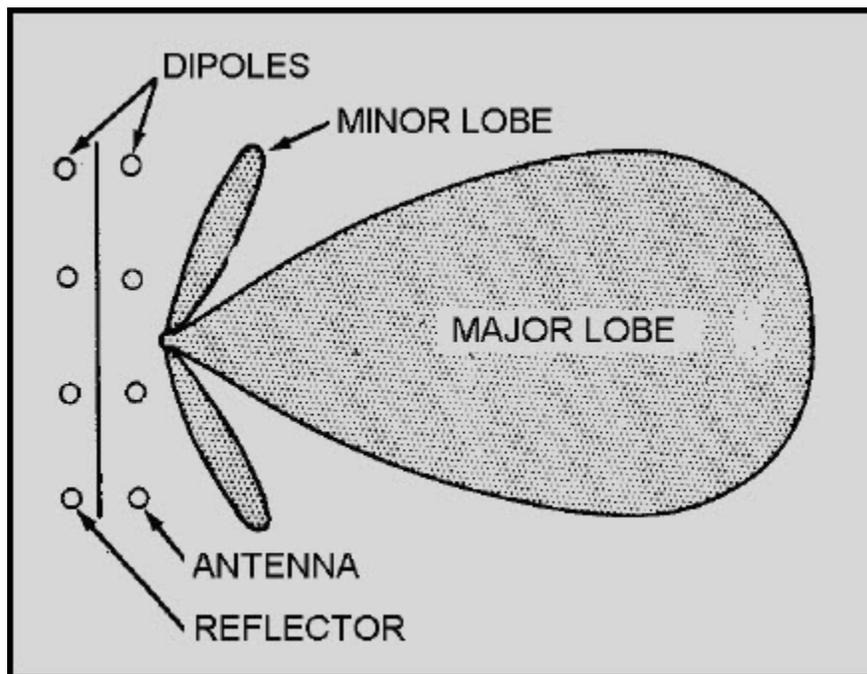


Figure 3-13 Field pattern of an antenna array

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You will encounter two basic types of antenna arrays, PARASITIC and DRIVEN. Both types of antenna arrays were explained in *NEETS*, Module 10, *Introduction to Wave Propagation, Transmission Lines, and Antennas*. Only a brief review is presented in this chapter.

The parabolic reflector antennas previously discussed and the antenna shown in figure 3-13 are examples of parasitic arrays. Notice that the reflector in figure 3-13 is not directly connected to the energy source. Driven arrays, in which all the radiating elements are connected to the energy source, have smaller losses than parasitic arrays while retaining some of the narrow-beam characteristics. Parasitic arrays, such as the parabolic reflector, are used primarily as antennas in fire control radars and other installations, such as microwave communication systems, that require very accurate (narrow) beams. Driven arrays are used primarily as search-radar antennas because extremely narrow beams are less critical than low losses.

If you position a number of driven half-wave antenna elements with respect to each other so that energy from the individual elements will add in certain directions and cancel in other directions, then the antenna system is directional.

Signals from a number of different sources may contribute to or subtract from the overall effect. By properly phasing the energy fed to the antenna elements, and by properly locating the elements, you can control the direction of the energy. You can cause the energy to add in the desired direction and to be out of phase (cancel) in the undesired direction.

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Driven arrays are usually made up of a number of half-wave dipoles positioned and phased so that the desired directional pattern will be achieved. Figure 3-14, view (A), shows a simple antenna array consisting of two horizontally mounted elements, each a half wavelength long and fed in phase. The resulting radiation pattern is in a direction at right angles to the plane containing the antenna conductor.

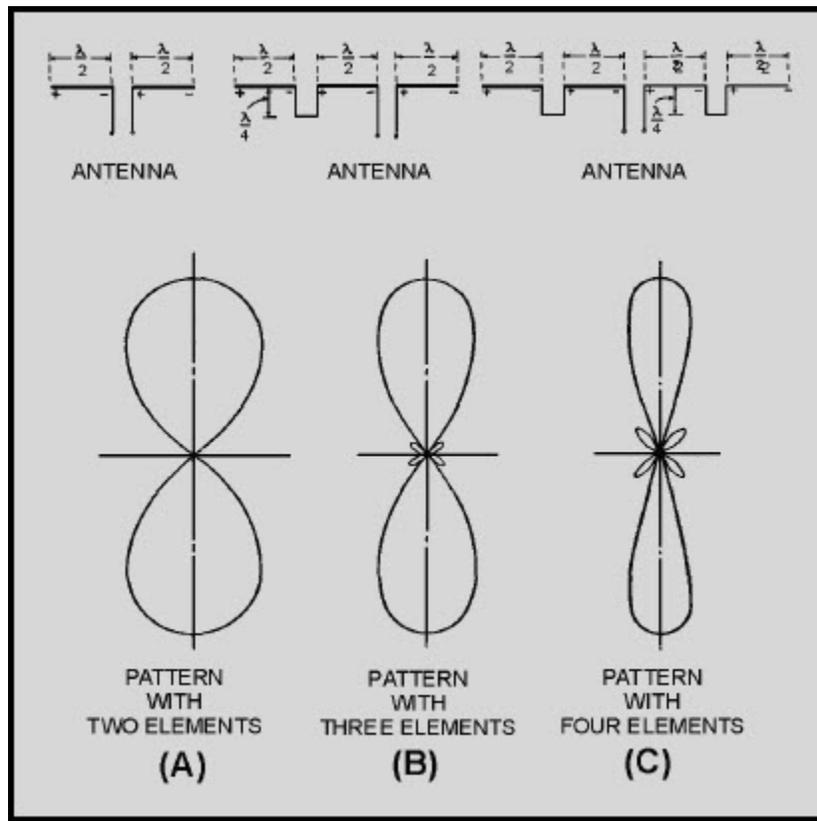


Figure 3-14 Horizontal array field patterns

Three- and four-element arrays are shown in figure 3-14, views (B) and (C), respectively. The field pattern of each array is shown beneath it. Note that the beam becomes sharper as the number of elements is increased. If a still-narrower beam is desired, you may add additional elements. The field patterns of the antennas in the figure are bidirectional. Unidirectional patterns may be obtained with a parasitic reflector mounted behind the driven antenna elements.

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The BEDSPRING ARRAY (figure 3-15), so called because of its resemblance to a bedspring, is an example of a unidirectional antenna. It consists of a stacked dipole array with an untuned reflector. The more dipoles that are used or stacked in one dimension (horizontal, for example), the more narrow the beam of radiated energy becomes in that plane. Consequently, the size of the antenna is not the same for all installations. Antennas such as the bedspring array are commonly used in TWO-DIMENSIONAL SEARCH RADARS that obtain the range and bearing information of a target.

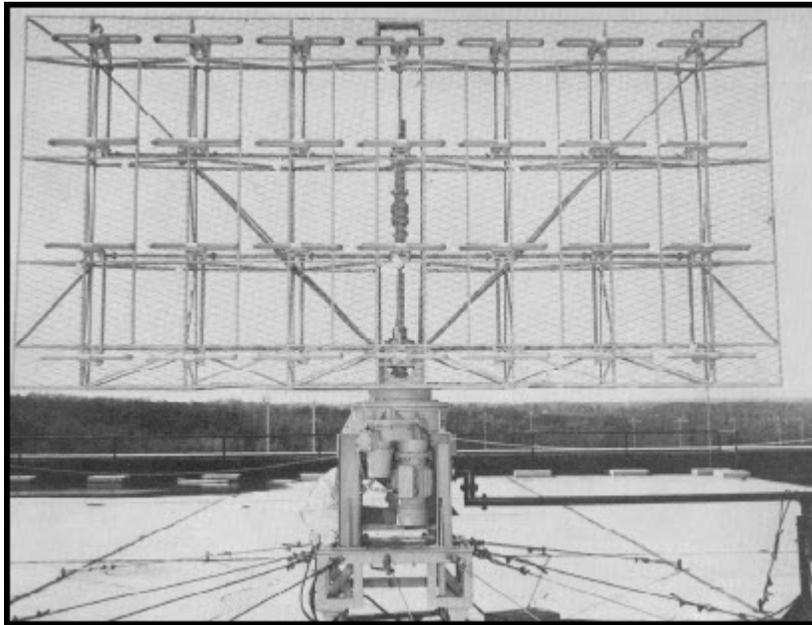


Figure 3-15 Bedspring array

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3.7 FREQUENCY-SENSITIVE ANTENNA

The radar antenna in figure 3-16 uses a feed section to drive horizontally stacked array sections which radiate the applied rf pulses. The same array sections receive the target returns. Each array contains slots cut to radiate and receive a particular frequency. Bearing data is obtained by mechanically rotating the antenna 360 degrees. Elevation data is obtained by electronic scanning of the beam in elevation. The radar antenna is frequency sensitive and radiates pulses at an elevation angle determined by the applied frequency. When the frequency is increased, the beam elevation angle decreases. Conversely, when the applied frequency is decreased, the beam elevation angle increases. The beam elevation angle is therefore selected by the application of a frequency corresponding to the desired angle of elevation.

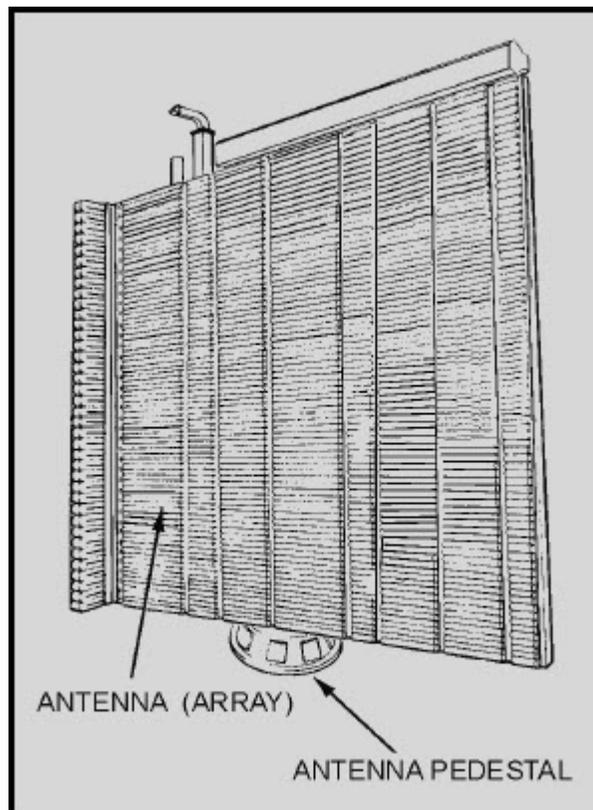


Figure 3-16 Frequency-sensitive antenna

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The physical length of the antenna feed section, called the SERPENTINE SECTION (figure 3-17), in relation to the wavelength of the applied energy determines the direction of the radiated beam. You may understand this more clearly if you consider how the beam is shifted. The shift occurs with a change in frequency because the positive and negative peaks of the energy arrive at adjacent slotted arrays at different times. The change in the field pattern is such that the angle of departure (angle at which the radiated beam leaves the antenna) of the beam is changed. Note that a change in phase of the applied rf energy would cause the same effect.

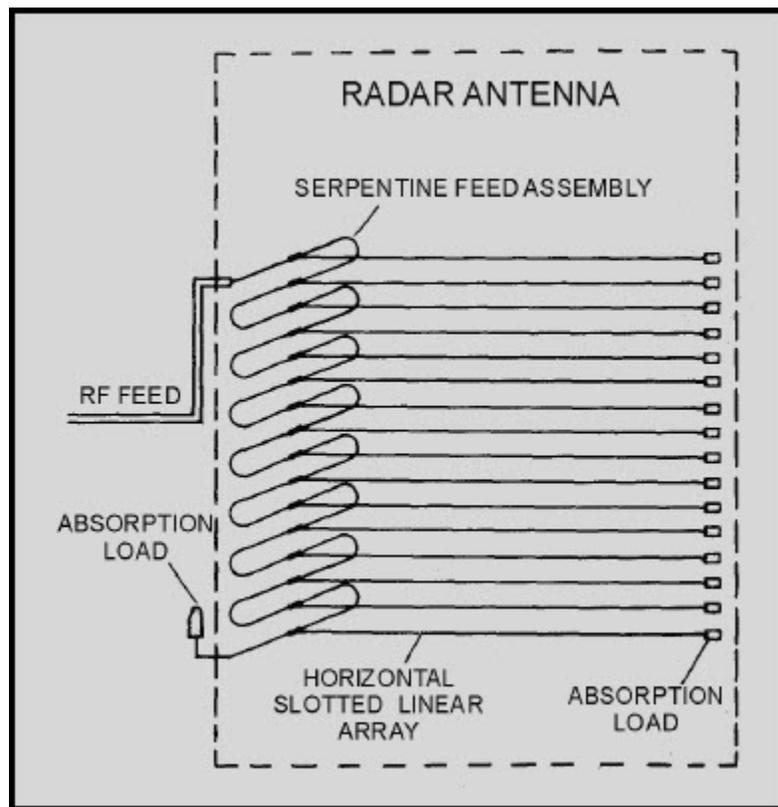


Figure 3-17 Serpentine feed

A SLOT ANTENNA exhibits many of the characteristics of a conventional dipole antenna. When arranged in arrays, a high degree of directivity can be obtained. Also, the beam can be caused to scan a volume of space by changing either the frequency or phase of the energy driving the antenna elements.

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3.7.1 Basic Slot Antenna and Its Complementary Dipole

The slot antenna consists of a radiator formed by cutting a narrow slot in a large metal surface. Such an antenna is shown in figure 3-18. The slot length is a half wavelength at the desired frequency and the width is a small fraction of a wavelength. The antenna is frequently compared to a conventional half-wave dipole consisting of two flat metal strips. The physical dimensions of the metal strips are such that they would just fit into the slot cut out of the large metal sheet.

This type of antenna is called the COMPLEMENTARY DIPOLE.

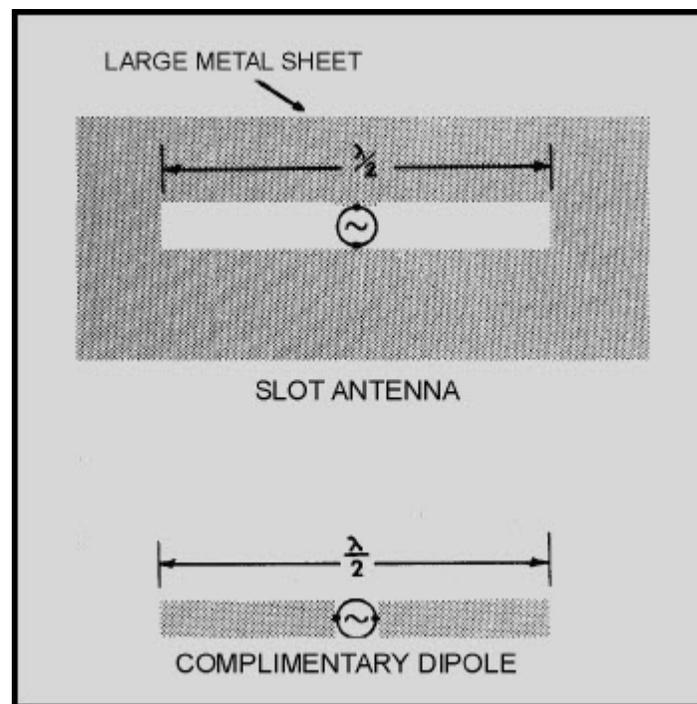


Figure 3-18 Slot antenna and complementary dipole

The slot antenna is compared to its complementary dipole to illustrate that the radiation patterns produced by a slot antenna cut into an infinitely large metal sheet and that of the complementary dipole antenna are the same.

Several important differences exist between the slot antenna and its complementary antenna. First, the electric and magnetic fields are interchanged. In the case of the dipole antenna shown in figure 3-18, the electric lines are horizontal while the magnetic lines form loops in the vertical plane. With the slot antenna, the magnetic lines are horizontal and the electric lines are vertical. The electric lines are built up across the narrow dimensions of the slot. As a result, the polarization of the radiation produced by a horizontal slot is vertical. If a vertical slot is used, the polarization is horizontal.

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A second difference between the slot antenna and its complementary dipole is that the direction of the lines of electric and magnetic force abruptly reverse from one side of the metal sheet to the other. In the case of the dipole, the electric lines have the same general direction while the magnetic lines form continuous closed loops.

When energy is applied to the slot antenna, currents flow in the metal sheet. These currents are not confined to the edges of the slot but rather spread out over the sheet. Radiation then takes place from both sides of the sheet. In the case of the complementary dipole, however, the currents are more confined; so a much greater magnitude of current is required to produce a given power output using the dipole antenna.

The current distribution of the dipole resembles the voltage distribution of the slot. The edges on the slot have a high voltage concentration and relatively low current distribution; the complementary dipole has a high current concentration and relatively low voltage.

Slot antennas are adaptable for the vhf and uhf ranges. One of their practical advantages is that the feed section which energizes the slot may be placed below the large metal surface in which the slot is cut. Thus, nothing needs to extend from the surface. In addition, the slot itself may be covered by a section of insulating material to provide a seal so that the antenna can be pressurized with dry air. Dry air pressurization reduces moisture in the waveguide and prevents arcing.

Many of the new radar systems reaching the fleet over the next few years will use frequency- or phase-sensitive antennas. Some of the new radars will use antennas that electronically scan the azimuth as well as elevation, eliminating the moving antenna.

Q-15. What is a set of antenna elements called?

Q-16. What type of antenna has all elements connected to the same energy source?

Q-17. What determines the beam elevation angle of an antenna that is electronically scanned in elevation?

Q-18. What is the polarization of the energy radiated by a vertical slot?

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3.8 SUMMARY

This chapter has presented information on the characteristics of microwave antennas. The information that follows summarizes the important points of this chapter.

The **ANTENNA CHARACTERISTICS** of microwave and low-frequency antennas are essentially the same. The efficiency of an antenna is expressed as a **POWER GAIN** or **POWER RATIO** as compared to a standard reference antenna.

The **STANDING WAVE RATIO (swr)** is a measurement of the impedance mismatch between a transmission line and its load and is an indicator of overall system efficiency.

DIRECTIVITY refers to the direction in which an antenna radiates and the narrowness of the radiated beam in **DIRECTIONAL ANTENNAS**.

OMNIDIRECTIONAL ANTENNAS radiate and receive in all directions at once.

RECIPROCITY is the ability of an antenna to both transmit and receive electromagnetic energy.

REFLECTOR ANTENNAS are antennas that use a reflector to focus electromagnetic energy into a beam that is directional in either the vertical plane, the horizontal plane, or both planes at once. The basic **PARABOLIC REFLECTOR** shown in the illustration, or one of its variations, is most often used.

LENS ANTENNAS use a **COLLIMATING LENS** to force the spherical components of a wavefront into parallel (focused) paths by delaying or accelerating portions of the wavefronts, as shown in the illustration.

An **ANTENNA ARRAY** is a set of antenna elements and may be one of two basic types, the **DRIVEN ARRAY** or the **PARASITIC ARRAY**.

FREQUENCY-SENSITIVE ANTENNAS use frequency-sensitive slots as radiation sources to achieve directivity. The angle at which the radiated beam leaves the antenna is determined by the frequency of the radiated energy. Currently the most common frequency-sensitive antennas use this feature to achieve elevation coverage while azimuth coverage is achieved by rotating the antennas. New systems will use stationary frequency-sensitive antennas to achieve both azimuth and elevation coverage.

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ANSWERS TO QUESTIONS Q1. THROUGH Q18.

A-1. Operating principles and electrical characteristics.

A-2. Power gain or power ratio.

A-3. Standing-wave ratio (swr).

A-4. Omnidirectional.

A-5. Antenna directivity.

A-6. Reciprocity.

A-7. Parabolic.

A-8. Light waves.

A-9. One.

A-10. Determine elevation.

A-11. Vertical.

A-12. Forces the radial segments of a wavefront into parallel paths.

A-13. Some wavefronts are accelerated so that all wavefronts exit the lens at the same time.

A-14. Delay lens.

A-15. Antenna Array.

A-16. Driven Array.

A-17. Frequency or phase of radiated energy.

A-18. Horizontal.

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APPENDIX A

Glossary

APERTURE - See slot.

BOUNDARY CONDITIONS - The two conditions that the E-field and H-field within a waveguide must meet before energy will travel down the waveguide. The E-field must be perpendicular to the walls and the H-field must be in closed loops, parallel to the walls, and perpendicular to the E-field.

BEARING - An angular measurement that indicates the direction of an object in degrees from true north. Also called azimuth.

BUNCHER CAVITY - The input resonant cavity in a conventional klystron oscillator.

BUNCHER GRID - In a velocity-modulated tube, the grid which concentrates the electrons in the electron beam into bunches.

CATCHER GRID - In a velocity-modulated tube, a grid on which the spaced electron groups induce a signal. The output of the tube is taken from the catcher grid.

CAVITY RESONATOR - A space totally enclosed by a metallic conductor and supplied with energy in such a way that it becomes a source of electromagnetic oscillations. The size and shape of the enclosure determine the resonant frequency.

CHOKE JOINT - A joint between two sections of waveguide that provides a good electrical connection without power losses or reflections.

COOKIE-CUTTER TUNER - Mechanical magnetron tuning device that changes the frequency by changing the capacitance of the anode cavities.

COPPER LOSS - Power loss in copper conductors caused by the internal resistance of the conductors to current flow. Also called I^2R loss.

CROWN-OF-THORNS TUNER - See Sprocket Tuner.

CUTOFF FREQUENCY - The frequency at which the attenuation of a waveguide increases sharply and below which a traveling wave in a given mode cannot be maintained. A frequency with a half wavelength that is greater than the wide dimension of a waveguide.

DIELECTRIC CONSTANT - The ratio of a given dielectric to the dielectric value of a vacuum.

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DIELECTRIC LOSSES - The electric energy that is converted to heat in a dielectric subjected to a varying electric field.

DIRECTIONAL COUPLER - A device that samples the energy traveling in a waveguide for use in another circuit.

DIRECTIVITY - The narrowness of the radiated beam from an antenna.

DIRECTIONAL COUPLER - A device that samples the energy traveling in a waveguide for use in another circuit.

DIRECTIVITY - The narrowness of the radiated beam from an antenna.

DOMINANT MODE - The easiest mode to produce in a waveguide, and also, the most efficient mode in terms of energy transfer.

DRIFT SPACE - In an electron tube, a region free of external fields in which relative electron position depends on velocity.

DUMMY LOAD - A device used at the end of a transmission line or waveguide to convert transmitted energy into heat so no energy is radiated outward or reflected back.

E-FIELD - Electric field that exists when a difference in electrical potential causes a stress in the dielectric between two points.

E-TYPE T-JUNCTION - A waveguide junction in which the junction arm extends from the main waveguide in the same direction as the E-field in the waveguide.

ELECTRIC FIELD - See E-field.

ELECTRONIC TUNING - In a reflex klystron, changing the frequency and output power of the tube by altering the repeller voltage.

ELECTROLYSIS - Chemical changes produced by passing an electrical current from one substance (electrode) to another (electrolyte).

ELECTRON ORBITAL MOVEMENT - The movement of an electron around the nucleus of an atom.

ELECTRON SPIN - The movement of an electron around its axis.

ELEVATION ANGLE - The angle between the line of sight to an object and the horizontal plane.

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FARADAY ROTATION - The rotation of the plane of polarization of electromagnetic energy when it passes through a substance influenced by a magnetic field that has a component in the direction of propagation.

FERRITE - A powdered and compressed ferric oxide material that has both magnetic properties and resistance to current flow.

FERRITE SWITCH - A ferrite device that blocks the flow of energy through a waveguide by rotating the electric field 90 degrees. The rotated energy is then reflected or absorbed.

GRID-GAP TUNING - A method of changing the center frequency of a resonant cavity by physically changing the distance between the cavity grids.

GROUP VELOCITY - The forward progress velocity of a wave front in a waveguide.

H-FIELD - Any space or region in which a magnetic force is exerted. The magnetic field may be produced by a current-carrying coil or conductor, by a permanent magnet, or by the earth itself.

H-TYPE T-JUNCTION - A waveguide junction in which the junction arm is parallel to the magnetic lines of force in the main waveguide.

HELIX - A spirally wound transmission line used in a traveling-wave tube to delay the forward progress of the input traveling wave.

HORIZONTAL PLANE - An imaginary plane tangent to and touching the Earth's surface as established by a stable element, such as a gyroscope.

HORN - A funnel-shaped section of waveguide used as a termination device and as a radiating antenna.

HOT CARRIER - A current carrier, which may be either a hole or an electron, that has relatively high energy with respect to the current carriers normally found in majority-carrier devices.

HOT-CARRIER DIODE - A semiconductor diode in which hot carriers are emitted from a semiconductor layer into the metal base. Also called a hot-electron diode. An example is the Schottky-Barrier diode.

HYBRID JUNCTION - A waveguide junction that combines two or more basic T-junctions.

HYBRID RING - A hybrid-waveguide junction that combines a series of E-type T-junctions in a ring configuration.

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IDLER FREQUENCY - In a parametric amplifier, the difference between the input signal and the pump signal frequency. Also called the lower-sideband frequency.

INTERACTION SPACE - The region in an electron tube where the electrons interact with an alternating electromagnetic field.

INTERELECTRODE CAPACITANCE - The capacitance between the electrodes of an electron tube.

I_R LOSS - See Copper Loss.

IRIS - A metal plate with an opening through which electromagnetic waves may pass. Used as an impedance matching device in waveguides.

LEAD INDUCTANCE - The inductance of the lead wires connecting the internal components of an electron tube.

LOAD ISOLATOR - A passive attenuator in which the loss in one direction is much greater than that in the opposite direction. An example is a ferrite isolator for waveguides that allows energy to travel in only one direction.

LOOP - A curved conductor that connects the ends of a coaxial cable or other transmission line and projects into a waveguide or resonant cavity for the purpose of injecting or extracting energy.

LOOSE COUPLING - Inefficient coupling of energy from one circuit to another that is desirable in some applications. Also called weak coupling.

MAGIC-T JUNCTION - A combination of the H-type and E-type T-junctions.

MAGNETIC FIELD - See H-field.

METALLIC INSULATOR - A shorted quarter-wave section of transmission line.

MICROWAVE REGION - The portion of the electromagnetic spectrum from 1,000 megahertz to 100,000 megahertz.

MODULATOR - A device that produces modulation; i.e., varies the amplitude, frequency, or phase of an ac signal.

NEGATIVE-RESISTANCE ELEMENT - A component having an operating region in which an increase in the applied voltage increases the resistance and produces a proportional decrease in current. Examples include tunnel diodes and silicon unijunction transistors.

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NONDEGENERATIVE-PARAMETRIC AMPLIFIER - A parametric amplifier that uses a pump signal frequency that is higher than twice the frequency of the input signal.

PHASE SHIFTER - A device used to change the phase relationship between two ac signals.

POWER GAIN - The ratio of the radiated power of an antenna compared to the output power of a standard antenna. A measure of antenna efficiency usually expressed in decibels. Also referred to as **POWER RATIO**.

POWER RATIO - See Power Gain.

PROBE - A metal rod that projects into, but is insulated from, a waveguide or resonant cavity and used to inject or extract energy.

PUMP - Electrical source of the energy required to vary the capacitance of a parametric amplifier.

RANGE - Distance, as measured from a point of reference, such as a radar, to a target or other object.

REACTANCE AMPLIFIER - A low-noise amplifier that uses a nonlinear variable reactance as the active element instead of a variable resistance. Also called a parametric amplifier.

RECIPROCITY - The ability of an antenna to both transmit and receive electromagnetic energy.

REFLEX KLYSTRON - A klystron with a reflector (repeller) electrode in place of a second resonant cavity to redirect the velocity-modulated electrons back through the cavity which produced the modulation.

REFRACTIVE INDEX - The ratio of the phase velocity of a wave in free space to the phase velocity of the wave in a given substance (dielectric).

REPELLER - Sometimes called a reflector. An electrode in a reflex klystron with the primary purpose of reversing the direction of the electron beam.

ROTATING JOINT - A joint that permits one section of a transmission line or waveguide to rotate continuously with respect to another while passing energy through the joint. Also called a rotary coupler.

SKIN EFFECT - The tendency for alternating current to concentrate in the surface layer of a conductor. The effect increases with frequency and serves to increase the effective resistance of the conductor.

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SLOT - Narrow opening in a waveguide wall used to couple energy in or out of the waveguide. Also called an aperture or a window.

SPROCKET TUNER - Mechanical tuning device for magnetron tubes that changes the frequency of the cavities by changing the inductance. Also called a crown-of-thorns tuner.

STAGGER TUNING - A method of klystron tuning in which the resonant cavities are tuned to slightly different frequencies to increase the bandwidth of the amplifier.

STANDING WAVE RATIO - The ratio of the maximum to the minimum amplitudes of corresponding components of a field, voltage, or current along a transmission line or waveguide in the direction of propagation measured at a given frequency.

SYNCHRONOUS TUNING - In a klystron amplifier, a method of tuning which tunes all the resonant cavities to the same frequency. High gain is achieved, but the bandwidth is narrow.

TRANSIT TIME - The time an electron takes to cross the distance between the cathode and anode.

TRANSVERSE ELECTRIC MODE - The entire electric field in a waveguide is perpendicular to the wide dimension and the magnetic field is parallel to the length. Also called the TE mode.

TRANSVERSE MAGNETIC MODE - The entire magnetic field in a waveguide is perpendicular to the wide dimension ("a" wall) and some portion of the electric field is parallel to the length. Also called the TM mode.

TUNNELING - The piercing of a potential barrier in a semiconductor by a particle (current carrier) that does not have sufficient energy to go over the barrier.

TUNNEL DIODE - A heavily doped junction diode that has negative resistance in the forward direction over a portion of its operating range. See NEGATIVE-RESISTANCE ELEMENT.

VARACTOR - A pn-junction semiconductor designed for microwave frequencies in which the capacitance varies with the applied bias voltage.

VARIABLE ATTENUATOR - An attenuator for reducing the strength of an ac signal either continuously or in steps, without causing signal distortion.

VELOCITY MODULATION - Modification of the velocity of an electron beam by the alternate acceleration and deceleration of electrons.

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VERTICAL PLANE - An imaginary plane that is perpendicular to the horizontal plane.

WAVEGUIDE - A rectangular, circular, or elliptical metal pipe designed to transport electromagnetic waves through its interior.

WAVEGUIDE MODE OF OPERATION - Particular field configuration in a waveguide that satisfies the boundary conditions. Usually divided into two broad types: the transverse electric (TE) and the transverse magnetic (TM).

WAVEGUIDE POSTS - A rod of conductive material used as impedance-changing devices in waveguides.

WAVEGUIDE SCREW - A screw that projects into a waveguide for the purpose of changing the impedance.

WINDOW - See Slot.

WOBBLE FREQUENCY - The frequency at which an electron wobbles on its axis under the influence of an external magnetic field of a given strength.