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Radar Antennas

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INTRODUCTION

RADAR proved to be one of the most important technical achievements of World War II. It has many sources, some as far back as the nineteenth century, yet its rapid wartime growth was the result of military necessity. This development will continue, for radar has increasing applications in a peacetime world.

In this paper we will discuss an indispensable part of radar—the antenna. In a radar system the antenna function is two-fold. It both projects into space each transmitted radar pulse, and collects from space each received reflected signal. Usually but not always a single antenna performs both functions.

The effectiveness of a radar is influenced decisively by the nature and quality of its antenna. The greatest range at which the radar can detect a target, the accuracy with which the direction to the target can be determined and the degree with which the target can be discriminated from its background or other targets all depend to a large extent on electrical properties of the antenna. The angular sector which the antenna can mechanically or electrically scan is the sector from which the radar can provide information. The scanning rate determines the frequency with which a tactical or navigational situation can be examined.

Radar antennas are as numerous in kind as radars. The unique character and particular functions of a radar are often most clearly evident in the design of its antenna. Antennas must be designed for viewing planes from the ground, the ground from planes and planes from other planes. They must see ships from the shore, from the air, from other ships, and from submarines. In modern warfare any tactical situation may require one or several radars and each radar must have one or more antennas.

Radar waves are almost exclusively in the centimeter or microwave region, yet even the basic microwave techniques are relatively new to the radio art. Radar demanded antenna gains and directivities far greater than those previously employed. Special military situations required antennas with beam shapes and scanning characteristics never imagined by communication engineers.

It is natural that war should have turned our efforts so strongly in the direction of radar. But that these efforts were so richly and quickly rewarded was due in large part to the firm technical foundations that had been laid in the period immediately preceding the war. When, for the common good, all privately held technical information was poured into one pool, all ingredients of radar, and of radar antennas in particular, were found to be present.

A significant contribution of the Bell System to this fund of technical knowledge was its familiarity with microwave techniques. Though Hertz himself had performed radio experiments in the present microwave region, continuous wave techniques remained for decades at longer wavelengths. However, because of its interest in new communication channels and broader bands the Bell System has throughout the past thirty years vigorously pushed continuous wave techniques toward the direction of shorter waves. By the middle nineteen-thirties members of the Radio Research Department of the Bell Laboratories were working within the centimeter region.

Several aspects of this research and development appear now as particularly important. In the first place it is obvious that knowledge of how to generate and transmit microwaves is an essential factor in radar. Many lower frequency oscillator and transmission line techniques are inapplicable in the microwave region. The Bell Laboratories has been constantly concerned with the development of generators which would work at higher and higher frequencies. Its broad familiarity with coaxial cable problems and in particular its pioneering work with waveguides provided the answers to many radar antenna problems.

Another telling factor was the emphasis placed upon measurement. Only through measurements can the planners and designers of equip-

ment hope to evaluate performance, to choose between alternatives or to see the directions of improvement. Measuring techniques employing double detection receivers and intermediate frequency amplifiers had long been in use at the Holmdel Radio Laboratory. By employing these techniques radar engineers were able to make more sensitive and accurate measurements than would have been possible with single detection.

Antennas are as old as radio. Radar antennas though different in form are identical in principle with those used by Hertz and Marconi. Consequently experience with communication antennas provided a valuable background for radar antenna design. As an example of the importance of this background it can be recalled that a series of experi-

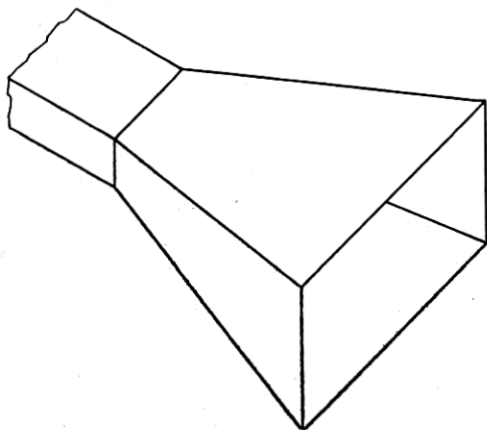


Fig. 1—An Electromagnetic Horn.

ments with short wave antennas for Transatlantic radio telephone service had culminated in 1936 in a scanning array of rhombic antennas. The essential principles of this array were later applied to shipborne fire control antenna which was remarkable and valuable because of the early date at which it incorporated modern rapid scanning features.

In addition to the antenna arts which arose directly out of communication problems at lower frequencies some research specifically on microwave antennas was under way before the war. Early workers in waveguides noticed that an open ended waveguide will radiate directly into space. It is not surprising therefore that these workers developed the electromagnetic horn, which is essentially a waveguide tapered out to an aperture (Fig. 1).

One of the first used and simplest radio antennas is the dipole (Fig.

2). Current oscillating in the dipole generates electromagnetic waves which travel out with the velocity of light. A single dipole is fairly non-directive and consequently produces a relatively weak field at a distance. When the wave-length is short the field of a dipole in a

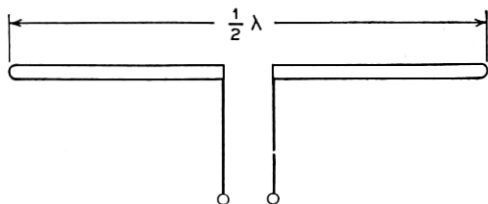


Fig. 2—A Microwave Dipole.

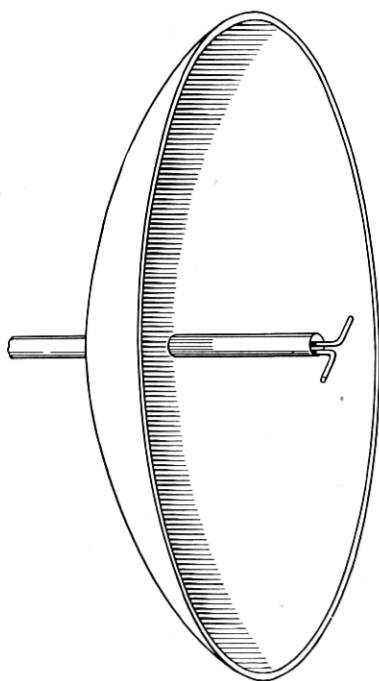


Fig. 3—A Dipole Fed Paraboloid.

chosen direction can be increased many times by introducing a reflector which directs or 'focusses' the energy.

In communication antennas the focussing reflector is most commonly a reflecting wire array. Even at an early date in radar the wave-length was so short that 'optical' reflectors could be used. These were

sometimes paraboloids similar to those used in searchlights (Fig. 3). Sometimes they were parabolic cylinders as in the Mark III, an early shipborne fire control radar developed at the Whippany Radio Laboratory.

From these relatively simple roots, the communication antenna, the electromagnetic horn and the optical reflector, radar antennas were developed tremendously during the war. That this development in the Bell Laboratories was so well able to meet demands placed on it was due in large part to the solid foundation of experience possessed by the Research and Development groups of the Laboratories. Free interchange of individuals and information between the Laboratories and other groups, both in the United States and Great Britain, also contributed greatly to the success of radar antenna development.

Because of its accelerated wartime expansion the present radar antenna field is immense. It is still growing. It would be impossible for any single individual or group to master all details of this field, yet its broad outline can be grasped without difficulty.

The purpose of this paper is two-fold, both to provide a general discussion of radar antennas and to summarize the results of radar antenna research and development at the Bell Laboratories. Part I is a discussion of the basic electrical principles which concern radar antennas. In Part II we will outline the most common methods of radar antenna construction. Practical military antennas developed by the Bell Laboratories will be described in Part III.

The reader who is interested in general familiarity with the over all result rather than with technical features of design may proceed directly from this part to Part III.

PART I

ELECTRICAL PRINCIPLES

1. GENERAL

Radar antenna design depends basically on the same broad principles which underlie any other engineering design. The radar antenna designer can afford to neglect no aspect of his problem which has a bearing on the final product. Mechanical, chemical, and manufacturing considerations are among those which must be taken into account.

It is the electrical character of the antenna, however, which is connected most directly with the radar performance. In addition it is through attention to the electrical design problems that the greatest number of novel antennas have been introduced and it is from the electrical viewpoint that the new techniques can best be understood.

An antenna is an electromagnetic device and as such can be understood

through the application of electromagnetic theory. Maxwell's equations provide a general and accurate foundation for antenna theory. They are the governing authority to which the antenna designer may refer directly when problems of a fundamental or baffling nature must be solved.

It is usually impracticable to obtain theoretically exact and simple solutions to useful antenna problems by applying Maxwell's Equations directly. We can, however, use them to derive simpler useful theories. These theories provide us with powerful analytical tools.

Lumped circuit theory is a tool of this sort which is of immense practical importance to electrical and radio engineers. As the frequency becomes higher the approximations on which lumped circuit theory is based become inaccurate and engineers find that they must consider distributed inductances and capacitances. The realm of transmission line theory has been invaded.

Transmission line theory is of the utmost importance in radar antenna design. In the first place the microwave energy must be brought to the antenna terminals over a transmission line. This feed line is usually a coaxial or a wave-guide. It must not break down under the voltage which accompanies a transmitted pulse. It must be as nearly lossless and reflectionless as possible and it must be matched properly to the antenna terminals.

The importance of a good understanding of transmission line theory does not end at the antenna terminals. In any antenna the energy to be transmitted must be distributed in the antenna structure in such a way that the desired radiation characteristics will be obtained. This may be done with transmission lines, in which case the importance of transmission line theory is obvious. It may be done by 'optical' methods. If so, certain transmission line concepts and methods will still be useful.

While it is true that transmission line theory is important it is not necessary to give a treatment of it in this paper. Adequate theoretical discussions can be found elsewhere in several sources.¹ It is enough at this point to indicate the need for a practical understanding of transmission line principles, a need which will be particularly evident in Part II, *Methods of Antenna Construction*.

We may, if we like, think of the whole radar transmission problem in terms of transmission line theory. The antenna then appears as a transformer between the feed line and transmission modes in free space. We cannot, however, apply this picture to details with much effectiveness unless we have some understanding of radiation.

In the sections to follow we shall deal with some theoretical aspects of radiation. We shall begin with a discussion of fundamental *transmission*

¹ See, for example, S. A. Schelkunoff, *Electromagnetic Waves*, D. Van Nostrand Co., Inc., 1943, in particular, Chapters VII and VIII, or F. E. Terman, *Radio Engineer's Handbook*, McGraw-Hill Book Co., Inc., 1943, Section 3.

principles. This discussion is applicable to all antennas regardless of how they are made or used. When applied to radar antennas it deals chiefly with those properties of the antenna which affect the radar range.

Almost all microwave radar antennas are large when measured in wavelengths. When used as transmitting antennas they produce desired radiation characteristics by distributing the transmitted energy over an area or 'wave front'. The relationships between the phase and amplitude of electrical intensity in this wave front and the radiation characteristics of the antenna are predicted by *wave front analysis*. Wave front analysis is essentially the optical theory of diffraction. Although approximate it applies excellently to the majority of radar antenna radiation problems. We shall discuss wave front analysis in Section 3.

2. TRANSMISSION PRINCIPLES

2.1 Gain and Effective Area of an Antenna

An extremely important property of any radar antenna is its ability to project a signal to a distant target. The *gain* of the antenna is a number which provides a quantitative measure of this ability. Another important property of a radar antenna is its ability to collect reflected power which is returning from a distant target. The *effective area* of the antenna is a quantitative measure of this ability. In this section these two quantities will be defined, and a simple relation between them will be derived. Their importance to radar range will be established.

Definition of Gain. When power is fed into the terminals of an antenna some of it will be lost in heat and some will be radiated. The *gain* G of the antenna can be defined as the ratio

$$G = P/P_o \quad (1)$$

where P is the power flow per unit area in the plane linearly polarized electromagnetic wave which the antenna causes in a distant region usually in the direction of maximum radiation and P_o is the power flow per unit area which would have been produced if all the power fed into the terminals had been radiated equally in all directions in space.

Definition of Effective Area. When a plane linearly polarized electromagnetic wave is incident on the receiving antenna, received power P_R will be available at the terminals of the antenna. The *effective area* of the antenna is defined, by the equation

$$A = P_R/P' \quad (2)$$

where P' is the power per unit area in the incident wave. In other words the received power is equal to the power flow through an area that is equal to the effective area of the antenna.

2.2 Relationship between Gain and Effective Area

Figure 4 shows a radio circuit in free space made up of a transmitting antenna T and a receiving antenna R . If the transmitted power P_T had

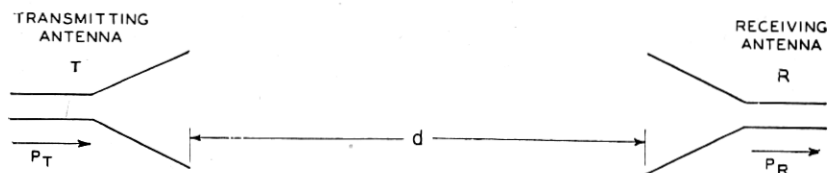


Fig. 4—Radio Circuit in Free Space.

been radiated equally in all directions, the power flow per unit area at the receiving antenna would be

$$P_0 = P_T \frac{1}{4\pi d^2} \quad (3)$$

Definition (1) gives, therefore, for the power flow per unit area at the receiving antenna

$$P = P_0 G_T = \frac{P_T G_T}{4\pi d^2} \quad (4)$$

and definition (2) gives for the received power

$$P_R = P A_R = \frac{P_T G_T A_R}{4\pi d^2} \quad (5)$$

From the law of reciprocity it follows that the same power is transferred if the transmitting and receiving roles are reversed. By (5) it is thus evident that

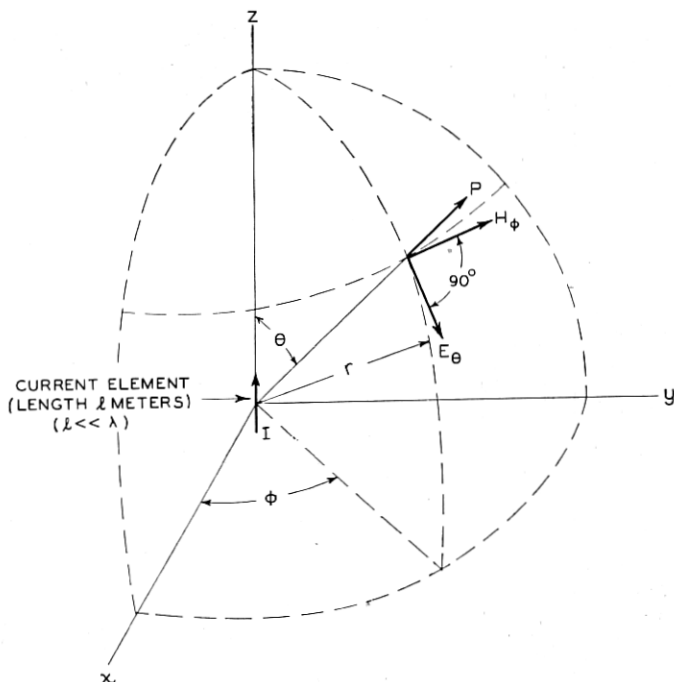
$$\begin{aligned} G_T A_R &= G_R A_T \\ \text{or} \\ G_T/A_T &= G_R/A_R \end{aligned} \quad (6)$$

Equation (6) shows that the ratio of the gain and effective area has the same constant value for all antennas at a given frequency. It is necessary, therefore, to calculate this ratio only for a simple and well known antenna such as a small dipole or uniform current element.

2.3 The Ratio G/A for a Small Current Element

In Fig. 5 are given formulas² in M.K.S. units for the free space radiation from a small current element with no heat loss. We have assumed that

² See S. A. Schelkunoff, *Electromagnetic Waves*, D. Van Nostrand Co., Inc., 1943, p. 133



FOR $r \gg \lambda$:

$$\text{MAGNETIC INTENSITY} = H_{\phi} = i \frac{I l}{2 \lambda r} e^{-i \frac{2 \pi}{\lambda} r} \sin \theta \quad \frac{\text{AMPERES}}{\text{METER}} \quad (1)$$

$$\text{ELECTRIC INTENSITY} = E_{\theta} = 120 \pi H_{\phi} \quad \frac{\text{VOLTS}}{\text{METER}} \quad (2)$$

$$\text{POWER FLOW} = P = |H_{\phi} E_{\theta}| = 30 \pi \left[\frac{I l}{\lambda r} \right]^2 \sin^2 \theta \quad \frac{\text{WATTS}}{\text{METER}^2} \quad (3)$$

$$P \text{ IS MAXIMUM FOR } \theta = 90^{\circ}. \text{ i.e., } P_{\text{MAX}} = 30 \pi \left[\frac{I l}{\lambda r} \right]^2 \quad \frac{\text{WATTS}}{\text{METER}^2} \quad (4)$$

POWER FLOW ACROSS SPHERE OF RADIUS r OR

$$\text{TOTAL RADIATION} = W = \int_0^{\pi} P 2 \pi r \sin \theta r d\theta = 80 \pi^2 \left[\frac{I l}{\lambda} \right]^2 \text{ WATTS} \quad (5)$$

$$\text{RADIATION RESISTANCE} = R_{\text{RAD}} = \frac{W}{I^2} = 80 \pi^2 \left(\frac{l}{\lambda} \right)^2 \text{ OHMS} \quad (6)$$

$$\text{BY (4) AND (5): } P_{\text{MAX}} = \frac{3}{8 \pi r^2} W \quad \frac{\text{WATTS}}{\text{METER}^2} \quad (7)$$

Fig. 5—Free Space Radiation from a Small Current Element with Uniform Current I Amperes over its Entire Length.

this element is centered at the origin of a rectangular coordinate system and that it lies along the Z axis. At a large distance r from the element

the maximum power flow per unit area occurs in a direction normal to it and is given by

$$P_{\max} = \frac{3W}{8\pi r^2} \frac{\text{watts}}{\text{meter}^2} \quad (7)$$

where W is the total radiated power. If W had been radiated equally in all directions the power flow per unit area would be

$$P_0 = \frac{W}{4\pi r^2} \frac{\text{watts}}{\text{meters}^2} \quad (8)$$

It follows that the gain of the small current element is

$$G_{\text{dipole}} = \frac{P_{\max}}{P_0} = 1.5 \quad (9)$$

The effective area of the dipole will now be calculated. When it is used to receive a plane linearly polarized electromagnetic wave, the available output power is equal to the induced voltage squared divided by four times the radiation resistance. Thus

$$P_R = \frac{E^2 \ell^2}{4R_{\text{rad}}} \text{Watts} \quad (10)$$

where E is the effective value of the electric field of the wave, ℓ is the length of the current element and R_{rad} is the radiation resistance of the current element. From Fig. 5 we see that $R_{\text{rad}} = \frac{80\pi^2 \ell^2}{\lambda^2}$ ohms. Since the power flow per unit area is equal to the electric field squared divided by the impedance of free space, in other words $P_0 = \frac{E^2}{120\pi}$ we have

$$A_{\text{dipole}} = \frac{P_R}{P_0} = \frac{3\lambda^2}{8\pi} \text{meter}^2 \quad (11)$$

We combine formulas (9) and (11) to find that

$$\frac{G_{\text{dipole}}}{A_{\text{dipole}}} = \frac{4\pi}{\lambda^2}$$

Since, as proved in 2.2 this ratio is the same for all antennas, it follows that for any antenna

$$\frac{G}{A} = \frac{4\pi}{\lambda^2} \quad (12)$$

2.4 The General Transmission Formula

Transmission loss between transmitter and receiver through the radio circuit shown in Fig. 4 was given by equation (5). By substituting the relation (12) into (5) we can obtain the simple free space transmission formula:

$$P_R = P_T \frac{A_T A_R}{\lambda^2 d^2} \text{ watts} \quad (13)$$

Although this formula applies to free space only it is believed to be as useful in radio engineering as Ohm's law is in circuit engineering.

2.5 The Reradiation Formula

One further relation, the radar reflection formula is of particular interest. Consider the situation illustrated in Fig. 6. Let P_T be the power radiated

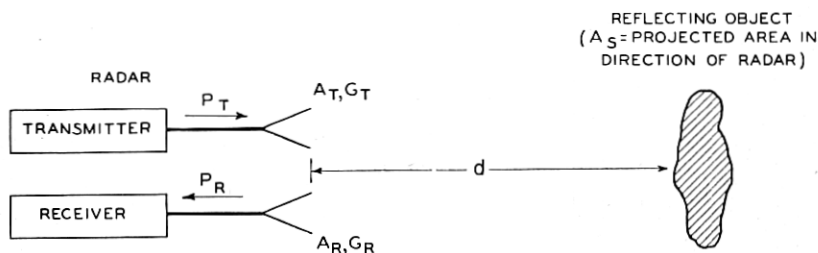


Fig. 6—Radar with Separate Receiving and Transmitting Antennas.

from an antenna with effective area A_T , A_S the area of a reflecting object at distance d from the antenna and P_R the power received by an antenna of effective area A_R . By equation (13) the power striking A_S is $\frac{P_T A_T A_S}{\lambda^2 d^2}$. If this power were reradiated equally in all directions the reflected power flow at the receiving antenna would be $\frac{P_T A_T A_S}{4\pi d^2 \lambda^2}$ but since the average reradiation is larger toward the receiving antenna, the power flow per unit area there is usually $K \frac{P_T A_T A_S}{4\pi d^2 \lambda^2}$ where $K \gg 1$. It follows from (2) that

$$P_R = K \frac{P_T A_T A_R A_S}{4\pi \lambda^2 d^4} \quad (14)$$

Formula (14) shows clearly why the use of large and efficient antennas will greatly increase the radar range.

Formula (14) applies to free space only. Application to other conditions

may require corrections for the effect of the "ground", and for the effect of the transmission medium, which are beyond the scope of this paper.

2.6 *The Plane, Linearly Polarized Electromagnetic Wave*

In the foregoing sections we have referred several times to 'plane, linearly polarized electromagnetic waves'. These waves occur so commonly in antenna theory and practice that it is worth while to discuss them further here.

Some properties of linearly polarized, plane electromagnetic waves are illustrated in Fig. 7. At any point in the wave there is an electric field and a magnetic field. These fields are vectorial in nature and are at right angles to each other and to the direction of propagation. It is customary to give the magnitude of the electric field only.

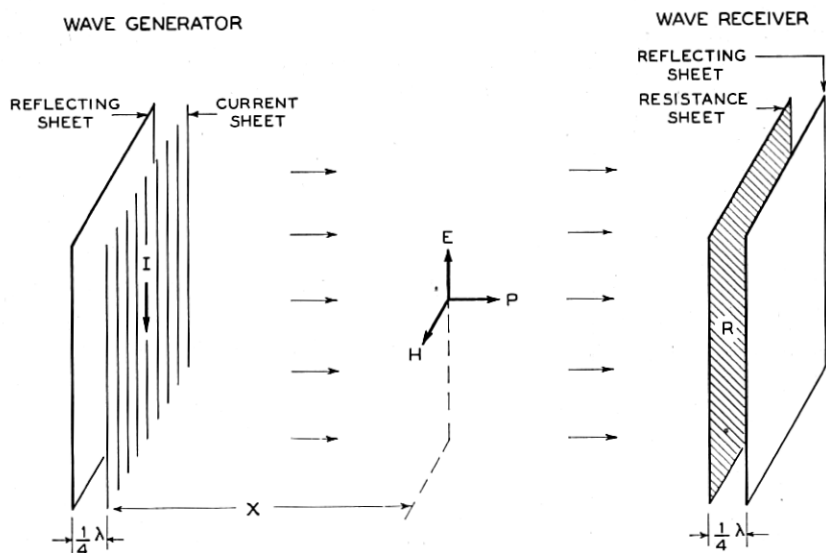
If we use the M.K.S. system of units the magnitudes of the fields are expressed in familiar units. Electric intensity appears as volts per meter and magnetic intensity as amperes per meter. The ratio of electric to magnetic intensity has a value of 120π or about 377 ohms. This is the 'impedance' of free space. The power flow per unit area is expressed in watts per square meter. We see, therefore, that the electromagnetic wave is a means for carrying energy not entirely unlike a familiar two wire line or a coaxial cable.

Electromagnetic waves are generated when oscillating currents flow in conductors. We could generate a plane linearly polarized electromagnetic wave with a uniphase current sheet consisting of a network of fine wires backed up with a conducting reflector as shown in Fig. 7. This wave could be absorbed by a plane resistance sheet with a resistivity of 377 ohms, also backed up by a conducting sheet. The perfectly conducting reflecting sheets put infinite impedances in parallel with the current sheet and the resistance sheet, since each of these reflecting sheets has a zero impedance at a spacing of a quarter wavelength.

A perfectly plane electromagnetic wave can exist only under certain ideal conditions. It must be either infinite in extent or bounded appropriately by perfect electric and magnetic conductors. Nevertheless thinking in terms of plane electromagnetic waves is common and extremely useful. In the first place the waves produced over a small region at a great distance from any radiator are essentially plane. Arguments concerning receiving antennas therefore generally assume that the incident waves are plane. In the second place an antenna which has dimensions of many wavelengths can be analyzed with considerable profit on the basis of the assumption that it transmits by producing a nearly plane electromagnetic wave across its aperture. This method of analysis can be applied to the majority of microwave radar antennas, and will be discussed in the following sections.

3. WAVE FRONT ANALYSIS

The fundamental design question is "How to get what we want?" In a radar antenna we want specified radiation characteristics; gain, pattern and polarization. Electromagnetic theory tells us that if all electric and magnetic currents in an antenna are known its radiation characteristics may be derived with the help of Maxwell's Equations. However, the essence of electromagnetic theory insofar as it is of use to the radar antenna



$$\begin{aligned} \text{MAGNETIC INTENSITY} &= H = Ie^{-i\frac{2\pi}{\lambda}x} \frac{\text{AMPERES}}{\text{METER}} \\ \text{ELECTRIC INTENSITY} &= E = 120\pi H \frac{\text{VOLTS}}{\text{METER}} \\ \text{POWER FLOW} &= P = EH \frac{\text{WATTS}}{\text{METER}^2} \\ \text{CURRENT DENSITY} &= I \frac{\text{AMPERES}}{\text{METER}} \\ \text{RESISTIVITY} &= R = 120\pi \text{ OHMS} \end{aligned}$$

Fig. 7—Linearly Polarized Plane Electromagnetic Waves.

designer can usually be expressed in a simpler, more easily visualized and thus more useful form. This simpler method we call *wave front analysis*.

In a transmitting microwave antenna the power to be radiated is used to produce currents in antenna elements which are distributed in space. This distribution is usually over an area, it may be discrete as with a dipole array or it may be continuous as in an electromagnetic horn or paraboloid. These currents generate an advancing electromagnetic wave over the aperture of

the antenna. The amplitude, phase and polarization of the electric intensity in portions of the wave are determined by the currents in the antenna and thus by the details of the antenna structure. This advancing wave can be called the 'wave front' of the antenna.

When the wave front of an antenna is known its radiation characteristics may be calculated. Each portion of the wave front can be regarded as a secondary or 'Huygens' source of known electric intensity, phase and polarization. At any other point in space the electric intensity, phase and polarization due to a Huygens source can be obtained through a simple expression given in the next section. The radiation characteristics of the antenna can be found by adding or integrating the effects due to all Huygens sources of the wave front.

This procedure is based on the assumption that the antenna is transmitting. A basic law of reciprocity assures us that the receiving gain and radiation characteristics of the antenna will be identical with the transmitting ones when only linear elements are involved.

This resolution of an antenna wave front into an array of secondary sources can be justified within certain limitations on the basis of the induction theorem of electromagnetic theory.³ These limitations are discussed in a qualitative way in section 3.13.

3.1 The Huygens Source

Consider an elementary Huygens source of electric intensity E_0 polarized parallel to the X axis with area dS in the XY plane (Fig. 8). This can be thought of as an element of area dS of a wave front of a linearly polarized plane electromagnetic wave which is advancing in the positive z direction.³ From Maxwell's Equations we can determine the field at any point of space due to this Huygens Source. The components of electric field, are found to be

$$\begin{aligned} E_r &= 0 \\ E_\theta &= i \frac{E_0 dS}{2\lambda r} e^{-i(2\pi/\lambda)r} (1 + \cos \theta) \cos \phi \\ E_\phi &= -i \frac{E_0 dS}{2\lambda r} e^{-i(2\pi/\lambda)r} (1 + \cos \theta) \sin \phi \end{aligned} \quad (15)$$

where λ is the wavelength.

We see at once that this represents a vector whose absolute magnitude at all points of space is given by

$$|E| = \frac{E_0 dS}{2\lambda r} (1 + \cos \theta). \quad (16)$$

³ S. A. Schelkunoff, Loc. Cit., Chap. 9.

Here $\frac{E_0 dS}{\lambda}$ is an amplitude factor which depends on the wavelength, intensity and area of the elementary source and $1/r$ is an amplitude factor which specifies the variation of field with distance. $(1 + \cos \theta)$ is an amplitude factor which shows that the directional pattern of the elementary source is a cardioid with maximum radiation in the direction of propagation and no radiation in the reverse direction.

When we use the properties of the Huygens source in analyzing a micro-

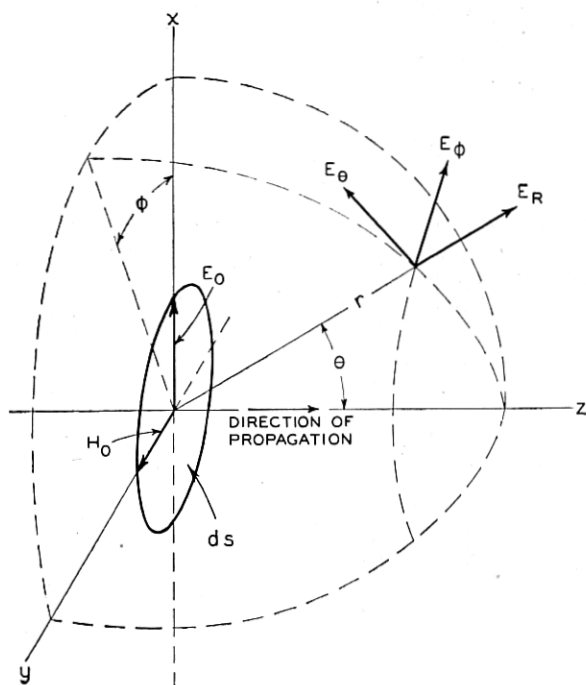


Fig. 8—The Huygens Source.

wave antenna we are usually concerned principally with radiation in or near the direction of propagation. For such radiation Equation 16 takes a particularly simple form in Cartesian Coordinates

$$E_x \cong i \frac{E_0 dS}{\lambda r} e^{-i(2\pi/\lambda)r}; E_y \cong 0; E_z \cong 0. \quad (17)$$

This represents an electric vector nearly parallel to the electric vector of the source. The amplitude is given by the factor $\frac{E_0 dS}{\lambda r}$ and the phase by the

factor $i e^{-i(2\pi/\lambda)r}$. With this equation as a basis we will now proceed to study some relevant matters concerning radar antennas.

3.2 Gain and Effective Area of an Ideal Antenna

On the basis of (17) we can now determine the gain of an ideal antenna of area S ($S \gg \lambda^2$). This antenna is assumed to be free of heat loss and to transmit by generating an advancing wave which is uniform in phase and amplitude in the XY plane. Let the electric intensity in the wave front of

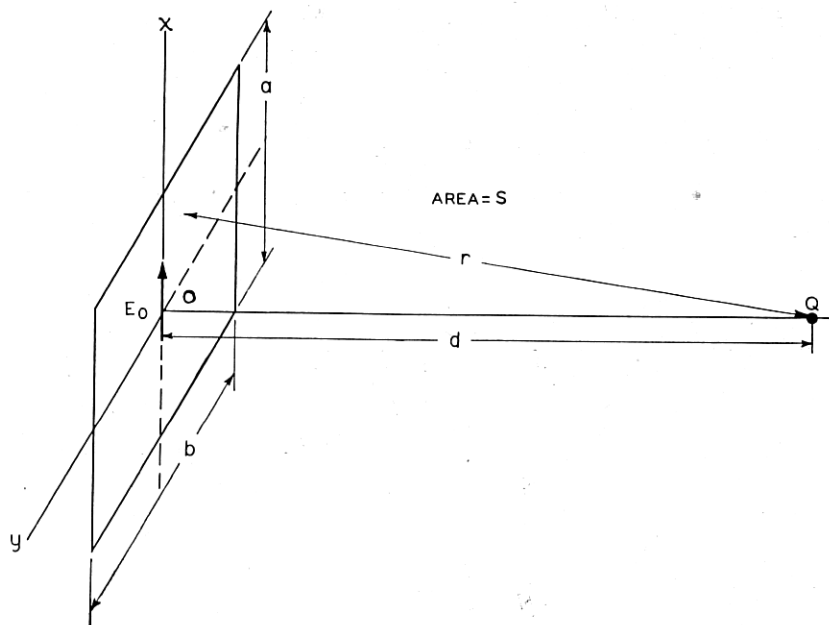


Fig. 9—An Ideal Antenna.

the ideal antenna be E_0 polarized parallel to the X axis (Fig. 9). The transmitted power P_T is equal to the power flow through S and is given by

$$P_T = \frac{E_0^2}{120\pi} S. \quad (18)$$

At a point Q on the Z axis the electric intensity is obtained by adding the effects of all the Huygens sources in S . If the distance of Q from O is so great that

$$r = d + \Delta$$

where Δ is a negligibly small fraction of a wavelength for every point on S then we see from (17) that the electric vector at Q is given by

$$E_x = \int_S i \frac{E_0 dS}{\lambda r} e^{-i(2\pi/\lambda)r} = i e^{-i(2\pi/\lambda)d} \frac{E_0 S}{\lambda d}; E_y = 0; E_z = 0. \quad (19)$$

The power flow per unit area at Q is therefore

$$P = \frac{1}{120\pi} \frac{E_0^2 S^2}{\lambda^2 d^2} = \frac{P_T S}{\lambda^2 d^2}$$

P_0 the power flow per unit area at Q when power is radiated isotropically from O is found by assuming that P_T is spread evenly over the surface of a sphere of radius d .

$$P_0 = \frac{P_T}{4\pi d^2} \quad (20)$$

The gain of a lossless, uniphase, uniamplitude, linearly polarized antenna is, by the definition of equation 1, the ratio of 19 and 20.

$$G_s = \frac{P_T S}{\lambda^2 d^2} \bigg/ \frac{P_T}{4\pi d^2} = \frac{4\pi S}{\lambda^2}. \quad (21)$$

It follows from 12 that the effective area of the ideal antenna is

$$A = S \quad (22)$$

In other words in this ideal antenna the effective area is equal to the actual area. This is a result which might have been obtained by more direct arguments.

3.3 Gain and Effective Area of an Antenna with Aperture in a Plane and with Arbitrary Phase and Amplitude

Let us consider an antenna with a wave front in the XY plane which has a known phase and amplitude variation. Let the electric intensity in the wave front be

$$E(x, y) = E_0 a(x, y) e^{i\phi(x, y)} \quad (23)$$

polarized parallel to the x axis. The radiated power is equal to the power flow through S and is given by

$$P_{\text{rad}} = \frac{E_0^2 \int a^2(x, y) dS}{120\pi}. \quad (24)$$

The input power to the antenna is

$$P_T = P_{\text{rad}}/L \quad (25)$$

where L is a loss factor (< 1). At a point Q on the Z axis the electric intensity is obtained by adding the effects of all the Huygens sources in S . If OQ is as great as in the above derivation for the gain of an ideal antenna then we see from 17 that the electric intensity at Q is

$$E_x = i \frac{e^{i(2\pi/\lambda)d}}{\lambda d} E_0 \int a(x, y) e^{i\phi(x, y)} dS; E_y = 0; E_z = 0. \quad (26)$$

The power flow per unit area at Q is given by

$$P = \frac{1}{120\pi} |E_x|^2 \quad (27)$$

and P_0 the power flow per unit area at Q when P_T is radiated isotropically is given by equation (3).

The power gain of the antenna, by definition 1 is therefore

$$G = \frac{P}{P_0} = \frac{|E_x|^2}{120\pi} \bigg/ \frac{P_T}{4\pi d^2} = \frac{4\pi L}{\lambda^2} \frac{\left| \int_S a(x, y) e^{i\phi(x, y)} dS \right|^2}{\int_S a^2(x, y) dS}. \quad (28)$$

The gain expressed in db is given by

$$G_{db} = 10 \log_{10} G \quad (29)$$

We combine 12 and 28 to obtain

$$A = L \frac{\left| \int_S a(x, y) e^{i\phi(x, y)} dS \right|^2}{\int_S a^2(x, y) dS} \quad (30)$$

a formula for the effective area of the antenna.

3.4 The Significance of the Pattern of a Radar Antenna

The accuracy with which a radar can determine the directions to a target depends upon the *beam widths* of the radar antenna. The ability of the radar to separate a target from its background or distinguish it from other targets depends upon the beam widths and the *minor lobes* of the radar antenna. The efficiency with which the radar uses the available power to view a given region of space depends on the *beam shape* of the antenna. These quantities characterize the antenna pattern. In the following sections means for the calculation of antenna patterns in terms of wave front theory will be developed, and some illustrations will be given.

3.5 Pattern in Terms of Antenna Wave Front

If the relative phase and amplitude in a wave front are given by

$$E(x, y) = a(x, y)e^{i\phi(x, y)} \quad (31)$$

the relative phase and amplitude at a distant point Q not necessarily on the Z axis (Fig. 10) in the important case where the angle QOZ between the direction of propagation and the direction to the point is small, is given from (17) by adding the contributions at Q due to all parts of the wave front. This gives

$$E_Q = \frac{i}{\lambda d} \int_S e^{-i(2\pi/\lambda)r} e^{i\phi(x, y)} a(x, y) dS. \quad (32)$$

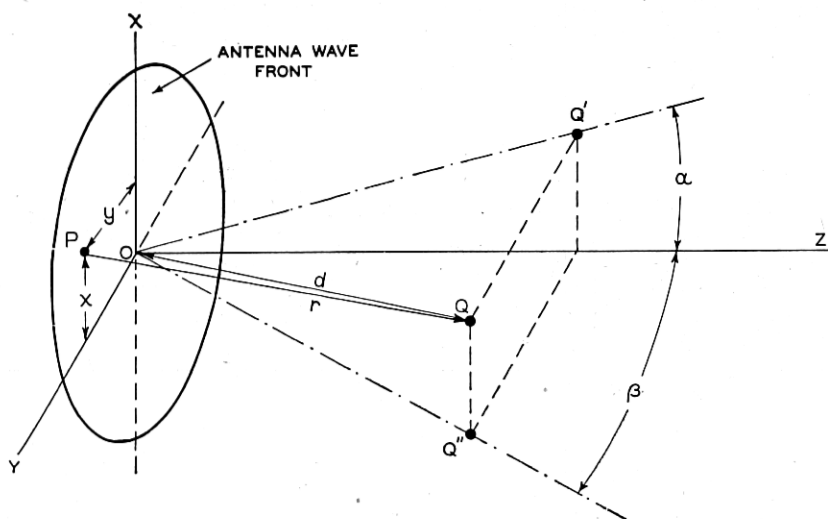


Fig. 10—Geometry of Pattern Analysis.

The quantity r in (32) is the distance from any point P with coordinates x, y , in the XY , plane to the point Q (Fig. 10). Simple trigonometry shows that when OQ is very large

$$r = d - x \sin \alpha - y \sin \beta \quad (33)$$

where d is the distance OQ , α is the angle ZOQ' between OZ and OQ' the projection of OQ on the XZ plane and β is similarly the angle ZOQ'' . The substitution of 33 into 32 gives

$$E_Q = \frac{i e^{-i(2\pi/\lambda)d}}{\lambda d} \int_S e^{i(2\pi/\lambda)(x \sin \alpha + y \sin \beta) + i\phi(x, y)} a(x, y) dS. \quad (34)$$

In most practical cases this equation can be simplified by the assumptions

$$\begin{aligned}\phi(x, y) &= \phi'(x) + \phi''(y) \\ a(x, y) &= a'(x)a''(y)\end{aligned}$$

from which it follows that

$$|E_Q| = F(d)F(\alpha)F(\beta) \quad (35)$$

where $F(d)$ is an amplitude factor which does not depend on angle,

$$F(\alpha) = \int e^{i(2\pi/\lambda)xs\sin\alpha + i\phi'(x)} a'(x) dx \quad (36)$$

is a directional factor which depends only on the angle α and not on the angle β or d , and $F(\beta)$ similarly depends on β but not on α or d . The pattern of an antenna can be calculated with the help of the simple integrals as in 36, and illustrations of such calculations will be given in the following sections.

3.6 Pattern of an Ideal Rectangular Antenna

Let the wave front be that of an ideal rectangular antenna of dimensions a, b ; with linear polarization and uniform phase and amplitude. The dimensions a and b can be placed parallel to the X and Y axes respectively as sketched in Fig. 9. Equation 36 then gives

$$F(\alpha) = \int_{-a/2}^{a/2} e^{i(2\pi/\lambda)xs\sin\alpha} dx = a \frac{\sin \psi}{\psi} \quad (37)$$

$$\text{where } \psi = \frac{\pi a \sin \alpha}{\lambda}.$$

Similarly

$$F(\beta) = b \frac{\sin \psi'}{\psi'} \quad (38)$$

$$\text{where } \psi' = \frac{\pi b \sin \beta}{\lambda}.$$

The pattern of the ideal rectangular aperture, in other words the distribution of electrical field in angle is thus given approximately by

$$F(\alpha)F(\beta) = ab \frac{\sin \psi}{\psi} \frac{\sin \psi'}{\psi'}. \quad (39)$$

The function $\frac{\sin \psi}{\psi}$ is plotted in Fig. 11. It is perhaps the most useful function of antenna theory, not because ideal antennas as defined above are particularly desirable in practice but because they provide a simple stand-

ard with which more useful but more complex antennas can profitably be compared.

3.7 Effect on Pattern of Amplitude Taper

The $\frac{\sin \psi}{\psi}$ pattern which results from an ideal wave front has undesirably high minor lobes for most radar applications. These minor lobes will be reduced if the wave front of constant amplitude is replaced by one which retains a constant phase but has a rounded or 'tapered' amplitude distribution.

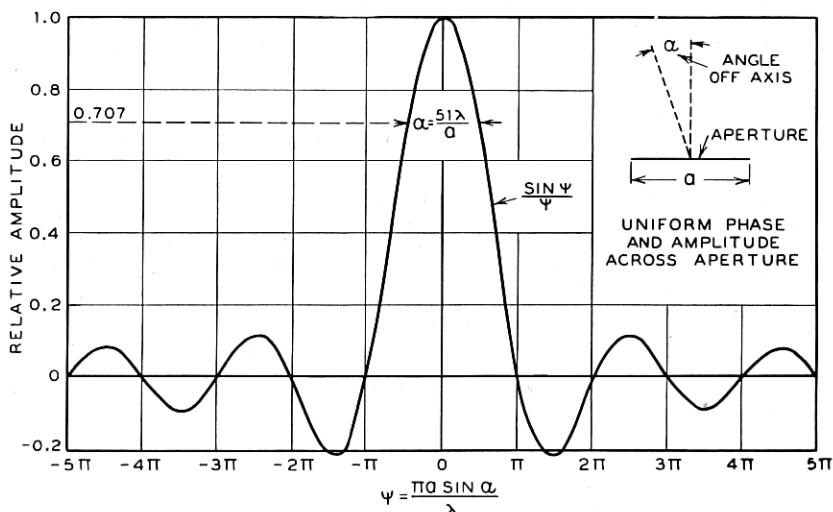


Fig. 11—Pattern of Ideal Rectangular Antenna.

If such an amplitude taper is represented analytically by the function

$$a'(x) = C_1 + C_2 \cos \frac{\pi x}{a} \quad (40)$$

then equation (36) is readily integrable. To integrate it we utilize the identity

$$\cos \frac{\pi x}{a} = \frac{e^{+i\pi x/a} + e^{-i\pi x/a}}{2}$$

upon which the integral becomes the sum of three simple integrals of the form

$$\int_{-a/2}^{a/2} e^{ikx} dx = a \left[\frac{\sin \frac{ka}{2}}{\frac{ka}{2}} \right] \quad (41)$$

We therefore obtain

$$F(\alpha) = aC_1 \frac{\sin \psi}{\psi} + a \frac{C_2}{2} \left[\frac{\sin \left(\psi + \frac{\pi}{2} \right)}{\left(\psi + \frac{\pi}{2} \right)} + \frac{\sin \left(\psi - \frac{\pi}{2} \right)}{\left(\psi - \frac{\pi}{2} \right)} \right] \quad (42)$$

The patterns resulting from two possible tapers are given by substituting $C_1 = 0, C_2 = 1$ and $C_1 = 1/3, C_2 = 2/3$ in (42). These patterns are evidently calculable in terms of the known function $\frac{\sin \alpha}{\alpha}$. They are plotted in Figs. 12 and 13.

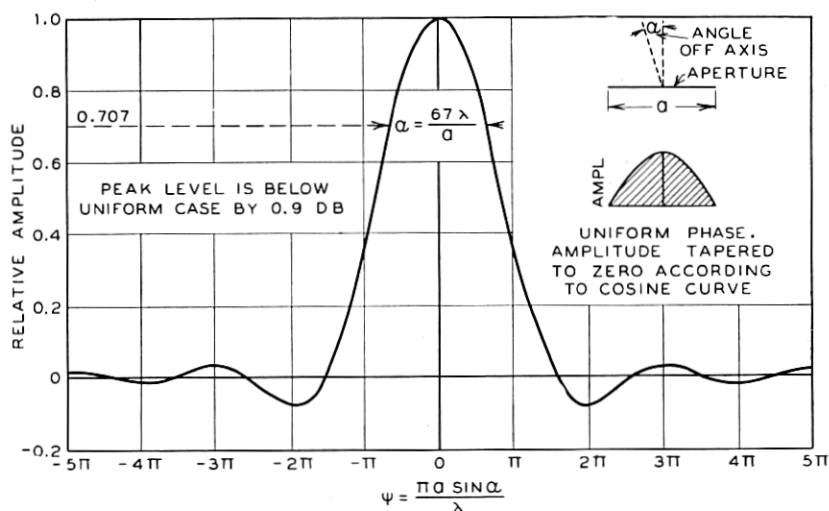


Fig. 12—Pattern of Tapered Rectangular Antenna.

It will be observed that minor lobe suppression through tapering is obtained at the expense of beam broadening. In addition to this the gain is reduced by tapering, as could have been calculated from 28. These undesirable effects must be contended with in any practical antenna design. The choice of taper must be made on the basis of the most desirable compromise between the conflicting factors.

3.8 Effect on Pattern of Linear Phase Variation

If we assume a constant amplitude and a linear phase variation

$$\phi'(x) = -k_1 x$$

over an aperture $-a/2 < x < a/2$ then 36 becomes a simple integral of the form (41) and we obtain

$$F(\alpha) = a \frac{\sin \psi''}{\psi''} \quad \text{where} \quad \psi'' = \frac{\pi a}{\lambda} \sin \alpha - \frac{k_1 a}{2} \quad (43)$$

The physical interpretation of (43) is simply that the pattern is identical to the pattern of an antenna with constant amplitude and uniform phase but rotated through an angle θ where

$$\sin \theta = \frac{k_1 \lambda}{2\pi}$$

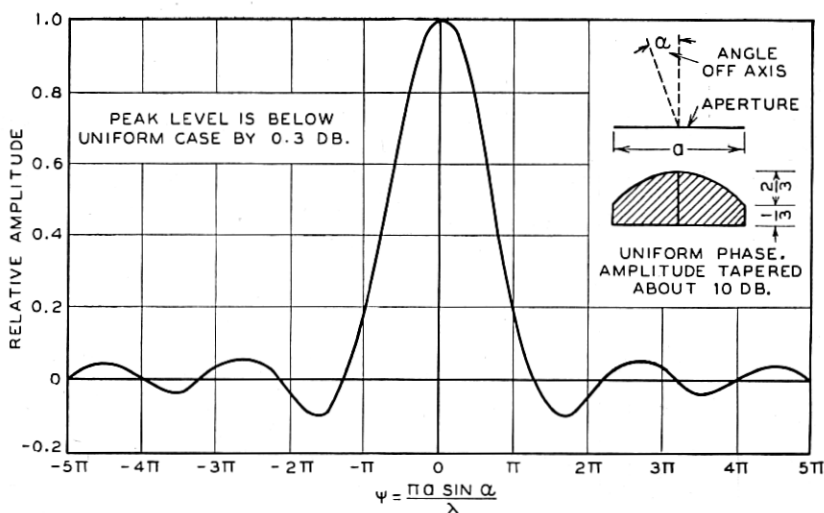


Fig. 13—Pattern of Tapered Rectangular Antenna.

Simple examination shows that the new direction of the radiation maximum is at right angles to a uniphase surface, as we would intuitively expect. This phenomenon has particular relevance to the design of scanning antennas.

3.9 Effect on Pattern of Square Law Phase Variation

If we assume a constant amplitude and a square law phase variation

$$\phi'(x) = -k_2^2 x^2$$

over the aperture $-a/2 < x < a/2$ then the substitution

$$x = \frac{1}{k_2} \left[X + \frac{2\pi}{\lambda} \frac{\sin \alpha}{2k_2} \right] \quad (44)$$

reduces (36) to the form

$$F(\alpha) = \frac{1}{k_2} e^{i\left(\frac{2\pi}{\lambda}\right)^2 \sin^2 \alpha / 4k_2^2} \int e^{-iX^2} dX \quad (45)$$

Equation (45) can be evaluated with the help of Fresnel's Integrals

$$\int \cos X^2 dX, \quad \int \sin X^2 dX$$

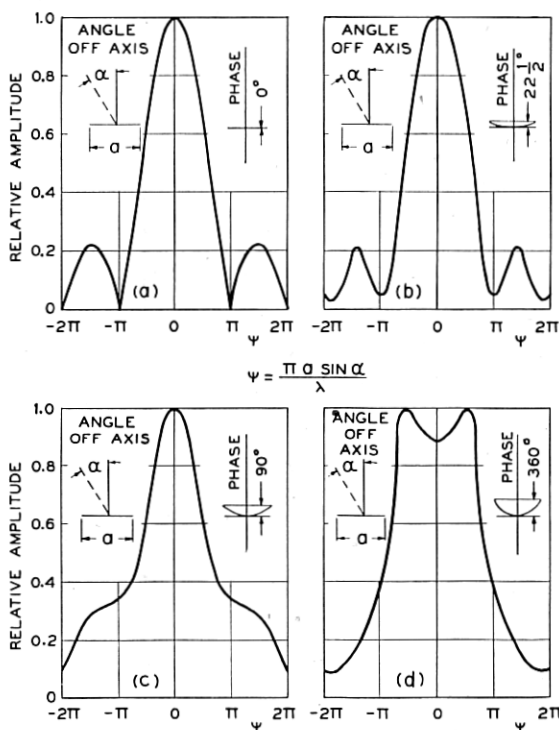


Fig. 14—Patterns of Rectangular Apertures with Square Law Phase Variation.

which are tabulated⁴, or from Cornu's Spiral which is a convenient graphical representation of the Fresnel Integrals.

Typical computed patterns for apertures with square law phase variations are plotted in Fig. 14. These theoretical curves can be applied to the following important practical problems.

(1) The pattern of an electromagnetic horn.

⁴ For numerical values of Fresnel's Integrals and a plot of Cornu's Spiral see Jahnke and Emde, *Tables of Functions* B. G. Teubner, Leipzig, 1933, or Dover Publications, New York City, 1943.

- (2) The defocussing of a reflector or lens due to improper placing of the primary feed.
- (3) The defocussing of a zoned reflector or lens due to operation at a frequency off mid-band.

In addition to providing distant patterns of apertures with curved wave fronts (44) provides theoretical 'close in' patterns of antennas with plane wave fronts. This arises from the simple fact that a plane aperture appears as a curved aperture to close in points. The degree of curvature depends on the distance and can be evaluated by extremely simple geometrical considerations. When this has been done we find that Fig. 14 represents the so-called Fresnel diffraction field.

With this interpretation of square law variation of the aperture we can examine several additional useful problems. We can for instance justify the commonly used relation

$$D = \frac{2a^2}{\lambda}$$

for the minimum permissible distance of the field source from an experimental antenna test site. This distance produces an effective phase curvature of $\lambda/16$. We can examine optical antenna systems employing large primary feeds, in particular those employing parabolic cylinders illuminated by line sources.

3.10 Effect on Pattern of Cubic Phase Variation

If we assume a constant amplitude and a cubic phase variation $\phi'(x) = -k_3^3 x^3$ over the aperture from $-a/2 < x < a/2$ then equation (36) becomes

$$F(\alpha) = \int_{-a/2}^{a/2} e^{-ik_3^3 x^3} \cdot e^{i(2\pi/\lambda)x \sin \alpha} \cdot dx \quad (46)$$

If $k_3^3 x^3 \leq \frac{\pi}{2}$ then it is a fairly good approximation to write

$$e^{-ik_3^3 x^3} = 1 - ik_3^3 x^3 - \frac{k_3^6 x^6}{2} + \dots \quad (47)$$

from which it follows that (46) can be integrated since it reduces to a sum of three terms each of which can be integrated.

Typical computed patterns for apertures with cubic phase variation are plotted in Figs. 15 and 16. Cubic phase distortions are found in practice when reflectors or lenses are illuminated by primary feeds which are off axis either because of inaccurate alignment or because beam lobing or scanning through feed motion is desired. The beam distortion due to cubic phase variation is known in optics as 'coma' and the increased unsymmetrical lobe which is particularly evident in Fig. 16 is commonly called a 'coma lobe'.

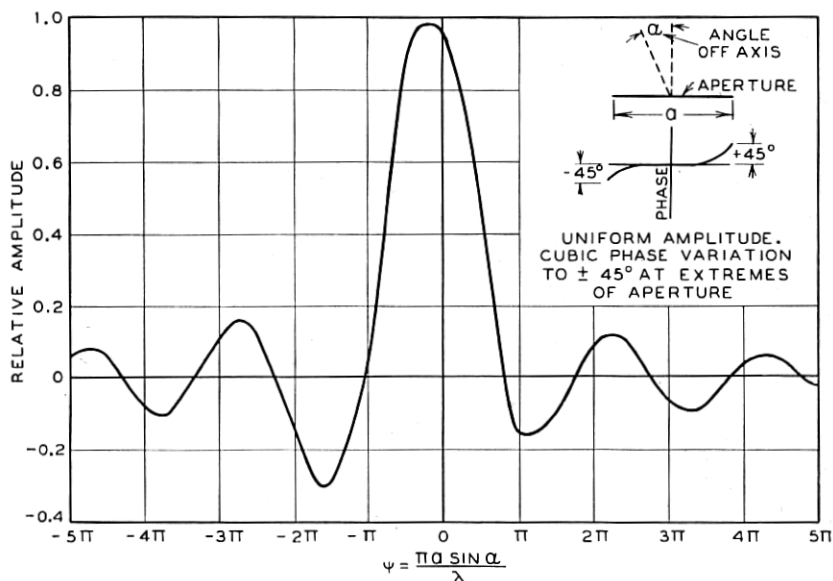


Fig. 15—Pattern of Rectangular Antenna with Cubic Phase Variation.

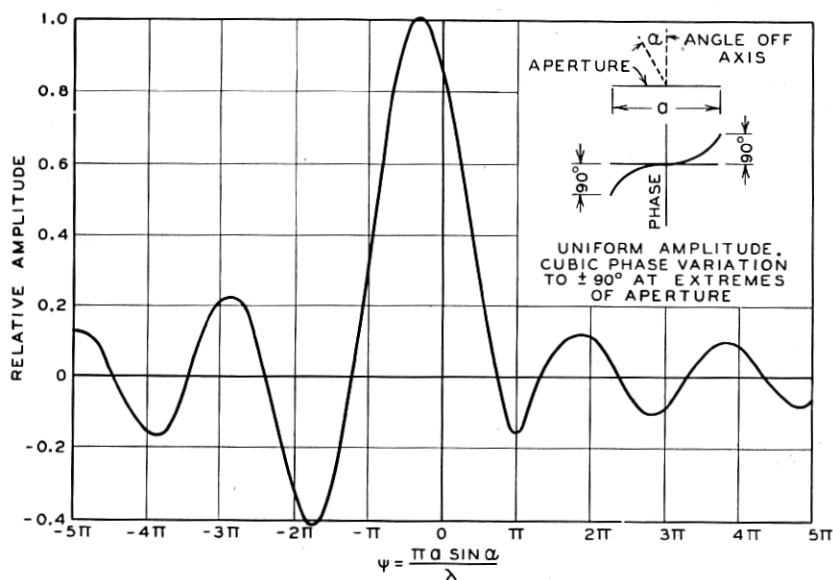


Fig. 16—Pattern of Rectangular Antenna with Cubic Phase Variation.

3.11 Two General Methods

In sections 3.7 and 3.8 we integrated (36) by expressing $a'(x)e^{i\phi'(x)}$ as a sum of terms of the form e^{ikx} . Since $a'(x)e^{i\phi'(x)}$ for finite amplitudes in a finite

aperture can always be expressed as a Fourier sum of this form this solution can in principle always be found.

Alternatively in section 3.10 the integral was evaluated as a sum of integrals of the general type $\int x^n e^{ikx} dx$. Since $a'(x)e^{i\phi'(x)}$ for finite amplitudes in a finite aperture can always be expressed in terms of a power series, this solution can also in principle always be found.

3.12 Arrays

When the aperture consists of an array of component or unit apertures the evaluation of (36) must be made in part through a summation. When all of the elementary apertures are alike this summation can be reduced to the determination of an 'Array Factor'. The pattern of the array is given by multiplying the array factor by the pattern of a single unit.

The pattern of an array of identical units spaced equally at distances somewhat less than a wavelength can be proved to be usually almost equivalent to the pattern of a continuous wave front with the same average energy density and phase in each region.

3.13 Limitations to Antenna Wave Front Analysis

Through the analysis of antenna characteristics by means of wave front theory as based on equation (17) we have been able to demonstrate some of the fundamental theoretical principles of antenna design. The use of this simple approach is justified fully by its relative simplicity and by its applicability to the majority of radar antennas. Nevertheless it cannot always be used. It will certainly be inaccurate or inapplicable in the following cases:

- (1) When any dimension of the aperture is of the order of a wavelength or smaller (as in many primary feeds).
- (2) Where large variations in the amplitude or phase in the aperture occur in distances which are of the order of a wavelength or smaller (as in dipole arrays).
- (3) Where the antenna to be considered does not act essentially through the generation of a plane wave front (as in an end fire antenna or a cosecant antenna).

When the wave front analysis breaks down alternative satisfactory approaches based on Maxwell's equation are sometimes but not always fruitful. Literature on more classical antenna theory is available in a variety of sources. For much fundamental and relevant theoretical work the reader is referred to Schelkunoff.⁵

⁵ S. A. Schelkunoff, Loc. Cit.

4. APPLICATION OF GENERAL PRINCIPLES

In the foregoing sections we have provided some discussion of what happens to a radar signal from the time that the pulse enters the antenna on transmission until the time that the reflected signal leaves the antenna on reception. We have for convenience divided the principles which chiefly concern us into three groups, transmission line theory, transmission principles and wave front theory.

With the aid of transmission line theory we can examine problems concerning locally guided or controlled energy. The details of the problems of antenna construction, such as those to be discussed in Part II frequently demand a grasp of transmission line theory. With it we can study local losses, due to resistance or leakage, which affect the gain of the antenna. We can examine reflection problems and their effect on the match of the antenna. Special antennas, such as those employing phase shifters or transmission between parallel conducting plates, introduce many special problems which lie wholly or partly in the transmission line field.

An understanding of the principles which govern transmission through free space aids us in comprehending the radar antenna field as a whole. Through a general understanding of antenna gains and effective areas we are better equipped to judge their significance in particular cases, and to evaluate and control the effects of particular methods of construction on them.

Wave front theory provides us with a powerful method of analysis through which we can connect the radiation characteristics produced by a given antenna with the radiating currents in the antenna. Through it we can examine theoretical questions concerning beam widths and shape, unwanted radiation and gain.

An understanding of theory is necessary to the radar antenna designer, but it is by no means sufficient. It is easy to attach too much importance to theoretical examination and speculation while neglecting physical facts which can 'make or break' an antenna design. Theory alone provides no substitute for the practical 'know how' of antenna construction. It cannot do away with the necessity for careful experiment and measurement. Least of all can it replace the inventiveness and aggressive originality through which new problems are solved and new techniques are developed.

PART II

METHODS OF ANTENNA CONSTRUCTION

5. GENERAL

Techniques are essential to technical accomplishment. An understanding of general principles alone is not enough. The designing engineer must have

at his disposal or develop practical methods which can produce the results he requires. The effectiveness and simplicity of these methods are fair measures of the degree of technical development.

The study of methods of radar antenna construction is the study of the means by which radar antenna requirements are met. In a broader sense this includes an examination of mechanical structures, of the metals and plastics from which antennas are made, of the processes by which they are assembled, and of the finishes by which they are protected from their environment. It might include a study of practical installation and maintenance procedures. But these matters, which like the rest of Radar have unfolded widely during the war, are beyond the scope of this paper. An adequate discussion of them would have to be based on hundreds of technical reports and instruction manuals and on thousands of manufacturing drawings. The account of methods which is to follow will therefore be restricted to a discussion, usually from the electrical point of view, of the more useful and common radar antenna configurations.

6. CLASSIFICATION OF METHODS

During the history of radar, short as it is, many methods of antenna construction have been devised. To understand the details of all of these methods and the diverse applications of each is a task that lies beyond the ability of any single individual. Nevertheless most of the methods fall into one or another of a limited collection of groups or classifications. We can grasp most of what is generally important through a study of these groups.

In order to provide a basis for classification we will review briefly, from a transmitting standpoint, the action of an antenna. Any antenna is in a sense a transformer between a transmission line and free space. More explicitly, it is a device which accepts energy incident at its terminals, and converts it into an advancing electromagnetic wave with prescribed amplitude, phase and polarization over an area. In order to do this the antenna must have some kind of energy distributing system, some means of amplitude control and some means of phase control. The distributed energy must be suitably controlled in phase, amplitude and polarization.

All antennas perform these functions, but different antennas perform them by different means. Through an examination of the means by which they are performed and the differences between them we are enabled to classify methods of antenna construction.

To distribute energy over its aperture an antenna can use a branching system of transmission lines. When this is done the antenna is an array. Arrays are particularly common in the short wave communication bands, but somewhat less common in the microwave radar bands. In a somewhat simpler method the antenna distributes energy over an area by radiating it

from an initial source or 'primary feed'. This distribution can occur in both dimensions at once, as from a point source. Alternatively the energy can be radiated from a primary source but be constrained to lie between parallel conducting plates so that it is at first distributed only over a long narrow aperture or 'line source'. Distribution over the other dimension occurs only after radiation from the line source.

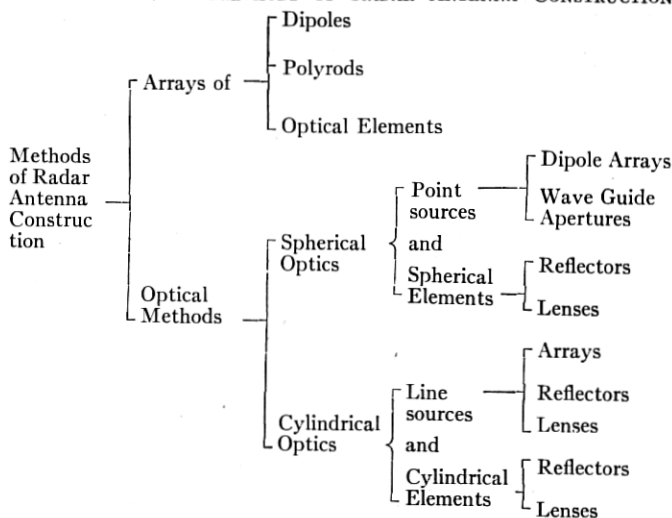
In order to control the amplitude across the aperture of an array antenna we must design the branching junctions so that the desired power division occurs in each one. When the energy is distributed by radiation from a primary source we must control the amplitude by selecting the proper primary feed directivity.

We can control the phase of an array antenna by choosing properly the lengths of the branching lines. Alternatively we can insert appropriate phase changers in the lines.

When the energy is distributed by primary feeds, methods resembling those of optics can be used to control phase. The radiation from a point source is spherical in character. It can be 'focussed' into a plane wave by means of a paraboloidal reflector or by a spherical lens. The radiation from a point source between parallel plates or from a uniphase line source is cylindrical in character. It can be focussed by a parabolic cylinder or a cylindrical lens.

In Table A we have indicated a possible classification of methods of radar antenna construction. This classification is based on the differences discussed in the foregoing paragraphs.

TABLE A
CLASSIFICATION OF METHODS OF RADAR ANTENNA CONSTRUCTION



7. BASIC DESIGN FORMULATION

Certain design factors are common to almost all radar antennas. Because of their importance it would be well to consider these factors in a general way before proceeding with a study of particular antenna techniques.

Almost every radar antenna, regardless of how it is made, has a well defined aperture or wave front. Through wave front analysis we can often examine the connections between the Huygens sources in the antenna aperture and the radiation characteristics of the antenna. We can, in other words, use wave front analysis to study the fundamental antenna design factors, provided the analysis does not violate one of the conditions of section 3.13.

7.1 Dimensions of the Aperture

The dimensions of the aperture of a properly designed antenna are related to its gain by simple and general approximate relations. If the aperture is uniphase and has an amplitude distribution that is not too far from constant the relation

$$G = \frac{4\pi A}{\lambda^2}$$

is useful in connecting the gain of an antenna with the area of its aperture. The effective area is related to the area of the aperture by the equation

$$A = \eta S$$

where η is an efficiency factor. In principle η could have any value but in practice for microwave antennas η has always been less than one. Its value for most uniphase and tapered amplitude antennas is between 0.4 and 0.7. In special cases, e.g. for cosecant antennas or for some scanners its value may be less than 0.4.

The necessary dimensions for the aperture may be determined from the required beam widths in two perpendicular directions. Beam widths are usually specified as half power widths, that is by the number of degrees between directions for which the one way response is 3 db below the maximum response. Figure 11 shows that for an ideal rectangular antenna with uniform phase, polarization and amplitude $\alpha_{P/2} = 51 \frac{\lambda}{a}$ degrees where $\alpha_{P/2}$ = half power width in degrees, a = aperture dimension and λ = wavelength. The relation $\alpha_{P/2} = 65 \frac{\lambda}{a}$ degrees is more nearly correct for the majority of practical antennas with round or elliptical apertures and with uniform phase and reasonably tapered amplitudes.

7.2 Amplitude Distribution

Except where special, in particular cosecant, patterns are desired the principle factors affecting amplitude distribution are efficiency and required minor lobe level. The amplitude distribution or taper of an ideal uniphase rectangular wave front affects the minor lobe level as indicated by Figures 11, 12 and 13. Practical antennas tend to fall somewhat below this ideal picture because of non-uniform phase and because of variations from the ideal amplitude distribution due to discontinuities in the aperture and undesired leakage or spillover of energy. Nevertheless a commonly used rule of thumb is that minor lobes 20 *db* or more below the peak radiation level are tolerable and will not be exceeded with a rounded amplitude taper of 10 or 12 *db*.

7.3 Phase Control

Uniphase wave fronts are used whenever a simple pattern with prescribed gain, beam widths and minor lobes is to be obtained with minimum aperture dimensions. When special results are desired such as cosecant patterns or scanning beams the phase must be varied in special ways.

Mechanical tolerances in the antenna structure make it impossible to hold phases precisely to the desired values. The accuracy with which the phases can be held constant in practice varies with the technique, the antenna size and the wave length. Undesired phase variations increase the minor lobes and reduce the gain of an antenna. The extent to which phase variations can be expected to reduce the gain is indicated in Fig. 17.

8. PARABOLIC ANTENNAS

The headlights of a car or the searchlights of an anti-aircraft battery use reflectors to produce beams of light. Similarly the majority of radar antennas employ reflectors to focus beams of microwave energy. These reflectors may be exactly or approximately parabolic or they may have special shapes to produce special patterns. If they are parabolic they may be paraboloids which are illuminated by point sources and focus in both directions, or they may be parabolic cylinders which focus in only one direction. If they are parabolic cylinders they may be illuminated by line sources or they may be confined between parallel conducting plates and illuminated by point sources to produce line sources.

8.1 Control of Phase

A simple and natural way to distribute energy smoothly in space is to radiate it from a relatively nondirectional 'primary' source such as a dipole array or an open ended wave guide. This energy will be formed into a directive beam if a reflector is introduced to bring it to a plane area or wave front with constant phase. If the primary source is effectively a point as far as

phase is concerned, that is if the radiated energy has the same phase for all points which are the same distance from a given point, then the reflector should be parabolic. This can be proved by simple geometrical means.

In Fig. 18 let the point source S coincide with the point $x = f, y = 0$ of a coordinate system and let the uniphase wave front coincide with the line $x = f$. Let us assume that one point O of the reflector is at the origin. Then it can be shown that any other point of the reflector must lie on the curve

$$y^2 = 4fx$$

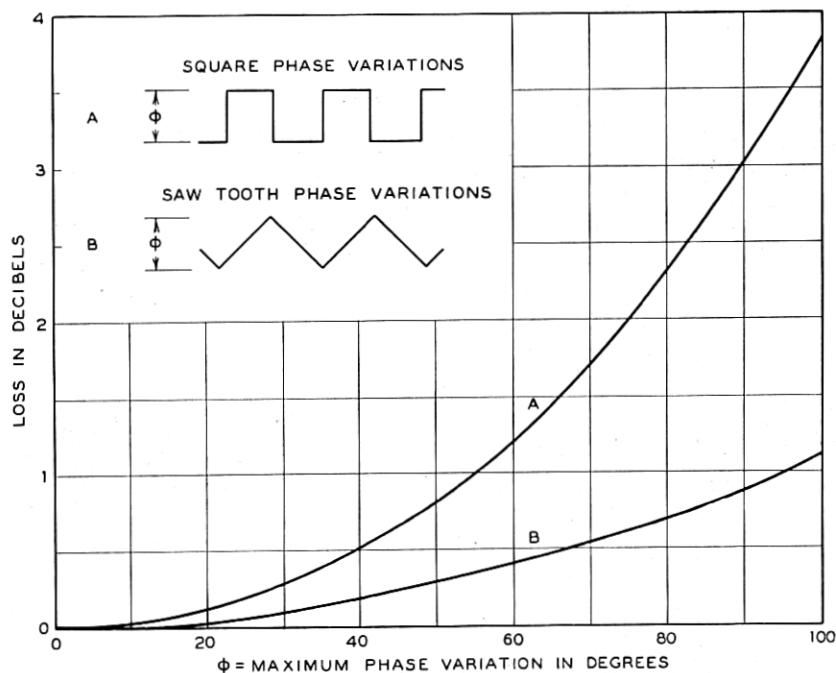


Fig. 17—Loss due to Phase Variation in Antenna Wave Front.

This is a parabola with focus at $f, 0$ and focal length f .

The derivation based on Fig. 18 is two dimensional and therefore in principle applies as it stands only to line source antennas employing parabolic cylinders bounded by parallel conducting planes (Fig. 24 and 25). If Fig. 18 is rotated about the X axis the parabola generates a paraboloid of revolution (Fig. 3). This paraboloid focusses energy spreading spherically from the point source at S in such a way that a uniphase wave front over a plane area is produced. Alternatively Fig. 18 can be translated in the Z direction perpendicular to the XY plane. The parabola then generates a

parabolic cylinder and the point source S generates a line source at the focal line of the parabolic cylinder (Fig. 19). The energy spreading cylindrically from the line source is focussed by the parabolic cylinder in such a way that a uniphase wave front over a plane area is again produced. Parabolic cylinders and paraboloids are both used commonly in radar antenna practice.

In the discussion so far it has been assumed that the primary source is effectively a point source and that the reflector is exactly parabolic. If the primary source is not effectively a point source, in other words if it produces waves which are not purely spherical, then the reflector must be distorted from the parabolic shape if it is to produce perfect phase correction. When

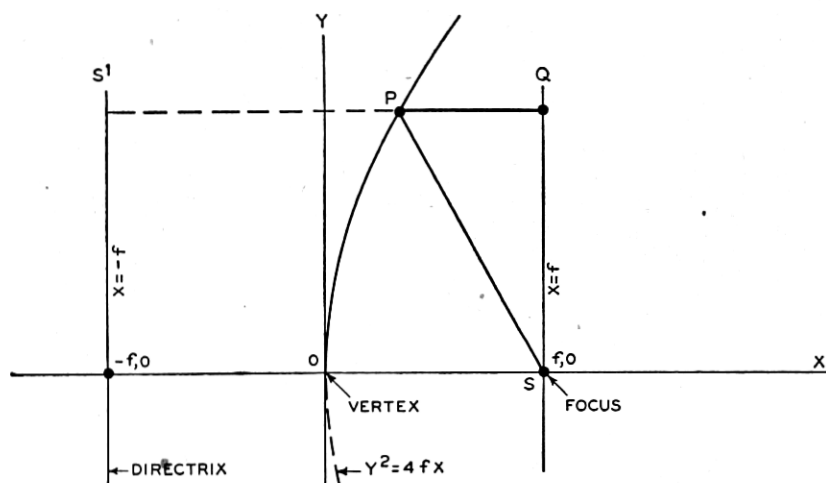


Fig. 18—Parabola.

this occurs the correct reflector shape is sometimes specified on the basis of an experimental determination of phase.

8.2 Control of Amplitude

When a primary source is used to illuminate a parabolic reflector there are two factors which affect the amplitude of the resulting wave front. One of these is of course the amplitude pattern of the primary source. The other is the geometrical or space attenuation factor which is different for different parts of the wave front. In most practical antennas each of these factors tends to taper the amplitude so that it is less at the edges of the antenna than it is in the central region. The effective area of the antenna is reduced by this taper.

In any finite parabolic antenna some of the energy radiated by the primary

source will fail to strike the reflector. The effective area must also be reduced by the loss of this 'spill-over' energy.

The maximum effective area for a parabolic antenna is obtained by designing the primary feed to obtain the best compromise between loss due to taper and loss due to spill-over. It has been shown theoretically that this best compromise generally occurs when the amplitude taper across the aperture is about 10 or 12 *db* and that in the neighborhood of the optimum the efficiency is not too critically dependent on the taper.⁶

This theoretical result is well justified by experience and has been applied to the majority of practical parabolic antennas. It applies both when the reflector is paraboloidal so that taper in both directions must be considered

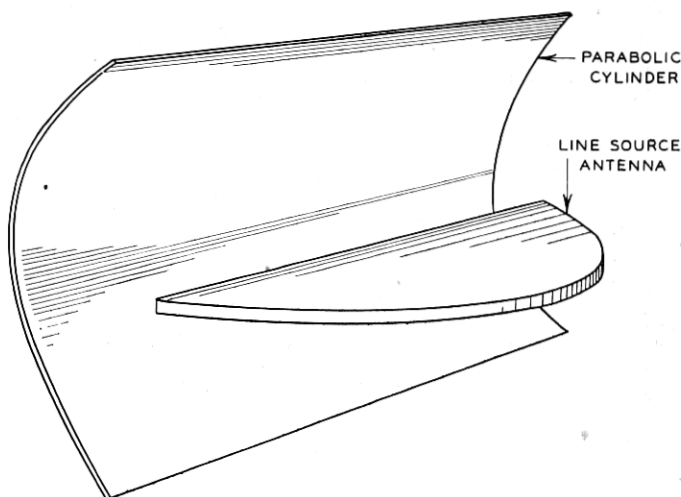


Fig. 19—A Parabolic Cylinder with Line Feed.

and when the reflector is a parabolic cylinder with only a single direction of taper. It is a fortunate by-product of a 10 or 12 *db* taper that it is generally sufficient to produce satisfactory minor lobe suppression.

8.3 Choice of Configuration

We have shown how a simple beam can be obtained through the use of a paraboloidal reflector with a point source or alternatively through the use of a reflecting parabolic cylinder and a line source. The line source itself can be produced with the help of a parabolic cylinder bounded by parallel conducting plates. We will now outline certain practical considerations. These considerations may determine which of the two reflector types will be

⁶ C. C. Cutler, Parabolic Antenna Design for Microwaves, paper to be published in Proc. of the I. R. E.

used for a particular job. They may help in choosing a focal length and in determining which finite portion of a theoretically infinite parabolic curve should be used. Finally they may assist in determining whether reflector technique is really the best for the purpose at hand or whether we could do better with a lens or an array.

In designing a parabolic antenna it must obviously be decided at an early stage whether a paraboloid or one or more parabolic cylinders are to be employed. This choice must be based on a number of mechanical and electrical considerations. Paraboloids are more common in the radar art than parabolic cylinders and are probably to be preferred, yet a categorical a priori judgment is dangerous. It will perhaps be helpful to compare the two alternatives by the simple procedure of enumerating some features in which each is usually preferable to the other.

Paraboloidal antennas

- (a) are simpler electrically, since point sources are simpler than line sources.
- (b) are usually lighter.
- (c) are more efficient.
- (d) have better patterns in the desired polarization.
- (e) are more appropriate for conical lobing or spiral scanning.

Antennas employing parabolic cylinders

- (a) are simpler mechanically since only singly curved surfaces are required.
- (b) have separate electrical control in two perpendicular directions.

This last advantage of parabolic cylinders is important in special antennas, many of which will be described in later sections. It is useful where antennas with very large aspect ratios (ratio of dimensions of the aperture in two perpendicular directions) are desired. It is highly desirable where control in one direction is to be achieved through some special means, as in cosecant antennas, or in antennas which scan in one direction only.

Let us suppose that we have selected the aperture dimensions and have decided whether the reflector is to be paraboloidal or cylindrical. The reflector is not yet completely determined for we are still free in principle to use any portion of a parabolic surface of any focal length. In order to obtain economy in physical size the focal length is generally made between $0.6a$ and $0.25a$ where 'a' is the aperture. For the same reason a section of the reflecting surface which is located symmetrically about the vertex is often chosen (Figures 3 and 19).

When a symmetrically located section of the reflector is used certain difficulties are introduced. These difficulties, if serious enough so that their removal justifies some increase in size can be bypassed through the use of an

offset section as shown in Fig. 20. We can comment on these difficulties as follows:

1. The presence of the feed in the path of the reflected energy causes a region of low intensity or 'shadow' in the wave front. The effect of this shadow on the antenna pattern depends on the size and shape of the feed and on the characteristics of the portion of the wave front where it is located. Its effect is to subtract from the undisturbed pattern a 'shadow pattern' component which is broad in angle. This decreases the gain and increases the minor lobes as indicated in Fig. 21.⁷

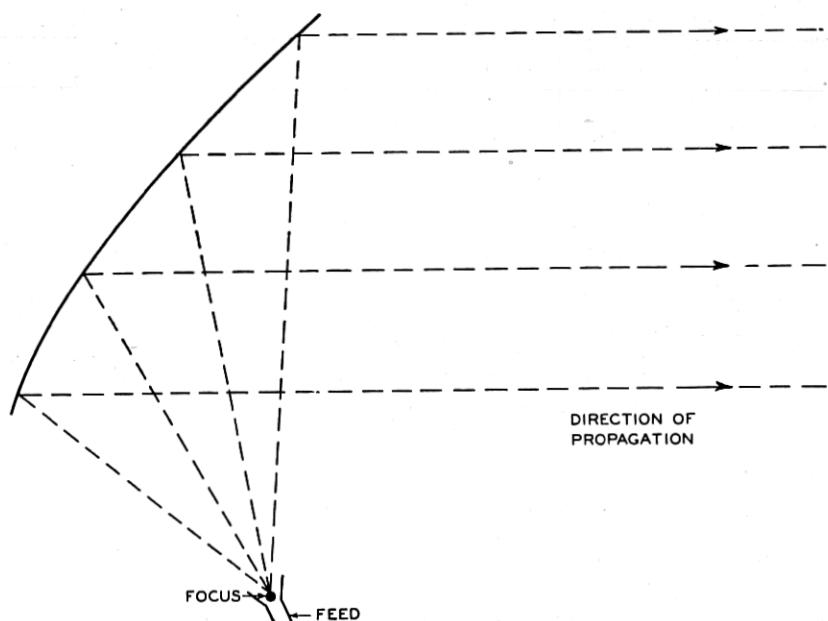


Fig. 20—Offset Parabolic Section.

2. Return of reflected energy into the feed introduces a standing wave of impedance mismatch in the feed line which is constant in amplitude but varies rapidly in phase as the frequency is varied. A mismatch at the feed which cancels the standing wave at one frequency will add to it at another frequency. A mismatch which will compensate over a band can be introduced by placing a raised plate of proper dimensions at the vertex of the reflector as indicated in Fig. 22, but such a plate produces a harmful effect on the pattern. In an antenna which must operate over a broad band it is consequently usually better to match

⁷ Figures 21, 22, and 23 are taken from C. C. Cutler, loc. cit.

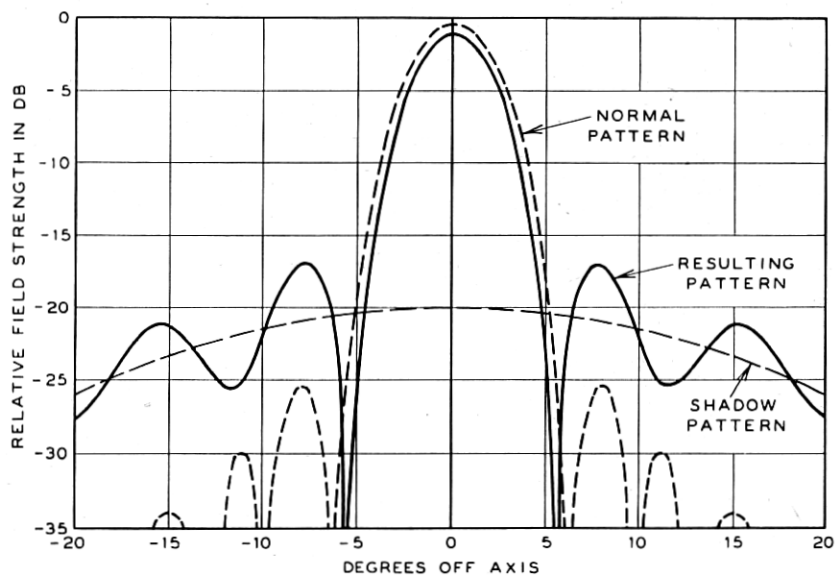


Fig. 21—Effect of Shadow on Paraboloid Radiation Pattern.

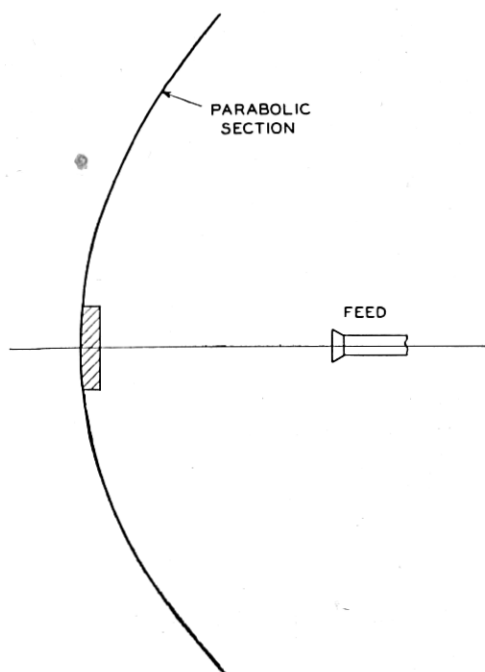


Fig. 22—Apex Matching Plate for Improving the Impedance Properties of a Parabola.

the feed to space and accept the residual standing wave, or if this is too great to use an offset section of the parabolic surface.

8.4 Feeds for Paraboloids

We have seen that an antenna with good wave front characteristics and consequently with a good beam and pattern can be constructed by illuminating a reflecting paraboloid with a properly designed feed placed at its focus. In this section we will examine the characteristics which the feed should have and some of the ways in which feeds are made in practice.

A feed for a paraboloid should

- a. be appropriate to the transmission line with which it is fed. This is sometimes a coaxial line but more commonly a waveguide.
- b. Provide an impedance match to this feed line. This match should usually be obtained in the absence of the reflector but sometimes, for narrow band antennas, with the reflector present.
- c. have a satisfactory phase characteristic. For a paraboloid the feed should be, as far as phase is concerned, a true point source radiating spherical waves. As discussed at the end of 8.1, if the wave front is not accurately spherical, a compensating correction in the reflector can be made.
- d. have a satisfactory amplitude characteristic. According to 8.2 this means that the feed should have a major radiation lobe with its maximum striking the center of the reflector, its intensity decreasing smoothly to a value about 8 to 10 *db* below the maximum in the direction of the reflector boundaries and remaining small for all directions which do not strike the reflector.
- e. have a polarization characteristic which is such that the electric vectors in the reflected wave front will all be polarized in the same direction.
- f. not disturb seriously the radiation characteristics of the antenna as a whole. The shadow effect of the feed, the feed line and the necessary mechanical supports must be small or absent. Primary radiation from the feed which does not strike the reflector or reflected energy which strikes the feed or associated structure and is then reradiated must be far enough down or so controlled that the antenna pattern is as required.

In addition to the electrical requirements for a paraboloid feed it must of course be so designed that all other engineering requirements are met, it must be firmly supported in the required position, must be connected to the antenna feed line in a satisfactory manner, must sometimes be furnished with an air tight or water tight seal, and so forth.

From the foregoing it is evident that a feed for a paraboloid is in itself a small relatively non-directive antenna. Its directivity is somewhat less than that obtained with an ordinary short wave array. It is therefore not surprising that dipole arrays are sometimes used in practice to feed paraboloids.

A simple dipole or half wave doublet can in itself be used to feed a paraboloid, but it is inefficient because of its inadequate directivity. It is preferable and more common to use an array in which only one doublet is excited directly and which contains a reflector system consisting of another doublet or a reflecting surface which is excited parasitically.

Dipole feeds although useful in practice have poor polarization characteristics and although natural when a coaxial antenna feed line is used are less convenient when the feed line is a waveguide. Since waveguides are more common in the microwave radar bands it is to be expected that waveguide feeds would be preferred in the majority of paraboloidal antennas.

The most easily constructed waveguide feed is simply an open ended waveguide. It is easy to permit a standard round or rectangular waveguide transmitting the dominant mode to radiate out into space toward the paraboloid. It will do this naturally with desirable phase, polarization and amplitude characteristics. It is purely coincidental, however, when this results in optimum amplitude characteristics. It is usually necessary to obtain these by tapering the feed line to form a waveguide aperture of the required size and shape. The aperture required may be smaller than a standard waveguide cross section so that its directivity will be less. In this case it may be necessary to 'load' it with dielectric material so that the power can be transmitted. It may be greater, in which case it is sometimes called an 'electromagnetic horn'. It may be greater in one dimension and less in the other, as when a paraboloidal section of large aspect ratio is to be illuminated.

If a single open ended waveguide or electromagnetic horn is used to feed a section of the paraboloid which includes the vertex, the waveguide feed line must partially block the reflected wave in order to be connected to the feed. To avoid this difficulty several rear waveguide feeds have been used. In this type of feed the waveguide passes through the vertex of the paraboloid and serves to support the feed at the focus. The energy can be caused to radiate back towards the reflector in any one of several ways, some of which involve reflecting rings or plates or parasitically excited doublets. The 'Cutler' feed⁸ is perhaps the most successful and common rear feed. It operates by radiating the energy back towards the paraboloid through two apertures located and excited as shown in Fig. 23.

⁸ C. C. Cutler, Loc. Cit.

8.5 Parabolic Cylinders between Parallel Plates

In 8.0 we saw that parabolic cylinders may be illuminated by line sources or that they may be confined between parallel plates and illuminated by point sources to produce line sources. In either of these two cases the characteristics which the feed should have are specified accurately by the conditions stated at the beginning of 8.4 for paraboloidal feeds with the exceptions that condition *c* must be reworded so that it applies to cylindrical rather than to spherical optics.

We will first consider parabolic cylinders bounded by parallel plates because in doing so we describe in passing one form of feed for unbounded parabolic cylinders. Two forms of transmission between parallel plates are used in practice.

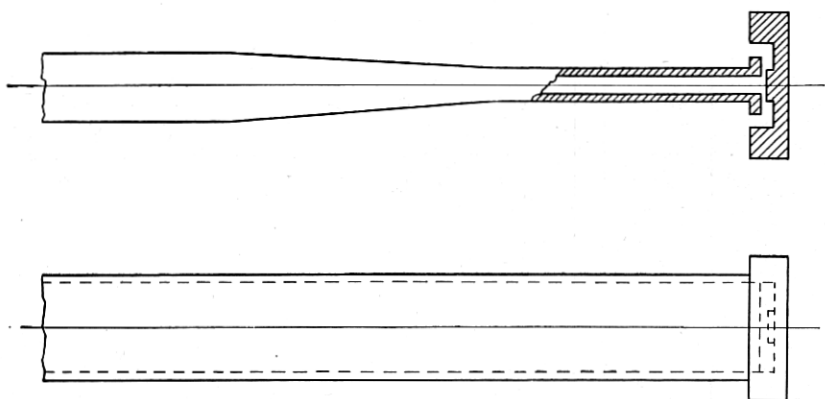


Fig. 23—Dual Aperture Rear Feed Horn.

- a. The transverse electromagnetic (TEM) mode in which the electric vector is perpendicular to the plates. This is simply a slice of the familiar free space wave and can be propagated regardless of the spacing between the plates. It is the only mode that can travel between the plates if they are separated less than half a wavelength. Its velocity of propagation is independent of plate spacing.
- b. The TE_{01} mode in which the electric vector is parallel to the plates. This mode is similar to the dominant mode in a rectangular waveguide and differs from it only in that it is not bounded by planes perpendicular to the electric vector. It can be transmitted only if the plate spacing is greater than half a wavelength, is the *only* parallel mode that can exist if the spacing is under a wavelength and is the only symmetrical parallel mode that can exist if the plate spacing is under three

halves of a wavelength. Its phase velocity is determined by the plate spacing in a manner given by the familiar waveguide formula

$$V_p = \frac{c}{\sqrt{\epsilon - \left(\frac{\lambda}{2a}\right)^2}}$$

where 'c' is the velocity of light, ϵ is the dielectric constant relative to free space of the medium between the plates, λ is the wavelength in air and 'a' is the plate spacing.

The TEM mode between parallel plates can be generated by extending the central conductor of a coaxial perpendicularly into or through the wave space and backing it up with a reflecting cylinder as indicated in Fig. 24.

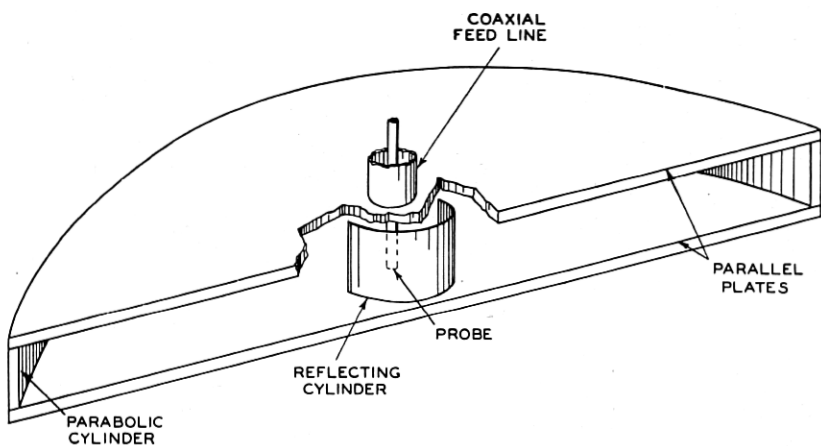


Fig. 24—Parabolic Cylinder Bounded by Parallel Plates. Probe Feed.

Alternatively this mode can be generated as indicated by Fig. 25 by a waveguide aperture with the proper polarization.

The TE_{01} mode, when used, is usually generated by a rectangular waveguide aperture set between the plates with proper polarization as indicated in Fig. 25. Care must be taken that only the desired mode is produced. The TEM mode will be unexcited if only the desired polarization is present in the feed. The next parallel mode is unsymmetrical and therefore even if it can be transmitted will be unexcited if the feed is placed symmetrically with respect to the two plates.

Parallel plate antennas as shown in figures 24 and 25 are useful where particularly large aspect ratios are required. The aperture dimension perpendicular to the plates is equal to the plate spacing and therefore small.

It can be increased somewhat by the addition of flares. The other dimension can easily be made large.

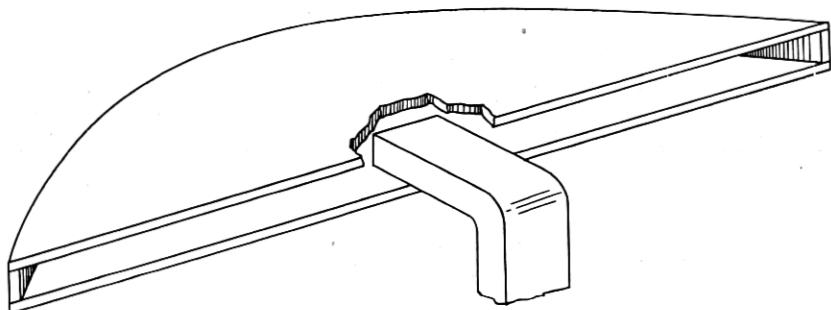


Fig. 25—Parabolic Cylinder Bounded by Parallel Plates. Wave Guide Feed.

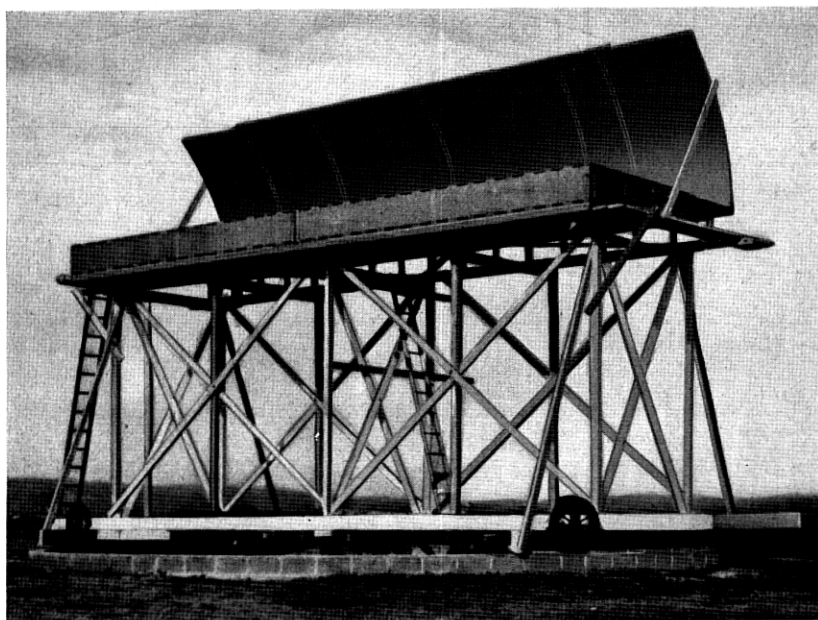


Fig. 26.—Experimental 7' x 32' Antenna.

8.6 Line Sources for Parabolic Cylinders

A line source for a parabolic cylinder is physically an antenna with a long narrow aperture. Any means for obtaining such an aperture can be used in producing a line source. Parallel plate systems as described in 8.5 have been used as line sources in several radar antennas. The large (7' x 32')

experimental antenna shown in Fig. 26 was one of the first to illustrate the practicality of this design.

The horizontal pattern of the 7' x 32' antenna is plotted in Fig. 27. The horizontal beam width is seen to be of the order of 0.7 degrees.

The antenna illustrated in Fig. 26 is interesting in another way for it is a good example of a type of experimental construction which was extremely useful in wartime antenna development. Research and development engi-

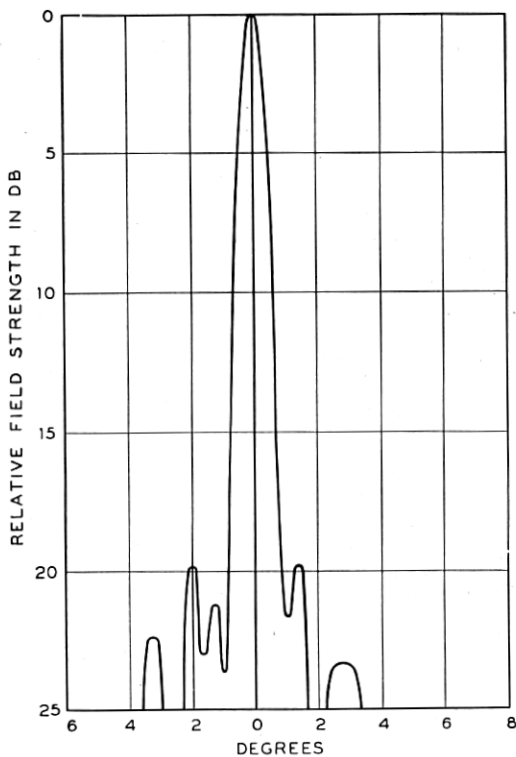


Fig. 27—7' x 32' Antenna, Horizontal Pattern.

neers found that they could often save months by constructing initial models of wood. Upon completion of a wooden model electrically important surfaces were covered with metal foil or were sprayed or painted with metal. Thus, where tolerances permitted, the carpenter shop could replace the relatively slow machine shop.

Another form of parallel plate line feed results when a plastic lens is placed between parallel plates and used as the focussing element. A linear array

of elements excited with the proper phase and amplitude can also be used. Some discussion of alternative approaches will appear in the section on scanning techniques.

8.7 Tolerances in Parabolic Antennas

The question of tolerances will always arise in practice. Ideal dimensions are only approximated, never reached. The ease of obtaining the required accuracy is an important engineering factor.

The tolerances in paraboloidal antennas or in parabolic cylinders illuminated by line sources can be divided into three general classes:

- (a) Tolerances on reflecting surfaces.
- (b) Tolerances on spacial relationships of feed and reflector.
- (c) Tolerances on the feed.

When the actual reflector departs from the ideal parabolic curve deviations in the phase will result. These will tend to reduce the gain and increase the minor lobes. The effects of such deviations on the gain can be estimated with the help of Fig. 17. We should recall that an error of σ in the reflector surface will produce an error of about 2σ in the phase front. Based on this kind of argument and on experience reflector tolerances are generally set in

practice to about $\pm \frac{\lambda}{16}$ or $\pm \frac{\lambda}{32}$ depending on the amount of beam deterioration that can be permitted.

In Fig. 28 are compared some electrical characteristics of two paraboloidal antennas, one employing a precisely constructed paraboloidal searchlight mirror and the other a carefully constructed wooden paraboloidal reflector with the same nominal contour. This comparison is revealing for it shows the harm that can be done even by small defects in the reflector surface. Although the two patterns are almost identical in the vicinity of the main beam, the general minor lobe level of the wooden reflector remains higher at large angles and its gain is less.

It must not, however, be assumed that a solid reflecting surface is necessary to insure excellent results. Any reflecting surface which reflects all or most of the power is satisfactory provided that it is properly located. Perforated reflectors, reflectors of woven material and reflectors consisting of gratings with less than half wavelength spacing are commonly used in radar antenna practice. These reflectors tend to reduce weight, wind or water resistance and visibility. Many of them will be described in Part III of this paper.

The feed of a parabolic reflector should be located so that its phase center coincides with the focus of the reflector. If it is located at an incorrect dis-

tance from the vertex a circular curvature of phase results and the system is said to be 'defocussed' (Sec. 3.9). As the feed is moved off the axis of the reflector the first effect is a shifting of the beam due to a linear variation of the phase (Sec. 3.8). For greater distances off axis a cubic component of phase error becomes effective (Sec. 3.10). Phase error, whether circular, cubic or more complex, results in a reduction in gain and usually in an increase of minor lobes. Although the effects of given amounts of phase curvature on the radiation characteristics of an antenna can be estimated by theoretical means, it is usually easier and quicker to find them experimentally.

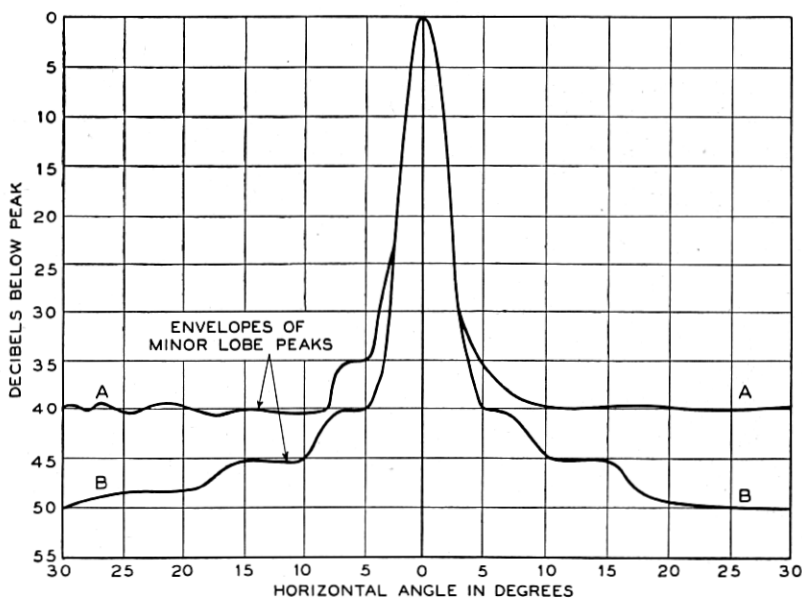


Fig. 28—Effect of Small Inaccuracies in Reflector.

The tolerances on the feed itself appear in various forms, many of which can be examined with the aid of transmission line theory and most of which are too detailed for discussion in this paper. It is generally true here also that experiment is a more effective guide than theory.

Experience has shown that when parallel plate systems are used, either as complete antennas or as line feeds for other elements, tolerances on the parallel conducting plates must be considered carefully. It is obvious that when the TE_{01} mode is used the plate spacing must be held closely, since the phase velocity is related to the spacing. This spacing can be controlled through the use of metallic spacers perpendicular to the plates. These

spacers, if small enough in cross section, do not disturb things unduly. The velocity of the TEM mode is, on the other hand, almost independent of the plate spacing. This mode is, however, more likely to cause trouble by leaks through joints and cracks in the plates.

9. METAL PLATE LENSES

At visible wavelengths lenses have, in the past, been far more common than in the microwave region, due chiefly to the absence of satisfactory lens materials. A solid lens of glass or plastic with a diameter of several feet is a massive and unwieldy object. By zoning, which will be discussed below, these difficulties can be reduced but they still remain.

A new lens technique, particularly effective in the microwave region was developed by the Bell Laboratories during the war.⁹ It is evident that any material in which the phase velocity is different from that of free space can be used to make a phase correcting lens. The material which is used in this new technique is essentially a stack of equally spaced metal plates parallel to the electric vector of the wave front and to the direction of propagation. Lenses made from this material are called 'Metal Plate Lenses'.

When the spacing between neighboring plates is between $\lambda/2$ and λ only one mode with electric vector parallel to the plates can be transmitted. This is the TE_{01} mode for which the phase velocity is given in Sec. 8.5. When the medium between the plates is air this equation can be converted into the expression

$$N = \sqrt{1 - \left(\frac{\lambda}{2a}\right)^2}$$

for the effective index of refraction. Here λ is the wavelength in air and a is the plate spacing.

As a varies between $\lambda/2$ and λ , N varies as indicated in Fig. 29. In the neighborhood of $a = \lambda$, N is not far from 1 and as a approaches $\lambda/2$, N approaches 0. Since N is always less than 1 we see that there is an essential difference between metal plate lenses and glass or plastic lenses for which N is always greater than 1. This difference is seen in the fact that a glass lens corrects phases by slowing down a travelling wave front, while a metal lens operates in the reverse direction by speeding it up. This means that a convergent lens with a real focus must be thinner in the center than the edge, the opposite of a convergent optical lens (Fig. 30).

Unless the value of N is considerably different from 1 it is evident that very thick lens sections must be used to produce useful phase corrections. For this reason values of ' a ' not far from $\lambda/2$ should be chosen. On the other hand values of ' a ' too close to $\lambda/2$ would cause undesirably large reflections

⁹ W. E. Kock, "Metal Lens Antennas", Proc. I. R. E., Nov., 1946.

from the lens surfaces and impose severe restrictions on the accuracy of plate spacings. The compromises that have been used in practice are $N = 0.5$ for which $a = 0.577\lambda$ and $N = 0.6$ for which $a = 0.625\lambda$.

Even with $N = 0.5$ or 0.6 lenses become thick unless inconveniently long focal distances are used. Thick lenses are undesirable not only because they occupy more space and are heavier but also because the plate spacing must be held to a higher degree of accuracy if the phase correction is to be as

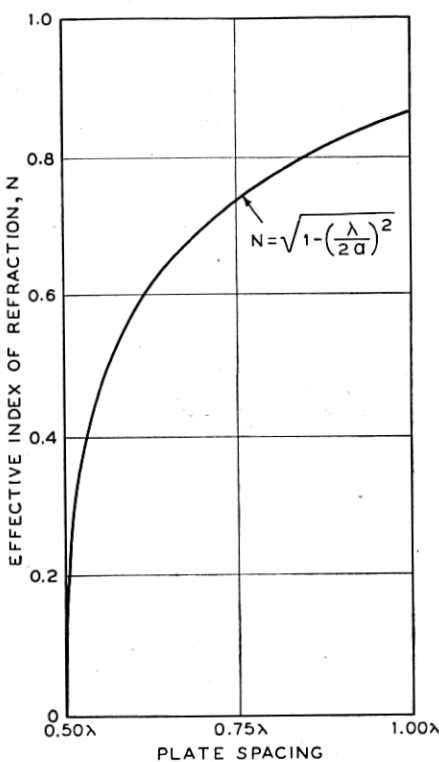


Fig. 29—Variation of Effective Index of Refraction with Plate Spacing in a Metal Plate Lens.

required. To get around these difficulties the technique of zoning is used.

Zoning makes use of the fact that if the phase of an electromagnetic vector is increased or decreased by any number of complete cycles the effect of the vector is unchanged. When applied to a metal plate lens antenna this means simply that wherever the phase correction due to a portion of the lens is greater than a wavelength this correction can be reduced by some integral number of wavelengths such that the residual phase correction is under one wavelength. If this is done it is evident that no portion of the

lens needs to correct the phase by more than one wavelength. It follows that no portion of the lens need to be thicker than $\lambda/(1 - N)$.

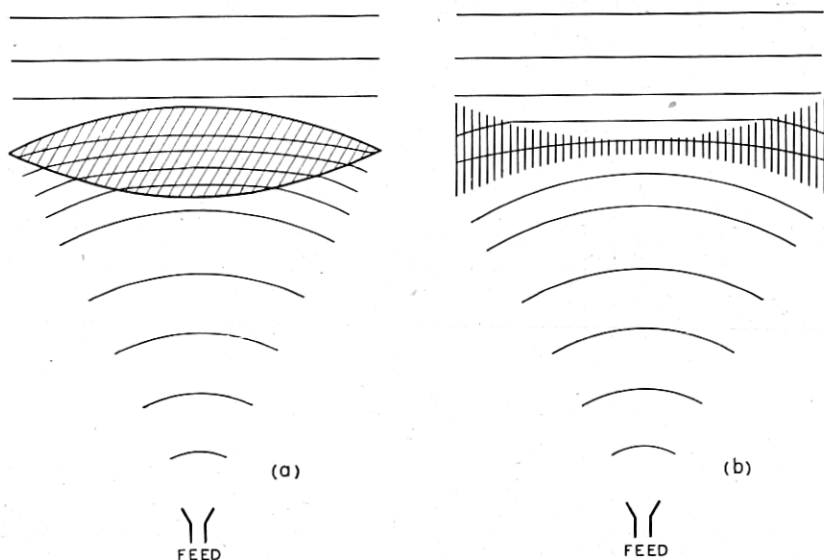


Fig. 30—Comparison of Dielectric and Metal Plate Lenses.

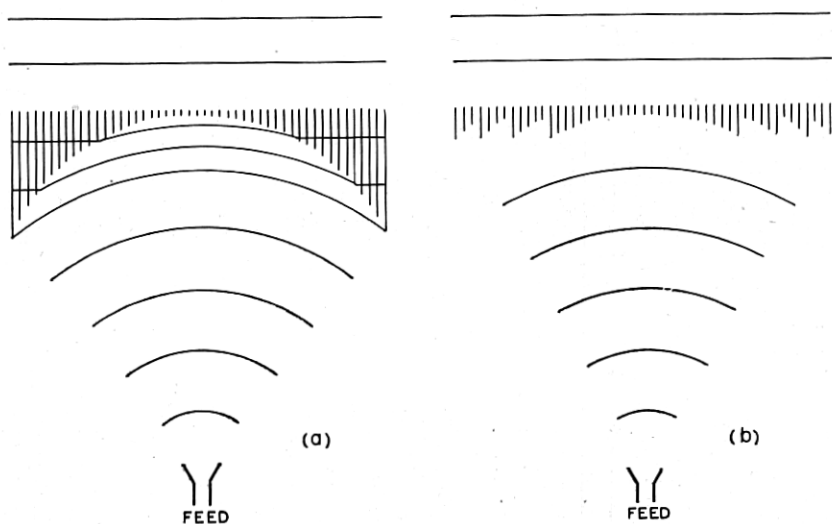


Fig. 31—Comparison of Unzoned and Zoned Metal Plate Lenses.

A cross section of a typical metal plate lens before and after zoning is illustrated in Fig. 31. This figure shows clearly why zoning reduces considerably the size and mass of a lens.

Zoning is not without disadvantages. One disadvantage is obviously that a zoned lens which is designed for one frequency will not necessarily work well at other frequencies. It is in principle possible to design a broad band zoned metal plate lens corresponding to the color compensated lenses used in good cameras. So far, however, this has not been necessary since band characteristics of simple lenses have been adequate.

Another difficulty that zoning introduces is due to the boundary regions between the zones. The wave front in this region is influenced partly by one zone and partly by the other and may as a result have undesirable phase and amplitude characteristics. This becomes serious only if especially short focal distances are used.

9.1 *Lens Antenna Configurations*

Any of the configurations which are possible with parabolic reflectors have their analogues when metal plate lenses are used. Circular lenses illuminated by point sources and cylindrical lenses illuminated by line sources are not only theoretically possible but have been built and used. Since a lens has two surfaces there is actually somewhat more freedom in lens design than in reflector design. Metal Plate Lenses have usually been designed with one surface flat, but the possibility of controlling both surfaces is emerging as a useful design factor where special requirements must be met.

Feeds for lenses should fulfill most of the same requirements as feeds for reflectors. We find a difference in size in lens feeds in that they must generally be more directive because of greater ratios of focal length to aperture. A difference in kind occurs because the feed is located behind the lens where none of the focussed energy can enter the feed or be disturbed by it. As a result some matching and pattern problems which arise in parabolic antennas are automatically absent when lenses are used.

In choosing a design for a lens antenna system with a given aperture one must compromise between the large size which is necessary when a long focal length is used and the more zones which result if the focal length is made short. Most metal plate lenses so far constructed have had focal lengths somewhere between 0.5 and 1.0 times the greatest aperture dimension.

9.2 *Tolerances in Metal Plate Lenses*

It is not difficult to see that phase errors resulting from small displacements or distortions of a metal plate lens are much less serious than those due to comparable distortions of a reflector surface. This follows from the fact that the lens operates on a wave which passes through it. If a portion of the lens is displaced slightly in the direction of propagation it is still operating on roughly the same portion of the wave front and gives it the same phase correction. If a portion of a reflector were displaced in the same way the error in the wave front would be of the order of twice the

displacement. Quantitative arguments show that less severe tolerances apply to all major structural dimensions of a metal lens antenna.

It is true of course that the dimensions of individual portions of the metal lens must be held with some accuracy. The metal plate spacing determines the effective index of refraction of the lens material. Where $N = 0.5$ it is customary to require that this be held to $\pm\lambda/75$, and where $N = 0.6$ to $\pm\lambda/50$. The thickness of the lens in a given region is less critical, and must be held to $\pm \frac{\lambda}{16(1-N)}$ where it is desired to hold the phase front to $\pm\lambda/16$.

Fig. 32 illustrates clearly the drastic way in which the location of a lens can be altered without seriously affecting the pattern. It shows, incidentally, how a lens may behave well when used as the focussing element in a moving feed scanning antenna.

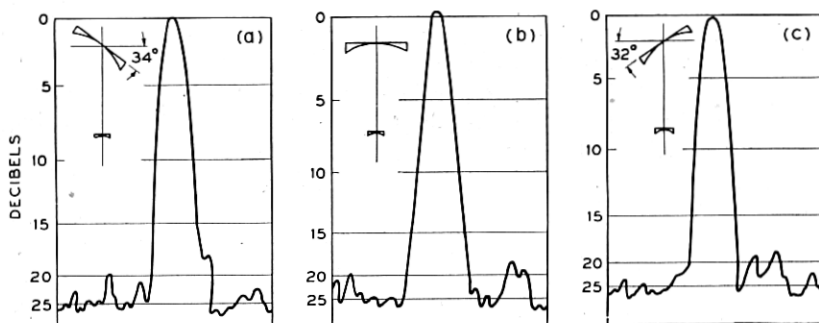


Fig. 32—Effect on Pattern of Lens Tilting.

9.3 Advantages of Metal Plate Lenses

On the basis of the above discussion we can see that metal plate lenses have certain considerable advantages. The most important of these is perhaps found in the practical matter of tolerances. It is a comparatively simple matter to hold dimensions of small objects to close tolerances but quite another thing to hold dimensions of large objects closely under the conditions of modern warfare. This advantage emerges with increasing importance as the wavelength is reduced.

Metal plate lenses have contributed a great degree of flexibility to radar antenna art. When they are used two surfaces rather than one may be controlled, and the dielectric constant can be varied within wide limits. Independent control in the two polarizations may be applied. We can confidently expect that they will become increasingly popular in the radar field.

10. COSECANT ANTENNAS

One of the earliest uses of radar was for early warning against aircraft.

The skies were searched for possible attackers with antennas which rotated continuously in azimuth. An equally important but later use appeared with the advent of great bombing attacks. Bombing radars 'painted' maps of the ground which permitted navigation and bombing during night and under even the worst weather conditions. In these radars also the antennas were rotated in azimuth, either continuously through 360° or back and forth through sectors.

The majority of radars designed to perform these functions provided vertical coverage by means of a special vertical pattern rather than a vertical scan. It can easily be seen that such a pattern would have to be 'special.' If we assume, for example, that a bombing plane is flying at an altitude of 10,000 feet, then the radar range must be $10,000 \csc 60^\circ = 11,500$ feet if a target on the ground at a bomb release angle of 60° from the horizontal is to be seen. Such a range would by no means be enough to pick up the target at say 10° in time to prepare for bombing, for then a range of $10,000 \csc 10^\circ = 57,600$ feet would be required. This range is far more than is necessary for the 60° angle. It appears then that in the most efficient design the radar range and therefore the radar antenna gain, must be different in different directions.

The required variation of gain with vertical direction could be specified in any one of several ways. It seems natural to specify that a given ground target should produce a constant signal as the plane flies towards it at a constant altitude. Neglecting the directivity of the target this will occur if the amplitude response of the antenna is given by $E = E_0 \csc \theta$. This same condition will apply by reciprocity to an early warning radar antenna on the ground which is required to obtain the same response at all ranges from a plane which is flying in at a constant altitude.

This condition is not alone sufficient to specify completely the vertical pattern of an antenna. For one thing it can obviously not be followed when $\theta = 0$, for this would require infinite gain in this direction. Therefore a lower limit to the value of θ for which the condition is valid must be set. In addition an upper limit less than 90° is specified whenever requirements permit, since control at high angles is especially difficult. When the limits have been set it still remains to specify the magnitude of the constant E_0 . This can be done by specifying the range in one particular direction. This specification must of course be consistent with all the factors that determine gain, including the reduction due to the required vertical spread of the pattern.

10.1 Cosecant Antennas based on the Paraboloid

It is evident that the standard paraboloidal antennas so far discussed will not produce cosecant patterns. These patterns being unsymmetrical will result only if the wave front phase and amplitude are especially controlled.

On the other hand, because paraboloidal antennas are simple and common it is natural that many cosecant designs should be based on them. These designs can be classified into two groups, those in which the reflector is modified and those in which the feed is modified.

Some early cosecant antennas were made by introducing discontinuities in paraboloidal reflectors as illustrated in Fig. 33. These controlled the radiation more or less as desired over the desired cosecant pattern but pro-

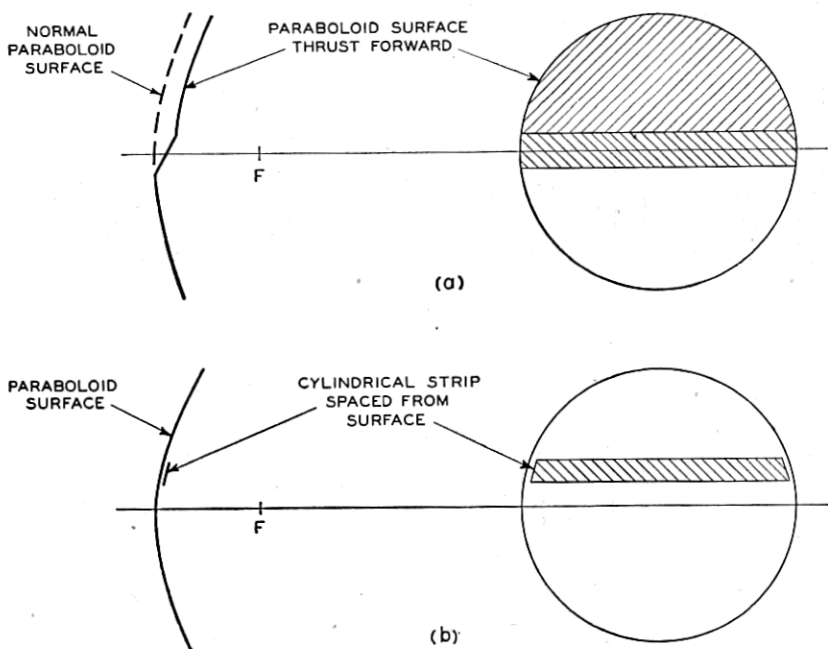


Fig. 33—Some Cosecant Antennas Based on the Paraboloid (Cosecant Energy Downward).

duced rather serious minor lobes elsewhere. This difficulty can be overcome through the use of a continuously distorted surface as illustrated in Fig. 34. This reflector, first used at the Radiation Laboratories, is a normal paraboloid in the lower part whereas the upper part is the surface that would be obtained by rotating the parabola through the vertex of the upper part about its focal point.

Several types of feed have been used in combination with paraboloids to produce cosecant patterns. These are usually arrays which operate on the principle that each element is a feed which contributes principally to one

region of the vertical pattern. The elements may be dipoles or waveguide apertures fed directly through the antenna feed line or they may be reflectors which reradiate reflected energy originating from a single primary source. No matter how excited they must be properly controlled in phase, amplitude and directivity.

Cosecant antennas based on the paraboloid are common and can sometimes fulfill all requirements with complete satisfaction. Nevertheless they

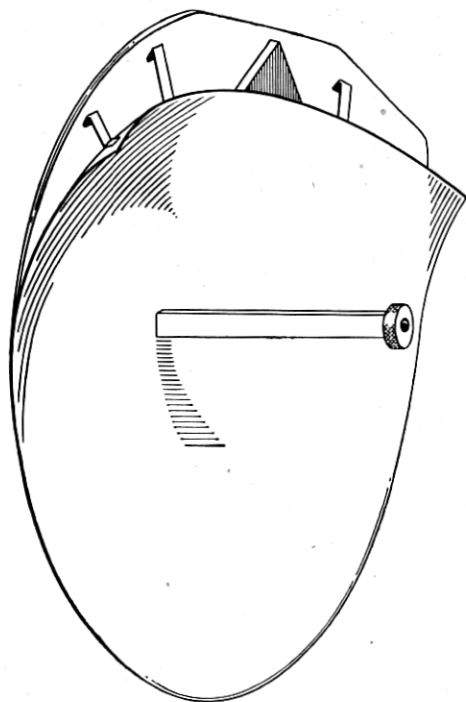


Fig. 34—Barrel Cosecant Antenna (Cosecant Energy Downward).

suffer from certain disadvantages. The most serious of these is that they lack resolution at high vertical angles, that is the beam is wider horizontally at high angles. This is to be expected for reasons of phase alone, for a paraboloidal reflector is, after all, designed to focus in only one direction. If phase difficulties were completely absent however, azimuthal resolution at high angles would still be destroyed because of cross polarized components of radiation. These components arise naturally from doubly curved reflectors, even simple paraboloids. They are sometimes overlooked when antennas are measured in a one way circuit with a linearly polarized test field, but must obviously be considered in radar antennas.

10.2 Cylindrical Cosecant Antennas

Harmful cross polarized radiation is produced by doubly curved reflectors. This radiation is difficult to control and therefore undesirable where a closely controlled cosecant characteristic at high angles is required. Although not at first evident, it seems natural now to bypass polarization difficulties through the use of singly curved cylindrical reflectors. These reflectors if illuminated with a line source of closely controlled linear polarization provide a beam which is linearly polarized. This beam has also in azimuth approximately the directivity of the line source at all vertical angles. It is thus superior in two significant respects to cosecant beams produced by doubly curved reflecting surfaces.

A cylindrical cosecant antenna consists of a cylindrical reflector illuminated by a line source. Part of the cylinder is almost parabolic and contributes chiefly to the strong part of the beam which lies closest to the horizontal. This part is merged continuously into a region which departs considerably from the parabolic and contributes chiefly to the radiation at higher angles.

Although wave front principles can be used and certainly must not be violated, the principles of geometrical optics have been particularly effective in the determination of cosecant reflector shapes. The detailed application of these principles will not be discussed here. While applying the geometrical principles the designer must be sure that the over-all size and configuration of the antenna can produce the results he wants. He must design a line source with the desired polarization and horizontal pattern and a vertical pattern which fits in with the cosecant design. In addition he must take particular care to reduce sources of pattern distortion to a level at which they cannot interfere significantly with the lowest level of the cosecant 'tail'.

11. LOBING

In many of the tactical situations of modern war radar can be used to provide fire control information. Radar by its nature determines range and microwave radar with its narrow well defined beams is a natural instrument for finding directions to a target, whether the missile to be sent to that target is a shell, a torpedo or a bomb. In fire control radar, as opposed to search or navigational radar, two properties of the antenna deserve particular attention. These are the *accuracy* and the *rate* with which direction to a target can be measured.

Lobing is a means which utilizes to the fullest extent the accuracy available from a given antenna aperture and which increases, usually as far as is desired, the rate at which this information is provided and corrected.

A lobing antenna which is to provide information concerning one angle only, azimuth for example, is capable of producing two beams, one at a time, and of switching rapidly from one to the other. This process is called *Lobe Switching*. The two beams are nearly coincident, differing in direction by about one beam width. When the signals from the two beams are compared, they will be equal only if the target lies on the bisector between the beams (Fig. 35). The two signals can be compared visually on an indicator screen of the radar or they can be compared electrically and fed directly into circuits which control the direction of fire.

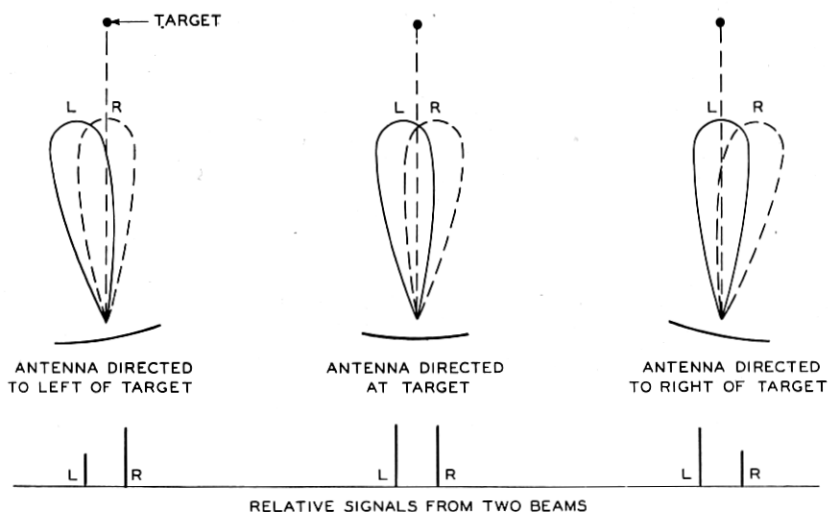


Fig. 35—Lobe Switching.

When two perpendicular directions are to be determined, such as the elevation and azimuth required by an anti-aircraft battery, four or in principle three discrete beams can be used. Radar antennas designed for solid angle coverage more commonly, however, produce a single beam which rotates rapidly and continuously around a small cone. This rotation is known as *conical lobing*. A comparison of amplitudes in a vertical plane can then be used to give the elevation of the target and a similar comparison in a horizontal plane to give its azimuth. Here too the electrical signals can be compared visually on an indicator screen, but an electrical comparison will provide continuous data which can be used to aim the guns and at the same time to cause the radar antenna to follow the target automatically.

11.1 Lobe Switching

Two methods of lobe switching are common. In one of these the lobing antenna is an array of two equally excited elements. Each of these ele-

ments occupies one half of the final antenna aperture, and provides a uniphase front across this half. If the two elements were excited with the same phase the radiation maximum of the resulting antenna beam would occur in a direction at right angles to the combined phase front. If the phase of one element is made to lag behind that of the other by a small amount, 60° say, the phase of the combined aperture will of course be discontinuous with a step in the middle. This discontinuous phase front will approximate with a small error, a uniphase wave front which is tilted somewhat with respect to the wave fronts of the elements. The phase shift will therefore result in a slight shift of the beam away from the normal direction. When the phase shift is reversed the beam shift will be reversed. Two properly designed elementary antennas in combination with a means for rapidly changing the phase will therefore constitute a lobe switching antenna. Such an antenna is described more in detail in Sec. 14.6.

Another method of lobe switching is more natural for antennas based on optical principles. In this method two identical feeds are placed side by side in the focal region of the reflector. When one of these feeds illuminates the reflector a beam is produced which is slightly off the normal axial direction. Illumination by the other feed produces a second beam which is equally displaced in the opposite direction. The lobe of the antenna switches rapidly when the two feeds are activated alternately in rapid succession. The antenna must use some form of rapid switching appropriate to the antenna feed line. In several applications switches are used which depend on the rapid tuning and detuning of resonant cavities or irises.

11.2 Conical Lobing

A conically lobing antenna produces a beam which nutates rapidly about a fixed axial direction. This is usually accomplished by rotating or nutating an antenna feed in a small circle about the focus in the focal plane of a paraboloid or lens. This antenna feed can be a spinning asymmetrical dipole or a rotating or nutating waveguide aperture. It can result in a beam with linear polarization which rotates as the feed rotates, or preferably in a beam for which the polarization remains parallel to a fixed direction. The beam itself must be nearly circularly symmetric so that the radar response from a target in the axial direction will not vary with the lobing. The reflector or lens aperture is consequently usually circular.

When the antenna is small it is sometimes easier to leave the feed fixed and to produce the lobing by moving the reflector.

12. RAPID SCANNING

A lobing radar can provide range and angular information concerning a single target rapidly and accurately but these things are not always enough.

It is sometimes necessary to obtain accurate and rapid information from all regions within an angular sector. It may be necessary to watch a certain region of space almost continuously in order to be sure of picking up fast moving targets such as planes. To accomplish any of these ends we must use a *rapid scanning* radar. A rapid scanning radar antenna produces a beam which scans continuously through an angular sector. The beam may sweep in azimuth or elevation alone or it may sweep in both directions to cover a solid angle. An azimuth or elevation scan may be sinusoidal or it may occur linearly and repeat in a sawtooth fashion. Solid angle scanning may follow a spiral or flower leaf pattern or it might be a combination of two one way scans. A combination of scanning in one direction and lobing in the other is sometimes used.

Scanning antennas must, unfortunately, be constructed in obedience to the same principles which regulate ordinary antennas. The same attention to phase, amplitude, polarization and losses is necessary if comparable results are to be obtained. When scanning requirements are added to these ordinary ones new problems are created and old ones made more difficult.

An antenna in order to scan in any specified manner must act to produce a wave front which has a constant phase in a plane which is always normal to the required beam direction. This can be done in several different ways. The simplest of these, electrically, is to rotate a fixed beam antenna as a whole in the required fashion. This can be called *mechanical scanning*. Alternatively an antenna array can be scanned if it is made up of suitable elements and the relative phases of these elements can be varied appropriately. This can be called *array scanning*. Thirdly, *optical scanning* can be produced by moving either the feed or the focussing element of a suitably designed optical antenna.

12.1 Mechanical Scanning

Electrical complexities of other types of rapid scanners are such that it is probably not going too far to say that the required scan should be accomplished by mechanical means wherever it is at all practical. This applies to radar antenna scans which occur at a slow or medium rate. Search antennas, whether they rotate continuously through 360° or back and forth over a sector are scanners in a sense but the scan is usually slow enough to be performed by rotating the antenna structure as a whole. As the scan becomes more rapid, mechanical problems become more severe and electrically scanning antennas appear more attractive.

Mechanical ingenuity has during the war extended the range in which mechanical scanners are used. One important and eminently practical mechanical rapid scanner, the 'rocking horse' is now in common use (Fig. 36). This antenna is electrically a paraboloid of elliptical aperture illu-

minated by a horn feed, a combination which produces excellent electrical characteristics. The paraboloid and feed combination is made structurally strong and is pivoted to permit rotational oscillation in a horizontal plane. It is forced to oscillate by a rigid crank rod which is in turn driven by an eccentric crank on a shaft. The shaft is belt driven by an electric motor and its rotational rate is held nearly constant by a flywheel. The mechanical arrangement described so far would oscillate rotationally in an approximately sinusoidal fashion. Since every action has an equal and opposite reaction it would, however, react by producing an oscillatory torque on its

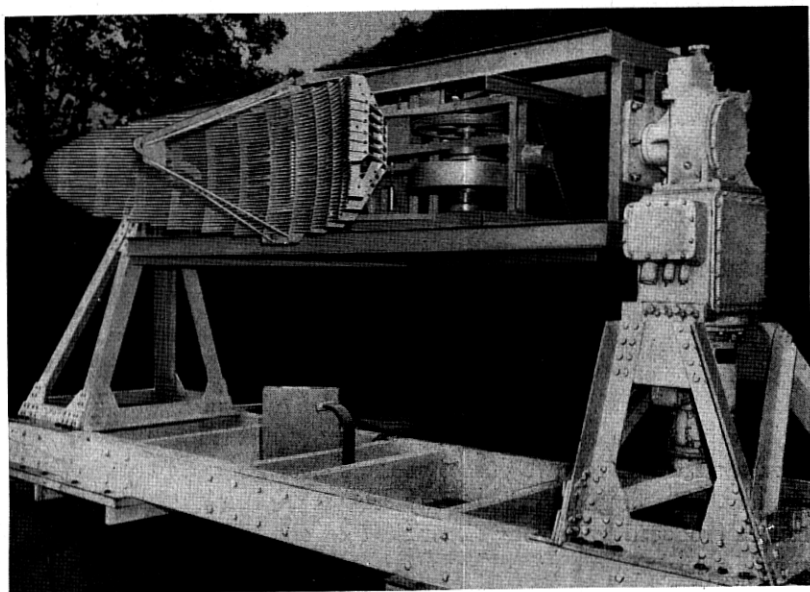


Fig. 36—Experimental Rocking Horse Antenna.

mounting. Since the antenna is large and the oscillation rapid this would produce a severe and undesirable vibration. To get around this difficulty an opposite and balancing rotating moment is introduced into the mechanical system. This appears in the form of a pivoted and weighted rod which is driven from the same eccentric crank by another and almost parallel crank arm.

Although not theoretically perfect the rotational 'dynamic' balancing described permits the antenna to scan without serious vibration. One form of this antenna will be described in a later section.

12.2 Array Scanning

During our discussion of general principles in Part II, we saw that an antenna wave front can be synthesized by assembling an array of radiating

elements and distributing power to it through an appropriate transmission line network. If the radiation characteristics of the array are to be as desired the electrical drive of each element must have a specified phase and amplitude. In addition each element must in itself have a satisfactory characteristic and the elements must have a proper spacial relationship to each other.

Such array antennas have been extremely useful in the 'short wave' bands where wavelengths and antenna sizes are many times larger than at most radar wavelengths but for fixed beam radar antennas they have been largely superseded by the simpler optical antennas. Where a rapidly scanning beam is desired, however, they possess certain advantages which were put to excellent use in the war. These advantages spring from the possibility of scanning the beam of an array through the introduction of rapidly varying phase changes in its transmission line distributing system.

Let us first examine certain basic conditions that must be fulfilled if an array antenna is to provide a satisfactory scan. The pattern of any array is merely the sum of the patterns of its elements taking due account of phase, amplitude and spacial relationships. If all elements are alike and are spaced equally along a straight line it is not difficult to show that a mathematical expression for the pattern can be obtained in the form of a product of a factor which gives the pattern of a single element and an array factor. The array factor is an expression for the pattern of an array of elements each of which radiates equally in all directions. Since each of the elements is fixed in direction it is only through control of the array factor that the scan can be obtained.

If we excite all points of a continuous aperture with equal phase and a smoothly tapered amplitude the aperture produces a beam with desirable characteristics at right angles to itself and no comparable radiation elsewhere. Similarly if we excite all elements of an array of identical equally spaced circularly radiating elements with equal phase and a smoothly tapered amplitude the array will produce a beam with desirable characteristics at right angles to itself. It will also produce a beam in any *other* direction for which waves from the elements can add up to produce a wave front. Such other directions will exist whenever the array spacing is greater than one wavelength.

In order to see this more clearly let us examine Fig. 37, where line XX' represents an array of elements. From each element to the line AA' is a constant distance, so AA' is obviously parallel to a wave front when the elements are excited with equal phase. If we can find a line BB' to which the distance from each element is exactly one wavelength more or less than from its immediate neighbors then it too is parallel to a wavefront, for energy reaching it from any element of the array will have the same phase

except for an integral number of cycles. The same will apply to a line CC' , to which the distance from each element is exactly two wave lengths more or less than from its immediate neighbors, or to any other line where this difference is any integral number of wavelengths.

Now in no radar antenna do we desire two or more beams for they will result in loss of gain and probably in target confusion. The array must therefore be designed so that for all positions of scan all beams except one will be suppressed. This will automatically occur if the array spacing is somewhat less than one wavelength. If the array spacing is greater than one wavelength these extra beams will appear in the array factor; they

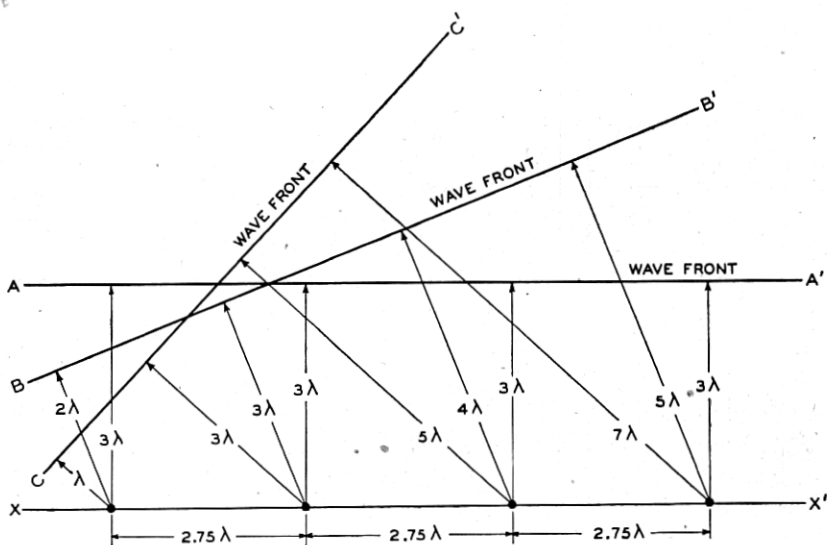


Fig. 37—Some Possible Wave Fronts of an Array of Elements Spaced 2.75λ .

must therefore be suppressed by the pattern of a single element. The pattern of an element must in other words, have no significant components in any direction where an extra beam can occur.

Where elements with only side fire directivity are spaced more than a wavelength apart in a scanning array it is almost impossible to obtain adequate extra lobe suppression. If these elements are spaced by the minimum amount, that is by exactly the dimensions of their apertures and all radiate in phase they may indeed just manage to produce a desirable beam. A little analysis shows however that an appreciable phase variation from element to element, even though linear, will introduce a serious extra lobe. To get around this difficulty elements with some end fire directivity must be used.

A simple end fire element, and one that has been used in practice, is the 'polyrod' (Fig. 38). A polyrod, is as its name implies, a rod of polystyrene. This rod, if inserted into the open end of a waveguide, and if properly proportioned and tapered, will radiate energy entering from the waveguide from points which are distributed continuously along its length. If the

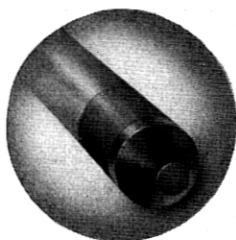


Fig. 38—A Polyrod.

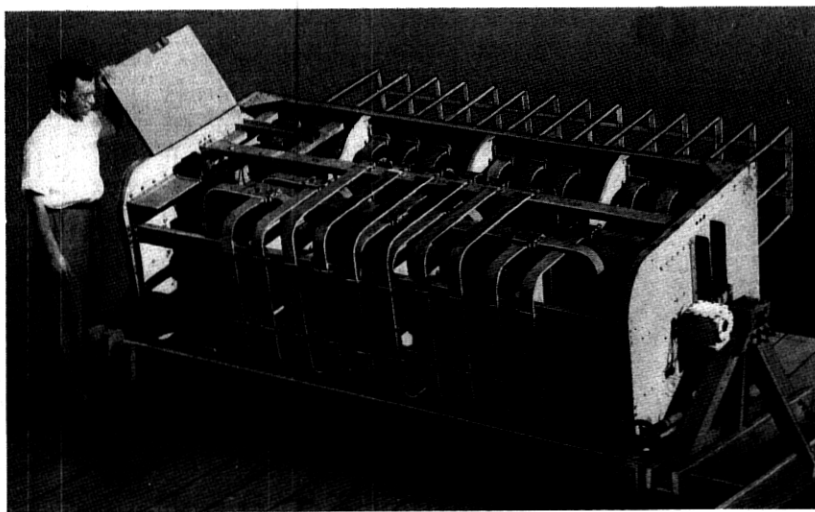


Fig. 39.—Experimental Polyrod Array.

wave in the polyrod travels approximately with free space velocity it will produce a radiation maximum in the direction of its axis. The radiation pattern of the polyrod will have a shape which is characteristic of end fire arrays, narrower and flatter topped than the pattern of a side fire array which occupies the same *lateral* dimension. This elementary pattern can be fitted in well with the array factor of a scanning array.

Such a scanning array is shown in Fig. 39 and will be described in

greater detail in section 14.8. Each element of this array consists of a fixed vertical array of three polyrods. This elementary array provides the required vertical pattern and has appropriate horizontal characteristics. Fourteen of these elements are arranged in a horizontal array with a spacing between neighbors of about two wavelengths. Energy is distributed among the elements with a system of branching waveguides. Thirteen rotary phase changers are inserted strategically in the distributing system. Each phase change is rotating continuously and shifts the phase linearly from 0° to 360° twice for each revolution. As the phase changers rotate the array produces a beam which sweeps repeatedly linearly and continuously across the scanning sector.

When elements of a scanning array are spaced considerably less than one wavelength it is a very simple matter to obtain a suitable elementary pattern, for the array factor itself has only a single beam. This advantage is offset by the greater number of elements and the consequent greater complexity of distributing and phase shifting equipment. In one useful type of scanning antenna however distributing and phase shifting is accomplished in a particularly simple manner. Here the distributing system is merely a waveguide which can transmit only the dominant mode. The wide dimension of the guide is varied to produce the phase shifts required for scanning. The elements are dipoles. The center conductor of each dipole protrudes just enough into the guide to pick up the required amount of energy.

It is evident from the above discussion that such a waveguide fed dipole array will produce a single beam in the normal direction only if the dipoles are all fed in phase and are spaced less than a wavelength. It is therefore not satisfactory to obtain constant phase excitation by tapping the dipoles into the guide at successive guide wavelengths for these are greater than free space wavelengths. Consequently the dipoles are tapped in at successive half wavelengths in the guide and reversed successively in polarity to compensate for the successive phase reversals due to their spacing.

This type of array provides a line source which can be scanned by moving the guide walls. In order to leave these mechanically free suitable wave trapping slots are provided along the length of the array.

A practical antenna of this type will be described in Sec. 16.3.

12.3 *Optical Scanning*

With a camera or telescope all parts of an angular sector or field are viewed simultaneously. We would like to do the same thing by radar means, but since this so far appears impossible we do the next best thing by looking at the parts of the field in rapid succession. Nevertheless certain points of similarity appear. These points are emphasized by a survey of the fixed

beam antenna field for there we find optical instruments in abundance, parabolic reflectors and even lenses.

It is not a very big step to proceed from an examination of optical systems to the suggestion that a scanning antenna can be provided by moving a feed over the focal plane of a reflector. Nevertheless experience shows that this will not be especially profitable unless done with due caution. The first effect of moving the feed away from the focus in the focal plane of a paraboloid is indeed a beam shift but before this process has gone far a third order curvature of the phase front is produced and is accompanied by a serious deterioration in the pattern and reduction in gain. This difficulty or aberration is well known in classical optical theory and is called coma. Coma is typified by patterns such as the one shown in Fig. 16. It is the first obstacle in the path of the engineer who wishes to design a good moving feed scanning antenna.

Coma is not an insuperable obstacle however. Its removal can be accomplished by the application of a very simple geometrical principle. This principle can be stated as follows: "The condition for the absence of coma is that each part of the focussing reflector or lens should be located on a circle with center at the focus."

This condition can be regarded as a statement of the spacial relationship required between the feed and all parts of the focussing element. It is a condition which insures that the phase front will remain nearly linear when the feed is moved in the focal plane. It can be applied approximately whether the focussing element is a reflector or a lens and to optical systems which scan in both directions as well as those which scan in one direction.

Coma is usually the most serious aberration to be reckoned with in a scanning optical system, but it is by no means the only one. Any defect in the phase and amplitude characteristic which arises when the feed is moved can cause trouble and must be eliminated or reduced until it is tolerable. Another defect in phase which arises is 'defocussing'. Defocussing is a square law curvature of phase and arises when the feed is placed at an improper distance from the reflector or lens. Its effect may be as shown in Fig. 14. It can in principle always be corrected by moving the feed in a correctly chosen arc, but this is not always consistent with other requirements on the system. In addition to troubles in phase an improper amplitude across the aperture of the antenna will arise when the feed is translated unless proper rotation accompanies this motion.

To combat the imperfections in an optical scanning system we can choose over-all dimensions in such a way that they will be lessened. Thus it is generally true that an increase in focal length or a decrease in aperture will increase the scanning capabilities of an optical system. This alone is usually not enough, however, we must also employ the degrees of free-

dom available to us in the designing of the focussing element and the feed motion to improve the performance. If the degrees of freedom are not enough we must, if we insist on an optical solution introduce more. This could in principle result in microwave lenses similar to the four and five element glass lenses found in good cameras, but such complication has not as yet been necessary in the radar antenna art.

Since military release has not been obtained as this article goes to press we must omit any detailed discussion of optically scanning radar antenna techniques.

PART III

MILITARY RADAR ANTENNAS DEVELOPED BY THE BELL LABORATORIES

13. GENERAL

In the final part of this paper we will describe in a brief fashion the end products of radar antenna technology, manufactured radar antennas. Without these final practical exhibits the foregoing discussion of principles and methods might appear academic. By including them we hope to illustrate in a concrete fashion the rather general discussion of Parts I and II.

The list of manufactured antennas will be limited in several ways. Severe but obviously essential are the limitations of military security. In addition we will restrict the list to antennas developed by the Bell Laboratories. In cases where invention or fundamental research was accomplished elsewhere due credit will be given. Finally the list will include only antennas manufactured by contract. This last limitation excludes many experimental antennas, some initiated by the Laboratories and some by the armed forces.

It is worthwhile to begin with an account of the processes by which these antennas were brought into production. The initiating force was of course military necessity. The initial human steps were taken sometimes by members of the armed forces who had definite needs in mind and sometimes by members of the Laboratories who had solutions to what they believed to be military needs.

With a definite job in mind conferences between military and Laboratories personnel were necessary. Some of these dealt with legal or financial matters, others were principally technical. In the technical conferences it was necessary at an early date to bring military requirements and technical possibilities in line.

As a result of the conferences a program of research and development was often undertaken by the Laboratories. An initial contract was signed which

called for the delivery of technical information, and sometimes for manufacturing drawings and one or more completed models. Usually the antenna was designed and manufactured as part of a complete radar system, sometimes the contract called for an antenna alone.

After preliminary work had been undertaken the status of the job was reviewed from time to time. If preliminary results and current military requirements warranted a manufacturing contract was eventually drawn up and signed by Western Electric and the contracting government agency. This contract called for delivery of manufactured radars or antennas according to a predetermined schedule.

Research and development groups of the Laboratories cooperated in war as in peace to solve technical problems and accomplish technical tasks. Under the pressure of war the two functions often overlapped and seemed to merge, yet the basic differences usually remained.

Members of the Research Department, working in New York and at the Deal and Holmdel Radio Laboratories in New Jersey were concerned chiefly with electrical design. It was their duty to understand fully electrical principles and to invent and develop improved methods of meeting military requirements. During the war it was usually their responsibility to prescribe on the basis of theory and experiment the electrical dimensions of each new radar antenna.

A new and difficult requirement presented to the Research Department was sometimes the cause of an almost personal competition between alternative schemes for meeting it. Some of these schemes were soon eliminated by their own weight, others were carried side by side far along the road to production. Even those that lost one race might reappear in another as a natural winner.

In the Development Groups working in New York and in the greatly expanded Whippany Radio Laboratory activity was directed towards coordination of all radar components, towards the establishment of a sound, well integrated mechanical and electrical design for each component and towards the tremendous task of preparing all information necessary for manufacture. It was the job of these groups also to help the manufacturer past the unavoidable snarls and bottlenecks which appeared in the first stages of production. In addition development personnel frequently tested early production models, sometimes in cooperation with the armed forces.

As we have intimated, research and development were indistinguishable at times during the war. Members of the research department often found themselves in factories and sometimes in aircraft and warships. Development personnel faced and solved research problems, and worked closely with research groups.

For several years when pressure was high the effort was intense; at times feverish. Judging by military results it was highly effective. Some of the material results of this effort are described in the following pages.

14. NAVAL SHIPBORNE RADAR ANTENNAS

14.1 *The SE Antenna*¹⁰

Very early in the war, the Navy requested the design of a simple search radar system for small vessels, to be manufactured as quickly as possible in order to fill the gap between design and production of the more complex search systems then in process of development. The proposed system was to be small and simple, to permit its use on vessels which otherwise would be unable to carry radar equipment because of size or power supply capability. This class of vessel included PT boats and landing craft.

The antenna designed for the SE system is housed as shown in Fig. 40. It was adapted for mounting on the top or side of a small ship's mast, and is rotated in azimuth by a mechanical drive, hand operated. The paraboloid reflector is 42 inches wide, 20 inches high, and is illuminated by a circular aperture 2.9 inches in diameter. In the interests of simplicity, the polarization of the radiated beam was permitted to vary with rotation of the antenna.

The SE antenna was operated at 9.8 cm, and fed by $1\frac{1}{2} \times 3$ rectangular waveguide. At the antenna base, a taper section converted from the rectangular waveguide to 3" round guide, through a rotating joint directly to the feed opening.

Characteristics of the SE antenna are given below:

Wavelength	9.7 to 10.3 cm	
Reflector	42" W x 20" H	
Gain	25 db	
Horizontal Beam Width	6°	
Vertical Beam Width	12°, varying somewhat with polarization	
Standing Wave	9.7 to 10.0 cm	4.0 db
	10.0 to 10.3 cm	6.0 db

14.2 *The SL Radar Antenna*¹¹

The SL radar is a simple marine search radar developed by Bell Telephone Laboratories for the Bureau of Ships. During the war, over 1000 of these radars were produced by the Western Electric Company and installed on Navy vessels of various categories. The principal field for installation was destroyer escort craft ("DE"s). Figure 41 shows an SL antenna installation aboard a DE. The antenna is covered, for wind and

¹⁰ Written by R. J. Phillips.

¹¹ Written by H. T. Budenbom.

weather protection, in a housing which can transmit 10 cm radiation. Visible also is the waveguide run down the mast to the r.f. unit.

The SL radar provides a simple non-stabilized PPI (Plan Position Indicator) display. The antenna is driven by a synchronous motor at 18 rpm. Horizontal polarization is used to minimize sea clutter. The

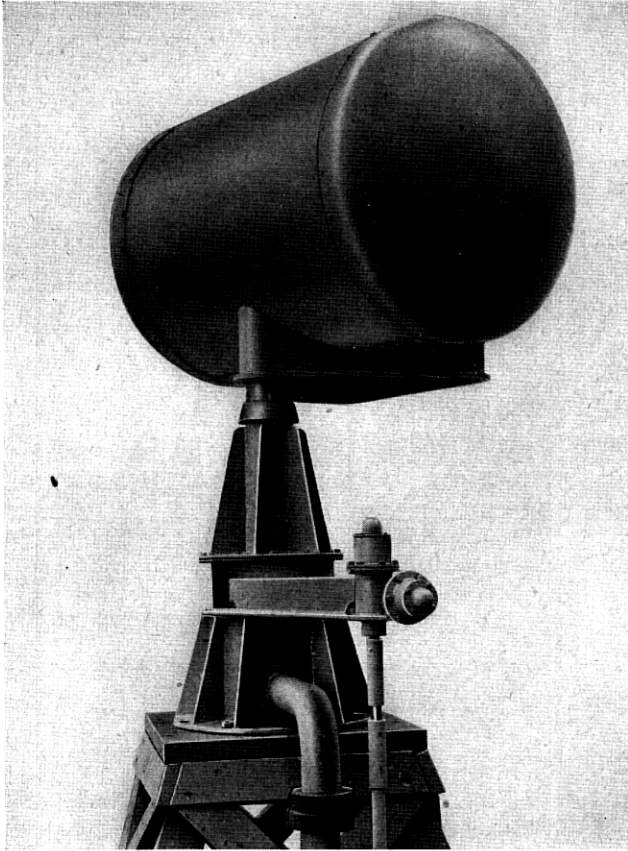


Fig. 40—SE Antenna.

radiating structure, shown in Figure 42, consists of a 20" sector of a 42" paraboloid. The resulting larger beam width in the vertical plane is provided in order to improve the stability of the pattern under conditions of ship roll. Figure 43 illustrates the path of the transmitted wave from the SL r.f. unit to the antenna. It also illustrates the manner in which horizontally polarized radiation is obtained. The diagram shows the position of

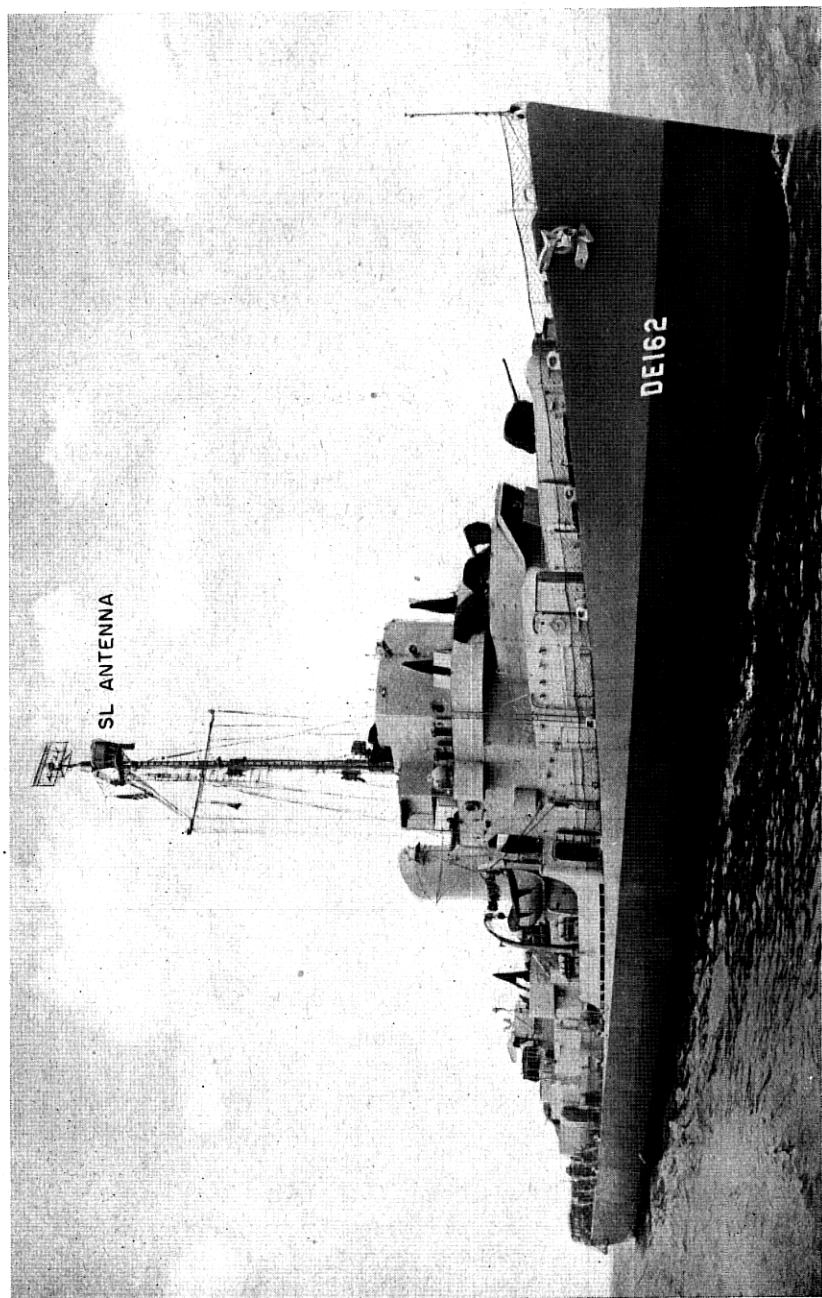


Fig. 41—SL Antenna Aboard DE.

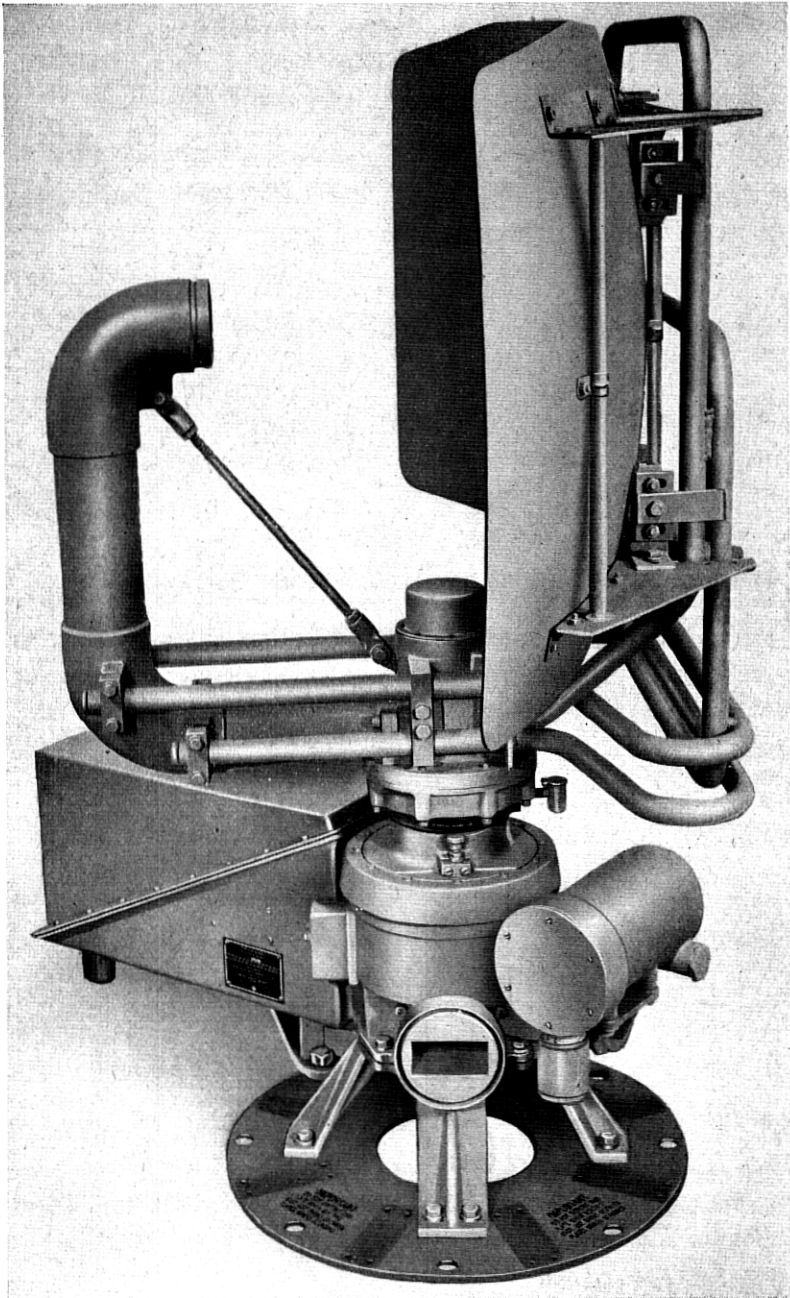


Fig. 42—SL Antenna.

the electric force vector in traversing the waveguide run. The path from the r.f. unit is in rectangular guide ($TE_{1,0}$ mode) through the right angle bend, to the base of the rotary joint. A transducer which forms the base portion of the joint converts to the TM_{01} mode in circular pipe. For this mode, the electric field has radial symmetry, much as though the waveguide were a coaxial line of vanishingly small inner conductor diameter.

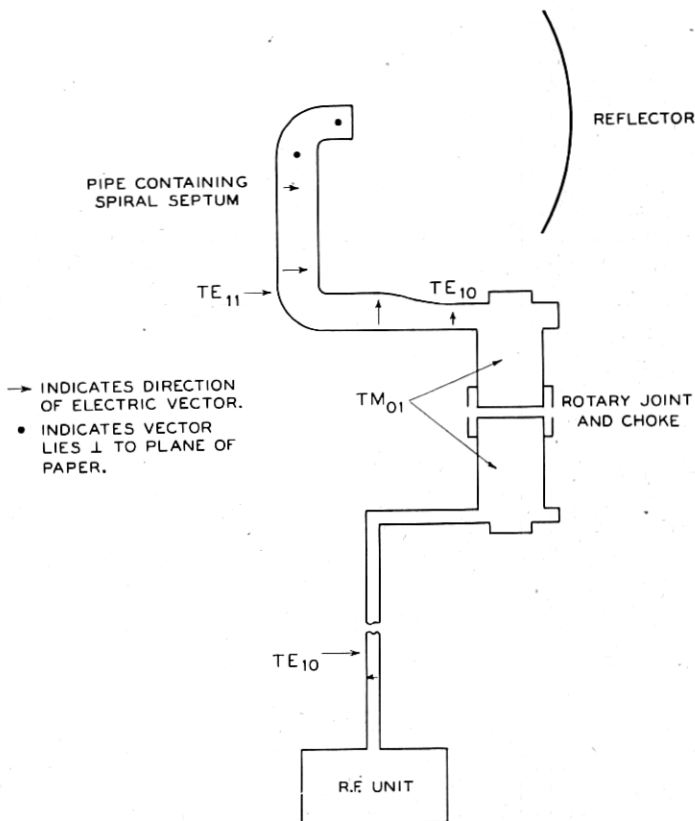


Fig. 43—SL Radar Antenna—Wave Guide Path.

The energy passes the rotary joint in this mode; choke labyrinths are provided at the joint to minimize radio frequency leakage. The energy then flows through another transducer, from TM_{01} mode back to TE_{10} mode. The lower horizontal portion of the feed pipe immediately tapers to round guide, the mode being now TE_{11} . Next the energy transverses a 90° elbow, which is a standard 90° pipe casting, and enters the vertical section im-

mediately below the feed aperture. The E vector is in the plane of the paper at this point. However, the ensuing vertical section is fitted with a spiral septum. This gradually rotates the plane of polarization until at the top of this pipe the E vector is perpendicular to the plane of the paper. Thus, after transversing another 90° pipe bend, the energy emerges horizontally polarized, to feed the main reflector.

Specific electrical characteristics of the SL antenna are:

Polarization—Horizontal
Horizontal Half Power Beamwidth—6°
Vertical Half Power Beamwidth—12°
Gain—about 22 db.

14.3 *The SJ Submarine Radar Antenna*

It had long been expected that one of the early offensive weapons of the war would be the submarine. It was therefore natural that early in the history of radar the need for practical submarine radars was felt. The principal components of this need were twofold, to provide warning of approaching enemies and to obtain torpedo fire control data. The SJ Submarine Radar was the first to be designed principally for the torpedo fire control function.

Work on the SJ system was under way considerably before Pearl Harbor. When this work was initiated the advantages of lobing fire control systems were clearly recognized, but no lobing antennas appropriate for submarine use had been developed. Requirements on such an antenna were obviously severe, for in addition to fulfilling fairly stringent electrical conditions, it would have to withstand very large forces due to water resistance and pressure.

The difficulties evident at the outset of the work were overcome by an ingenious adaptation of the simple waveguide feed. It was recognized that a shift of the feed in the focal plane of a reflector would cause a beam shift. Why not, then, use two waveguide feeds side by side to produce the two nearly coincident beams required in a lobing antenna? When this was tried it was found to work as expected.

It remained to devise a means of switching from one waveguide feed to the other with the desired rapidity. This in itself was no simple problem, but was solved by applying principles learned through work on waveguide filters. The switch at first employed was essentially a branching filter at the junction of the single antenna feed line and the line to each feed aperture. Both branches of this filter were carefully tuned to the same frequency, that of the radar. The switching was performed by the insertion of small rapidly rotating pins successively into the resonant cavities of the

two filters (Fig. 44). Presence of the pins in one of the filters detuned it and therefore prevented power from flowing through it. Rotation of the pins accordingly produced switching as desired.

In a later modification of this switch the same general principles were used but resonant irises rather than resonant cavities were employed.

The SJ Submarine Radar was in use at a comparatively early date in the war and saw much service with the Pacific submarine fleet. Despite some early doubts, submarine commanders were soon convinced of its powers.

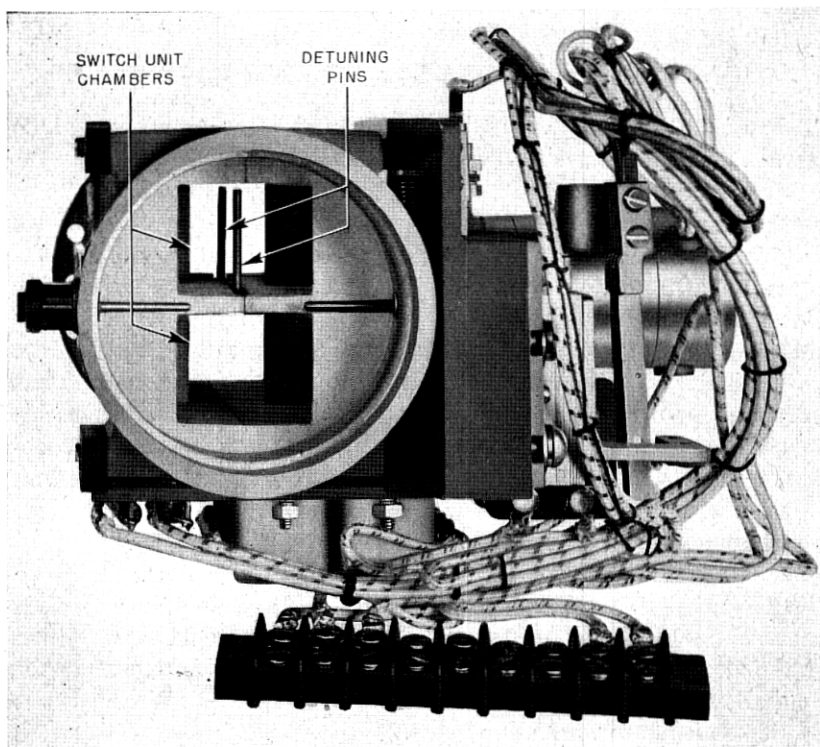


Fig. 44—The SJ Tuned Cavity Switch.

It is believed that in the majority of cases it replaced the periscope as the principle fire control instrument. In addition it served as a valuable and unprecedented aid to navigation.

It is interesting and relevant to quote from two letters to Laboratories engineers concerning the SJ. One dated October 3, 1943, from the radar officer of a submarine stated that there were twenty "setting sun" flags painted on the conning tower and asked the engineer to "let your mind dwell on the fact that you helped to put more than 50% of those flags there".

The commander of another submarine wrote in a similar vein, "You can rest assured that we don't regard your gear as a bushy-brain space taker, but a very essential part of our armament".

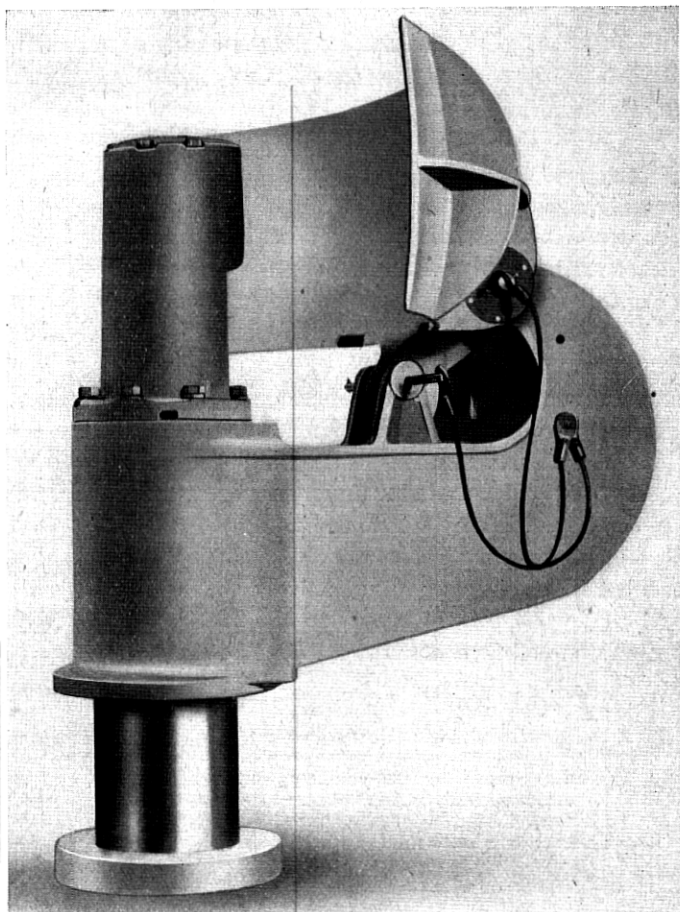


Fig. 45—The SJ Submarine Radar Antenna.

Figure 45 is a photograph of an SJ antenna. Its principal electrical characteristics are as follows:

Gain > 19 db
Horizontal Half Power Beamwidth 8°
Vertical Half Power Beamwidth 18°
Vertical Beam Character—Some upward radiation
Lobe Switching Beam Separation—approximately 5°
Gain reduction at beam cross-over < 1 db
Polarization—Horizontal

14.4 The Modified SJ/Mark 27 Radar Antenna

The SJ antenna described above performed a remarkable and timely fire control job as a lobing antenna but was found to be unsatisfactory when rotated continuously to produce a Plan Position Indicator (PPI) presentation. In the PPI method of presentation range and angle are presented as radius and angle on the oscilloscope screen. Consequently a realistic map of the strategic situation is produced. This map is easily spoiled by false signals due to large minor lobes of the antenna.

Since it had been established that the PPI picture was valuable for navigation and warning as well as for target selection it was decided to modify the antenna in a way that would reduce these undesirably high minor lobes. These were evidently due principally to the shadowing effect of the massively built double primary feed. Accordingly a new reflector was designed which in combination with a slightly modified feed provided a much improved pattern.

The new reflector was different in configuration principally in that it was a partially offset section of a paraboloid. The reflector surface was also markedly different in character since it was built as a grating rather than a solid surface. This reduced water drag on the antenna. In addition the grating was less visible at a distance, an advantage that is obviously appreciable when the antenna is the only object above the water.

This modified antenna was used not only on submarines as part of the SJ-1 radar but also on surface vessels as the Mark 27 Radar Antenna. Figure 46 shows one of these antennas. Its electrical characteristics are as follows:

Gain > 20 db
 Horizontal Half Power Beamwidth = 8°
 Vertical Half Power Beamwidth = 17°
 Vertical Beam Character—Some upward radiation
 Lobe Switching Beam Separation—approximately 5°
 Gain reduction at beam cross-over < 1 db
 Polarization—Horizontal

14.5 The SH and Mark 16 Antenna¹²

The antennas designed for the SH and Mark 16 Radar Equipments are practically identical. The SH system was a shipborne combined fire control and search system, and the Mark 16 its land based counterpart was used by the Marine Corps for directing shore batteries.

These systems operated at 9.8 cm. The requirement that the system, operate as a fire control as well as a search system imposed some rather stringent mechanical requirements on the antenna. For search purposes, the antenna was rotated at 180 rpm, and indications were presented on a plan position indicator. For fire control data, slow, accurately controlled motion was required. Bearing accuracy is attained by lobe switching in

¹² Written by R. J. Philipps.

much the same manner as in the SJ and SJ-1 antennas previously described.

The antenna is illustrated in Fig. 47. With the SH system, the unit is mast mounted; for the Mark 16, the unit is mounted atop a 50 foot steel

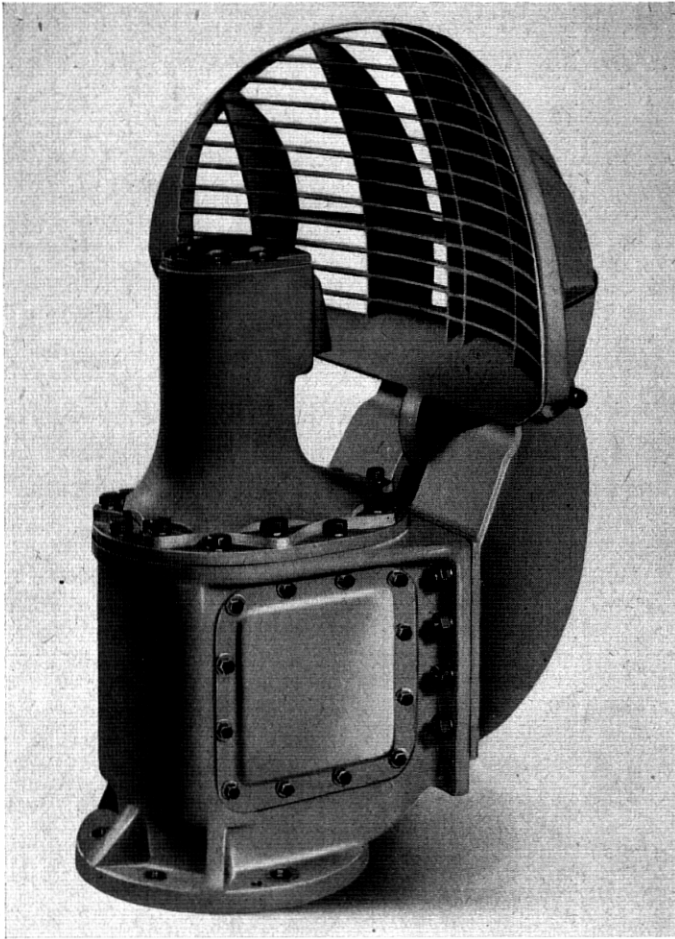


Fig. 46—The SJ-1/Mark 27 Submarine Radar Antenna.

tower which can be erected in a few hours with a minimum of personnel. The electrical characteristics are as follows:

Gain—21. db

Reflector Dimensions 30" W x 20" H

Horizontal beam width—7.5°

Vertical beam width—12°

Lobe separation—5° approximately

Loss in gain at lobe crossover—1 db approximately

Scan—(1) 360°, at 180 rpm for PPI operation

(2) 360°, at approximately 1 rpm for accurate azimuth readings, with lobe switching

SH systems were most successfully used in invasion operations in the Aleutians. They were installed on landing craft, and the use of the high

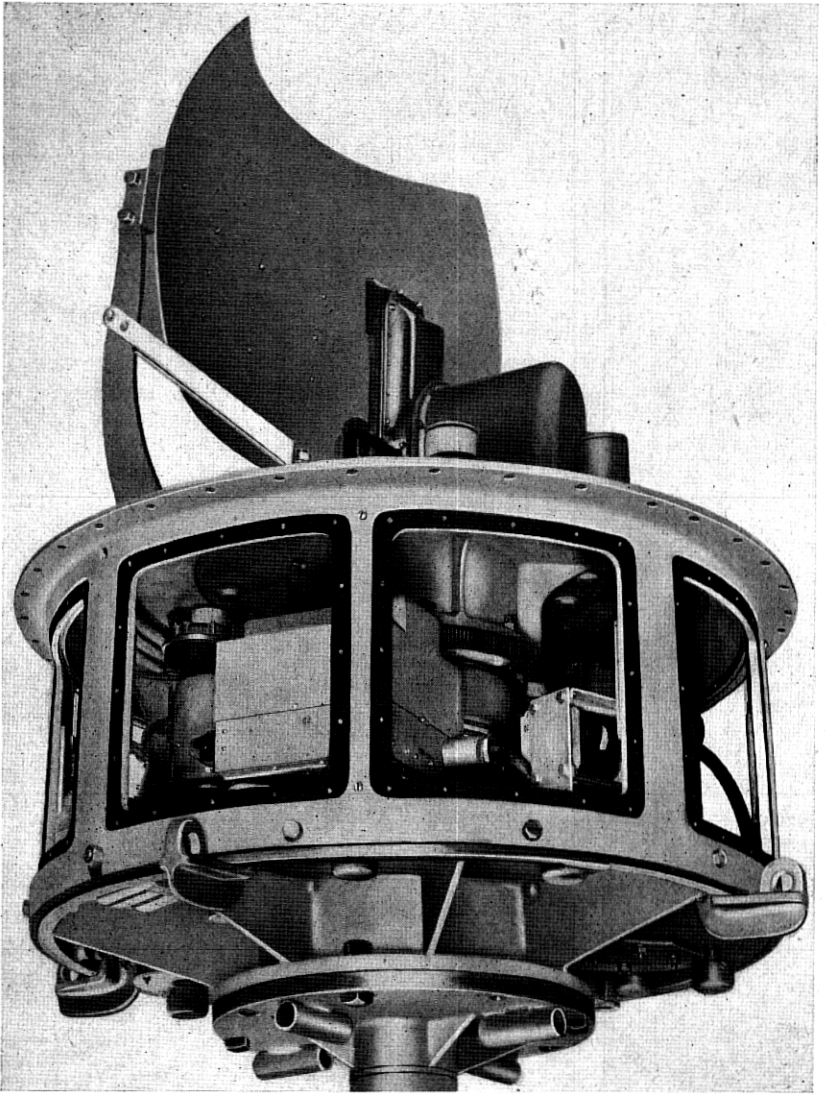


Fig. 47—SH Antenna.

speed scan enabled the craft to check constantly their relative positions in the dense fogs encountered during the landing operations.

14.6 Antennas for Early Fire Control Radars¹³

The first radars to be produced in quantity for fire control on naval vessels were the Mark 1, Mark 3 and Mark 4 (originally designated FA, FC and FD). These radars were used to obtain the position of the target with sufficient accuracy to permit computation of the firing data required by the guns. The first two (Mark 1 and Mark 3) were used against enemy surface targets while the Mark 4 Radar was a dual purpose system for use against both surface and aircraft targets. These radars were described in detail in an earlier issue.¹⁴ However, photographs of the antennas and pertinent information on the antenna characteristics are repeated herein for the sake of completeness. (See Table B and Figures 48, 49 and 50.)

TABLE B

Dimensions	Radar			
	Mark 1	Mark 3		Mark 4
	6' x 6'	3' x 12'	6' x 6'	6' x 7'
Operating Frequency	500 or 700 MC	680-720 MC		680-720 MC
Beam Width in Degrees (Between half power points one way.)				
Azimuth	12°	6°	12°	12°
Elevation	14°	30°	14°	12°
Antenna Gain	22 db	22 db	22 db	22.5 db.
Beam Shift in Degrees				
Azimuth	0°	±1.5°	±3°	±3°
Elevation	0°	0°	0°	±3°

An antenna quite similar to the Mark 3, 6 ft. x 6 ft. antenna, was also used on Radio Set SCR-296 for the Army. This equipment was similar to the Mark 3 in operating characteristics but was designed mechanically for fixed installations at shore points for the direction of coast artillery gun fire. For these installations the antenna was mounted on an amplidyne controlled turntable located on a high steel tower. The entire antenna and turntable was housed within a cylindrical wooden structure resembling a water tower. Equipments of this type were used as a part of the coastal defense system of the United States, Hawaiian Islands, Aleutian Islands and Panama.

¹³ Written by W. H. C. Higgins.

¹⁴ "Early Fire Control Radars for Naval Vessels," W. C. Tinus and W. H. C. Higgins, B. S. T. J.

14.7 A Shipborne Anti-Aircraft Fire Control Antenna¹⁵

A Shipborne Anti-Aircraft Fire Control Antenna is shown in Fig. 51. This antenna consists of two main horizontal cylindrical paraboloids in each

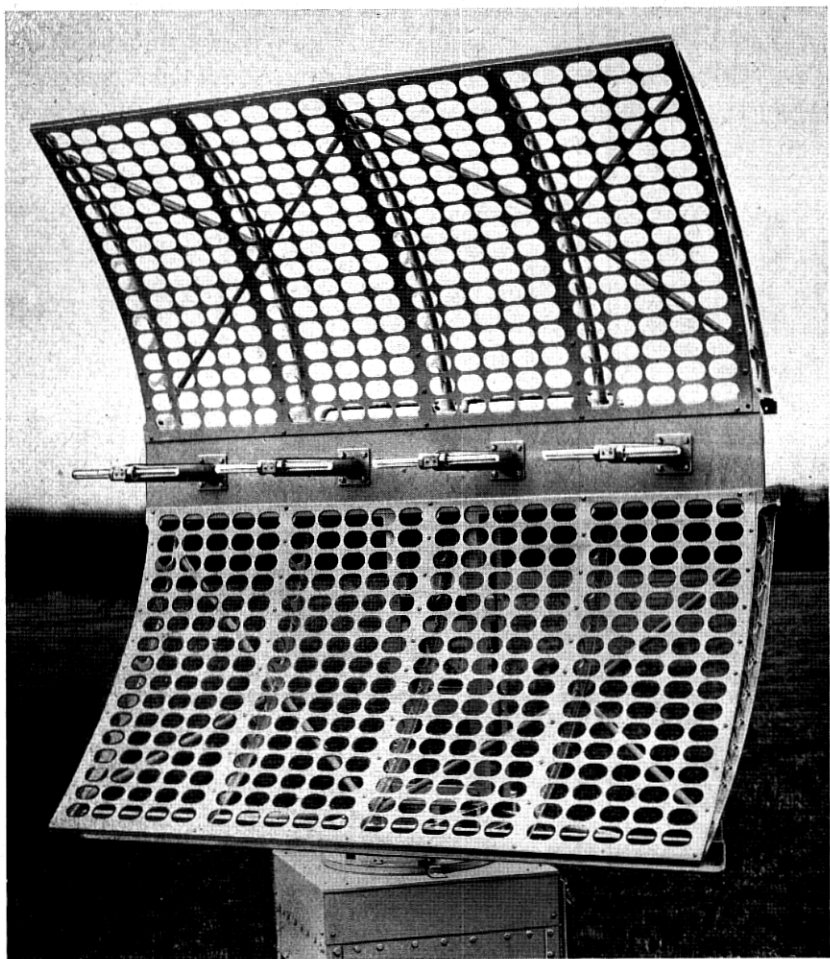


Fig. 48—Mark 1 Antenna.

of which two groups of four half-wave dipoles are mounted with their axes in a horizontal line at the focus of the parabolic reflectors. The four groups of dipoles are connected by coaxial lines on the back of the antenna to a lobe

¹⁵ Written by C. A. Warren.

switcher, which is a motor driven capacitor that has a single rotor plate and four stator plates, one for each group of dipoles. The phase shift introduced into the four feed lines by the lobe switching mechanism causes the antenna beam to be "lobed" or successively shifted to the right, up, left and down as the rotor of the capacitor turns through 360 degrees.

Mounted centrally on the front of the antenna at the junction of the two parabolic antennas is a smaller auxiliary antenna consisting of two dipole elements and a parabolic reflector, the purpose of which is to reduce the minor lobes that are present in the main antenna beam. The auxiliary

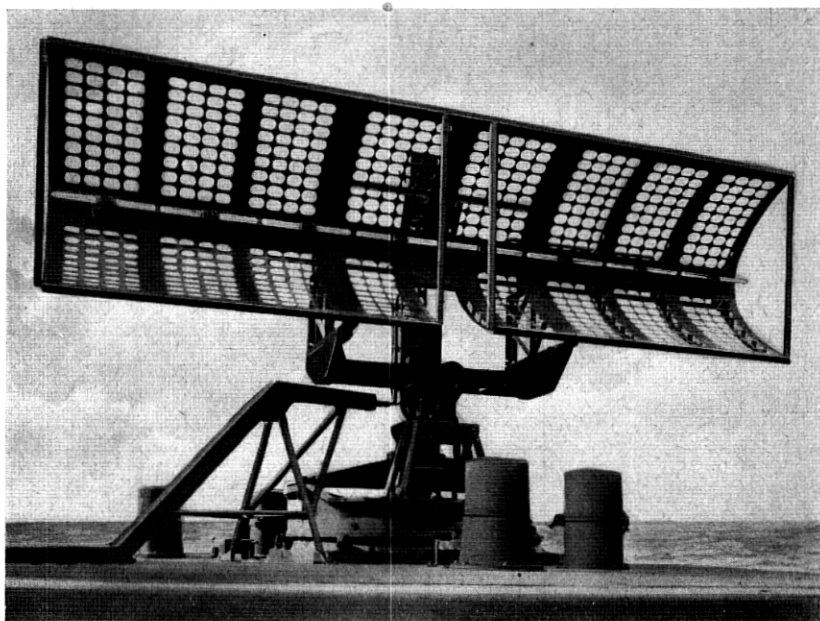


Fig. 49—Mark 3 Radar Antenna on Battleship New Jersey.

antenna beam is not lobe switched and is sufficiently broad in both the horizontal and vertical planes to overlap both the main antenna beam and the first minor lobes. The auxiliary antenna feed is so designed that its field is in phase with the field of the main beam of the main antenna. This causes the feed of the auxiliary antenna to "add" to the field of the main antenna in the region of its main beam, but to subtract from the field in the region of its first minor lobes. This occurs because the phase of the first minor lobes differs by 180 degrees from that of the main beam. As a result, the field of the main beam is increased and the first minor lobes are greatly

reduced. By reducing these minor lobes to a low value, the region around the main beam is free of lobes, thus greatly reducing the possibility of false tracking due to "cross overs" between the main beam and the minor lobes.

14.8 *The Polyrod Fire Control Antenna*

The Polyrod Fire Control antenna is an array scanner employing essentially the same principles as those used in the multiple unit steerable antenna

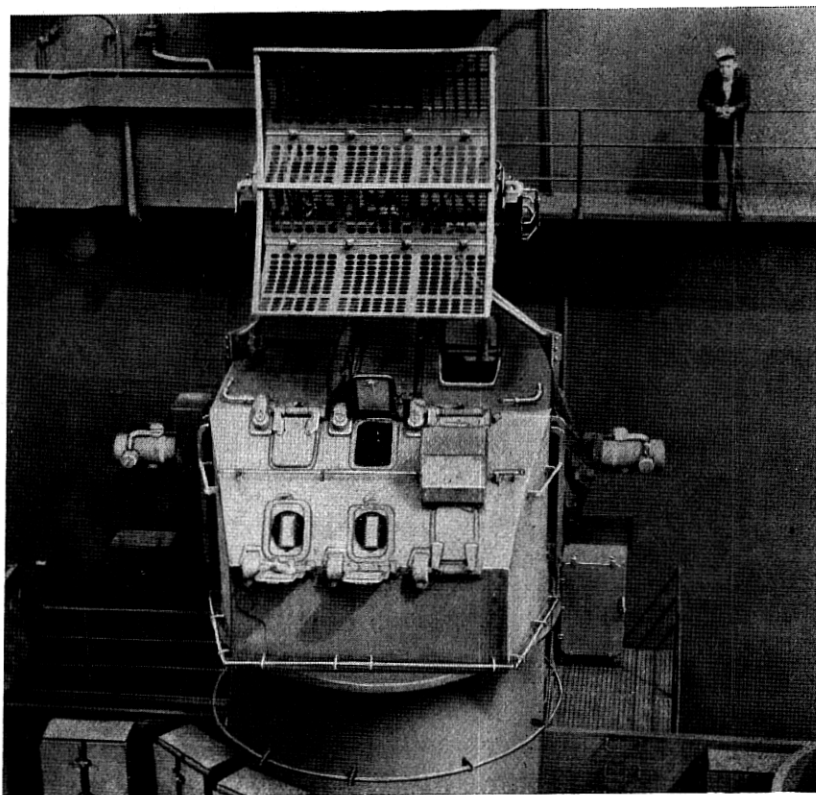


Fig. 50—Mark 4 Radar Antenna on Battleship Tennessee.

system (MUSA) developed before the war for short-wave transatlantic telephony. Some of these principles have been discussed in Sec. 12.2. That they could be applied with such success in the microwave region was due to a firm grounding in waveguide techniques, to the invention of the polyrod antenna and the rotary phase changer, and especially to excellent technical work on the part of research, development and production personnel. It is perhaps one of the most remarkable achievements of wartime.

radar that the polyrod antenna emerged to fill the rapid scanning need a early and as well developed as it did.

The Polyrod Fire Control antenna is a horizontal array of fourteen identical fixed elements, each element being a vertical array of three polyrods. Energy is distributed to the elements through a waveguide manifold. The phase of each element is controlled and changed to produce the desired scan by means of thirteen rotary phase changers. These phase shifters are

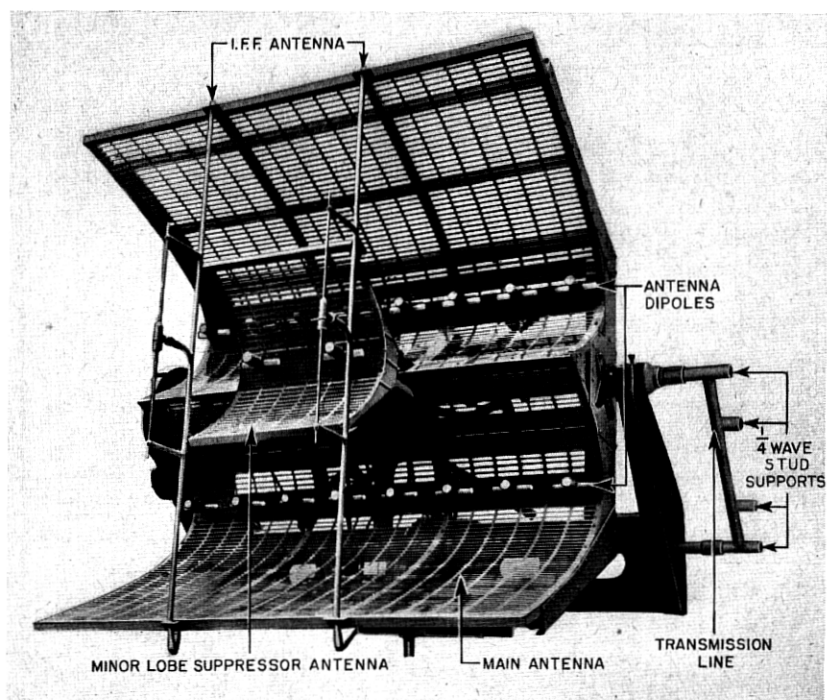


Fig. 51.—Shipborne Anti-Aircraft Fire Control Antenna

geared together and driven synchronously. Figure 52 is a schematic diagram of the waveguide and phase changer circuits.

Figure 39 shows an experimental polyrod antenna under test at Holmdel. Figure 53 is another view of the Polyrod antenna.

14.9 The Rocking Horse Fire Control Antenna

It was long recognized that an important direction of Radar development lay towards shorter waves. This is particularly true for fire control antennas where narrow, easily controlled beams rather than great ranges are needed. The Polyrod antenna had pretty thoroughly demon-

strated the value of rapid scanning, yet the problem of producing a rapid scanning higher frequency antenna of nearly equal dimensions was a new and different one.

Several possible solutions to this problem were known. The array technique applied so effectively to the polyrod antenna could have been applied here also, but only at the expense of many more elements and greater complexity.

After much preliminary work it was finally concluded that a mechanically scanning antenna, the "rocking horse," provided the best solution to the higher frequency scanning problem. This solution is practical and relatively simple.

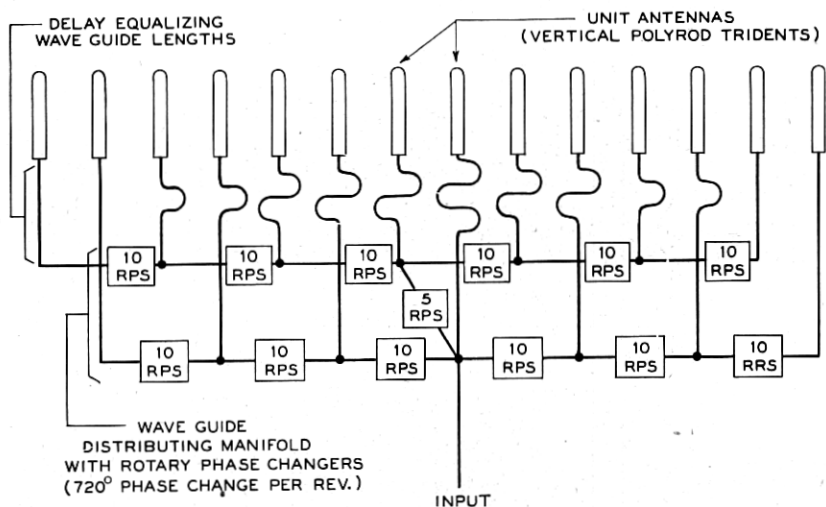


Fig. 52.—Schematic Diagram of Polyrod Fire Control Antenna.

The operation of the rocking horse is described in Sec. 12.1. It is essentially a carefully designed and firmly built paraboloidal antenna which oscillates rapidly through the scanning sector. Its oscillation is dynamically balanced to eliminate undesirable vibration.

Figure 54 is a photograph of a production model of the rocking horse antenna.

14.10 The Mark 19 Radar Antenna¹⁶

In Anti-aircraft Fire Control Radar Systems for Heavy Machine Guns it is necessary to employ a highly directive antenna and to obtain continuous rapid comparison of the received signals on a number of beam positions

¹⁶ Sections 14.10, 14.11 and 14.12 were written by F. E. Nimmcke.

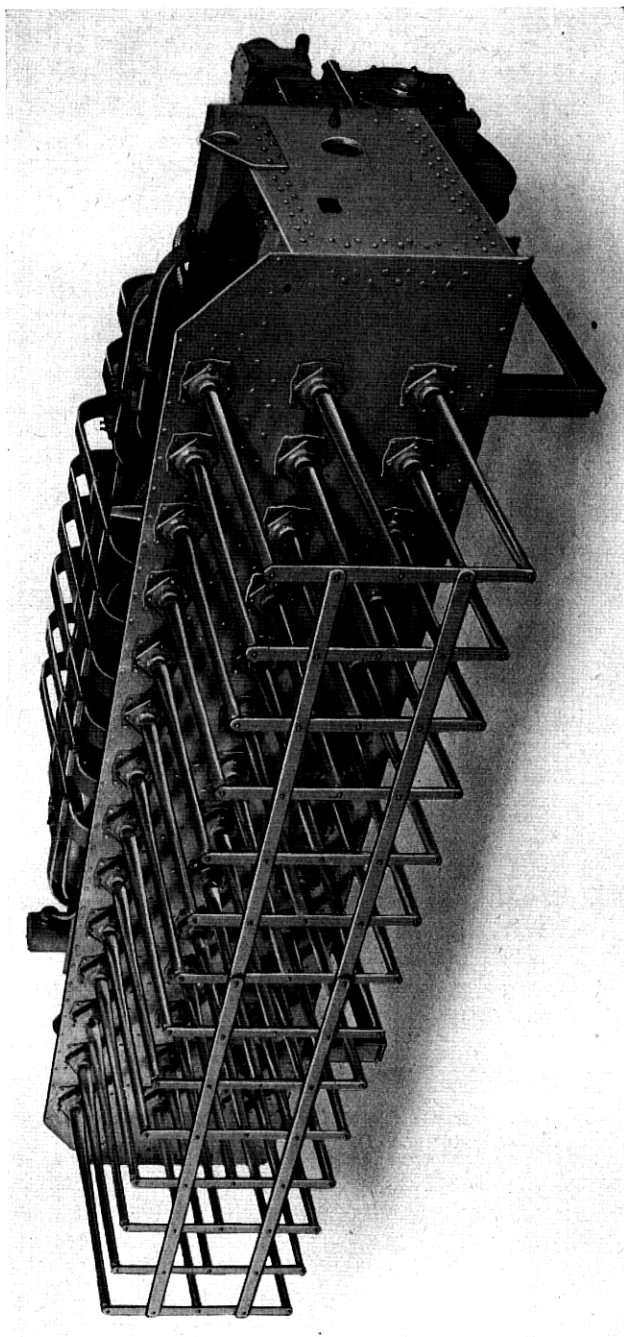


Fig. 53.—Polyrod Fire Control Antenna.

as discussed in Section 11.2. Such an antenna is also required to obtain the high angular precision for anti-aircraft fire control. These requirements are achieved by the use of a conical scanning system. The beam from the antenna describes a narrow cone and the deviation of the axis of the cone from the line of sight to the target can be determined and measured by the phase difference between the amplitude modulated received signal and the frequency of the reference generator associated with the

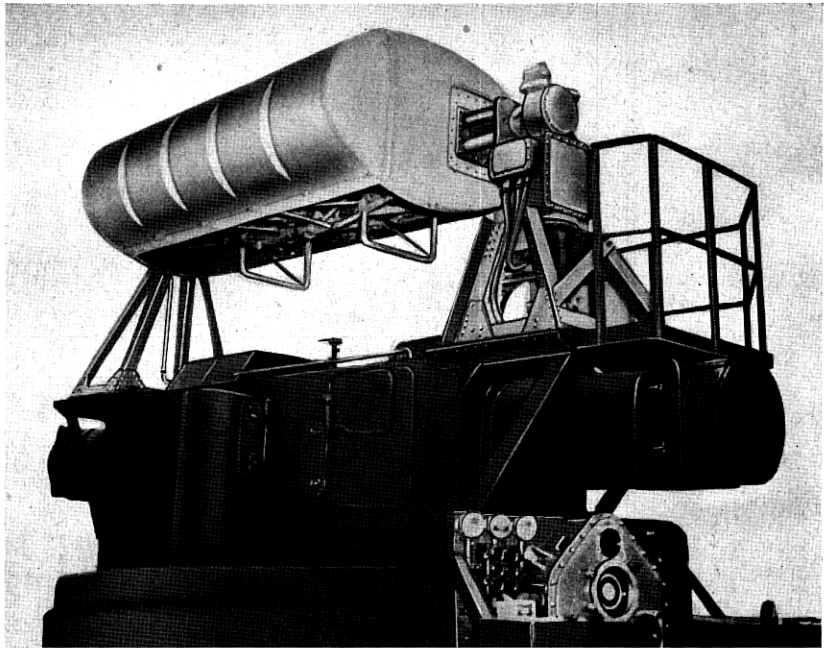


Fig. 54.—Rocking Horse Fire Control Antenna.

antenna. This information is presented to the pointer-trainer at the director in the form of a wandering dot on an oscilloscope.

The antennas described in sections 14.10, 14.11 and 14.12 were all designed by the Bell Laboratories as anti-aircraft fire control radar systems, particularly for directing heavy machine guns. They were designed for use on all types of Naval surface warships.

In Radar Equipment Mark 19, the first system to be associated with the control of 1.1 inch and 40 mm anti-aircraft machine guns, the antenna was designed for operation in the 10 cm region. This antenna consisted of a spinning half dipole with a coaxial transmission line feed. The antenna

was driven by 115-volt, 60 cycle, single phase motor to which was coupled a two-phase reference voltage generator. The motor rotated at approximately 1800 rpm which resulted in a scanning rate of 30 cycles per second. This antenna was used with a 24-inch spun steel parabolic reflector which provided, at the 3 db point, a beam width of approximately 11° and a beam shift of 8.5° making a total beam width of approximately 20° when scanning. The minor lobes were down more than 17 db (one way) from the maximum; and the gain of this antenna was 21 db. This antenna assembly

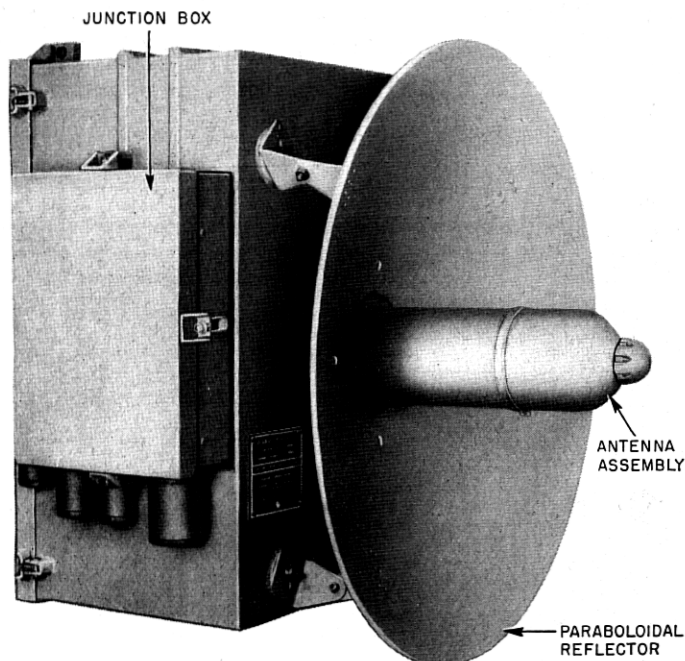


Fig. 55—Mark 19 Antenna.

was integral with a transmitter-receiver (Fig. 55) which was mounted on the associated gun director. Consequently, the size of the reflector was limited by requirements for unobstructed vision for the operators in the director. As a matter of fact, for this type of radar system serious consideration must be given to the size and weight of the antenna and associated components.

14.11 *The Mark 28 Radar Antenna*

The beam from the antenna used in Radar Equipment Mark 19 was relatively broad and to improve target resolution, the diameter of the

reflector for the antenna in Mark 28 was approximately doubled. The Mark 28 is a 10 cm system and employs a conical scanning antenna similar to that described for Mark 19. The essential difference is that the spun steel parabolic reflector is 45 inches in diameter which provides a beam width of approximately 6.5° and a beam shift of 4.5° making a total of 11° .

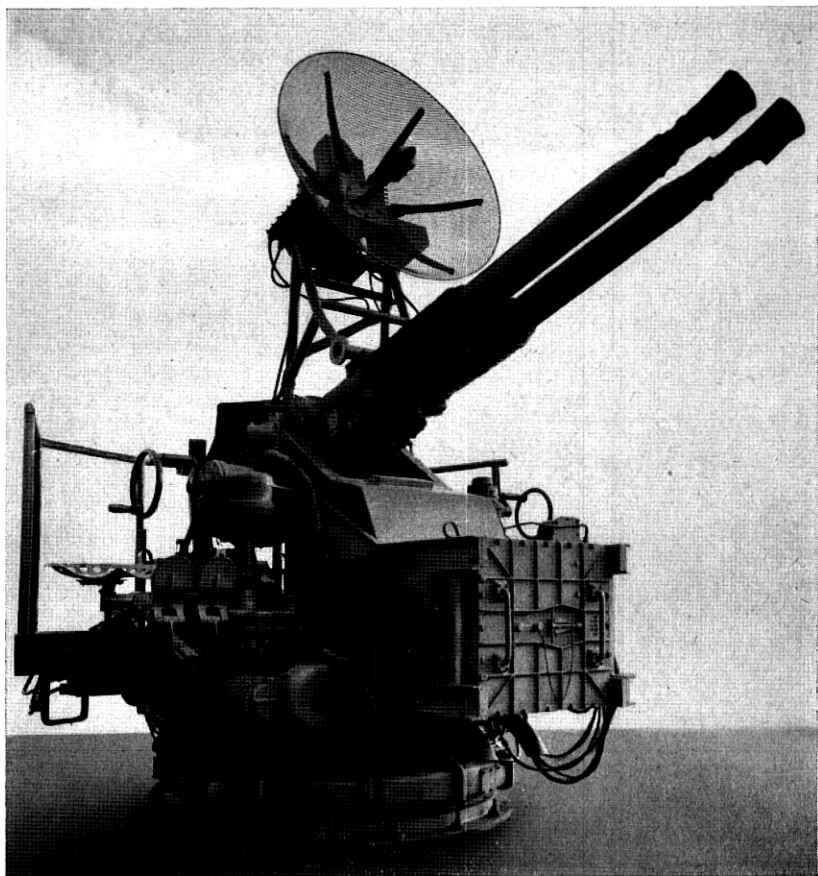


Fig. 56—Mark 28 Antenna Mounted on 40 MM Gun.

The minor lobes are down more than 17 db (one way) from the maximum; and the gain of this antenna is 26 db. It was found necessary to perforate the reflector of this dimension in order to reduce deflection caused by gun blast and by wind drag on the antenna assembly. The antenna assembly for Radar Equipment Mark 28 is shown in Fig. 56. This assembly is shown mounted on a 40 mm Gun.

14.12 A 3 CM Anti-Aircraft Radar Antenna.

To obtain greater discrimination between a given target and other targets, or between a target and its surroundings, the wavelength was reduced to the 3 cm region. An antenna for this wavelength was designed to employ the conical scan principle. In this case the parabolic reflector was 30 inches in diameter and transmitted a beam approximately 3° wide at the 3db point with a beam shift of 1.5° making a total of 4.5° with the antenna scanning. The minor lobes are down more than 22 db (one way) from the maximum; and the gain of this antenna is 35 db.

In the 3 cm system in which a Cutler feed was used, the axis of the beam was rotated in an orbit by "nutation" about the mechanical axis of the antenna. This was accomplished by passing circular waveguide through the hollow shaft of the driving motor. The rear end of the feed (choke coupling end) was fixed in a ball pivot while the center (near the reflector) was off set the proper amount to develop the required beam shift. This off set was produced by a rotating eccentric driven by the motor. The latter was a 440 volt, 60 cycle, 3 phase motor rotating at approximately 1800 rpm which resulted in a scanning rate of 30 cycles per second. The two-phase reference voltage generator was integral with the driving motor.

It was found necessary at these radio frequencies to use a cast aluminum reflector and to machine the reflecting surface to close tolerances in order to attain the consistency in beam width and beam direction required for accurate pointing. An antenna assembly for the 3 cm anti-aircraft radar is shown in Fig. 57.

15. LAND BASED RADAR ANTENNAS

15.1 *The SCR-545 Radar "Search" and "Track" Antennas*¹⁷

The SCR-545 Radar Set was developed at the Army's request to meet the urgent need for a radar set to detect aircraft and provide accurate target tracking data for the direction of anti-aircraft guns.

This use required that a narrow beam tracking antenna be employed to achieve the necessary tracking accuracy, furthermore, a narrow beam antenna suitable for accurate tracking has a very limited field of view and requires additional facilities for target acquisition. This was provided by the search antenna which has a relatively large field of view and is provided with facilities for centering the target in its field of view. These two antennas are integrated into a single mechanical structure and both radar axes coincide.

The "Search" antenna operates in the 200 mc band and is com-

¹⁷ Section 15.1 was written by A. L. Robinson.

posed of an array of 16 quarter wave dipoles spaced 0.1 wave-length in front of a flat metal reflector. All feed system lines and impedance matching devices are made up of coaxial transmission line sections. The array is divided into four quarters, each being fed from the lobe switching mechanism. This division is required to permit lobe switching in both horizontal and vertical planes. The function of the lobe switching mecha-

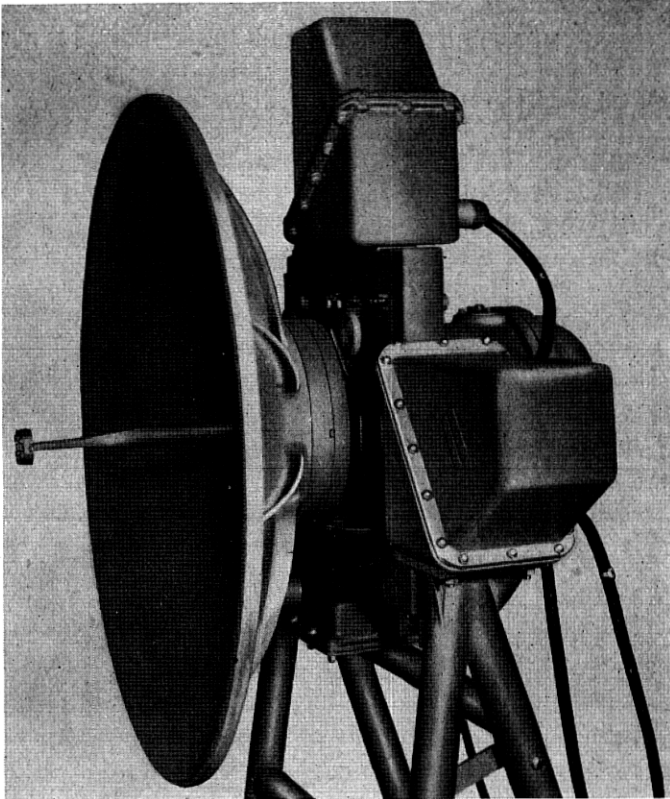


Fig. 57.—3CM Anti-Aircraft Radar Antenna.

nism is to introduce a particular phase shift in the excitation of the elements of one half of the antenna with respect to the other half. The theory of this type of lobe switching is discussed in section 11.1. The antenna beam spends approximately one quarter of a lobing cycle in each one of the four lobe positions. Each of the four lobe positions has the same radiated field intensity along the antenna axis and therefore when a target is on axis equal signals will be received from all four lobe positions.

The "Track" antenna operates in the 10 cm. region and consists of a reflector which is a parabola of revolution, 57 inches in diameter, illuminated by a source of energy emerging from a round waveguide in the lobing mechanism. Conical lobing is achieved by rotating the source of energy around the parabola axis in the focal plane of the parabola. Conical lobing is discussed in section 11.2. The round waveguide forming the source is filled with a specially shaped polystyrene core to control the illumination of the parabola and to seal the feed system against the weather. The radio frequency power is fed through coaxial transmission line to a coaxial-waveguide transition which is attached to the lobing mechanism.

The "Search" and "Track" antenna lobing mechanisms are synchronized and driven by a common motor.

The radio frequency power for both antennas is transmitted through a single specially constructed coaxial transmission line to the common antenna structure, where a coaxial transmission line filter separates the power for each antenna.

Figure 58 is a photograph of a production model of the SCR-545 Radar Set. The principal electrical characteristics of the antennas are tabulated below:

	Antennas	
	Search	Track
Gain	14.5 db	30 db
Horizontal Beamwidth	23.5°	5°
Vertical Beamwidth	25.5°	5°
Polarization	Horizontal	Vertical
Type of Lobing	Lobe switching	Conical lobing
Angle between lobe positions	10°	3°
Lobing rate	60 cycles/sec.	60 cycles/sec.

The SCR-545 played an important part in the Italian campaign, particularly in helping to secure the Anzio Beach Head area, as well as combating the "V" bombs in Belgium. However the majority of SCR-545 equipments were sent to the Pacific Theater of Operations and played an important part in operations on Leyte, Saipan, Iwo Jima, and Okinawa.

15.2 The AN/TPS-1A Portable Search Antenna¹⁸

In order to provide early warning information for advanced units, a light weight, readily transportable radar was designed under Signal Corps contract.

¹⁸ Written by R. E. Crane.

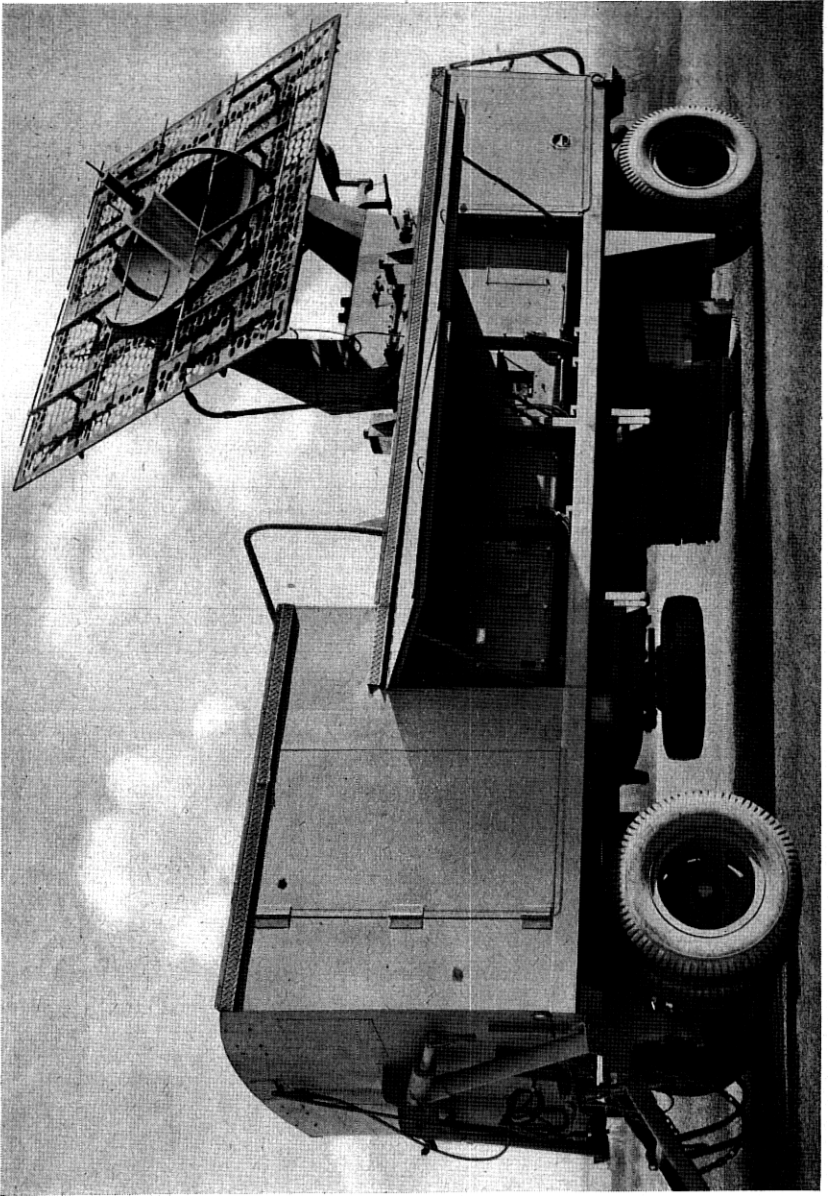


Fig. 58—SCR/545 Antenna.

The objective was to obtain as long range early warning as possible with moderate accuracy of location. Emphasis was placed on detection of low flying planes.

The objectives for the set indicated that the antenna should be built as large as reasonable and placed as high as reasonable for a portable set. Some latitude in choice of frequency was permitted at first. For ruggedness and reliability reasons which seemed controlling at the time, the frequency was pushed as high as possible with vacuum tube detectors and R.F. amplifiers. This was finally set at 1080 mc.

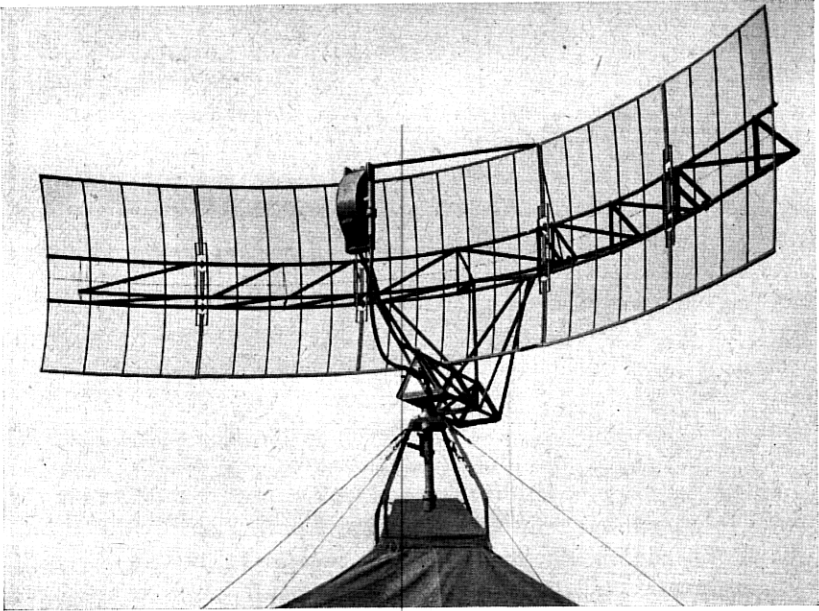


Fig. 59—AN/TPS-1A Antenna.

The antenna as finally produced was 15 ft. in width and 4 ft. in height. The reflecting surface was paraboloidal. The mouth of the feed horn was approximately at the focus of the generating parabola. The feedhorn was excited by a probe consisting of the inner conductor of the coaxial transmission line extended through the side of the horn and suitably shaped. To reduce side lobes and back radiation the feedhorn was dimensioned to taper the illumination so that it was reduced about 10 db in the horizontal and vertical planes at the edges of the reflector. Dimensions of probe and exact location of feed, etc. were determined empirically to secure acceptable impedance over the frequency band needed. This band, covered by spot frequency magnetrons, was approximately $\pm 2.5\%$ from mid frequency.

Figure 59 shows the antenna in place on top of the set.

The characteristics of this antenna are summarized below:

Gain	27.3 db.
Horizontal Half Power Beamwidth	4.4°
Vertical Half Power Beamwidth	12.6°
Vertical Beam Characteristic	Symmetrical
Polarization	Horizontal
Impedance (SWR over $\pm 2.5\%$ band)	<4.0 db

16. AIRBORNE RADAR ANTENNAS

16.1 *The AN/APS-4 Antenna*¹⁹

AN/APS-4 was designed to provide the Navy's carrier-based planes with a high performance high resolution radar for search against surface and airborne targets, navigation and interception of enemy planes under conditions of fog and darkness. For this service, weight was an all important consideration and throughout a production schedule that by V-J day was approaching 15,000 units, changes to reduce weight were constantly being introduced. In late production the antenna was responsible for 19 lbs. out of a total equipment weight of 164 lbs. The military requirements called for a scan covering 150° in azimuth ahead of the plane and 30° above and below the horizontal plane in elevation. To meet this requirement a Cutler feed and a parabolic reflector of 6.3" focal length and 14½" diameter was selected. Scanning in azimuth was performed by oscillating reflector and feed through the required 150° while elevation scan was performed by tilting the reflector. Beam pattern was good for all tilt angles. In early flight tests the altitude line on the B scope due to reflection from the sea beneath was found to be a serious detriment to the performance of the set. To reduce this, a feed with elongated slots designed for an elliptical reflector was tried and found to give an improvement even when used with the approximately round reflector. The elliptical reflector was also tried, but did not improve the performance sufficiently to justify the increased size.

As will be noted in Fig. 60, the course of the mechanical development brought the horizontal pivot of the reflector to the form of small ears projecting through the parabola. No appreciable deterioration of the beam pattern due to this unorthodox expedient was noted.

The equipment as a whole was built into a bomb-shaped container hung in the bomb rack on the underside of the wing. Various accidents resulted in this container being torn off the wing in a crash landing in water or dropped on the deck of the carrier. After these mishaps, the equipment was frequently found to be in good working order with little or no repair required.

¹⁹ Written by F. C. Willis.

Gain	28 db
Beamwidth	6° approx. circular
Polarization	Horizontal
Scan	Mechanical
Scanning Sector—Azimuth	150°
Scanning Sector—Elevation	60°
Scanning Rate	one per sec.
Total weight	19 lbs.

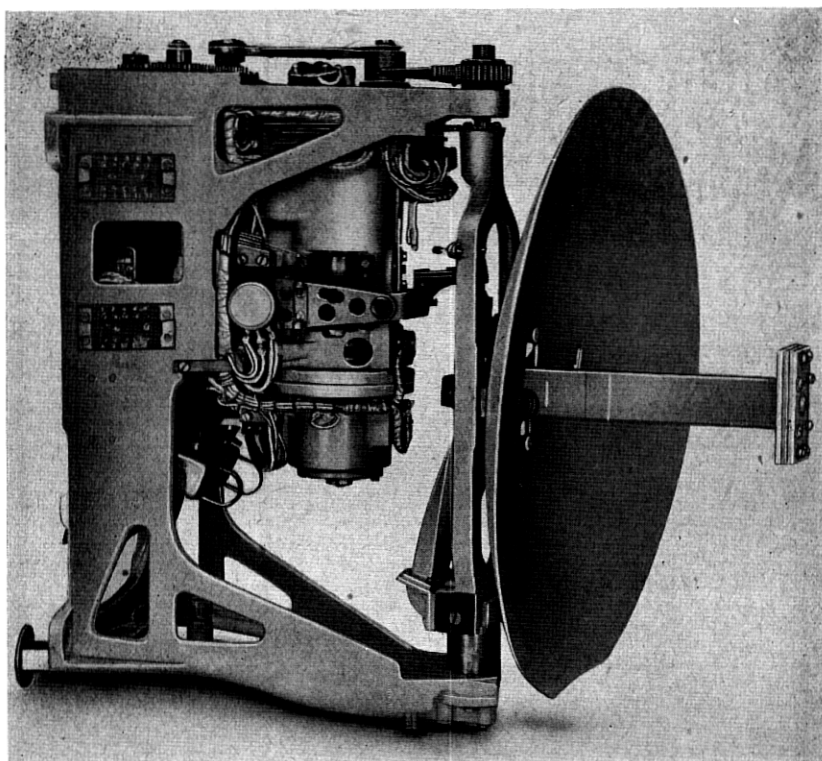


Fig. 60—AN/APS-4 Antenna.

16.2 The SCR-520, SCR-717 and SCR-720 Antennas²⁰

The antenna shown in Fig. 61 is typical of the type used with the SCR-520 and SCR-720 aircraft interception (night fighter) airborne radar equipment, as well as the SCR-717 sea search and anti-submarine airborne radar equipment. The parabolic reflector is 29 inches in diameter and produces a radiation beam about 10° wide. The absolute gain is approximately 25 db. RF energy is supplied to a pressurized emitter through a pressurized transmission line system which includes a rotary joint located on the ver-

²⁰ Written by J. F. Morrison.

tical axis and a tilt joint on the horizontal axis. Either vertical or horizontal polarization can be used by rotating the mounting position of the emitter. Vertical polarization is preferred for aircraft interception work and horizontal polarization is preferred for sea search work.

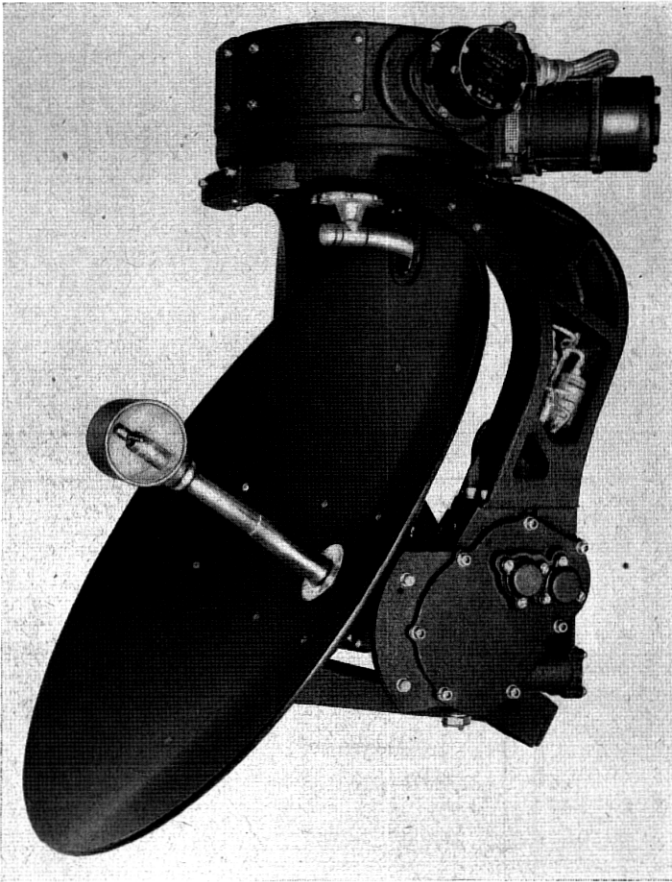


Fig. 61—SCR-520 Antenna.

For aircraft interception the military services desired to scan rapidly a large solid angle forward of the pursuing airplane, i.e. 90° right and left, 15° below and 50° above the line of flight. The data is presented to the operator in the form of both "B" and "C" presentations and for this purpose potentiometer data take-offs are provided on the antenna. The reflector is spun on a vertical axis at a rate of 360 rpm and at the same time it is

made to nod up and down about its horizontal axis by controllable amounts up to a total of 65° and at a rate of 30° per second.

In the sea search SCR-717 equipment, selsyn azimuth position data take-offs are provided which drive a PPI type of indicator presentation. The rotational speed about the vertical axis in this case is either 8 or 20 rpm as selected by the operator. The reflector can also be tilted about its horizontal axis above or below the line of flight as desired by the operator.

It will be noted that the emitter moves with the reflector and accordingly it is always located at the focal point throughout all orientations of the antenna.

16.3 *The AN/APQ-7 Radar Bombsight Antenna*²¹

Early experience in the use of bombing-through-overcast radar equipment indicated that a severe limitation in performance was to be expected as the result of the inadequate resolution offered by the then available airborne radar equipments. This lack of resolution accounted for gross errors in bombing where the target area was not ideal from a radar standpoint.

To meet this increased resolution requirement in range, the transmitted pulse width was shortened considerably. In attempting to increase the azimuthal resolution, higher frequencies of transmission were employed. This enabled an improvement in azimuthal resolution without resorting to larger radiating structures, a most important consideration on modern high speed military aircraft.

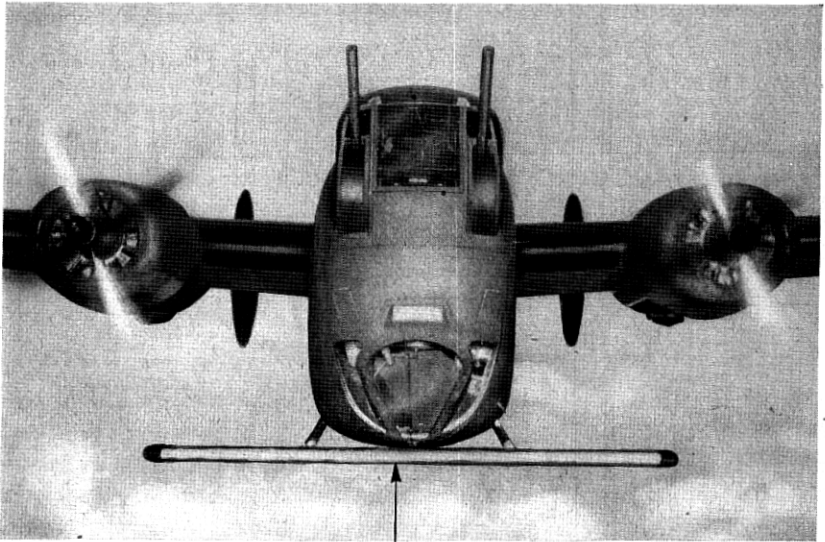
To extend the size of the radiating structure without penalizing the aircraft performance, the use of a linear scanning array which would exhibit high azimuthal resolution was considered. This array was originally conceived in a form suitable to mount within the existing aircraft wing and transmit through the leading edge. As development proceeded, the restrictions imposed on the antenna structure as well as the aircraft wing design resulted in the linear array scanner being housed in an appropriate separate air foil and attached to the aircraft fuselage (Fig. 62).

The above study resulted in the development of the AN/APQ-7 radar equipment, operating at the X-band of frequencies.²² This equipment provided facilities for radar navigation and bombing.

The AN/APQ-7 antenna consisted of an array of 250 dipole structures spaced at $\frac{1}{2}$ wavelength intervals and energized by means of coupling probes extending into a variable width waveguide. The vertical pattern was arranged to exhibit a modified csc^2 distribution by means of accurately shaped "flaps" attached to the assembly.

²¹ Written by L. W. Morrison.

²² A large part of the antenna development was carried out at the M. I. T. Radiation Laboratory.



ANTENNA AIRFOIL ASSEMBLY

Fig. 62—AN/APQ-7 Antenna Mounted on B24 Bomber.

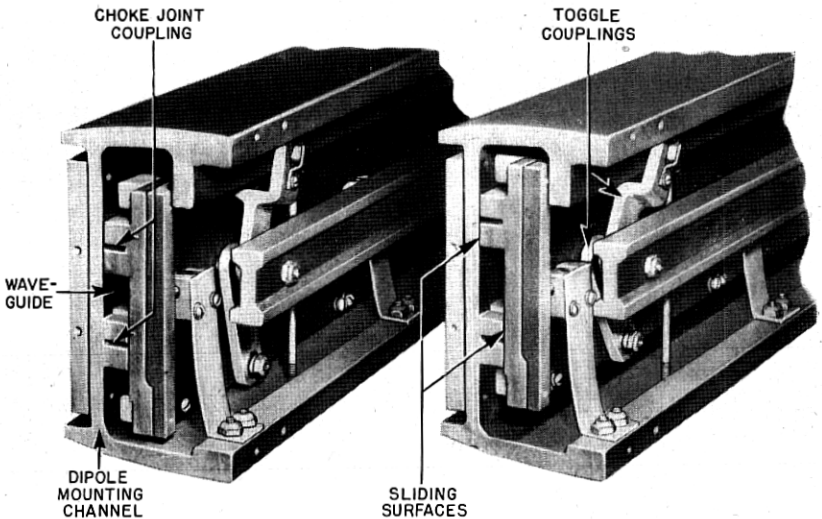


Fig. 63—AN/APQ-7 Antenna. Left—Contracted Wave Guide Assembly. Right—Expanded Wave Guide Assembly.

The scanning of the beam is accomplished by varying the width of the feed waveguide. This is accomplished by means of a motor driven actuated cam which drives a push rod extending along the waveguide assembly back

and forth. Toggle arms are attached to this push rod at frequently spaced intervals which provides the motion for varying the width of the waveguide while assuring precise parallelism of the side walls throughout its length (Fig. 63).

The normal range of horizontal scanning exhibited by this linear array, extends from a line perpendicular to the array to 30° in the direction of the feed. By alternately feeding each end, a total scanning range of $\pm 30^\circ$ from the perpendicular is achieved. Appropriate circuits to synchronize the indicator for this range are included.

The use of alternate end feed on the AN/APQ-7 antenna requires that the amount of energy fed to the individual dipoles is somewhat less than if a single end feed is employed.

The AN/APQ-7 antenna is $16\frac{1}{2}$ feet in length and weighs 180 pounds exclusive of air foil housing.

The following data applies:

Gain = 32.5 db
Horizontal beamwidth = 0.4°
Vertical beam characteristic = modified csc²
Scan—Array scanning
Scanning Sector— $\pm 30^\circ$ Horizontal
Scanning Rate = $45^\circ/\text{second}$

ACKNOWLEDGMENTS

Contributors to the research and development of the radar antennas described in this paper included not only the great number of people directly concerned with these antennas but also the many people engaged in general research and development of microwave components and measuring techniques. A complete list of credits, therefore, will not be attempted.

In addition to the few individuals mentioned in footnotes throughout the paper, the authors would like to pay special tribute to the following co-workers in the Radio Research Department: C. B. H. Feldman who with the assistance of D. H. Ring made an outstanding contribution in the development of the polyrod array antenna; W. A. Tyrrell for his work on lobe switches; A. G. Fox, waveguide phase changers; A. P. King, paraboloids and horn antennas; A. C. Beck, submarine antennas; G. E. Mueller, polyrods.