THEORETICAL BASIS FOR EXPERIMENTAL DEVELOPMENT TECHNIQUES FOR AIRCRAFT ANTENNA DESIGN

By

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THEORETICAL BASIS FOR EXPERIMENTAL DEVELOPMENT

TECHNIQUES FOR AIRCRAFT ANTENNA DESIGN

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PREFACE

The subject of this thesis was chosen primarily because the author felt a special need for such an exposition in his work at the Engineering Radiation Laboratory, Douglas Aircraft Company, Inc., Tulsa, Oklahoma. Experimental work has been the basis of considerable progress since the time of Newton. Results of experimental measurements have confirmed theories, destroyed theories, and inspired the formulization of new theories. We have been very fortunate to inherit extremely good procedures for various experimental measurements used for obtaining design data. However, we have come to rely entirely too much on formulated procedure, and are inclined to forget the theoretical basis for the procedures. This thesis presents the theory, from an engineering viewpoint, necessary for a qualitative understanding of the experimental measurements associated with design and development of aircraft antennas. The material used in preparation of this thesis was collected from a variety of sources, most of which have not been published; therefore, the selected bibliography reflects only those published references which were of general use.

I wish to acknowledge Dr. H. L. Jones of Oklahoma State University for his guidance and encouragement, G. A. O'Reilly and H. H. Williams of Douglas Aircraft Company, Inc. for their inspiring discussion on the general subject presented herein, and the Douglas Aircraft Company, Inc. for their financial assistance throughout my graduate study program.

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CHAPTER I

INTRODUCTION

The rapid progress of the aircraft industry since World War II has brought about an increased demand for specialization in fields formerly considered foreign to the aircraft industry. Electronic industries, for example, have become so closely intergrated with aircraft enterprises that some of the leading aircraft companies have made significient contributions to the state of the art. At the present each industry contributes to the advancement and progress of the other by each creating new requirements for the other.

With the modernization of aircraft electronic equipment has come a greater need for improved connecting links between the aircraft equipment and the support equipment remotely located from the aircraft. These links are necessarily made by electromagnetic waves in free space. The region of transistion between the guided waves to or from the electronic equipment and free space is called the antenna.

In the past far too little emphasis was placed on the importance of the antenna. Aircraft engineers considered the antenna a necessary evil which could be "hung" on the airplane at any convenient mechanical location. Little regard was given to the possibility that the location chosen might seriously impair the electrical performance of the antenna. As the number of antennas required on certain military airplanes increased, the aircraft designers became aware of the importance of coordinated design of the airframe and the antenna.

This thesis describes the major principles involved in design and development of aircraft antennas. No attempt is made to present design details because they are usually applicable to only specific antennas. The principles presented herein are generally applicable to the entire electromagnetic specturum, from 25mc to 35,000mc, used by the military aircraft. The experimental test methods are equally applicable to antennas other than aircraft antennas, but emphasis is placed on the aircraft useage. The term aircraft is assumed to include space-craft as well as airplanes.

The purpose of this thesis is to present the relation between the theoretical background and the development techniques along with sufficient definitions to give the reader a comprohensive concept of some of the problems encountered by the aircraft antenna designer. The main text of the thesis is composed of six chapters. Chapter II points out the more important design considerations. It is followed by a chapter on scale model techniques which gives the basis of the scale model measuring techniques universally accepted as an accurate method of measuring radiation patterns. Chapters IV and V are devoted primarily to impedance measurements and radiation pattern measurements but they include definitions of the various parameters associated with impedance and radiation patterns. Chapter VI deals briefly with the methods of handling elliptically polarized antenna problems. The concluding chapter summarizes the principles presented in the previous chapters by outlining the procedure followed by the author in developing simple aircraft antennas.

CHAPTER II

DESIGN CONSIDERATIONS

Mechanical

The modern aircraft is ordinarily a high speed vehicle which has velocities ranging from mach 0.5 to mach 3 for airplanes and much higher for guided missiles. Accelerations can be extremely high. Associated with the high velocities are skin temperatures that may be as high as 500°F or more. Obstacles protruding into air stream which have insufficient thermoconductance to the heat sink can have temperatures which exceed the melting point of copper. The great accelerations cause exceedingly high bending moments and shear strain on structual members cantilevered out at right angles to the direction of acceleration.

The antenna designer has the problem of creating a device with certain specified electrical characteristics that will withstand the required mechanical strains. The antenna must not impair the dynamic behavior of the aircraft, and aerodynamic drag must be minimized. Finally, the antenna should be light weight. Flush mounted antennas are the answer to many of the mechanical problems, but often the electrical requirements will not permit flush mounting. A compromise then has to be made.

Electrical

Airborne transmitting and receiving equipment are designed, with few exceptions, to have an input or output impedance of 50 ohms. This impedance matches the widely used coaxial cables such as RG-9B/U, RG-8A/U, and RG-55A/U which have characteristic impedances ranging between 50 and 53 ohms.

The voltage standing wave ratio is a figure of merit used to indicate how well an antenna matches its transmission line. The VSWR is a real number, one or greater, the value of which depends on the degree of mismatch. This value is numerically defined by

$$\rho = \frac{|\Gamma| + 1}{|\Gamma| - 1}, \tag{1}$$

where ρ is the VSWR and $|\Gamma|$ is the magnitude of the voltage reflection coefficient. The inverse relationship is evident

$$|\Gamma| = \frac{\rho - 1}{\rho + 1} \tag{2}$$

The reflection coefficient is related to impedance by

$$\Gamma = \frac{Z_1 - Z_o}{Z_1 + Z_o}, \tag{3}$$

where Z_1 is the load or antenna impedance and Z_o is the characteristic impedance of the transmission line. The reflection coefficient is also defined as the ratio of the voltage of a wave reflected from the load to that incident on the load.

Some of the reasons antenna VSWR should be kept as low as possible are: 1) Efficiency is reduced because the energy, which is proportional to Γ^2, reflected from the mismatched load is mostly wasted; 2) The reflected power may seriously impair operation of a transmitter or even burn it out; 3) Power handling capacity of a transmission line is reduced because of hot spots at current maximum points and because of arc-over at voltage maximum points; 4) The "long line effect" can make it impossible to tune a given transmitter to its prescribed frequency.

Receiving equipment is affected only by the reduction in efficiency. For this reason the maximum allowable VSWR is usually higher for receiving antennas than for transmitting antennas. The general military specification for airborne antennas requires the VSWR of receiving antennas to be 5:1 or less and for transmitting antennas, 2:1 or less.

Impedance matching is one of the most difficult problems for the antenna designer. The distributed character of the parameters which determine the antenna impedance make the equivalent circuit somewhat more complicated than a simple RLC network. Like transmission lines, however, good use can be made of a simple equivalent circuit at or near resonance. Thus, if an antenna is at a minimum impedance resonant point, the equivalent circuit is that of a series RLC network. When frequency is increased, the antenna becomes inductive, and when frequency is decreased, the antenna becomes capacitive. Likewise, if the antenna is at a maximum impedance resonance the equivalent circuit is equivalent to a parallel RLC circuit. The rate at which the antenna impedance moves away from its resonant resistance when the frequency is varied from resonance is a function of the equivalent "Q" of the antenna. If the equivalent "Q" of the antenna is very low, the antenna is broad band; whereas, if the equivalent "Q" is high, the antenna is narrow band.

Design of narrow band antennas is usually less complicated than design of broad band antennas. If the initial VSWR of a narrow band antenna is less than 10:1, an efficient matching network is easily designed. For VHF and higher frequencies, a specific length of transmission line having proper characteristic impedance is used for a matching section. At lower frequencies lump constant L, T, or π sections can be used.

The problem is more complicated for broad band antennas. A broad band impedance matching device cannot be designed except in special cases. An example for one special case is an exponential tapper from one resistance level to another, such as a 600 ohm resistive load matched to 50 ohms. The impedance of an antenna is not usually a simple resistance, but instead, a complicated function of frequency. For this reason, broad band antennas must inherently have broad band characteristics. In other words the equivalent "Q" of the antenna must be low.

The antenna radiation pattern, which is a plot of the antenna's directional properties, is a very important electrical characteristic. For transmitting antennas, it gives the relative power density radiated in any direction, and for receiving antennas, it gives the relative effective receiving cross section area for plane waves impinging on the antenna from any direction. The receiving pattern of an antenna is identical in shape to its transmitting pattern. Occasionally the antenna pattern is plotted in terms of relative field strength instead of relative power density; however, one can always be found from the other. The magnitude of each point on the power pattern can be calculated by squaring the magnitude of the corresponding point on the field strength pattern.

Fig. 1 represents a spherical coordinate system in free space with a transmitting antenna located at its center. The radius of the sphere for observation is many times greater than the largest dimension of the antenna so that the antenna appears as a point compared to the sphere. The sphere is also large compared to the wave length of the

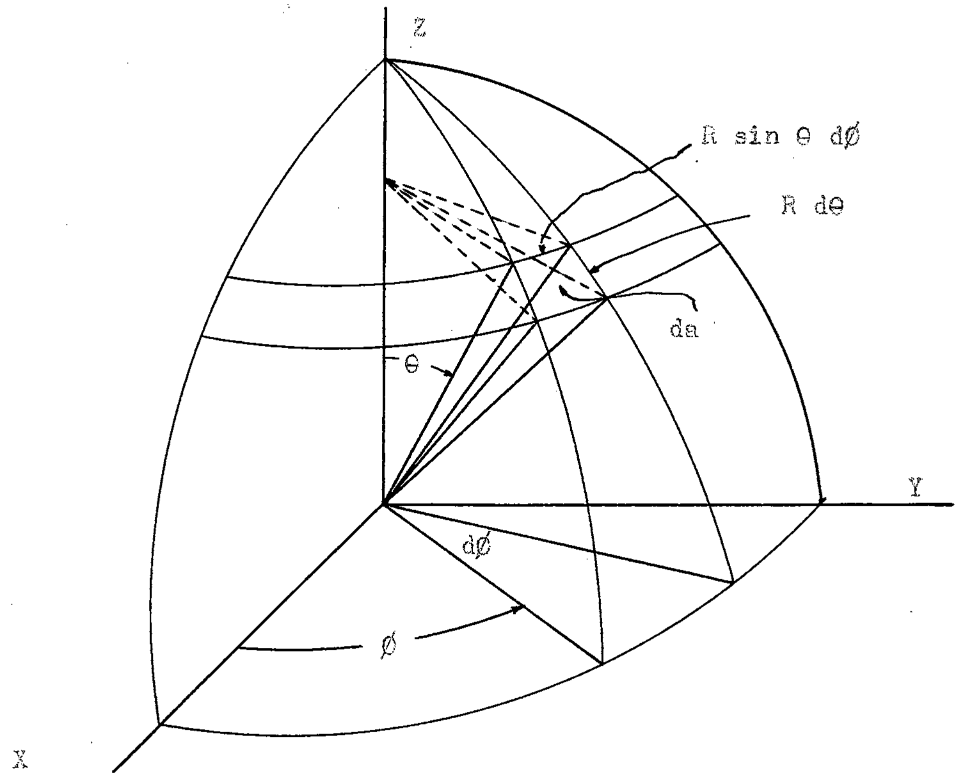

Fig. 1--Spherical coordinates for a radiation source in free space.

radiated energy from the antenna. The energy appears to stream from the source in radial lines. The time rate of energy flow per unit area is the Poynting vector or power density vector. Since the surface of the observation sphere is at a great distance from the source, the radiated waves can be considered plane over a small cross section. The Pounting vector then has only a radial component P_r, with no components in either the θ or $\emptyset$ direction. Thus, the magnitude of the Poynting vector is equal to the radial component ($|\bar{F}| = P_r$).

The graph of P_r or KP_r, where K is a constant, as a function of either θ or ϕ is called a specific power pattern. The pattern as a function of but one of the angles is only one small part of the total radiation pattern. The total radiation pattern is a three dimensional plot of P_r as a function of both θ and ϕ. The total power pattern for an isotropic source is a sphere. Any specific power pattern for the isotropic source is a circle with origin at the center.

The radiation pattern of a given antenna is most useful when it is referred or normalized to the hypothetical isotropic pattern. The radiation pattern so normalized is called the directivity function pattern. The maximum point on the directivity function pattern is called the antenna directivity(D_o). Mathematically D_o is defined as the ratio of maximum power density radiated, at a given distance from the source, to the average power density at that distance. The average power density is given by

$$P_r(av) = \frac{1}{4\pi R^2} \iint P_r(\theta,\phi)\ da, \qquad (4)$$

where R is the radius of the sphere for observation. Then

$$D_o = \frac{4\pi R^2\ P_r(max)}{\iint P_r(\theta,\phi)\ da} = \frac{4\pi\ P_r(max)}{\iint P_r(\theta,\phi)\ \sin\phi\ d\theta\ d\phi}, \qquad (5)$$

where $P_r(\theta,\phi)$ is the power density in the direction specified by θ and ϕ;

θ is the angle between the Z axis and the radius R;

ϕ is the angle that the projection of R on the X-Y plane makes with the X axis.

The radius R drops out because $da = R^2 \sin\theta \, d\theta \, d\phi$. The directivity function is then given by

$$D(\theta,\phi) = \frac{4\pi \, P_r(\theta,\phi)}{\iint P_r(\theta,\phi) \sin\theta \, d\theta \, d\phi}. \tag{6}$$

The gain of an antenna is very closely associated with the directivity, except it considers the efficiency of the antenna. Gain can be expressed either with respect to a known antenna or with respect to the hypothetical lossless isotropic antenna. The latter is called the absolute gain. The absolute gain and gain function are related directly to directivity and directivity function by the efficiency. Thus

$$G(\theta,\phi) = KD(\theta,\phi). \tag{7}$$

The directivity and gain functions have meaning only when applied to transmitting antennas; however, the ability of an antenna to absorb energy from a plane wave is directly related to its gain and directivity. The factor determining the antenna's receiving ability is its effective cross section area $A(\theta,\phi)$ or effective aperature. The relation of effective aperature to the gain function, when the antenna impedance is matched to the receiver, is given by

$$A(\theta,\phi) = \frac{G(\theta,\phi) \; \lambda^2}{4\pi}, \tag{8}$$

where λ is wave length, in free space, of the frequency received.

The gain function together with the effective aperature function forms the basis of the very important range equation

$$P_R = \frac{P_T G_T G_R \lambda^2}{16\pi^2 \; R^2}\,, \tag{9}$$

where P_R is the power received by a matched receiver in watts;

P_T is the power fed into the transmitting antenna in watts;

G_T is the gain of the transmitting antenna;

R is the distance between the transmitting and receiving antennas in same units as λ.

Equation (9) is more easily understood by considering the significance of various combinations of its factors. The factor $P_T G_T / 4\pi R^2$ is the power density of the radiated wave in space at the receiving antenna in watts per square unit, and $G_R \lambda^2 / 4$ is the effective aperature, in square units, of the receiving antenna in the direction of arrival of the signal.

For point to point communications, it is usually desirable to have very high gain transmitting and receiving antennas, so that little energy is radiated in directions other than toward the receiving antenna, and the receiving antenna's effective aperature is as large as possible in the direction of the transmitting antenna. The requirements for aircraft antennas, however, can be quite different. An antenna is a passive device; therefore, broad pattern coverage and high gain are incompatable requirements. One has to be sacrificed for the other.

Thus, the directional requirements for an antenna must be specified; then, the antenna is designed to have maximum gain in those required directions.

Polarization is a fundamental property of an antenna just as it is a fundamental property of an electromagnetic wave. The polarization of an antenna is the same as that of the electromagnetic field radiated from it. The polarization is specified by the direction of electric field. To simplify analysis the electric field is resolved into two orthogonal components. The plane wave solution of Maxwell's wave equation requires that the direction of polarization be normal to the direction of propagation. The R component of polarization is zero, leaving only the θ and ϕ components. If E_θ and E_ϕ, the amplitude of the orthogonal components, have a time phase difference, the polarization is elliptical.

The forementioned equations for gain and directivity have not considered polarization. The equations have been based on total power density. It is often desirable to include polarization in gain and directivity equations. Equation (6) then becomes

$$D(\theta,\phi,p) = \frac{4\pi\, P(\theta,\phi,p)}{\iint P(\theta,\phi) \sin\theta \, d\theta \, d\phi}, \qquad (10)$$

where the new parameter p is the polarization considered. For linearly polarized antennas, p is defined by

$$p = \tan^{-1} E_\theta / E_\phi. \qquad (11)$$

A method of handling elliptically polarized antenna problems is given in Chapter VI.

CHAPTER III

SCALE MODEL ANTENNAS

Because of the complicated nature of even the simplest type of antenna, it is desirable to do a great deal of experimental work when developing an antenna. For aircraft antennas a complete theoretical design is extremely difficult, if not impossible, because of the many parameters which cannot be evaluated. Even when theoretical design is practical, it is necessary to verify the theory be experimental measurement.

Fortunately, experimental measurements can be simplified by use of scale model antennas. It is well known that the impedance of a thin perfect conducting half-wave dipole in free space is approximately 73 ohms, regardless of the wave length. The radiation pattern is also the same for any wave length so long as the length of the antenna is one-half wave length. The fact that both antenna radiation pattern and impedance depend upon ratios of the physical dimensions to wave length suggests that measurements made from scale model antennas at scale wave lengths would be accurate representations of full scale measurements.

The proper scale factor may be obtained from Maxwell's equations;

$$\text{Curl } E = -j\omega\mu H, \quad \text{Curl } H = (g+j\omega e)E. \tag{12}$$

The various quanities in a model will be denoted by primes; then,

$$\text{Curl}' \, E' = -j\omega'\mu' H', \quad \text{Curl}' \, H' = (g'+j\omega' e')E', \tag{13}$$

where the prime after the curl denotes differentitation with respect to the scaled coordinates. For the measured parameters of the scale model to accurately represent those for full size, it is necessary that the equations of (12) and (13) be equal; thus,

$$\text{Curl}' E' = \text{Curl } E, \text{ and Curl}' H' = \text{Curl } H. \tag{14}$$

The following scale factors will be used to relate the quantities of (12) and (13):

$$E' = K_E E, \quad H' = K_H H, \quad \omega' = K_\omega \omega, \quad g' = K_g g, \text{ and } e' = K_e e; \tag{15}$$

then,

$$\left.\begin{aligned} j\omega'\mu' H' &= K_\omega K_\mu K_H \, \omega\mu H \\ (g' + j\omega' e')E' &= (K_g g + jK_\omega K_e \omega_e)\, K_E E. \end{aligned}\right\} \tag{16}$$

From (16), (14), and (13)

$$K_\omega K_\mu K_H = 1; \quad K_\omega K_e K_E = 1; \quad K_E K_g = 1;$$

then,

$$K_\omega K_E = \frac{1}{K_e}; \quad K_\mu K_\omega = \frac{1}{K_H}; \quad K_g = \frac{1}{K_E}. \tag{17}$$

In practice K_e and K_μ must be unity because both the full size and scale model antennas operate in the same medium(near free space condition). This means that the dielectric constants and relative permeability of material in the scale model antenna must be the same at the scaled frequency as those for the full sized antenna at the full scale frequency. In general, the dielectric constant of most insulating materials vary with frequency; therefore, the dielectric material used in the scale model is often different from that used for the full scale antenna.

K_E and K_H are equal to each other, and they are also equal to the scale factor for the model antenna dimensions. These relationships are proved by substituting the equation from (15) into (14). Thus,

$$K_E \text{ Curl}' E = \text{Curl } E, \quad K_H \text{ Curl}' H = \text{Curl } H. \tag{18}$$

Then the vector operator

$$\nabla' = \frac{1}{K_E}\nabla = \frac{1}{K_H}\nabla, \quad K_E = K_H, \tag{19}$$

where ∇ and ∇' are related to Curl E and Curl' E by

$$\text{Curl}' E = \nabla' X E; \text{ Curl } E = \nabla X E;$$

and,

$$\nabla' = i\frac{\partial}{\partial x'} + j\frac{\partial}{\partial y'} + k\frac{\partial}{\partial z'}; \tag{20}$$

$$\nabla = i\frac{\partial}{\partial x} + j\frac{\partial}{\partial y} + k\frac{\partial}{\partial z};$$

also,

$$\nabla' = i\frac{\partial}{\partial x}\frac{\partial x}{\partial x'} + j\frac{\partial}{\partial y}\frac{\partial y}{\partial y'} + k\frac{\partial}{\partial z}\frac{\partial z}{\partial z'}. \tag{21}$$

Now if

$$x' = K_c x, \quad y' = K_c y, \text{ and } z' = K_c z, \tag{22}$$

then

$$\frac{\partial x}{\partial x'} = \frac{\partial y}{\partial y'} = \frac{\partial z}{\partial z'} = \frac{1}{K_c}; \tag{23}$$

therefore,

$$\nabla' = \frac{1}{K_c}\nabla. \tag{24}$$

From equations (18) and (24),

$$K_c = K_E = K_H. \tag{25}$$

The relation of all the scale factors have been found; now it will be shown how they can be applied in a practical situation. It has been previously stated that K_e and K_H should be unity. The equation of (17) can be simplified by substitution of (25). Then

$$K_\omega = \frac{1}{K_c}; \quad K_g = \frac{1}{K_c}. \tag{26}$$

Thus, if a scale model antenna is 1/25 full size, the model frequency will be 100 times the full scale frequency. The conductance of all parts of the scale model should also be increased by the same factor; however, when the conductivity of antenna conductors is very high and that of the insulating material is extremely low, the scaling of the conductance can be omitted with little loss in accuracy. This is evident from the extreme values of $g=\infty$ for the conductors and $g=0$ for the insulators. The scaled model impedance measurements are more seriously affected by above assumptions than the radiation pattern measurements.

A scale model of the aircraft must be constructed for use when measuring the radiation patterns of the model antenna because the entire aircraft surface should actually be considered a part of the antenna. The models are usually constructed of a wood shell having a coating of good conducting material on the outside. The inside of the model is hollowed out in order to accommodate the receiving equipment required for radiation pattern measurement. The conducting coating is often put on by a metal-flame-spraying process. In this process, fine metal particles(usually powdered copper) is passed through a hot flame and blown, in a finely divided state, onto the wood model. The thickness of the coating is about 0.010 inch. Silver paint, also, has been used with reasonable success, but it has disadvantages. Most other metalic paint has too low of conductivity to be used with any degree of accuracy.

CHAPTER IV

IMPEDANCE MEASUREMENTS

At power line frequencies, the wave lengths are extremely long (5×10^6 MTRS at 60 cycles). The finite velocity of electromagnetic energy propagation is an important consideration only for very long transmission lines. Thus, the impedance of a load can be measured through a length of connecting wires without consideration for the slight phase shift caused by the finite velocity of propagation.

For frequencies in the VHF band and higher, the finite velocity of propagation in the transmission line must be considered when measuring impedances. The lowest frequency in the VHF band is 30mc; the wave length in free space is 10 meters. The velocity factor for a coaxial cable having polyethylene dielectric is approximately 0.66. The wave length for 30mc in the cable is $10 \times 0.66 = 6.6$ meters. The maximum transmission line length allowable without correction is from 0.01 to 0.05 wave lengths depending on the application. For 30mc the maximum cable length would be between 6.6 and 33cm.

The true impedance of the antenna is easily found from the measured impedance if the electrical lengths and attenuation of the transmission line are known. A Smith chart or similar impedance chart is usually used for making the line length correction to eliminate tedious calculations involving the transmission line equations.

At 3000mc and higher the wave length in the cable is so short that the impedance of an antenna has but little meaning. The terminals of the antenna cannot be clearly defined. The length of transmission line corresponding to .05 wave lengths at 3000mc is 0.33cm. Here, for impedance to have meaning, the antenna terminals must be specified at a particular point within the connector. The VSWR that the antenna causes on its transmission line is a more important factor because it is not a function of the transmission electrical length. It is a function of transmission line length only when the line has attenuation loss present.

When the transmission line has attenuation loss, the VSWR varies exponentially with distance from the antenna. The attenuation of the antenna connector is very small and the antenna VSWR is defined at the input to the connector.

It should be pointed out here that VSWR can be defined at a point, and the transmission line is not required to have a voltage maximum and minimum on it. This might seem contrary to the commonly used definition that the VSWR is equal to ratio of maximum to minimum voltage along the transmission line. The descrepancy is clearified by revising the definition as follows:

> The Voltage Standing Wave Ratio(VSWR) of any load with respect to a transmission line having a given characteristic impedance is equal to the ratio of the maximum to minimum voltage on a hypothetical lossless transmission line at least one-half wave length long having the same characteristic impedance.

The antenna VSWR, using the revised definition, is clearly a function of antenna impedance and the transmission line characteristic impedance as given by equations (1) and (3).

In general, the antenna VSWR cannot be measured at the antenna terminals. The measured VSWR will be less than the antenna VSWR by a factor which depends on the attenuation of the transmission line connecting the antenna to the measuring device.

A correction for line loss must be made to obtain the required accuracy. The correction is made by first determining the attenuation characteristics of the cable and then applying this factor to the measured VSWR to obtain the true VSWR of the antenna.

In Fig. 2 the transmission line is assumed to be terminated in a perfect short circuit.

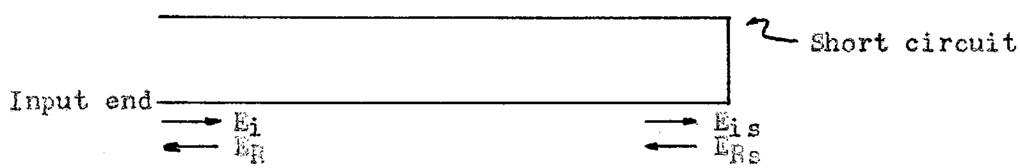

Fig. 2--Incident and reflected voltages on a short circuited transmission line.

The VSWR measured at the input end of the cable is

$$S_i = \frac{|E_i| + |E_R|}{|E_i| - |E_R|}, \tag{27}$$

and the magnitude of the reflection coefficient is

$$|\Gamma_i| = \frac{|E_R|}{|E_R|} = \frac{S_i - 1}{S_i + 1}, \tag{28}$$

where $|E_i|$ is the magnitude of the incident voltage wave;

$|E_R|$ is the magnitude of the reflected voltage.

The incident voltage at the short circuit, E_{is}, is given by

$$E_{is} = \alpha E_i , \tag{29}$$

where α is the attenuation factor for the transmission line. The short can be made nearly perfect; therefore,

$$\frac{E_{Rs}}{E_{is}} = 1 , \tag{30}$$

where E_{Rs} is the amplitude of the reflected wave at the short circuit. Since the transmission line is a bilateral,

$$|E_R| = \alpha |E_{Rs}| ; \tag{31}$$

$$|E_R| = \alpha |E_{is}| = \alpha^2 |E_i| ; \tag{32}$$

$$\alpha^2 = \frac{|E_R|}{|E_i|} = \frac{S_i - 1}{S_i + 1} . \tag{33}$$

After α is determined the short circuit is replaced by the antenna to be measured. The incident voltage at the antenna is

$$|E_{ia}| = \alpha |E_i| . \tag{34}$$

The reflected voltage is

$$|E_{Ra}| = |\Gamma_a| \; E_{ia} , \tag{35}$$

where $|\Gamma_a|$ is the magnitude of the antenna reflection coefficient;

$$|E_R| = \alpha^2 |\Gamma_a| \; |E_i| ; \tag{36}$$

$$|\Gamma_a| = \frac{1}{\alpha^2} \frac{|E_R|}{|E_i|} = \frac{1}{\alpha^2} \frac{S_m - 1}{S_m + 1} ; \tag{37}$$

then,

$$S_a = \frac{\alpha^2 \, (S_m + 1) + S_m - 1}{\alpha^2 \, (S_m + 1) + 1 - S_m} \,, \tag{38}$$

where S_a is the corrected standing wave ratio of the antenna;

S_m is the measured VSWR.

Combining (38) with (33) gives the desired relation of the antenna VSWR in terms of the measured VSWR and the short circuited cable VSWR:

$$S_a = \frac{\dfrac{S_{sc} - 1}{S_{sc} + 1} \, (S_m + 1) + S_m - 1}{\dfrac{S_{sc} - 1}{S_{sc} + 1} \, (S_m + 1) + 1 - S_m} \,. \tag{39}$$

The procedure for correcting impedance measurements for cable attenuation is essentially the same as that for VSWR measurements. The impedance at the input of the transmission line can be measured with a slotted line or some form of impedance bridge. The impedance is corrected for line length, the same as though the transmission line were lossless. The impedance is plotted on the Smith chart; then a constant VSWR circle is drawn through the point and its value noted. The corrected VSWR is calculated by equation (39) and drawn on the chart. The point of intersection of a line, drawn from the center of the chart through the measured impedance, with the corrected VSWR circle is the corrected impedance.

CHAPTER V

MEASUREMENTS OF RADIATION PATTERNS

In this chapter the term "radiation pattern" will be applied to any plot of field intensity about the antenna as a function of any of the coordinate system angles. Each radiation pattern is accompanied by sufficient constant information of the system parameters to determine the specific type of pattern.

The antenna under test is usually used as a receiving antenna when its pattern is being measured, because of the simplicity of the test receiver compared to the transmitter. Since the receiving and transmitting patterns are identical, the test antenna will be referred to as either a transmitting or a receiving antenna as necessary to simplify the discussion.

The coordinate system used for the aircraft antenna is a spherrical coordinate system oriented with respect to the aircraft without regard to the location of the antenna. Fig. 1 is an example of such a coordinate system. This system has been adopted by most aircraft companies for use with airplanes. The Z axis coinsides with the vertical axis of the airplane. The X axis coinsides with the fore-aft axis of the airplane, and is directed forward. The angle ϕ then becomes an azimuth angle and θ becomes an elevation angle.

The choice of the coordinate system in Fig. 3 is logical for airplanes because an airplane, in normal flight attitude, moves in a plane parallel to the ground. Thus, an azimuth radiation pattern is easily referred to the ground reference.

The coordinate system is usually oriented differently for missiles

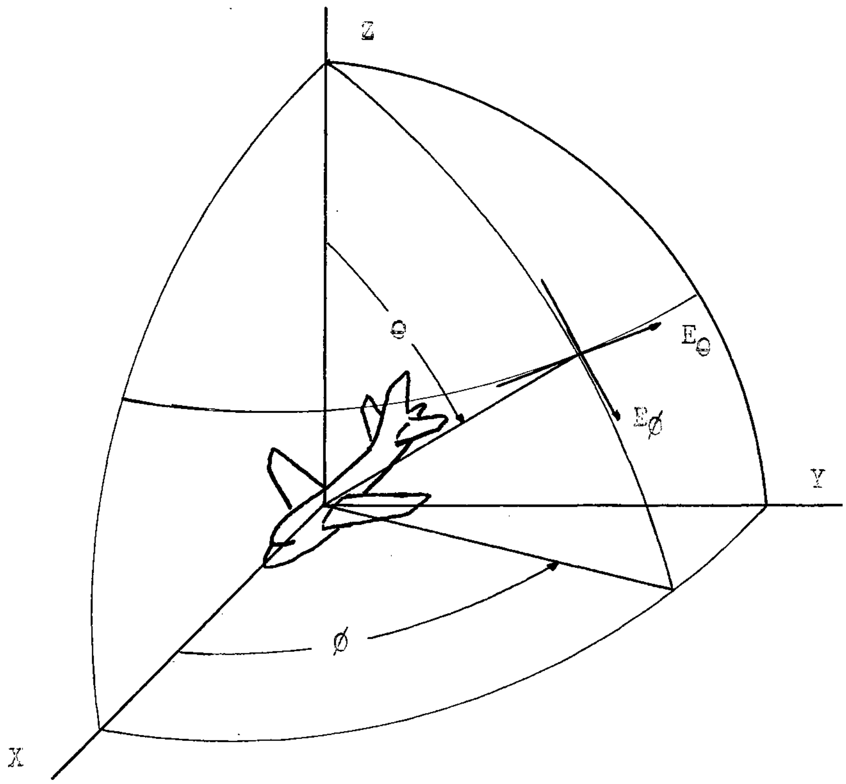

Fig. 3--Airplane coordinates for radiation pattern measurements.

because the missile does not necessarily have a fixed attitude with respect to the earth's surface. In any case, the coordinate system is referred to the vehicle carrying the antenna and not the antenna.

There are several radiation pattern cuts that are used in evaluating an antenna. The conic and major plane cuts are the most useful. There can be any number of conic cuts. The conic cut is a radiation pattern made at a constant lattitude; i.e., Θ = constant. There are only three major plane cuts; these are radiation patterns along great circles in the three coordinate planes.

The polarization of the plane wave used in measuring any radiation pattern must be indicated for the pattern to have meaning. Thus, a conic cut pattern could be made, for example at $\theta=100^{\circ}$, for E_{θ}, E_{ϕ}, or any polarization between E_{θ} and E_{ϕ}. Because of the close relation of E_{θ} and E_{ϕ} to vertical and horizontal polarization, respectively, with the airplane coordinate system, most pattern cuts for airplane antennas are made with E_{θ} and E_{ϕ} polarization.

Often E_{θ} polarization is mistakenly referred to as vertical polarization. Actually, an airplane antenna in flight can receive truly vertically polarized signals from the ground only in very special cases. Consider an airplane flying directly over a ground transmitting antenna. According to the wave equation, the polarization must be normal to the direction of propagation. Since the airplane was directly above the transmitting antenna the direction of propagation was vertical; therefore, any electromagnetic energy received in the airplane from the ground transmitter must have been horizontally polarized. Since the coordinate system used is spherical, an attempt to describe the polarization by a two dimensional rectangular system is erroneous.

The transmitting equipment used for a radiation pattern measurement is usually fixed in position. Energy radiated from the transmitting antenna illuminates the test antenna and the entire model aircraft in which the antenna is installed with a nearly plane wave. Scale model techniques are used so that the aircraft model can be easily rotated by a simple rotational fixture. Rotation of the model rotates the coordinate system also. This is equivalent to a stationary coordinate system and moveable observation point.

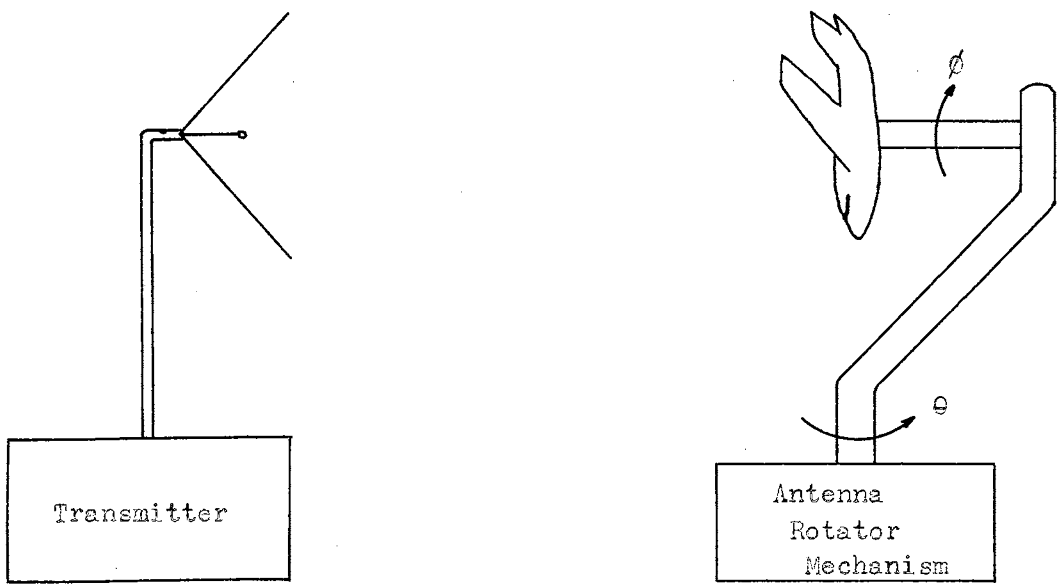

Fig. 4--Antenna pattern measuring range.

The rotational fixture, illustrated in Fig. 4, has two degrees of freedom so that all angles of Θ and Ø can be observed.

Synchro-transformers are geared to both the head and arm so that either Θ or Ø angular position information can be used to drive an automatic recorder such as the Airborne Instruments Laboratory Type 116R Polar Recorder.

A typical transmitter set-up for antenna pattern measurement is depected in Fig. 5. At the lower model frequencies, calibrated signal generators are usually available, but for higher frequencies, reflex klystron oscillator tubes with necessary power supplies, modulators, etc., are in wide use. The frequency range covered in a well equipped scale model antenna pattern range is from about 300 to 35,000mc. The R.F. power output from the signal source should be at least 100 milli-

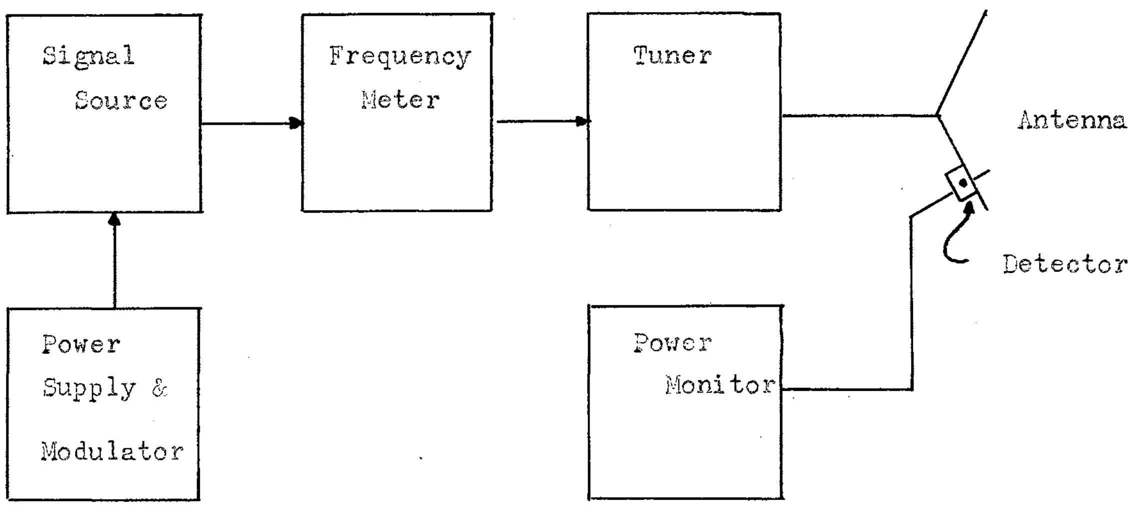

Fig. 5--Transmitter set-up for antenna radiation pattern measurements.

watts. The R.F. signal is amplitude modulated with an audio frequency so that the output from crystal receiver in the model will be an amplitude modulated audio signal.

The amplitude of the audio signal is proportional to power received by the antenna. The modulation frequency is often controlled by a tuning fork. This allows the use of a very narrow band(2 to 10 cycles) audio filter to be used in the signal channel of the recorder, which greatly reduces the noise. It is necessary, when using the narrow band filter, to record the antenna radiation pattern very slowly so that the recorder pen will follow any deep nulls that may exist in the pattern.

The choice of transmitting antenna is very important. Two fundamental requirements for the approximated plane wave front at the test antenna are: 1) that the phase change over the projected area the model must traverse is less than some predetermined value, and 2) that amplitude variation is less than a prescribed ratio.

The receiver consists of a simple crystal detector such as a 1N23C mounted in a suitable detector mount, and a tuner to match the detector to the test antenna. The entire unit is mounted inside the model so that it does not obstruct the radiation pattern of the test antenna. The only requirement for electrical wiring from the model is an audio cable to connect from the receiver output to the signal input of the recorder. A special high resistance(approximately 2000 ohms at d-c) audio cable is often used, especially when measuring patterns of nearly omni-directional antennas. The high resistance cable reflects very little R.F. energy; therefore, the radiation pattern distortion is minimized.

The crystal detector as well as other type detectors used for pattern range receivers are square law devices. That is, the output voltage is proportional to the input R.F. voltage squared. The voltage output is thus proportional to the power input. If the recorder is a linear recorder, the deflection of the pen will be proportional to the R.F. power received by the antenna. When it is desirable to make a field strength or voltage pattern, a square root amplifier can be inserted anywhere in the audio system.

To calibrate the radiation patterns it is necessary to perform a radiation pattern intergration. Approximate intergration can be performed either automatically or manually. The accuracy in either case depends largely on the number of conic cuts made.

A transmitting antenna is assumed to be located at the center of a coordinate system, such as Fig. 1, and the power density at distance R is assumed to be S. The power flow through the elemental area da is Sda since da is normal to R. A receiving antenna located on the observation sphere will transfer power to its load from the passing wave. This power will be proportional to S. Thus,

$$P_r = KS', \tag{40}$$

where P_r is the power delivered to the load from the receiving antenna;

S' is the component of power density of the wave, transmitted from the test antenna, having the same polarization as the receiving antenna;

K is a proportionality constant consisting of the power fed to the transmitting antenna, receiving antenna gain, receiver gain, recorder sensitivity, etc.

The deflection of a chart recording pen is assumed to be proportional to P_r so that

$$D = CP_r; \qquad D = CKS', \tag{41}$$

where D is deflection of the pen;

C is the pen sensitivity constant.

The total power radiated in the polarization is

$$P_T = \iint S'\,da; \tag{42a}$$

$$KP_T = \iint KS'da = KR^2 \int_{\theta=0}^{2\pi}\int_{\phi=0}^{\pi/2} S' \sin\theta \, d\phi \, d\theta;$$

$$KP_T = KR^2 \int_0^{\pi/2} \sin\theta \left[\int_0^{2\pi} S' d\phi \right] d\theta. \tag{42}$$

S', in general, is a complicated function of both θ and ϕ. Let

$$F(\theta) = \int_0^{2\pi} S' d\phi. \tag{43}$$

Then a close approximation to (42) is

$$KP_T = KR^2 \sum_{\substack{\theta=0 \\ i=1}}^{\substack{i=n \\ \theta=\pi}} F(\theta_i) \sin\theta \Delta\theta_i. \tag{44}$$

The problem reduces to finding $F(\theta)$ for n values of θ.

The conic radiation patterns are used to find $F(\theta)$. The conic cut can be either a polar or a rectangular plot. The area under the rectangular plot is proportional to $F(\theta)$. This area can be easily measured with a planimeter.

Let L be the length of paper corresponding to 2π;

A be the measured area. Then,

$$F(\theta) = \frac{2\pi A}{CKL}; \tag{45}$$

$$KP_T = \frac{R^2 K 2\pi}{CKL} \sum_{i=0}^{i=n} A_i \sin\theta_i \Delta\theta_i. \tag{46}$$

Further simplification can be accomplished by proper choice of the angles θ_i. For small increments of $\Delta\theta$,

$$\sin\theta\Delta\theta \approx (\cos\theta); \tag{47}$$

then,

$$KP_T = \frac{2\pi R^2}{CL} \sum A_i \Delta(\cos\theta_i). \tag{48}$$

If θ is chosen so that $\Delta\cos\theta$ = constant

$$KP_T = \frac{2\pi R^2}{CL} \Delta\cos\theta \sum_{i=1}^{i=n} A_i. \tag{49}$$

Equation (49) is proportional to the total power radiated in the polarization of the receiving antenna. To complete the calibration

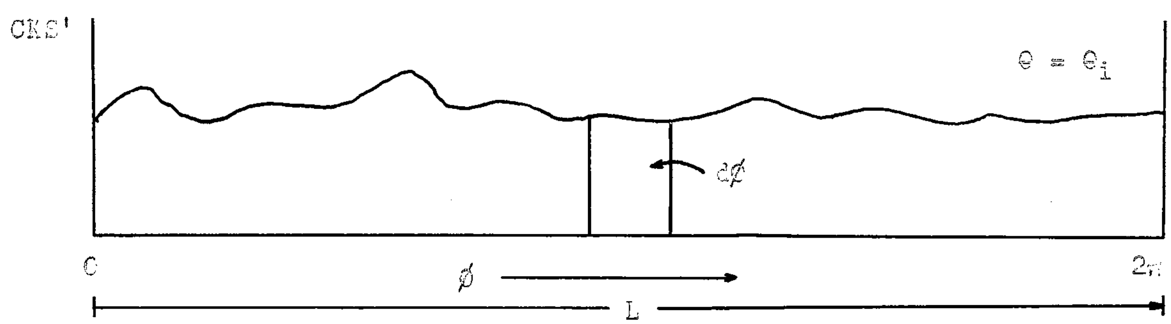

Fig. 6--Conic radiation pattern plotted in rectangular coordinates.

an intergration must be performed for an orthogonal polarization;

$$KP_T = \frac{2\pi R^2 \Delta\cos\theta}{CL} \sum A_{i\theta} + A_{i\phi}, \tag{50}$$

where $A_{i\theta}$ is the area of the E_θ conic;

$A_{i\phi}$ is the area of the E_ϕ conic;

P_T' is the total power radiated by the test antenna.

The average power density

$$S_a = \frac{\text{Total power}}{4\pi R^2}, \tag{51}$$

$$KS_a = \frac{K(\text{Total power})}{4\pi R^2}, \tag{52}$$

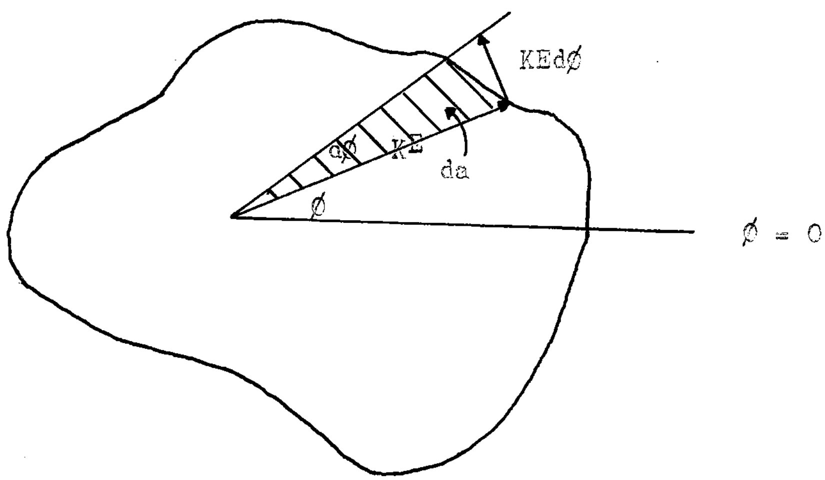

Fig. 7--Conic radiation pattern plotted in polar coordinates.

and

$$KS_a = \frac{\Delta \cos \theta}{2LC} \sum_{i=1}^{i=n} (A_{i\theta} + A_{i\phi}). \tag{53}$$

The value KS_a is the isotropic level calibration for all radiation patterns having K as a common parameter. The directivity in any direction of the test antenna can be found by the ratio of KS'/KS_a.

If a polar recorder is used for recording the radiation patterns, the deflection of the pen must be proportional to field strength when a manual pattern intergration is desired. Fig. 7 represents a polar conic radiation pattern for one polarization.

The elemental area

$$da = 1/2\ K^2E^2\ d\phi. \tag{54}$$

Then the total area of the pattern is

$$A' = 1/2\ K^2 \int_0^{2\pi} E^2\ d\phi = 1/2\ K' \int_0^{2\pi} S'\ d\phi, \tag{55}$$

where E is the field strength in volts per meter;

K and K' are proportionality constants.

Then,

$$F(\theta) = \frac{2A'}{K'}. \tag{56}$$

Substituting (56) into (44),

$$P_T = R^2 \sum_{\substack{\theta=0 \\ i=1}}^{\substack{i=n \\ \theta=\pi}} \frac{2A'_i}{K'} \sin \theta_i \Delta \theta_i. \tag{57}$$

The isotropic level should be defined in the same units as used to plot the radiation pattern. Thus, when the pattern is plotted in relative field strength,

$$\text{Isotropic level} = \sqrt{K' \frac{P_T}{4\pi R^2}}\ .$$

After making the simplification $\sin \theta \Delta\theta = \Delta\cos\theta$, and adding the power for the other polarization, the isotropic level becomes

$$\text{Isotropic level} = \sqrt{\frac{\Delta\cos\theta}{2\pi}} \sqrt{\sum_{i=1}^{i=n} A_{i\theta} + A_{i\phi}} . \qquad (59)$$

The foregoing mathematical development for manual pattern intergration can be summarized into a simple mechanical procedure:

1) First the value of $\Delta\cos\theta$ is chosen. The value of 0.1 is convenient and sufficiently accurate for most applications.

2) The conic angles are found from the value of $\Delta\cos\theta$. Thus, if

   $\Delta\cos\theta = 0.1$,

   $\theta_1 = \cos^{-1} 0.1$,

   $\theta_2 = \cos^{-1} 0.2$,

   $\theta_3 = \cos^{-1} 0.3$, etc.

3) A set of conic radiation patterns is made for the angles of θ in step 2. The set of patterns must include both polarizations, E_θ and E_ϕ. The parameter K must be constant for the full intergration set, as well as for any other patterns of interest such as the principle plane cuts.

4) The area of each conic of the intergration set is measured by a suitable planimeter.

5) The isotropic level is calculated by (53) or (59) depending on the type recorder used.

6) The radiation patterns of interest are calibrated by drawing the isotropic level on each pattern.

The process of pattern intergration is greatly simplified by use of an automatic intergrater. Unlike the manual intergration, the automatic operation uses constant increments of $\Delta\theta$. This is more accurate for two reasons. First, there is a greater number of samples, and the samples are made at equal increments of θ; second, the approximation that $\Delta\cos\theta = \sin\theta\Delta\theta$ does not have to be made.

$\Delta\theta_i$ in (44) can be placed outside the summation; then,

$$KP_T = 2\pi R^2 K\Delta\theta \sum_{\substack{\theta=0 \\ i=1}}^{\substack{i=n \\ \theta=2\pi}} F(\theta_i)\sin\theta_i. \tag{60}$$

The intergration of (60) is performed electronically by a simple antilog computer such as in Fig. 8. Although the details of the intergrater are basisically electronic circuit problems, the fundamental theory of operation is of great importance to the antenna engineer; therefore, a brief discussion on the theory of operation is in order.

The input voltage E_i in Fig. 8 is a d-c voltage proportional to radiated power density. R_1 is a gain calibration control. R_2 is a sinusodial voltage divider. This voltage divider is conveniently made by use of a rotary switch, having a position for each value of θ_i and several fixed resistors of the proper values. The input to the d-c amplifier is thus proportional to $S' \sin\theta_i$. S_1 is a commutator which is synchronized with the azimuth rotation. A typical sampling rate is 1000 units per revolution. The product

$$R_3C_1 \ggg \Delta t, \tag{61}$$

where Δt is the length of time that S_1 is closed when the azimuth rotation rate is a minimum. The voltmeter is a vacuum tube voltmeter that measures the accumulated voltage across the condenser. The diode

CR_1 eliminates discharge of the condenser when the output of the d-c amplifier is zero. S_2 is a reset switch which discharges the condenser.

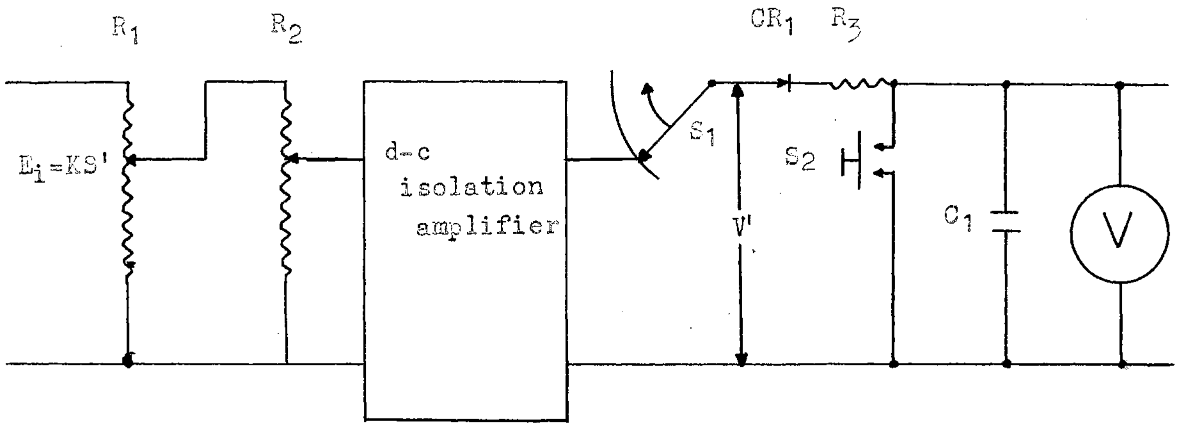

Fig. 8--Simplified schematic of an automatic intergrater.

The voltage V' is a series of positive pulses, the amplitude of each pulse being proportional to S'.

If V'_i is the amplitude of any general pulse, then the current through R_3 with the restriction imposed by (61) is

$$i_R = \frac{V'_i - V_{oi}}{R_3}, \tag{62}$$

where V_{oi} is the accumulated voltage across the condenser.

It will further be assumed that

$$V_i \gg V. \tag{63}$$

This restriction cannot practically be made valid for extremely low values of power densities approaching zero; however, it can be made valid for any value of power density that will contribute appreciably to the total intergration, by proper choice of R_3C_1 and amplifier gain. The restriction of (63) does not imply that V_{oi} is zero. The value of V_{oi} is

$$V_{oi} = \frac{1}{C} \int_0^t i_R dt = \frac{1}{R_3C} \sum V'_i \Delta t. \tag{64}$$

Now, since $\Delta t \propto \Delta \phi$, $V_i \propto S_i' \sin \theta$, the voltage added to the condenser for each conic is

$$\Delta V_{oi} = K \sin \theta \sum_{\substack{i=1 \\ \phi=0}}^{\substack{\phi=2\pi \\ i=n}} S_i \Delta \phi. \tag{65}$$

The total voltage added to the condenser for the complete set of conics becomes

$$V_{oi} = K_1 \sum_{\substack{j=1 \\ \theta=0}}^{\substack{\theta=\pi \\ j=m}} \left(\sum_{\substack{i=1 \\ \phi=0}}^{\substack{\phi=2\pi \\ i=n}} S_i \Delta \phi \right) \sin \theta_j. \tag{66}$$

Equation (66) is of the same form as (60), since $F(\theta) = \sum S_i \Delta \theta$. Therefore, V_{oi} is proportional to the total power radiated by the test antenna.

Calibration of the intergrater consists of obtaining the proper relation between V_{oi} and the set of radiation patterns, so that the isotropic radius can be drawn on the patterns. This calibration is a very simple procedure and will not be further discussed.

CHAPTER VI

ELLIPTICALLY POLARIZED ANTENNAS

The preceeding chapters have been primarily concerned with linearly polarized antennas. Only recently has the aircraft antenna designer been interested, to any extent, in elliptically polarized antennas. Most airplanes fly at attitudes parallel to the ground; therefore, radio links can be maintained between airplanes and from airplane to ground with linearly polarized antennas. The development of long range guided missiles, which fly in orbits at attitudes(with respect to the earth's surface) ranging from perpendicular to parallel, has forced the aircraft antenna designer to become concerned with the more complicated elliptically polarized antennas.

Circular polarization is the special case of elliptical polarization where the two orthogonal components of the electromagnetic wave are equal and in 90° time quadrature. This special case is normally the desired condition when linear polarization is not suitable; therefore, the general case of elliptical polarization will not be discussed. Actually linear polarization is the opposite extreme of elliptical polarization.

Usually a linearly polarized airborne antenna is used to receive energy from a circularly polarized transmitted signal; however, sometimes the airborne antenna is circularly polarized and the transmitted signal is linearly polarized. In either case the problem is to find the mutual response of a linearly and a circularly polarized system.

In Chapter II it was convenient to compare the radiation properties of an antenna to a hypothetical isotropic radiator. The only requirement for the isotropic antenna was that it radiate equal power densities in all directions. It was not necessary to specify the polarization of the energy radiated from the isotropic source. Here, it is favorable to specify that the isotropic antenna be circularly polarized in all directions. Gain referred to the circularly polarized isotropic antenna has a somewhat different meaning than the gain and directivity function referred to in Chapter II.

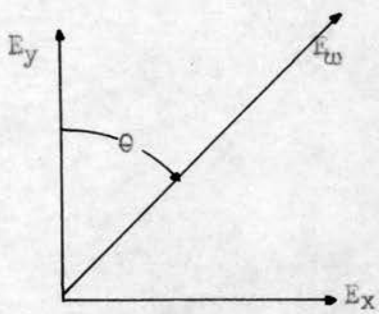

Fig. 9--Orthogonal space components of a circularly polarized wave.

Before further consideration of antenna gain problems with circular polarization, it is necessary to investigate the nature of a circularly polarized plane wave in space. Fig. 9 depects two orthogonal components of a circularly polarized wave. It is assumed that the wave is progressing out of the paper. The magnitude of wave components in the plane of the paper are such that

$$E_x = E_m \sin \omega t; \tag{67}$$

$$E_y = E_m \cos \omega t. \tag{68}$$

Then the resultant magnitude of the two components is

$$|E_\omega| = \sqrt{E_m^2 \sin^2 \omega t + E_m^2 \cos^2 \omega t};$$
$$|E_\omega| = E_m. \tag{69}$$

The direction of E_ω with respect to the X and Y direction is given by

$$\theta = \tan^{-1} \frac{E_x}{E_y} = \tan^{-1} \frac{E_m \sin \omega t}{E_m \cos \omega t};$$
$$\theta = \omega t. \tag{70}$$

Equations (69) and (70) indicate that the resultant electric field is a constant amplitude rotating at the angular velocity ω.

It is clear that the circularly polarized plane wave is composed of two linearly polarized plane waves; the amplitudes of the linear plane waves are in both time and space quadratures. One-half of the total energy is carried by each linear wave component.

Now, if "A" were a linearly polarized isotropic antenna placed in the field and polarized in the X direction, it would receive energy only from the X component. Thus, the presence of the Y component would have no effect on the energy received by the antenna. On the other hand, if "B", a circularly polarized isotropic antenna of proper rotational sense, were placed in the field, it would receive energy from both components. "A" would receive only one-half as much energy as "B". Under the conditions just described, the gain of "A" would have been -3db with respect to a circularly polarized isotropic antenna. Similar reasoning can be used to show that for a linearly polarized wave, "A" would have a +3db gain with respect to "B".

At first it seems awkward to use the circularly polarized isotropic antenna as a reference; however, the usefulness of the recourse can be more appreciated when the specification for an antenna is being calculated. For instance, if the power output of a missile telemetry transmitter, the ground receiver sensitivity, and the absolute gain of the linearly polarized ground receiving antenna were known, as well as the complicated trajectory of the missile, the radiation pattern specification can be calculated. The simplest approach would be to apply equation (9) for various points along the trajectory to find a gain function, just as though the transmitting antenna were always properly polarized. This gain function will be 3db greater at all points than the function with respect to a circularly polarized isotropic antenna, since the relative polarization of the circularly polarized wave and the linearly polarized antenna are not affected by the missile attitude. The factor of 3db is subtracted rather than added since the circularly polarized isotropic antenna, in this case, has a negative gain(-3db) with respect to the linearly polarized isotropic antenna.

Measurement of gain with respect to the circularly polarized isotropic antenna can become quite complicated, especially when the radiated electromagnetic wave is circularly polarized. For convenience of discussion the symbol "G*" will be used to refer to this gain.

The first step in measuring G* is the normal intergration using linear polarization as described in Chapter V. This intergration establishes a reference power density and provides a calibration of the measuring system parameters in terms of that reference power density. Next the linearly polarized transmitting antenna for the pattern range is replaced by a circularly polarized transmitting antenna. The power density at the test antenna is set, by adjusting transmitted power, to twice the value of the linearly polarized reference.

This power density is easily set up by use of a linearly polarized antenna located near the test antenna for field strength monitor. The power density of the circularly polarized wave is twice that of the linearly polarized reference when the power received by the monitor antenna is the same as that received from the reference field. This, of course, assumes that the monitor antenna was polarized in the same direction as the linearly polarized field when the reference reading was made.

Radiation patterns for all angles of interest are then recorded for the circularly polarized signal. The isotropic level calculated from the linear intergration will be the same as the circularly polarized isotropic level; therefore, $G^*(\Theta,\emptyset)$ will be the ratio of the radiation pattern amplitude to the isotropic level.

The reason for performing the intergration with linear polarization is that the polarization characteristics of the test antenna are not known. Thus, the test antenna could be circularly polarized of clockwise sense in one direction and circularly polarized of counter-clockwise sense in another direction. The rotational sense being wrong for circularly polarized antennas is equivalent to having linearly polarized antennas cross polarized; therefore, no signal is received.

CHAPTER VII

CONCLUSIONS AND RECOMMENDATIONS

The previous chapters have presented a theoretical basis for measurement techniques used in experimental development of aircraft antennas. Use of those principles and techniques will be summarized by discussion of an experimental development procedure for a simple aircraft antenna.

1) First, the electrical requirements for the antenna must be firmly established.

Usually the antenna system must be designed to match a 50 ohm transmission line, with the VSWR not to exceed a specified value over a given frequency range. The normal maximum VSWR is 2:1 for transmitting and 5:1 for receiving antennas; however, more rigid requirements are sometimes necessary. If the antenna system is composed of two or more antennas, the individual antenna impedance must be designed accordingly with the proper matching networks and power dividers so that the system VSWR is less than specified.

The radiation pattern requirement is either directly specified in terms of absolute gain as a function of coordinate angles, together with the polarization requirement or indirectly in terms of "radio-link" system parameters. When the latter is specified the direct radiation pattern requirements are calculated by the range formula,

$$P_R = \frac{P_T\ G_T\ G_R\ \lambda^2}{16\pi^2\ R^2}.$$

2) The available locations for the antenna on the aircraft are carefully studied. The merits and disadvantages of each potential location are tabulated for comparison. Some of the locations may have definite structual advantages, others may have favorable aerodynamic potentialities, but those locations may not be suitable from an electrical standpoint. Although the final antenna location will be determined by experimental measurement, many hours of valuable time can be saved by excluding most of the impractical locations before any experimental work is started. Often a less desirable mechanical location is best for an electrical consideration.

3) The general types of antennas for potential use are studied. From the electrical requirements and possible locations on the aircraft, the general type antenna(vertical stub, blade, slot, loop, spiral, etc.) compatible with each potential location are established.

4) Requirements for aircraft models and mock-up sections are established. The scale for the model aircraft is determined from the maximum model size the pattern range can conveniently accommodate; the frequency is scaled up by the same factor that the model dimensions are scaled down. By proper choice of the scale factor it is usually possible to make the scaled frequency range conform with a standard wave guide frequency band. Often either full or one-half scale mock-up sections of the aircraft skin at the proposed antenna locations are constructed for use in impedance measurements. The shape of the aircraft a few wave lengths from the antenna location has only a little effect on impedance; therefore, the aircraft sections of a few square wave lengths are sufficient for most impedance measurements.

5) Preliminary scale model radiation patterns are made. These patterns

should include the principle plane cuts and at least one conic near the center of the elevation sector of interest. The preliminary patterns should be made for each likely location with the general type of antenna considered practical for those locations.

6) Preliminary impedance measurements are begun on the types of antennas proved to be most promising by the preliminary radiation pattern study. From these studies the antenna general type and location should be determined.

7) A set of radiation patterns for intergration is made and the intergration performed. The isotropic level, determined from the intergration, is plotted on each of the radiation patterns of interest so that all points of deficiency in the radiation can be readily observed.

8) The terminal impedance of the prototype antenna is measured at several frequencies at close enough intervals to provide sufficient data for plotting an impedance vs frequency curve over the operating frequency range; from this curve a qualitative equivalent circuit can usually be approximated. The relation between the physical dimensions of the antenna and parameters of the equivalent circuit is established. Physical changes necessary for improved antenna performance can then be predicted. The physical changes are accomplished, and the impedance of the modified antenna measured.

9) A set of radiation patterns is made for the modified antenna. The cause of deficiency in the radiation patterns can usually be related to current or to the absence of current flow at various points along the antenna and the aircraft skin. Careful study of the effect the modification had on the antenna radiation pattern will be indicative of the kind of physical change that will improve performance of the antenna.

It is often necessary to make several modifications to the basic antenna before arriving at the final design. The importance of accuracy in measurement cannot be stressed too highly. Changes in impedance and radiation pattern caused by alterations of the antenna shape are often much less than predicted; therefore, errors due to hasty measurements often swamp out the desired data. A qualitative electrical effect caused by physical modification is often more important than quantitive effect. Thus, a very slight electrical change, if detected, caused by a single modification to the antenna can be extremely helpful in determining a better second modification. Careful analysis of the experimental data and correlation of these data always hasten, with minimum wasted effort, a good final antenna design.

Measurements associated with elliptical polarization problems are crude and difficult to perform. Future work should be directed toward improving these measuring techniques. The work should include development of new equipment for automatically recording polarization data, as well as standardization of the methods of presenting the data. The experts in the antenna and propagation field are not in agreement as to the best way to present and use elliptical polarization data. Progress along these lines should accelerate as space communication becomes more common.

A SELECTED BIBLIOGRAPHY

Hull, J. F., Novic, G., and Cordray, R. "How Long-line Effect Impairs Tunable Radar." Electronics, Vol. 27, No. 2 (February, 1954). pp. 168-173

King, R. W. P., Mimno, H. R., and Wing, A. H. Transmission Lines Antennas and Waveguides. New York, London: McGraw-Hill Book Company, Inc., 1945

Kraus, J. D. Antennas. New York, Toronto, London: McGraw-Hill Book Company, Inc., 1950

Markin, J. "Smith Chart Applications." Tele-Tech & Electronic Industries, Vol. 12, No. 5 (May, 1953). pp. 85-87, 116, 118

Schelkunoff, S. A. Advance Antenna Theory. New York: John Wiley & Son, Inc., 1955 London: Chapmain & Hull, Limited, 1955

Schelkunoff, S. A. and Fries, H. T. Antenna Theory and Practice. New York: John Wiley & Son, Inc., 1952 London: Chapmain & Hull, Limited, 1952

Sinclair, G., Jordan, E. C., and Vaughn, E. W. "Measurements of Aircraft Antenna Patterns Using Models." Proc. IRE, Vol. 35 (December, 1947). pp. 1451-1462

Sypherd, W. O., Fountain, A. M., and Brown, S. The Engineer's Manual of English. Chicago, Atlanta, Dallas, New York: Scott, Foresman and Company, 1943

Terman, F. E. "Antennas." Chapter 23, Electronics and Radio Engineering. New York, Toronto, London: McGraw-Hill Book Company, Inc., 1955

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