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THE UNIVERSITY OF MICHIGAN

COLLEGE OF ENGINEERING

DEPARTMENT OF ELECTRICAL ENGINEERING

Radiation Laboratory

SIMPLIFIED MODELING TECHNIQUES FOR AVIONIC

ANTENNA PATTERN SIGNATURES

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TABLE OF CONTENTS

ABSTRACT

Under a previous contract AF33(615)-1964, an investigation was made of measurement techniques to determine antenna radiation characteristics at the fundamental, spurious and harmonic frequencies. This study is continued, making further use of simplified models as an effective, low cost method for obtaining necessary antenna pattern signatures over the desired frequency ranges. Pattern data for a slot antenna is presented for a typical harmonic frequency using both the precision and a simplified aircraft model. A method for presenting this data in two different statistical forms is described and illustrated.

To obtain a meaningful spectrum signature of antennas, it is desirable to have information in regard to the antenna impedance, transmission line characteristics, and the output impedance of the transmitter available. For this reason a multiple generator model is examined to find the power transfer characteristics between the transmitter and the antenna. A need for further work on this problem is indicated. Preliminary tests on an ARC-27 transmitter to determine its operational characteristics outside its normal operating range are described.

An analog to digital converter for use with a standard antenna pattern recorder is described in Appendix A.

I. INTRODUCTION

The statement of problem as set forth in the contract which provides for the present investigation is as follows.

The problem is to determine practical methods for obtaining the desired antenna pattern spectrum signatures with minimum cost and complications by the use of simplified aircraft models or mockups and statistical methods for data presentation. The prediction of intersystem electromagnetic compatibility will necessitate a determination of the three dimensional spectral characteristics of all emitters and receptors on the flight vehicle.

To determine realistic pattern spectrum signature, careful consideration must be made of the spectral output of the emitter source, i.e. the typical AF transmitter being used with the antenna as well as the impedance properties of the actual antenna system employed on the vehicle.

The stated objective of the contract is: To make use of 'statistical' or 'digitalized' patterns to experimentally establish the fact that satisfactory antenna signature information can be obtained by the use of crude aircraft models or mockups by comparing with the results of data obtained using precision models. Early attention will be given to the selection of a statistical or digital method of pattern data presentation that would eliminate, if possible, the requirement for numerous antenna patterns. The program will be concerned with investigation, analysis and experimentation to determine feasible approaches for standardized measurement techniques to become a part of MIL-STD-449B.

Under a previous and related contract (AF33(615)-1964) an investigation was made of measurement techniques applicable for obtaining antenna radiation characteristics at the fundamental, spurious and harmonic frequencies. During that study it was found that a simplified modeling technique could be employed to obtain the antenna radiation characteristics at the desired frequencies. This technique was shown to be applicable to those antenna systems which are fed by a coaxial transmission line and operate in the frequency range of 100 Mc to 3 Gc or higher. The term "simplified modeling" derives its name from the fact that the aerodynamically shaped aircraft is replaced by simple geometric shapes. For example, the fuselage is replaced with a cylinder of the appropriate diameter and length and the wings and tail structure are replaced with flat planar surfaces.

As a further output of the original contract, the Radio Science Laboratory of The University of Michigan designed and fabricated a digital converter unit that enables one to record antenna patterns both in analog and digital form. The analog to digital converter designed and fabricated by the above laboratory can now be used to record antenna patterns in digital form for use by the Electromagnetic Compatibility Analysis Center (ECAC) for making interference predictions such as are needed for the various military organizations.

The present contract is being conducted from the systems aspect rather than the component aspect. In one task for example consideration is being given to the determination of the spectrum signature of a typical airborne transmitter (ARC-27). This information will be of value in helping to predict the expected level of interfering signals. The principle measurement technique that will be considered during this contract will be for the determination of the antenna impedance characteristics at the fundamental spurious and harmonic frequencies and its effect on the transmitter system operating characteristics.

II. SIMPLIFIED TECHNIQUES

In the previous contract the pattern characteristics of both a modified monopole and a quarterwave monopole were studied at the fundamental and harmonic frequencies. It was found advantageous to use simplified modeling techniques because of the savings in time and costs. Both conventional patterns and cumulative gain distributions of the radiation characteristics of the antenna showed that the simplified models provided sufficient accuracy and detail. Typical data is given in the final report (AFAL-TR-65-104). During the initial period of the present contract, this study was continued using a unilobe antenna structure. In addition to this new antenna type, a second statistical analysis technique was employed; this will be discussed in the section on sector scanning.

2.1 Slot Antenna

As a further justification of the simplified modeling technique, the radiation characteristics of a unilobe structure have been considered. For the purpose of this study an open ended section of waveguide as shown in Fig. 1 for the precision model and in Fig. 2 for the simplified model was used as the radiating element, simulating a typical L-band antenna. This antenna was operated at a scale fundamental frequency of 8 Gc and the second, third and fourth harmonics. Taper transitions were used between bands to minimize the generation of higher order modes in waveguides. Typical data for the first and fourth harmonics are shown.

FIG. 2: OPEN-ENDED WAVEGUIDE ON SIMPLIFIED T-33.

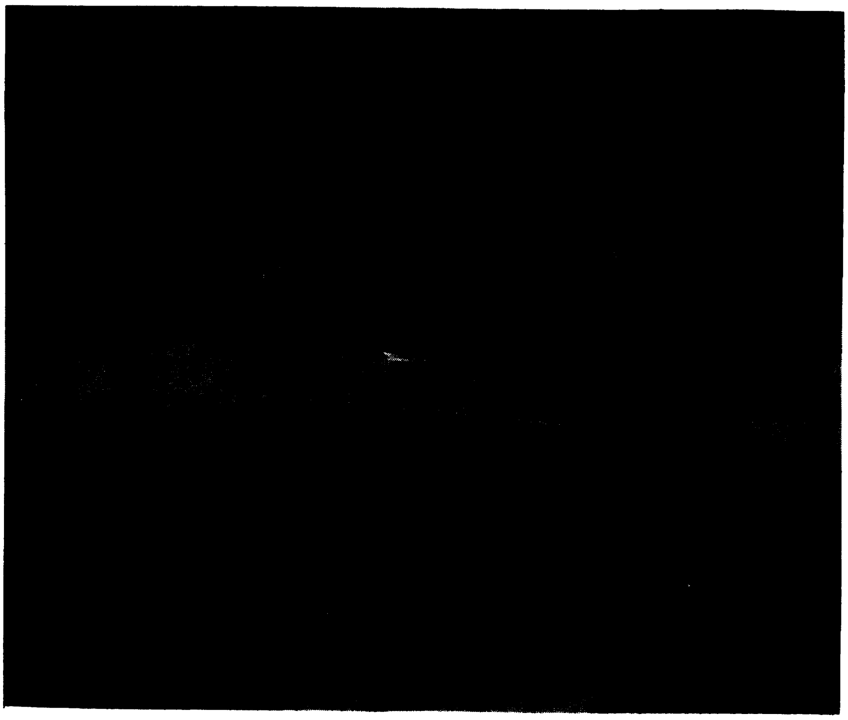


FIG. 1: OPEN-ENDED WAVEGUIDE ON PRECISION T-33.

Figure 3 is an analog plot at the fundamental frequency of the radiation characteristics of the antenna mounted on the precision model and Fig. 4 is a similar plot employing a simplified model. Figure 5, a composite cumulative gain distribution of the precision and simplified modeling data, demonstrates the type of agreement that can be achieved between the two models. Similar data for the fourth harmonic frequency 33.2 Gc is shown in Figs. 6, 7 and 8.

2.2 Sliding Sector Data

Initially, the cumulative gain distributions were used to compare the simplified and precision modeling techniques. It was suggested later by personnel of ECAC that an alternate data reduction technique be considered for airborne antennas. This alternate technique employs a sliding sector as illustrated in Fig. 9. Here it is seen that the sector width (in degrees) is denoted as α, and the sector slide increment (in degrees) as β. The sector α is slid in steps of β across the 360-degree width of the conventional antenna pattern. As α is slid, two sets of statistical data are calculated and plotted as follows: 1) the average gain within the sector α_1 is determined and recorded as $\alpha_1/2$; 2) the standard deviation of the individual gains withing α_1 sector is determined and recorded as $\alpha_1/2$. The sector α_1 is slid β^o becoming α_2 and the above calculations repeated and plotted at $\alpha_2/2 = \alpha_1/2+\beta$. The process is repeated for α_3, α_4, etc. across the pattern (Note: α_1=α_2=α_3, etc.). Figure 10 is a composite (precision vs. simplified model) of average data for the modified

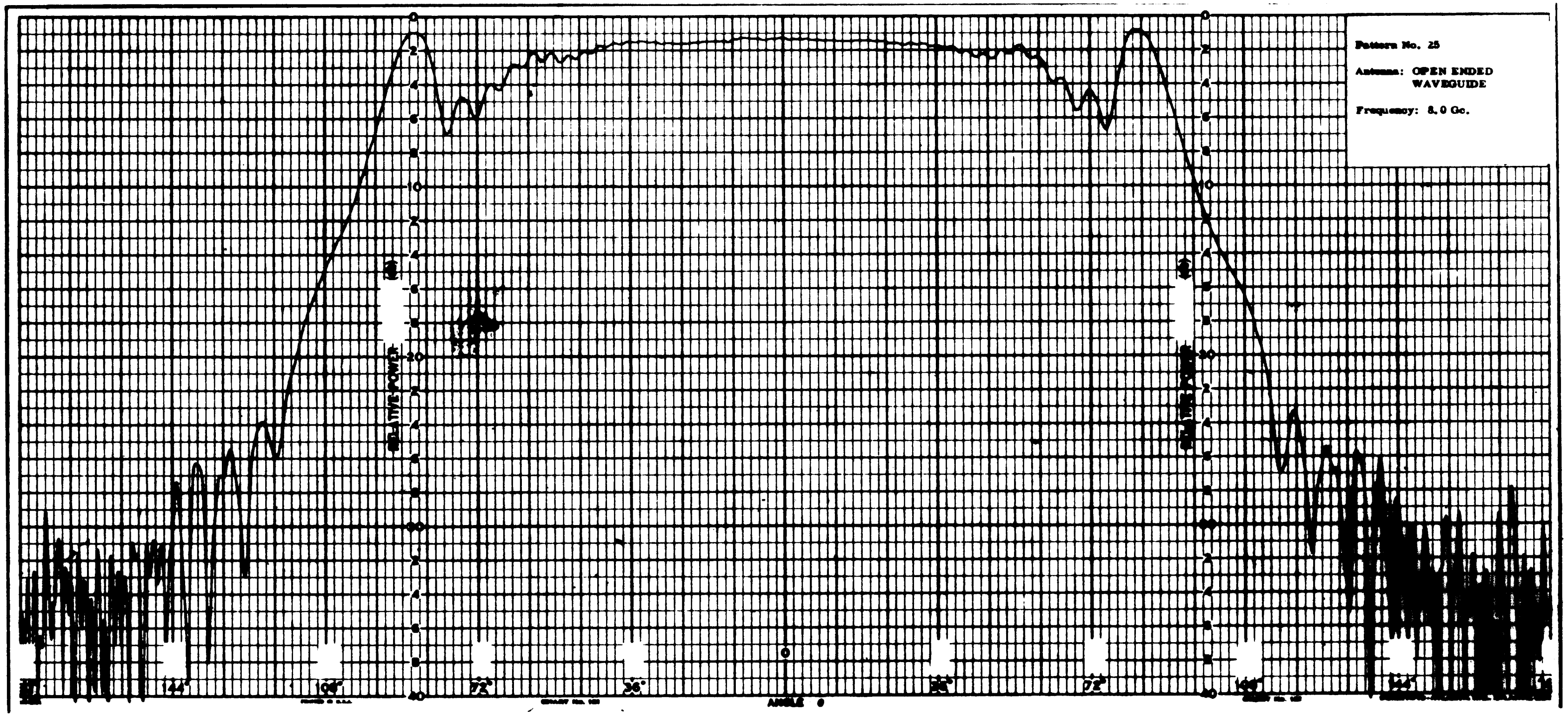


FIG. 3: **PRECISION T-33 MODEL (LESS WING TANKS). FREQ. = 8.0 Gc.**

FIG. 4: SIMPLIFIED T-33 MODEL. FREQ. = 8.0 Gc.

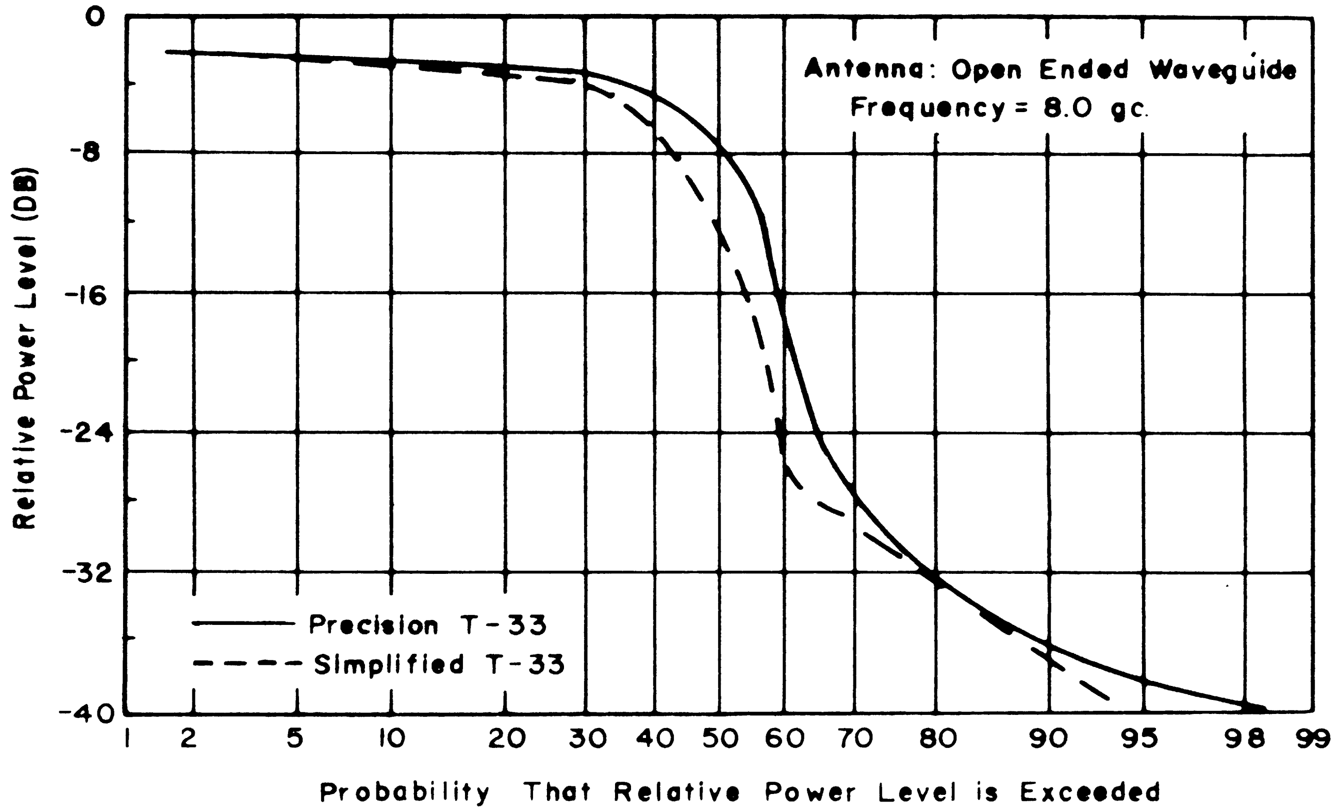


FIG. 5: CUMULATIVE GAIN DISTRIBUTIONS OF PRECISION AND SIMPLIFIED MODELS.

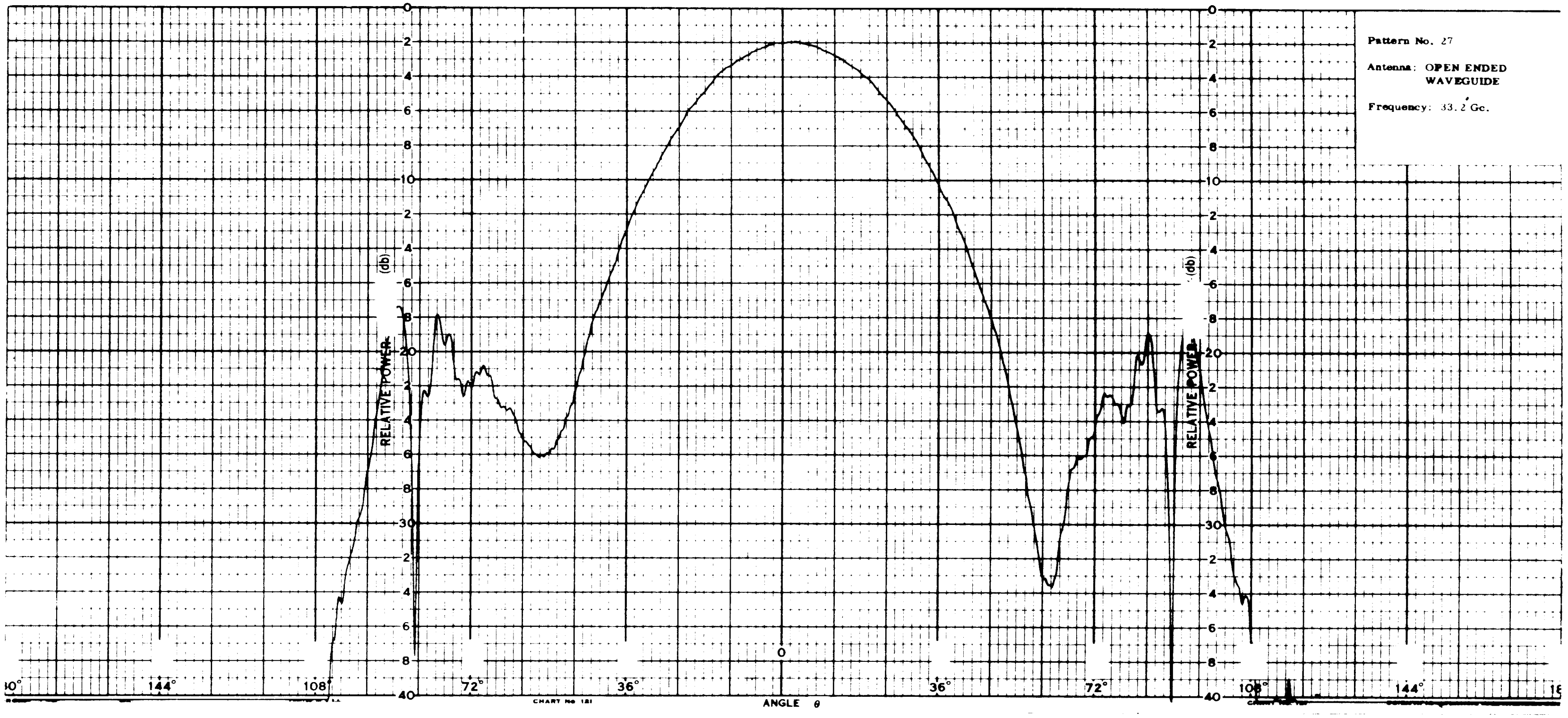


FIG. 6: **PRECISION T-33 MODEL (LESS WING TANKS). FREQ.** = 33.2 Gc.

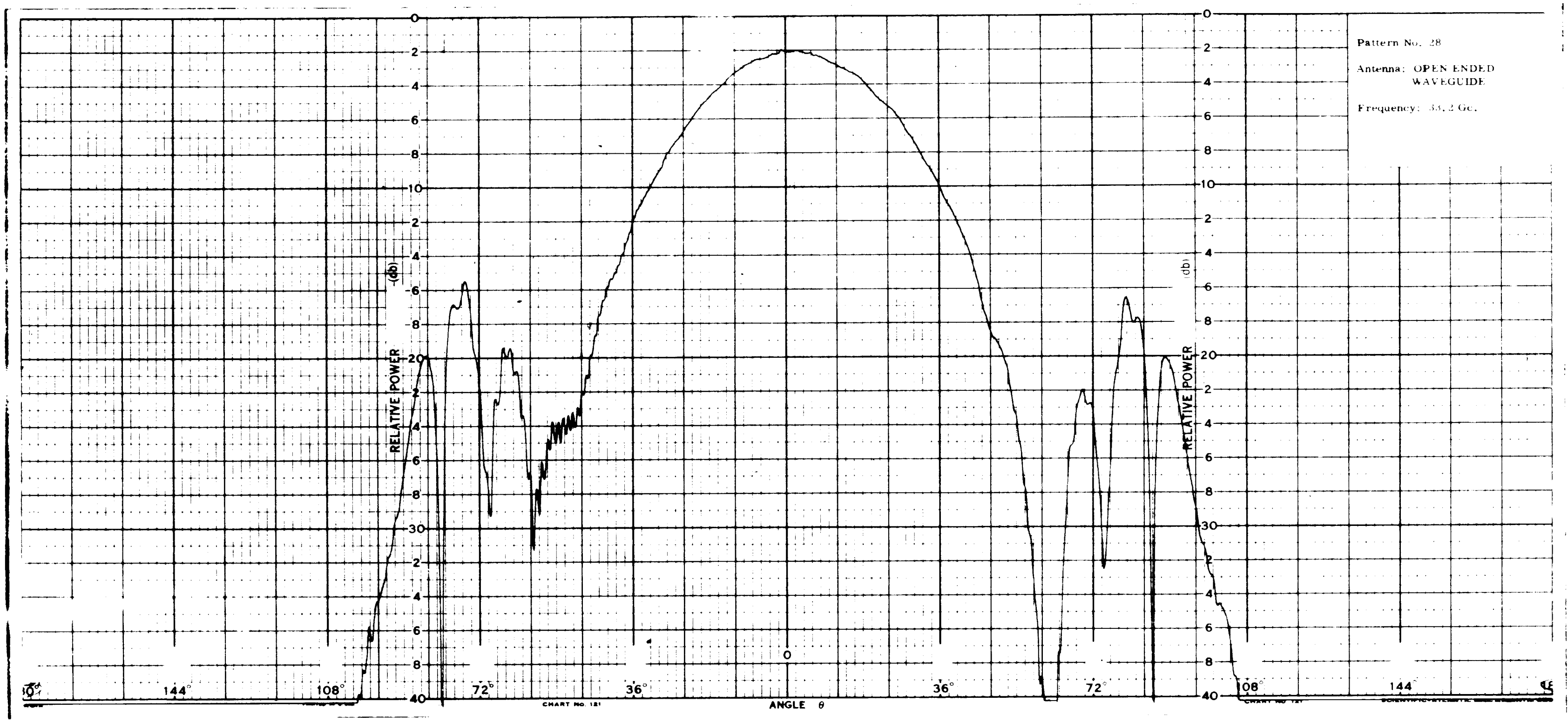


FIG. 7: SIMPLIFIED T-33 MODEL. FREQ. = 33.2 Gc.

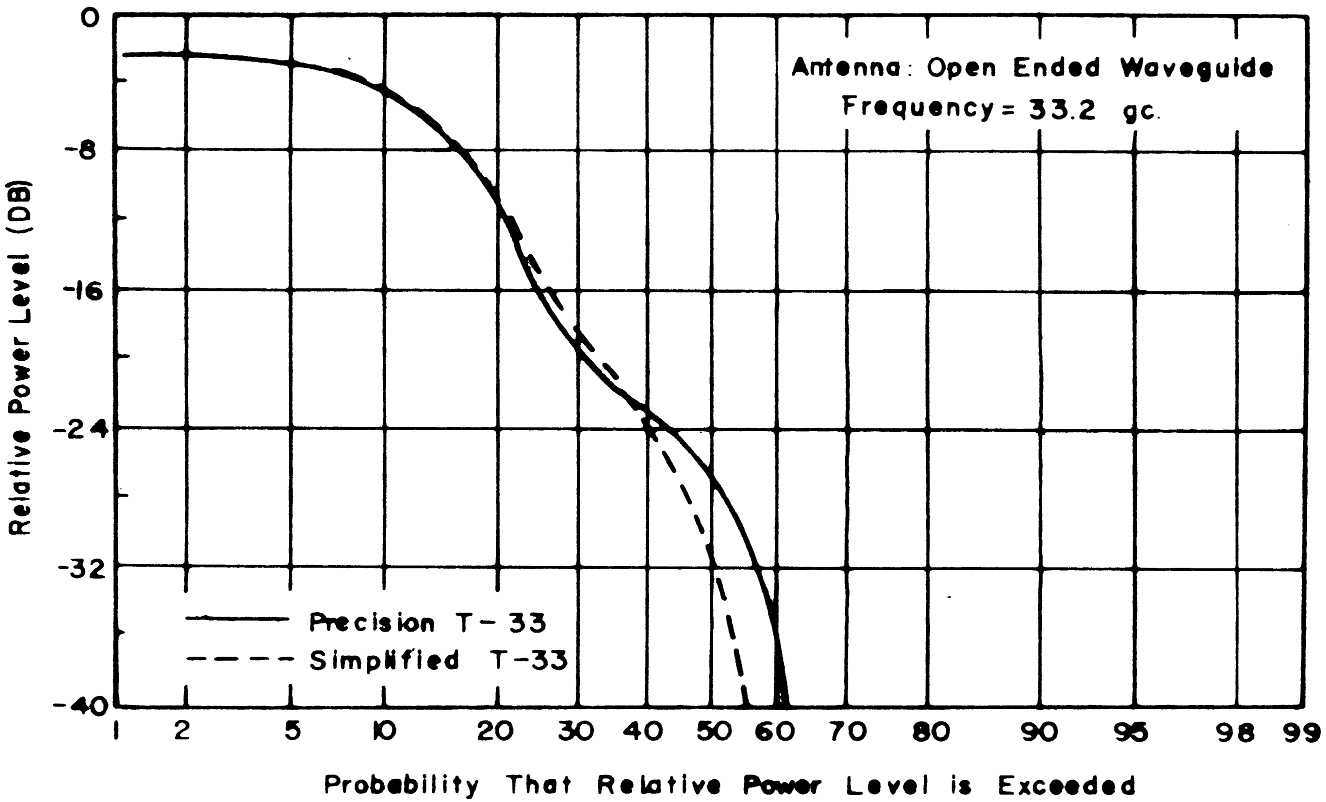


FIG. 8: CUMULATIVE GAIN DISTRIBUTIONS OF PRECISION AND SIMPLIFIED MODELS.

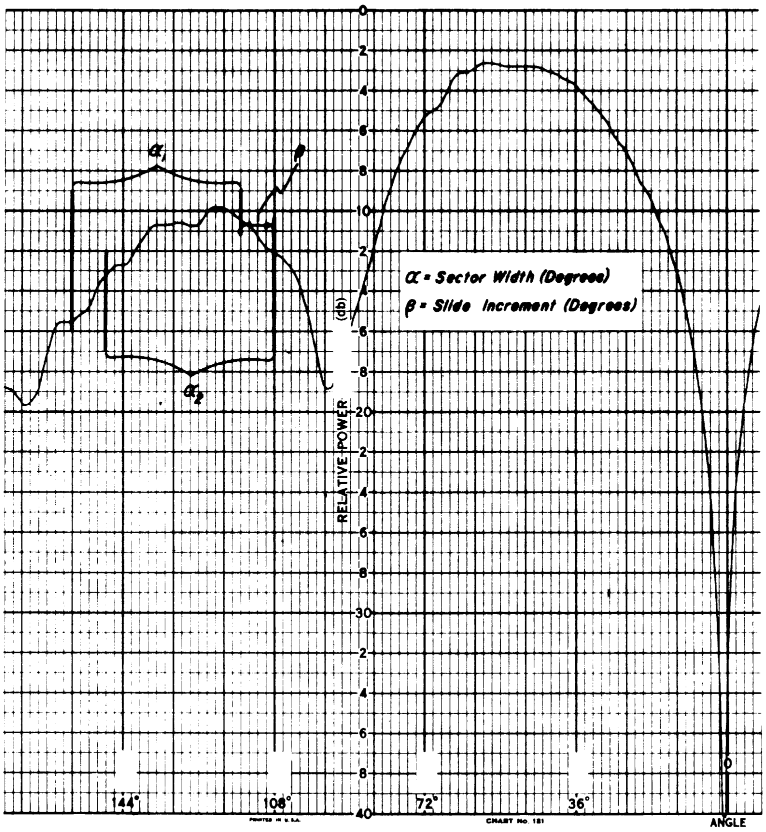


FIG. 9: SLIDING SECTOR DESCRIPTION

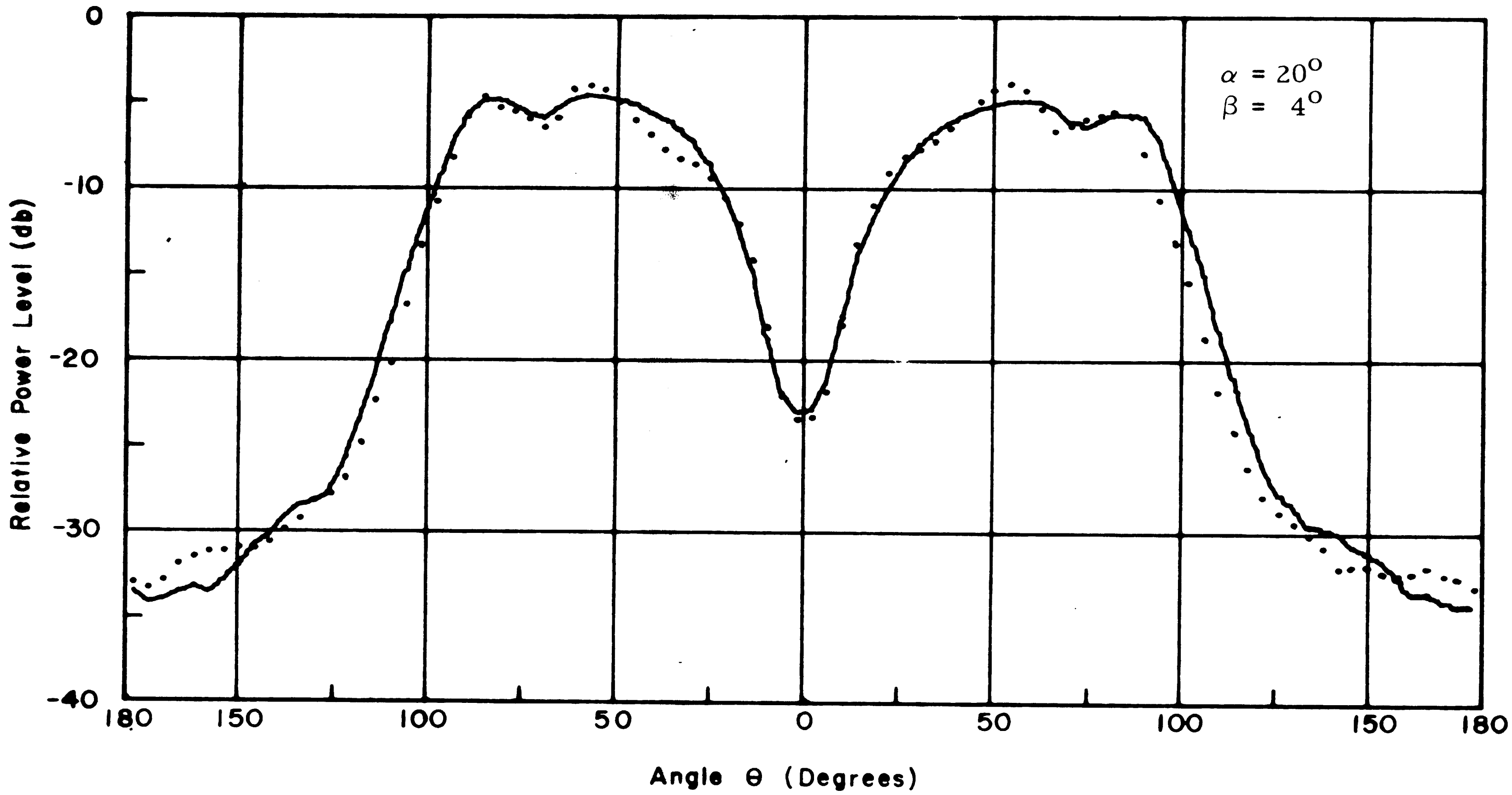


FIG. 10: AVERAGE OF SLIDING SECTOR OF PRECISION AND SIMPLIFIED MODELS. Antenna: Modified Monopole. Freq. = 7.2 Gc.

monopole at 7.2 Gc with $\alpha = 20^o$ and $\beta = 40^o$. Figure 11 is a composite of the standard deviation data. During discussions with personnel of ECAC it has been learned that data accurate to within $\pm$ 3 db is considered acceptable. Data obtained by our simplified modeling techniques is accurate to within these limits.

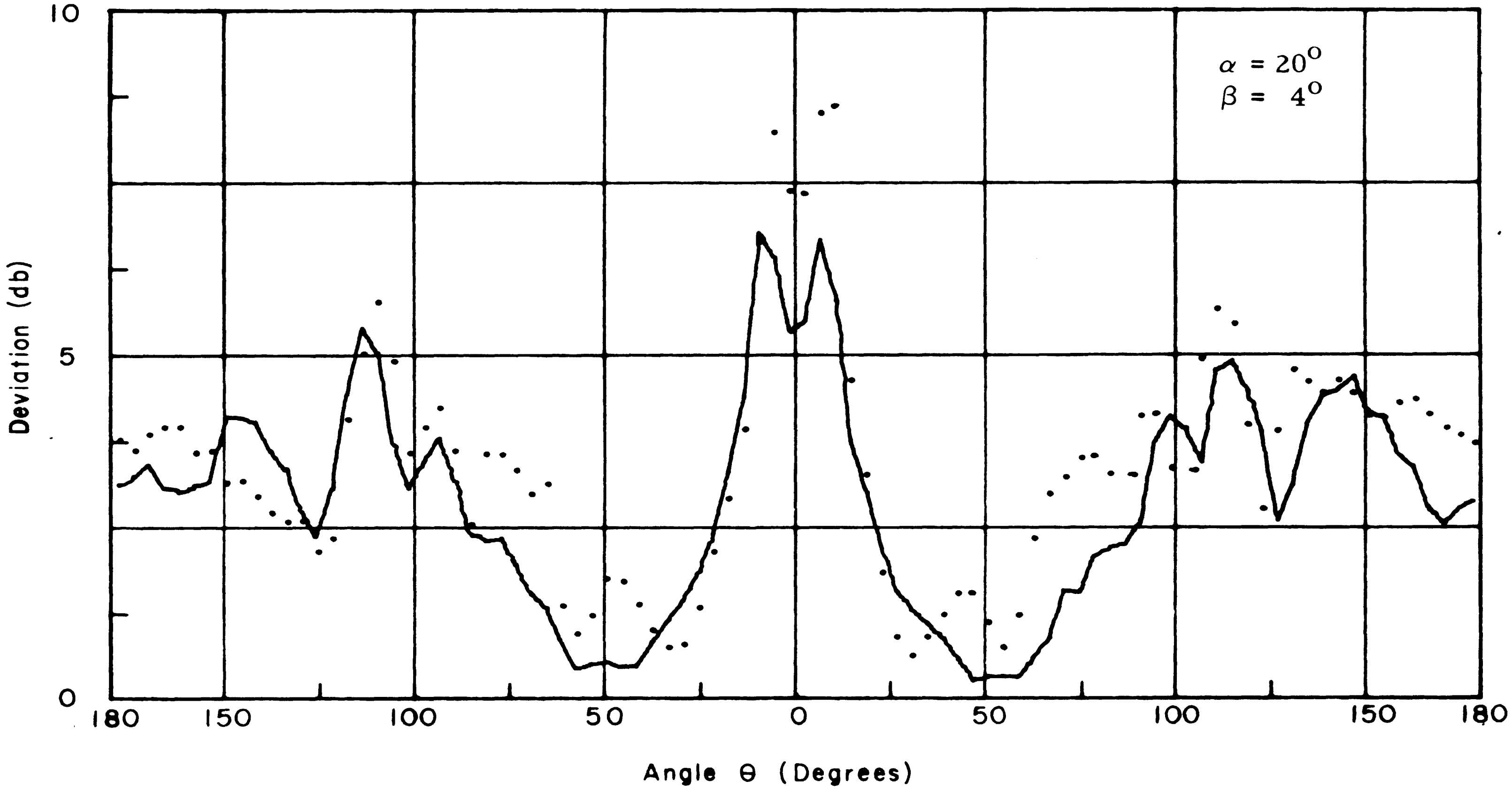


FIG. 11: STANDARD DEVIATION OF SLIDING SECTOR OF PRECISION AND SIMPLIFIED MODELS.
Antenna: Modified Monopole. Freq. = 7.2 Gc.

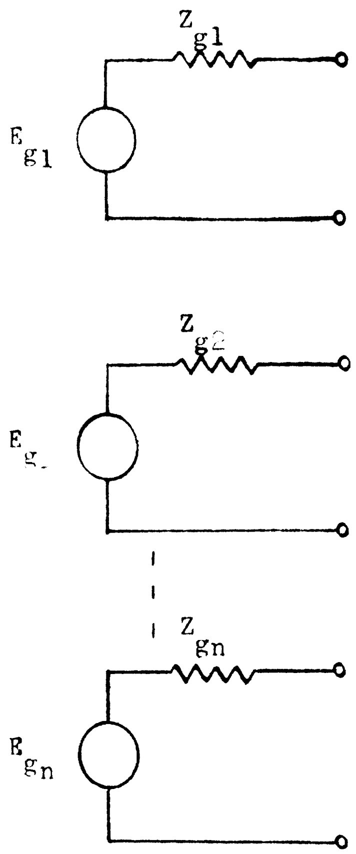
III. IMPEDANCE MEASUREMENT TECHNIQUES

Consideration is now being given to a measurement technique for obtaining impedance data on antennas at the fundamental spurious and harmonic frequencies. The feasibility of replacing a typical transmitter, that generates not only the designed frequency but also spurious harmonic frequencies, with several ideal frequency generators is now being investigated.

3.1 Multiple Generator Model

Consider the following problem: 1) A transmitter is to be connected to an antenna using a coaxial transmission line, and it is necessary to determine what percentage of the generated power is radiated by the antenna at the fundamental, harmonic, and spurious frequencies. 2) Various installations require different transmission line lengths and different antenna types. 3) What measurements of the transmitter, transmission line, and antenna must be made to provide sufficient information to determine the radiated power.

Since a satisfactory method has already been developed for determining radiation characteristics, emphasis will now be on the remaining problems. Consider an equivalent circuit for the transmitter, as shown in Fig. 12. The generator EMF's and impedances are the quantities that need to be determined to provide a quantitative model. With this model, the power radiated by the antenna can be calculated provided the transmission line impedance, length and the antenna impedance



E_{g_k} = EMF of the generator for the k-th spurious frequency

Z_{g_k} = Impedance of the generator for the k-th spurious frequency

FIG. 12: TRANSMITTER EQUIVALENT CIRCUIT

are known at each frequency of interest. Various installations will utilize slightly different lengths of transmission line which, in general, will have considerable effect on the radiated power. For many problems to be solved by ECAC, the expected (or average, together with the standard deviation) power for all lengths (after subtracting out the closest integral number of half wavelengths) will constitute the required information. However, this report will not discuss this part of the problem further. Note also that only coaxial transmission lines are being considered. The moding considerations of waveguides adds another dimension to the whole problem. The treatment of waveguide transmission lines will therefore be deferred at this time.

A key consideration of the multiple generator model is the aspect of interactions between the generator. By interactions we mean the following. Suppose a load Z_{L_K} is attached to generator (E_{g_k}, Z_{g_k}). Let Z_{L_K} vary and observe generator (E_{g_j}, Z_{g_j}). If (E_{g_j}, Z_{g_j}) is constant as Z_{L_K} is varied then the model is considered to be free of interaction.

If for a real transmitter this model is free of interaction, then the computation of power absorbed by an antenna is readily made. It is necessary to gather experimental evidence to determine the extent of interaction for a real transmitter. We know of no simple analytical treatment to prove or disprove this question.

If the model is not free of interaction, then perhaps a linear correction term added to E_{g_k} and Z_{g_k} will provide a sufficiently accurate model for ECAC purposes.

An experiment to determine the extent of this interaction may run into technical difficulty. The problem is how to construct a load that is constant at one or more frequencies but variable at another frequency. A scheme to accomplish this for two frequencies f_1 and f_2, except for an odd harmonic relation between them, has been devised. The scheme is realized as shown in Fig. 13.

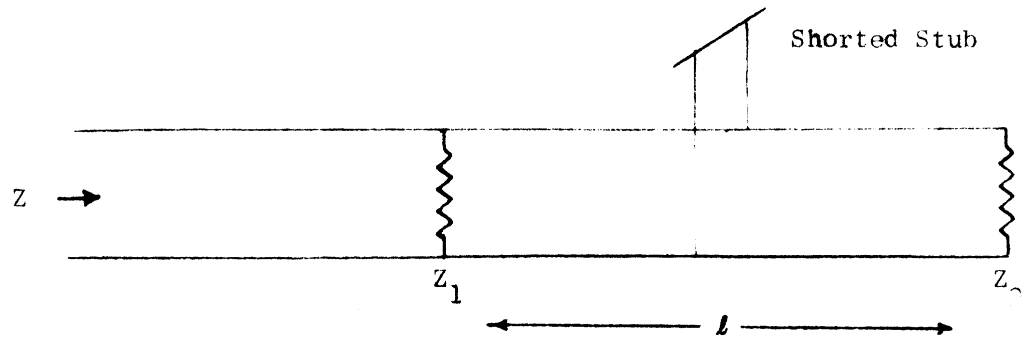


FIG. 13: VARIABLE LOAD CONFIGURATIONS

For this scheme, make the length of the shorted stub a quarter wavelength at f_1. Therefore at f_1, the position of the shorted stub has no effect on Z. However, at f_2, provided $f_2 \not\simeq (2n - 1) f_1$, Z will vary when the stub is moved along the length ℓ. By choosing appropriate values of Z_1 and Z_2 (which may be pure resistances), a large range of variations of Z can be obtained. No doubt some similar arrangement could be used for $f_1 \approx (2n - 1)f_2$.

3.2 Power Transfer Considerations

In the Electromagnetic Compatibility Program there is need for a simple method for computing the power absorbed by a load (an antenna) when the antenna impedance, Z_a, transmission line impedance, Z_o, and source impedance, Z_s, are far from equal. A general method is desired for computing the maximum and minimum power for a given Z_a, Z_o, and Z_s if the transmission line length is varied. For some purposes, a method is used for determining the probability that the power absorbed by Z_a will exceed a stated power if a random length of transmission line is used. A solution can be described for the first two problems and a suggested solution for the third will be indicated. It will be assumed that the transmission line is lossless, hence $Z_o = R_o$.

The general situation is as shown in Fig. 14.

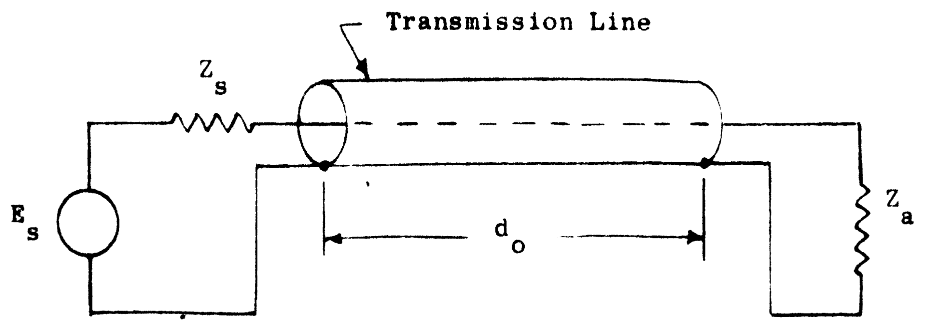


FIG. 14: GENERAL TRANSMISSION LINE PROBLEM

For a precise formulation of the problem, more dimensions should be added to Fig. 14, as shown in Fig. 15.

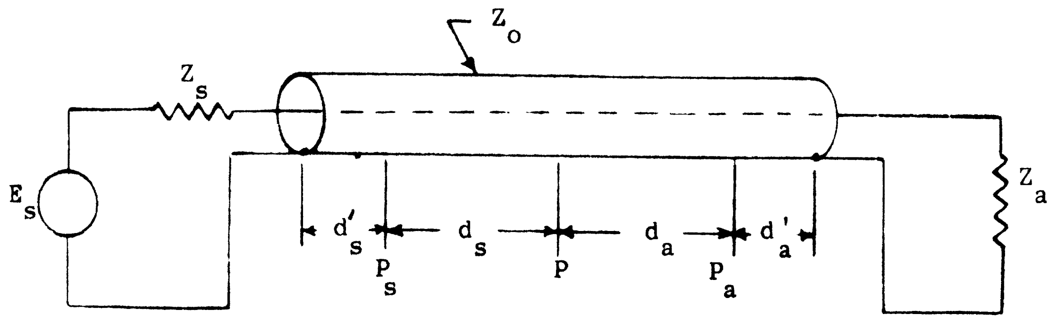


FIG. 15: MODIFIED TRANSMISSION LINE PROBLEM

In Fig. 15, the distance d_o has been broken up into four distances d'_s, d_s, d_a, and d'_a such that

$$d'_s + d_s + d_a + d'_a = d_o \tag{1}$$

The two distances d'_s and d'_a have been chosen in the following manner. Consider first d'_a. The antenna impedance Z_a will produce a standing wave on the transmission line; the point P_a is chosen at one of the minima of this standing wave. Then d'_a is the distance from the antenna to this point P_a. At P_a the antenna load impedance Z_a will be transformed to an impedance $Z'_a = R'_a$ and $R'_a < Z_o$. The distance d'_s to point P_s is determined in the same manner but using Z_s as the terminating impedance.

Now we can write equations for the impedance seen at points P, Z'_s, and Z'_a. Z'_s is the impedance at P seen looking toward the source, and Z'_a is the impedance at P seen looking toward the antenna. Expressions for the resistive and reactive components on transmission lines are well known. A particularly clear and compact form is given in Montgomery et al*.

Let

$$Z'_s = R'_s + X'_s \tag{2}$$

and

$$Z'_a = R'_a + X'_a \tag{3}$$

For the source

$$R'_s = Z_o \frac{r_s}{r_s^2 \cos^2 kd_s + \sin^2 kd_s} \tag{4}$$

$$X'_s = Z_o \frac{(1-r_s^2) \sin kd_s \cos kd_s}{r_s^2 \cos^2 kd_s + \sin^2 kd_s} \tag{5}$$

and for the antenna

$$R'_a = Z_o \frac{r_a}{r_a^2 \cos^2 kd_a + \sin^2 kd_a} \tag{6}$$

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*Montgomery, C.G., et al. (1948) "Waveguides as Transmission Lines," Chapter 3, p. 72, Principles of Microwave Circuits, 8, Radiation Laboratory Series, McGraw-Hill Book Co., Inc., New York, New York.

$$X_a' = Z_o \frac{(1-r_a^2)\sin kd_a \cos kd_a}{r_a^2 \cos^2 kd_a + \sin^2 kd_a} \tag{7}$$

In these equations, r_s is the standing wave ratio on a line of characteristic impedance Z_o terminated in Z_s, k is the propagation constant of this line. Likewise, r_a is the standing wave ratio on the same line terminated in Z_a.

Now let $kd_s = \pi/2$ to establish point P, then

$$R_s' = Z_o r_s \tag{8}$$

$$X_s' = 0 \tag{9}$$

For any arbitrary length of transmission line between the source and the antenna, the antenna impedance seen at point P will be

$$Z_a' = R_a' + jX_a' \tag{10}$$

Now let us compute the power absorbed by the antenna. The equivalent circuit is as shown in Fig. 16. Note that in Fig. 16 E_s' is not, in general, the same as E_s in Figs. 14 and 15. The equivalent source voltage has also been transformed by the transmission line. As will be shown presently, this transformation need not be calculated; a measurement of maximum power available from the source will suffice to define E_s' so that neither E_s or E_s' need be measured explicitly.

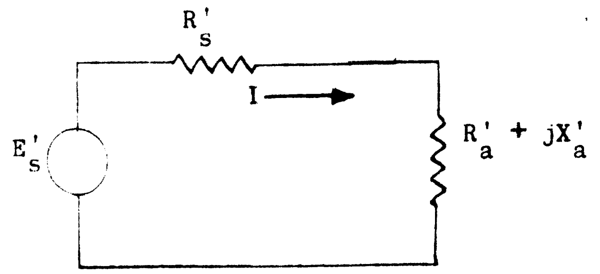

FIG. 16: EQUIVALENT CIRCUIT FOR THE TRANSMISSION LINE PROBLEM

From Fig. 16, the power absorbed by the antenna is

$$I = \frac{E'_s}{\sqrt{(R'_s + R_a)^2 + (X'_a)^2}} \tag{11}$$

$$P = I^2 R'_a = \frac{E'^2_s R_a}{(R'_s + R'_a)^2 + (X'_a)^2} \tag{12}$$

Equation 12 may be normalized by computing the maximum power that any load might draw from the source. For maximum power, the transmission line should be terminated by a load R'_s at $kd_s = \pi/2$. For this condition the maximum power is given by

$$P_M = I^2 R'_s \tag{13}$$

$$I = \frac{E'_s}{2R'_s} \tag{14}$$

$$P_M = \frac{E'^2_s}{4R'_s} \tag{15}$$

Now, Eq. 12 can be written in normalized form.

$$\frac{P}{P_M} = \frac{E_s'^2 R_a'}{(R_s' + R_a')^2 + (X_a')^2} \div \frac{4R_s'}{E_s'^2} \tag{16}$$

$$\frac{P}{P_M} = \frac{4R_s' R_a'}{(R_s' + R_a')^2 + (X_a')^2} \tag{17}$$

Substituting the relations 4, 6, and 7 (remembering that $kd_s = \pi/2$) Eq. 17 becomes

$$\frac{P}{P_M} = \frac{4 Z_o r_s \dfrac{Z_o r_a}{r_a^2 \cos^2 kd_a + \sin^2 kd_a}}{\left(Z_o r_s + \dfrac{Z_o r_a}{r_a^2 \cos^2 kd_a + \sin^2 kd_a}\right)^2 + \left(\dfrac{Z_o (1 - r_a^2) \cos kd_a \sin kd_a}{r_a^2 \cos^2 d_a + \sin^2 kd_a}\right)^2} \tag{18}$$

Simplifying Eq. 18 results in

$$\frac{P}{P_M} = \frac{4 r_s r_a (r_a^2 \cos^2 kd_a + \sin^2 kd_a)}{\left[r_s (r_a^2 \cos^2 kd_a + \sin^2 kd_a) + r_a\right]^2 + (1 - r_a^2)^2 \cos^2 kd_a \sin^2 kd_a} \tag{19}$$

Now for certain values of d_a P/P_M will be maximum or minimum.

If $kd_a = n\pi \qquad n = 0, 1, 2, \ldots\ldots$

$$P \Big|_{n\pi} = P_{min} = P_M \frac{4r_s r_a r_a^2}{(r_s r_a^2 + r_a)^2 + 0} \tag{20}$$

$$= P_M \frac{4r_s r_a}{(r_s r_a + 1)^2} \tag{21}$$

If $kd_a = \frac{2n+1}{2}\pi \qquad n = 0, 1, 2, \ldots\ldots$

$$P \Big|_{\frac{(2n+1)\pi}{2}} = P_{max} = P_M \frac{4r_s r_a}{(r_s + r_a)^2} \tag{22}$$

$$\frac{P_{max}}{P_{min}} = \alpha^2 = \left(\frac{r_s r_a + 1}{r_s + r_a}\right)^2 \tag{23}$$

Equation 23 is a simple equation that determines the ratio of the maximum power to the minimum power delivered to an antenna from a source using any arbitrary length of the transmission line. The only quantities that need to be measured are the standing wave ratios for the antenna and the source impedance. Equation 22 gives the ratio of the maximum power delivered to an arbitrary impedance to the maximum power available from the source using an impedance matched to the source.

The equations of greatest interest at this point are Eqs. 15, 19, and 23, First,

Eq. 15:

$$P_M = \frac{E_s'^2}{4R_s'} \tag{15}$$

This equation establishes the absolute maximum power available from the sources using a matched load; no other load can extract more power from the source.

Now consider Eq. 19:

$$\frac{P}{P_M} = \frac{4 r_s r_a (r_a^2 \cos^2 kd_a + \sin^2 kd_a)}{\left[r_s (r_a^2 \cos^2 kd_a + \sin^2 kd_a) + r_a\right]^2 + (1 - r_a^2)^2 \cos^2 kd_a \sin^2 kd_a} \qquad (19)$$

For fixed r_s and r_a, the power, P, absorbed by the load will vary between the two values given by Eqs. 21 and 22.

$$P_{min} = P_M \frac{4 r_s r_a}{(r_s r_a + 1)^2} \qquad (21)$$

$$P_{max} = P_M \frac{4 r_s r_a}{(r_s + r_a)^2} \qquad (22)$$

Thus, P can never be greater than P_{max} nor less than P_{min} regardless of the transmission line length.*

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*Recall the assumption that the transmission line is lossless and contains no reflection from connectors or other discontinuities. This lossless condition can be removed if the standing wave ratio r_s and the power available, P_M, are measured through a length of transmission line approximately equal to the length used in a typical installation. The reflectionless condition will also be removed for a given installation by this technique, but the variation from one installation to another will not. Normal connector reflections are small and should not affect the prediction appreciably.

These equations are relatively simple and will predict the maximum and minimum power radiated by an antenna when connected to a specified source. The data necessary from measurements consists of two standing wave ratio measurements and one power measurement.

The third problem indicated earlier is somewhat more complicated and requires additional work. However, from some limited computations, it appears that a set of normalized curves can be drawn to provide approximate answers to the question "what is the probability, under conditions of arbitrary transmission line lengths (within, say, a few wavelengths of some nominal length), that the power absorbed by the antenna will exceed a specified value P_1?" For this problem, we have to turn our attention to Eq. 19.

Let us solve Eqs. 21 and 22 for r_s and substitute in Eq. 19.

$$\alpha^2 = \frac{P_{max}}{P_{min}} = \left(\frac{r_s r_a + 1}{r_s + r_a}\right)^2 \tag{24}$$

$$\alpha(r_s + r_a) = r_s r_a + 1$$

$$\alpha r_s - r_s r_a = 1 - \alpha r_a$$

$$r_s = \frac{1 - \alpha r_a}{\alpha - r_a} \tag{25}$$

Using Eqs. 25 and 22, Eq. 19 becomes

$$\frac{P}{P_{max}} = \frac{\left(r_a + \frac{1-\alpha r_a}{\alpha - r_a}\right)^2 (r_a^2 \cos^2 kd_a + \sin^2 kd_a)}{\left[\frac{1-\alpha r_a}{\alpha - r_a}(r_a^2 \cos^2 kd_a + \sin^2 kd_a) + r_a\right]^2 + (1-r_a^2)^2 \sin^2 kd_a \cos^2 kd_a} \qquad (26)$$

Although Eq. 26 appears complicated and awkward, there is some evidence that the equation can be represented rather simply as a family of curves. To show this, it may be necessary to compute curves from Eq. 26 and plot a normalized family. This procedure is best handled by a computer program.

Some preliminary calculations have been made; they indicate that a group of curves from Eq. 26 with α = constant, plotted as a function of kd_a, $0 < kd_a < \pi/2$ will show little variation as r_a is varied.

Figure 17 shows the result of one such computation. The solid curve was drawn through points computed by a slide rule for $r_a = r_s = 2.5$, the points marked by x are for $r_a = 10$ and $r_s = 1.58$, in both cases $\alpha = 1.45$. These points are so close to the solid curve to lead one to conjecture that the shape of this curve for α = constant will be approximately invariant for a wide range of r_a.

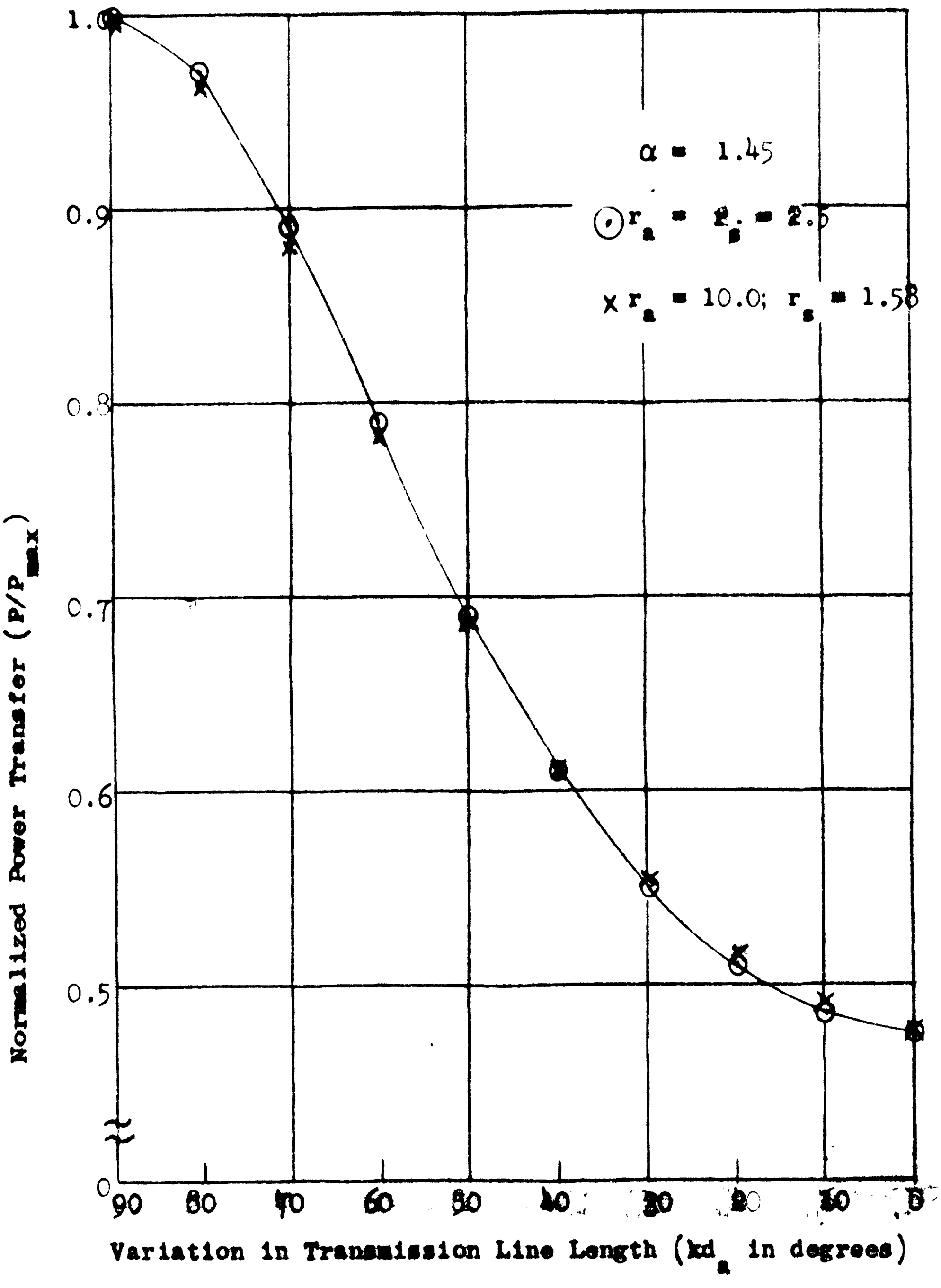

FIG. 17: POWER TRANSFER AS TRANSMISSION LINE LENGTH

IV. ARC-27 EXPERIMENT

An ARC-27 transmitter has been obtained from Wright Field and tested. The transmitter performance is below specifications in the VHF region and it does not function at all in the UHF region. These initial tests were performed at the Michigan Air National Guard at Metropolitan Airport in Detroit in order to take advantage of their special test facilities. Since the transmitter was obtained from Wright Field, minus a power supply, a 28-volt, 20 amp supply has been designed and is now being built. It is expected that this supply will be completed within a week's time and when the transmitter is operating properly further evaluation will be performed. A Polarad spectrum analyzer is available at the Radiation Laboratory, and this will be used to determine the transmitter power output over a frequency band extending well beyond its normal operating range.

APPENDIX A: DEVELOPMENT OF AN ANALOG DIGITAL CONVERTER

Early in this program, it became apparent that for the rapid reduction of antenna pattern data in statistical form it would be desirable to have the antenna data digitalized with data stored in digital form on magnetic tape. One would be able then to utilize a computer for data reduction rather than employ tedius hand techniques. To place the antenna data in digital form, an analog digital conversion unit was needed. This unit was subsequently designed and built and is now operating.

The function of the converter is to sample, at one-degree intervals, an analog waveform such as an antenna pattern, and to record the magnitude of the sampled value on a general-purpose tape recorder in the form of a binary-coded number. A block diagram of the converter is shown in Fig. A-1.

The amplitude of the sample is derived from a potentiometer which is mechanically ganged to the recording pen on the analog recorder. A fixed voltage is applied to this potentiometer to provide a linear voltage proportional to pen deflection. Pulses at one-degree intervals are generated by a light source, photocell, and rotating mirror mounted inside the antenna position indicator.

The process for recording the amplitude at a given angle is as follows: A one-degree pulse starts a 100 kcs, 7-stage binary counter. The output of this counter is coupled to a digital-to-analog converter which provides a voltage output linearly proportional to the count (128 levels). This voltage is compared to

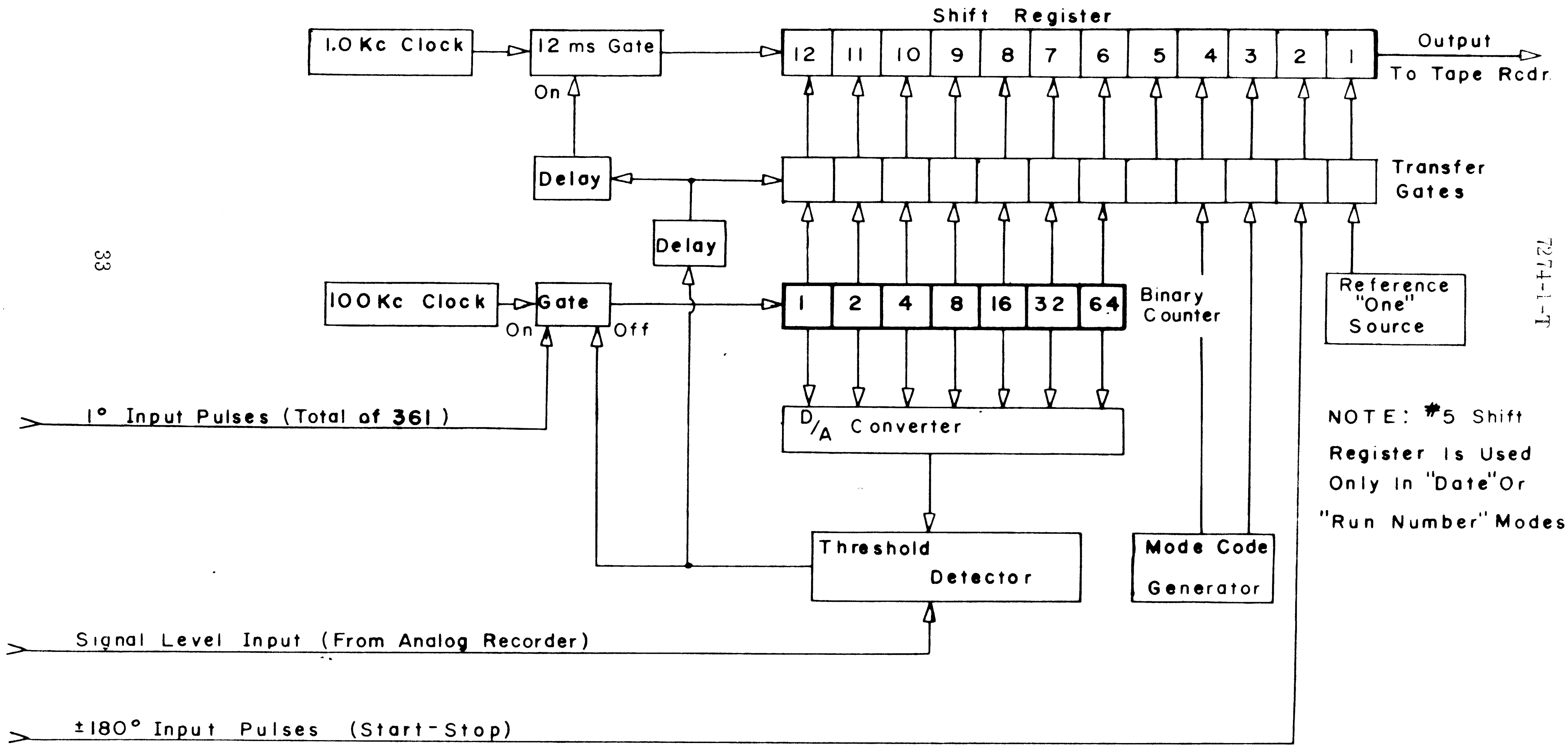

FIG. A-1: BLOCK DIAGRAM OF $^{A}/_{D}$ CONVERTER the potentiometer voltage in a threshold (zero-crossing) detector. When the two voltages are identical, the detector generates a pulse which is used to stop the counter. The final count is therefore proportional to the analog pen deflection, and hence, the instantaneous amplitude on the antenna pattern for that particular angle.

The final condition of the counter (either "0" or "1" for each of the seven stages) is then transferred to a shift register. A one-kilocycle "clock" then transfers the count to the output in the form of 300-microsecond pulses spaced 1000 microseconds apart. Each pulse will be a "0" (-6 volts) or a "1" (-12 Volts), corresponding to the final counter condition.

A total of twelve pulses are actually "clocked out" for each data point — the format for a data point taken at 180^o is shown in Fig. A-2. The first pulse (always a "1") is used as a computer start indication and reference level corresponding to a logic "1". For the first and the 361st Data points (180^o point of the antenna pattern), the second pulse clocked out is a "1"; for the interim 359 data points, it is a "0". The third and fourth pulses are used to indicate the mode of operation: "01" for a date, "10" for a run number, or "11" for a data point. The fifth pulse is always a zero for data points; it is used only for date and run number. The binary count occupies the remaining seven pulse positions, with the least significant bit clocked out last.

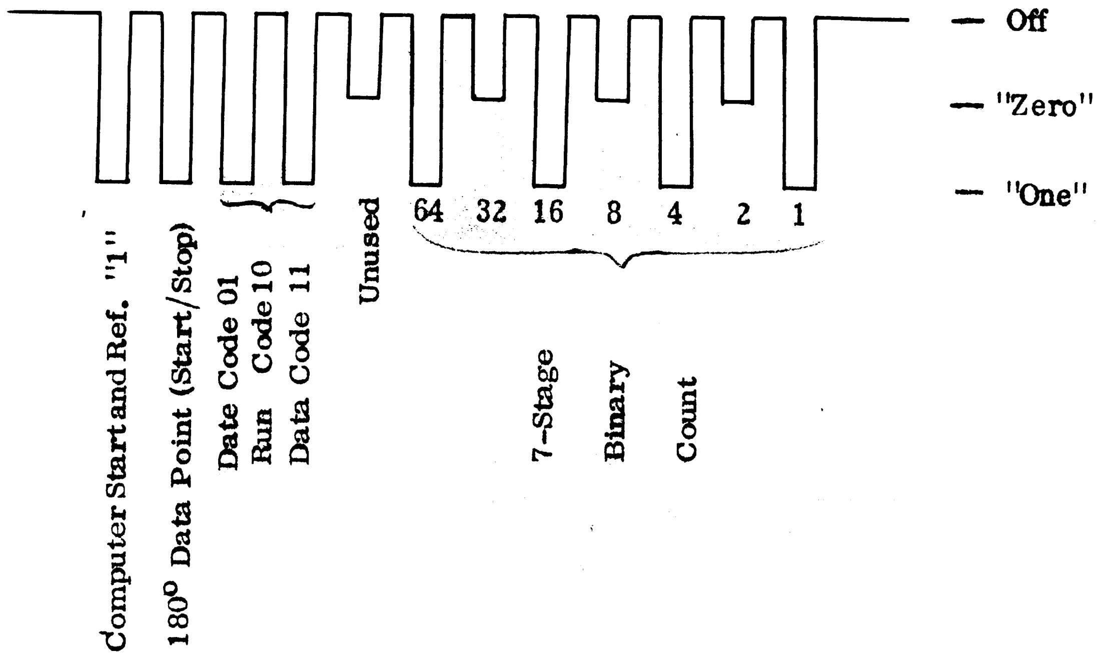

FIG. A-2: DATA POINT AT 180° ANGLE. AMPLITUDE AT 85.

For a date or a run number, the format is slightly different. Any number from 00 to 99 can be entered by setting two rotary switches to its component "tens" and "units." For the number "37", the "tens" switch is set to 3, and the "units" switch to "7". The last eight pulse positions are used to transfer this number to the tape according to the format shown in Fig. A-3.

A complete antenna pattern consists of 365 words (a "word" is one 12-pulse train). First, the date is recorded with three words, e.g., 3-18-65, then the run number (from 00 to 99), then 361 data points. A five-second burst of a one-kilocycle square wave is recorded at the end of each pattern as a distinct aural indication for the playback operator.

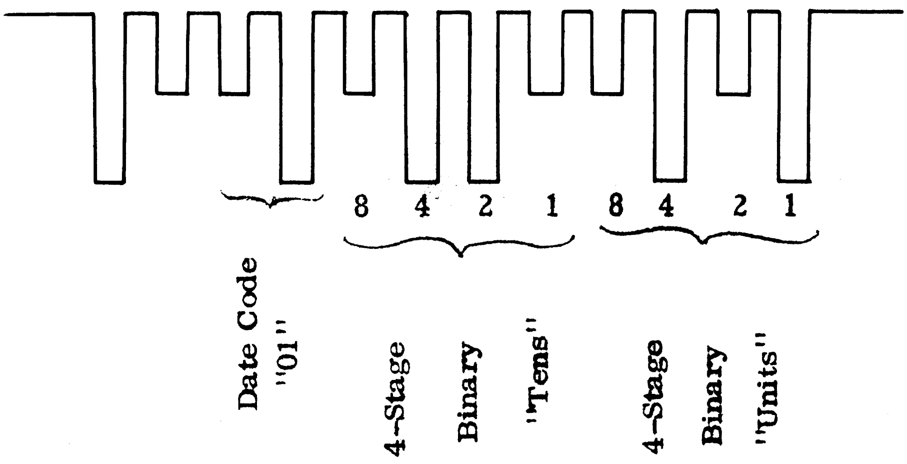

FIG. A-3: DATE ENTRY "65".