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ABSTRACT

This report discusses the construction and operation of the aximuth-elevation direction finder. It considers the proposed antennas and the antenna feed networks and the techniques being developed to achieve the design specifications. While the theory is well understood there are experimental design problems which are discussed along with their effect on the basic feed network. Several methods of signal detection are described with relative merit for each system presented. The report illustrates the function of the computer in the signal detection system and in the computation of aximuth and elevation angles of the monitored signal. Several small computers and analog-to-digital converters were considered during this period and the final choice is described. A brief survey of the effects of ground reflection on the operation of the system is given. However, due to the complexity of the problem no definite conclusions have been reached.

FOREWORD

This report was prepared by The University of Michigan Radiation Laboratory of the Department of Electrical Engineering under United States Army Electronics Command Contract No. DAAB07-67-C0547. This contract was initiated under United States Army Project No. 5A6 79191 D902-05-11 "Azimuth and Elevation Direction Finder Techniques". The work is administered under the direction of the Electronics Warfare Division, Advanced Techniques Branch at Fort Monmouth, New Jersey. Mr. S. Stiber is the Project Manager and Mr. E. Ivone is the Contract Monitor.

The authors wish to express their thanks to Dr. B. L. J. Rao for his theoretical contributions; to Messrs. A. J. Loudon, D.R. Marble, E.C. Bublitz and C. D. Spragg for their efforts in the experimental work performed during this period, and to Mr. P.H. Wilcox for preparing the computer program required for the three-dimensional analysis.

The material reported herein represents the results of the continuing investigation into the study of techniques designing a broadband circularly polarized azimuth and elevation direction finder antenna.

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I

INTRODUCTION

This report describes the construction and operation of the components in the azimuth and elevation direction finder presently being designed and fabricated by the Radiation Laboratory. The azimuth-elevation direction finder consists of an antenna system, electromechanical switch, receiver (GFE), Automatic Signal Recognition unit (GFE), an A to D converter, computer, and a visual display.

The antenna system consists of 16 to 17 antennas mounted on an arbitrary surface. Each antenna will have associated with it a particular θ and ϕ coordinate of the spherical coordinate system. Direction finding is accomplished by assigning to each antenna a vector whose amplitude is that of the received signal with the direction given by the outward pointing normal for the antenna receiving the signal. The signal information from the 17 antennas is fed into the computer where it is vectorially added, and the sum vector has θ and ϕ coordinates in the direction of the incoming signal.

The antennas are multiplexed by a rotating electromechanical switch. Tentatively, the switch will have three rotational rates, 10, 100, and 1000 rpm. The operator will then be able to select the optimum switching rate, which will depend upon the types of signals being interrogated by the azimuth and elevation direction finder. At present, signals which are considered to be of the greatest concern are those from search radars. Since the antennas are time multiplexed, one would like to have a very high switch rotational rate to assure interception of all radar pulses, especially those of a long range ground based rotating antenna. However, one must be careful not to switch at a rate faster than the computer cycling rate or the response time of the receiver. The cycling rate of the computer determines the switching rate since the inverse of the switch rotational rate times the number of antennas gives the time allowed for the computer to accept data from each antenna. Consideration must also be given to direction finding signals with CW, AM, and FM modulation. However, it is generally agreed that these type signals will not present as severe a problem to this system as those of the radar pulses and do not directly affect the switch rotational rate.

The switch will consist of 16 or 17 input ports with one output port. In addition to the signal input, an identification or reference signal must be available in the switch to identify to the computer which antenna is being interrogated. This identification is conceived to be in the form of a light beam to be interrupted as the switch output port slews past each of the input ports. It is anticipated that the switch will interrogate each of the antennas such that information will be collected and stored in the computer in a sequential format. For example, the first antenna would be associated that $\theta = 0^o$, $\phi = 0^o$ in the spherical coordinate system of Fig. 1-1; the second antenna would be at $\theta = 45^o$, $\phi = 0^o$; the third antenna at position $\theta = 45^o$, $\phi = 45^o$, etc., to the last antenna. However, at the present time the effect of picking antennas that are widely separated in space orientation is being considered. This sequence of interrogation may have advantages for radar pulse detection in that a greater portion of the sky would be scanned in a shorter period than the cycle time of the switch, thereby increasing detection probability for fast moving beams. The accuracy per single scan probably would not be within contract requirements of 2^o for azimuth and 5^o for elevation, but consolidating information for several scans should give improved accuracy. At present the mounting surface has not been chosen but as is shown in Section III, on mounting surfaces and Section VI on ground reflections, there may be improved accuracy by using a surface that is planar rather than hemispherical as originally proposed.

The electromechanical switch has a dual function to perform: 1) switch the microwave energy, and 2) provide a reference signal for the computer. The reference signal identifying the coordinates of the interrogated antenna is to be transferred from the switch directly to the computer. After passing through the switch, the microwave signals are amplified and reduced to an IF frequency by a receiver.

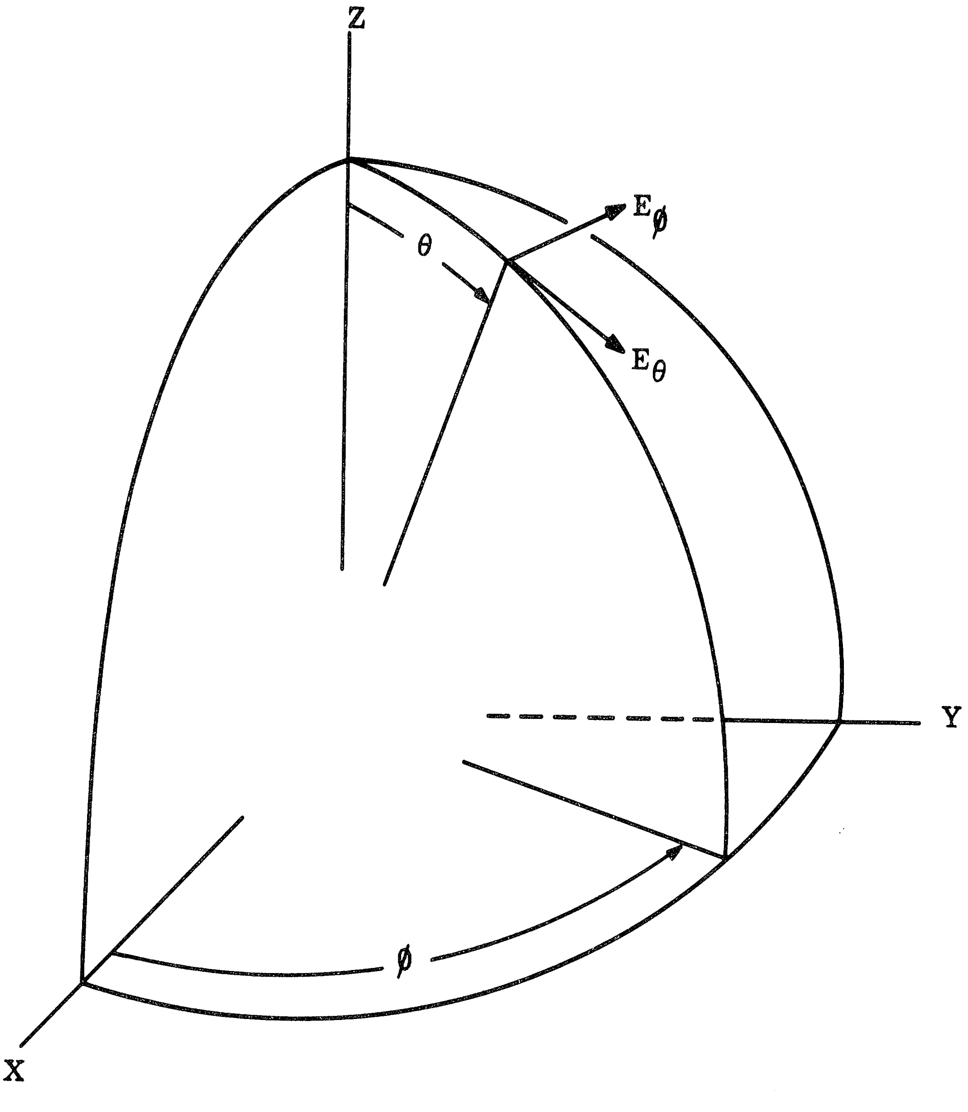


FIG. 1-1: Spherical Coordinate System Used For Azimuth-Elevation Direction Finder.

A video amplifier may be required between the receiver and the analog-to-digital converter to amplify the IF signal to the voltage level required by the analog-to-digital converter. In the event that more than one frequency is received within the bandwidth of the receiver, an automatic signal recognition unit (ASRU) would be required. It is understood that such units are available and have been employed by the military for this purpose. The data from the ASRU is then transferred to the analog-to-digital converter where it is digitalized in the proper format for the computer. Due to the requirements for detecting radar pulses which may be quite narrow in time, a standard analog-to-digital converter will probably not be used. Further discussion of the detection method for detecting narrow pulses during the entire switch aperture will be discussed in a later section (SectionV on the computer and the analog-to-digital converter).

In the computer the data is stored sequentially similar to that discussed above. For example, the first storage compartment would consist of data for the pole antenna, or the first antenna. The second storage compartment would have data from the second antenna or as in the example, $\theta = 45^{o}$, $\phi = 0^{o}$. Each compartment would then have the data stored for a definite space orientation of that antenna throughout all 17 antennas. When all 17 antennas have been interrogated the computer will then vectorially add all the 17 vectors in storage and take the arc tangents to obtain the θ and ϕ direction of the arriving signal.

Accuracy of this system depends upon a well controlled antenna pattern with the limitations being discussed by Ferris, et al (1967). The present status of this antenna development and the feed system for the antenna is discussed in Section II on the antenna and feed system design.

II

ANTENNA SYSTEM AND PHASING NETWORK

During the first part of the experimental program a cavity-backed spiral was tested over the 5:1 band from 600 to 3000 MHz. The cavity-backed spiral was used to achieve the circular polarization required by the contract, but early in the program it became evident that a cavity-backed spiral could not give the antenna pattern control necessary over the required 5:1 frequency band.

A cavity-backed spiral with the $\lambda/4$ deep cavity has the proper boundary condition for reflected wave reinforcement only for a narrow range of frequencies. Operating such antennas over more than one octave results in pattern deterioration with the main beam of the antenna pointing off-axis. It is desired that the antenna for this direction finding system should have a cosine pattern and that the maximum of the beam should be normal to the axis of the antenna. Because of the poor pattern control over the 5:1 band, the cavity-backed spiral was discarded and an investigation of the log conical was started. As was reported in the previous quarterly (Ferris, et al, 1967), a bifilar log spiral will set up moding at the higher frequencies. This moding results from higher order current bands that formed when the spirals electrical circumference is equal to $(n + m) \lambda$ where n is equal to 1 for the summation mode and 2 for the delta mode and m is equal to the number of filaments. A bifilar antenna operated in the sum mode (i.e., pattern maximum on-axis) has higher order modes at the third and fifth harmonics of the fundamental. Of course at all frequencies above the third harmonic, there would be some region of the antenna which will support the third harmonic mode. For a multifilar element the radiating current bands become more randomly distributed as the number of filaments increases. The quadrafilar spiral, whose four terminals are fed $0^{\circ}, 90^{\circ}, 180^{\circ}, 270^{\circ}$, will have a radiating band when the electrical circumference is 1 wavelength in diameter and the next higher order radiating band occurs when the spiral is 5 wavelengths in diameter. Although the quadrifilar spiral will be operating at the high frequency limit of the

5:1 frequency band near the region of the 5 wavelength radiating band, it is felt that for a feasibility study this antenna is a compromise between desired operating characteristics and cost for phasing networks. Better pattern control can be achieved by a further increase of the number of radiating windings, but a more complex phasing network and increased research and development time is required with the additional number of radiating arms.

The quadrifilar feed network requires a power split and a phase shift which is more easily accomplished over these broadbands by stripline type transmission lines. A stripline phasing network that would give 0^{o}, 90^{o}, 180^{o} and 270^{o} phase distribution across the 5:1 band, would employ three 3 db couplers and one 90^{o} phase shifter as shown in Fig. 2-1. A 3 db coupler in stripline has the property that the output arm has a 90^{o} phase shift from the non-coupled port, and this phase shift is relatively constant across the bandwidth of operation of the 3 db coupler as long as the 3 db coupling is maintained. By going through a 3 db coupler the outputs display equal power and have a 90^{o} phase shift with respect to each other. A total phase shift of 180^{o} is achieved by placing a 90^{o} phase shifter in the output ports of the 3 db coupler. The addition of the 90^{o} phase shifter causes the 3 db hybrid to display the characteristics of a magic tee. If 3 db hybrid couplers are placed in each of the two outputs from the first 3 db coupler and 90^{o} phase shifter, the resulting four outputs from the directional couplers are now divided equally in power and the phase of these four equal amplitudes is 0^{o}, 90^{o}, 180^{o} and 270^{o}.

The 3 db coupler and 90^{o} phase shifter are constructed from the data that has been documented by Shelton and Mosko (1966). By using a technique originally conceived by Shiffman (1958), a broadband phase shifter or hybrid coupler can be constructed to have an arbitrarily large frequency band. As the bandwidth increases, there is a similar increase in the number of quarterwave sections that are required in the directional coupler and in the phase shifter. These sections are a quarter wavelength at the mid frequency of the operating band. It is well known that a 3 db

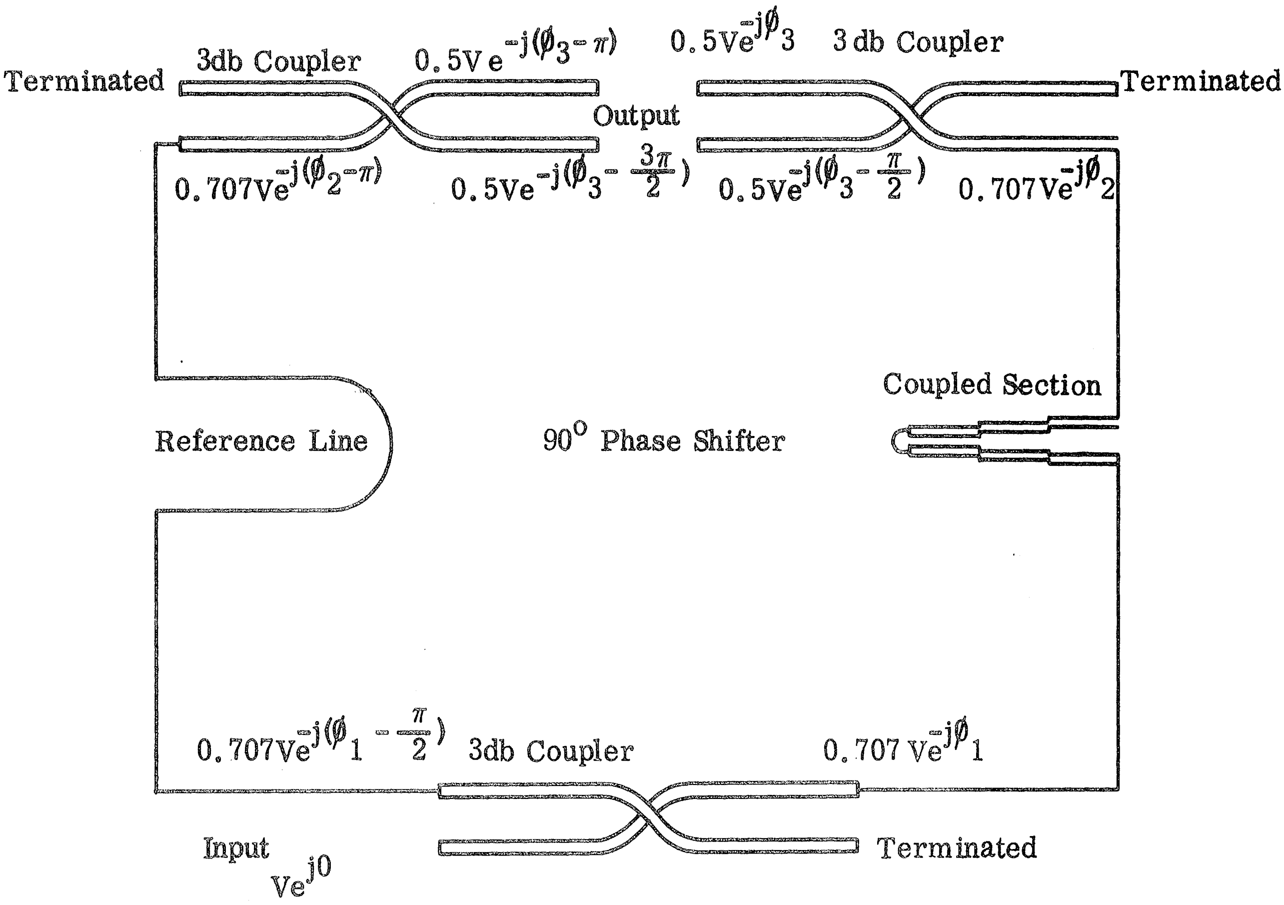


FIG. 2-1: Antenna Feed Network

coupler can be made with a single quarterwave length section. However, the bandwidth at most for such a single quarterwave section is approximately one octave. Figure 2-2 is a drawing of a simple quarterwave length coupler. If energy is fed into Port 1, it will flow out of Port 2; Port 4 is the coupled port and Port 3 is an isolation port. When the section is one quarterwave length long, the induced voltages will add at Port 4 to give the coupled port, but the equations show that at Port 3 the voltages cancel giving an isolation port.

Coupling in the directional coupler is accomplished by perturbing the even and odd mode of fields. Equation (2.1) shows that the coupling coefficient Γ is proportional to the impedance discontinuities of either the even or the odd mode.

$$\Gamma_j = \frac{\sqrt{\frac{Z_{oej+1}}{Z_{ooj+1}}} - \sqrt{\frac{Z_{oej}}{Z_{ooj}}}}{\sqrt{\frac{Z_{oej+1}}{Z_{ooj+1}}} + \sqrt{\frac{Z_{oej}}{Z_{ooj}}}} = \frac{Z_{oej+1} - Z_{oej}}{Z_{oej+1} + Z_{oej}} \quad . \tag{2.1}$$

It makes no difference whether we talk of the even mode or odd mode, since the couplers are designed to have a constant characteristic impedance, Z_o, according to equation (2.2). From equation (2.1) we see that a wave entering from a 50 Ω line into the coupler at Port 1 would be going from matched even and odd mode impedance into the same characteristic impedance with a mode impedance mismatch. The even

$$Z_o = \sqrt{Z_{oe}\, Z_{oo}} \tag{2.2}$$

mode impedance of the quarterwave line is higher than the feed line so that the Γ or the induced voltage at Port 4 will be a positive Γ while at Port 2 the fields are going from a higher even mode impedance to a lower even mode impedance and Γ will be negative as shown in Eq. (2.3). Each of these induced voltages will cause a wave to travel to the right and to the left at Ports 3 and 4. Equation (2.3) shows the

$$\Gamma_j = \frac{Z_{oej+1} - Z_{oej}}{Z_{oej+1} + Z_{oej}}$$

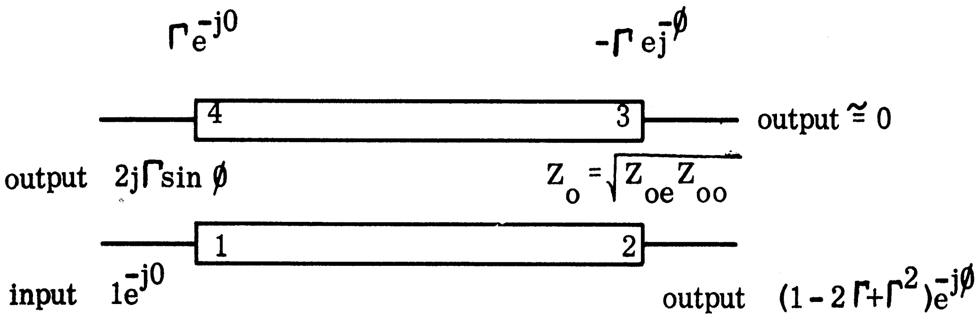


FIG. 2-2: Single Section Quarterwave Directional Coupler

magnitude and phase of these induced voltages.

$$\text{at Port 4} \quad +\Gamma_1 \, e^{-jo}, \qquad \text{at Port 3} \quad -\Gamma_1 e^{-j\phi} \tag{2.3}$$

ϕ = electrical length of the stripline

Γ_1 = voltage coupling coefficient .

Equation (2.4) indicates the result of summing these two traveling voltages waves at Port 3.

$$+\Gamma_1 \, e^{-j\phi} - \Gamma_1 \, e^{-j\phi} = 0 \quad . \tag{2.4}$$

The summation of the voltages at Port 3 causes cancellation of the voltage, therefore, no energy will appear at this port. However, in equation (2.5) the voltages add at Port 4. It is evident from equation (2.5) that the voltages will add giving a maximum voltage at Port 4 when the line is $\lambda/4$ long, i.e., the magnitude of the induced voltage is proportional to sin ϕ where ϕ is the electrical length of the line. Note that the voltage output from Port 4 is $+2j\Gamma_1 \, e^{-j\phi} \sin\phi$ and that of Port 2 is $(1-2\Gamma_1 + \Gamma_1^2)e^{-j\phi}$.

$$+\Gamma_1 \, e^{-jo} - \Gamma_1 \, e^{-j2\phi} = \Gamma_1 \, e^{-j\phi} \, (e^{+j\phi} - e^{-j\phi})$$

$$= +2j\,\Gamma_1 \, e^{-j\phi} \sin\phi \quad . \tag{2.5}$$

There is a net phase difference of 90° between these two as indicated by the j in the Port 4 term. This $\lambda/4$ coupler gives the desired 3 db coupling and the 90° phase shift between the output ports, but has a limited range of 3 db coupling due to the sin ϕ term which is proportional to frequency.

When one wants to build a square wave with a Fourier series, it takes higher order terms to build up the flat response across the top and extend the width of the square wave. Frequency response of a directional coupler is broadbanded in much the same way by adding more terms or more sections to the primary center section.

Figure 2-3 is a schematic of a three section directional coupler. Equation (2.6) gives the voltage response at Port 3 and Port 4. Here again Γ_2 at Port 4 is positive and Γ_2 at Port 3 is negative for the same reason as discussed above for the single section coupler:

$$\text{at Port 4} \quad +\Gamma_2 e^{-j0} \,, \qquad \text{at Port 3} \quad -\Gamma_2 e^{-j3\phi} \;. \tag{2.6}$$

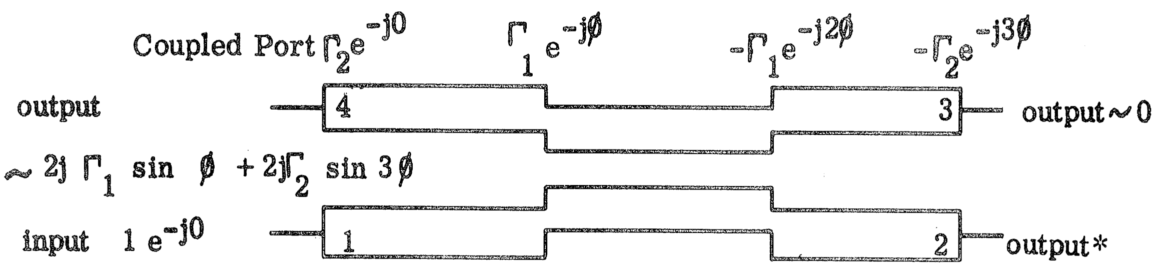
Equation (2.7) shows the summation of these voltages waves at Port 3. As in the equations for the single quarterwave coupler, there is a cancellation of these voltage waves giving no power out of Port 3:

$$+\Gamma_2 e^{-j3\phi} - \Gamma_2 e^{-j3\phi} = 0 \;. \tag{2.7}$$

Equation (2.8) illustrates the addition of the voltages at Port 4. The wave induced at Port 3 traveled an electrical distance of 3ϕ from the reference at Port 1, and an additional 3ϕ from Port 3 to Port 4 for a total distance of 6ϕ. A combination of these two waves, the one induced at Port 4 and one that traveled from Port 3 to Port 4, gives a term that is proportional to sin 3ϕ. From this, it is evident that each time two symmetrical sections are added to a directional coupler, the coupling is increased by adding odd harmonics in ϕ. In this way, the coupling will start with sin ϕ for the central term, the next symmetric sections couple proportional to sin 3ϕ, the next sections couple proportional to sin 5ϕ, etc.

$$\Gamma_2 e^{j0} - \Gamma_2 e^{-j6\phi} = \Gamma_2 e^{-j3\phi} \left(e^{j3\phi} - e^{-j3\phi}\right) = 2j\,\Gamma_2 e^{-j3\phi} \sin\phi \;. \tag{2.8}$$

Looking at equation (2.8) it is also apparent that this three section coupler has the desirable quality that the two output ports are separated in phase by 90°. By choosing the proper mode impedance mismatches, therefore giving the proper reflection coefficients at these various junctions, it is possible to design uniform coupling across a broad frequency band in the same manner that a square wave is constructed with a Fourier series. Coupling between sections depends upon reflections established



$$* \left(1 - 2\Gamma_1 - 2\Gamma_2 + 4\Gamma_1\Gamma_2 - 2\Gamma_1^{\,2}\Gamma_2 - 2\Gamma_1\Gamma_2^{\,2} + \Gamma_1^{\,2} + \Gamma_2^{\,2} + \Gamma_1^{\,2}\Gamma_2^{\,2}\right) e^{-j3\phi}$$

FIG. 2-3: Three Section Quarterwave Coupler by impedance mismatches caused by the variation of strip width and overlap of the coupled section. All calculations are based on the assumption of no higher order reflections, i.e., no mutual coupling between sections. For this reason it is imperative that all sections be designed with a constant characteristic impedance.

As in the case of Fourier series, the first term has the largest coefficient. The succeeding terms build up the square wave response at the edges and reduce the ripple. For the 7 section 8:1 band 3 db directional coupler, the major contour is established by the center section and the first section outside the center section. The remaining terms are added to decrease the ripple.

To cover the 5:1 bandwidth we have chosen an 8:1 band directional coupler listed in the design tables by Mosko and Shelton. These design tables were derived by Mosco and Shelton using a computer iteration technique that would choose the values of the coefficients, Γ_1, Γ_2, Γ_3, etc. An error curve was obtained by summing the response from the Γ's in terms of ϕ and comparing with a desired curve. Computer analysis of the error curve changed the Γ's and the process was repeated until the response of the coupler was within the specified requirements in the computer program. These variables specified in the program are bandwidth, ripple, number of sections allowed and tolerance. The ripple is the average deviation from the desired coupling. Of course, some peaks will be higher than this and the tolerance indicates how much higher than the average these will be.

A 3 db coupler is more difficult to build than those of lower coupling, as they require closer coupling and greater construction accuracy. For this tighter coupling it is generally the accepted practice to design the coupler as two tandem -8.3 db couplers. Two similar -8.3 db couplers connected in tandem will display -3 db coupling over the bandwidth of the operation of the -8.3 db couplers. The tandem design allows a larger separation between the ground planes and less coupling in the center sections. Increased distance between the two ground planes minimizes variation in the coupler due to mechanical deformation in the two ground planes.

Coupling tolerances become quite critical for broadband high voltage coupling coefficients. The coupler is designed in terms of reflections from the even and odd mode impedances. Knowing these reflections, one is able to estimate the proper strip width and strip overlap to give the 3 db coupling. This strip width is approximate for several reasons: 1) the dielectric has variations in dielectric constant that become apparent at microwave frequencies, 2) at microwave frequencies the stripline tends to radiate so the walls of the stripline are lined with screws forming a rectangular TEM transmission line, 3) not only do the screws in the sidewalls perturb the waves but they also deform the ground plane locally and possibly compress the dielectric, and 4) these theoretical curves are for exact tolerances which one is not able to achieve in practice. The effect of the screws upon the impedance is shown by a plot of a 50 Ω stripline or a coupler using the Time Domain Reflectometer (TDR). Such a plot reveals a series of small impedance discontinuities along the line. A similar plot of an air line with the TDR system shows the air line has a nearly constant impedance level. Upon investigating this effect, it was found the discontinuities due to the screws could be increased or decreased according to the tension on the screws. Another effect to be studied is the effect on the impedance caused by varying the distance between the screws and the center conductor. The design of the coupler is based upon maintaining a characteristic 50 Ω impedance at each section throughout the entire coupler, and any undesired small perturbations can change the broadband response.

Another very important aspect of the coupler design is the requirement for reflection free connectors. Stripline connectors should have a very low VSWR, if not they will act as an impedance mismatch producing unwanted reflections. Early work on the coupler used modified Type N UG/58-A panel connectors. The large center pin introduced capacitance into the junction with the stripline masking some of the problems in the stripline itself. These reflections were reduced by changing to Omni-Spectra* (OSM) stripline connectors.

Mosko and Shelton have a very good theoretical representation of the problem of stripline coupler design. However, they do not give a systematic technique for correction of fabrication errors that result as noted above. It is most important to recognize the type of error in the coupler and source of this error. At the present time we know how to correct for these errors and this will be illustrated in Table 2-1. It is noticed there that for the first try we were wanting a Γ_1 of .559 and that we had an actual Γ of .422. In the second try, the actual Γ's of the first attempt were determined by taking four points in ϕ, i.e., frequency, and measuring the coupling at these points. Given the four coupling values and with four unknown coupling coefficients it was possible to determine the individual coupling coefficients that were actually achieved. However, by using only four points any experimental error would also be calculated. To obtain a more accurate representation of the actual coupling, a large number of coupling values are measured as a function of frequency and by this over-determination of the coupling data, the best or most probable coupling coefficients are calculated by the computer. For the second try, a linear error analysis was performed and determined that Γ_1 should be .700 on the design curve to obtain a .559 coupling. This represented 100 per cent overlap of the center section and consequently the maximum coupling that could be achieved for this ratio of center conductor separation to distance between ground planes, s/b = 1/9. Notice the effect of coupling tolerances on the 4th try, the coupling Γ_1 was again given for .700 with the same strip width and overlap but measured Γ_1 was .519 in this case. When the maximum coupling is required small variations in the strip widths produce large variations in the impedance introducing undesired reflections in the line. A more critical factor is that two center sections must be fully overlapped. Errors of a few mills of the overlapping of the center conductors will produce a large variation in the coupling between the two lines. Shelton (1965) also indicates that a variation

* Trademark brand name

of 1 mil for the center conductor separation will produce 0.14 db error in coupling.

Figure 2-4 shows the individual sin ϕ, sin 3ϕ, sin 5ϕ, and sin 7ϕ components, their relative amplitudes and the total coupling desired across the frequency band.

TABLE 2-1

First Try

Design	$2\Gamma_1 = .559$	$2\Gamma_2 = .168$	$2\Gamma_3 = .058$	$2\Gamma_4 = .011$
Actual	$2\Gamma_1 = .422$	$2\Gamma_2 = .069$	$2\Gamma_3 = .028$	$2\Gamma_4 = .013$

Second Try

Design	$2\Gamma_1 = .700$	$2\Gamma_2 = .261$	$2\Gamma_3 = .085$	$2\Gamma_4 = .007$
Actual	$2\Gamma_1 = .501$	$2\Gamma_2 = .249$	$2\Gamma_3 = .036$	$2\Gamma_4 = .029$

Fourth Try

Design	$2\Gamma_1 = .700$	$2\Gamma_2 = .298$	$2\Gamma_3 = .073$	$2\Gamma_4 = .001$
Actual	$2\Gamma_1 = .519$	$2\Gamma_2 = .161$	$2\Gamma_3 = .041$	$2\Gamma_4 = .001$

The overall coupling of a -8.3 db represents a voltage coupling of 0.385. By this simple graphical technique, we do predict, with reasonable accuracy, the ripple quoted by Shelton (1965), i.e., approximately 0.2 db. Figure 2-5 is a measured coupling curve for one -8.3 db section of the tandem coupler and the second curve is the one produced from the summation of the calculated Γ's that were achieved.

In Fig. 2-5 we see the values that were obtained by calculating Γ's from the experimental curves and then reconstructing this curve to show a slight error due to the experimental error of the four values measured. In general the overall curve is fairly close and is good enough to give an indication of the coupling error due to

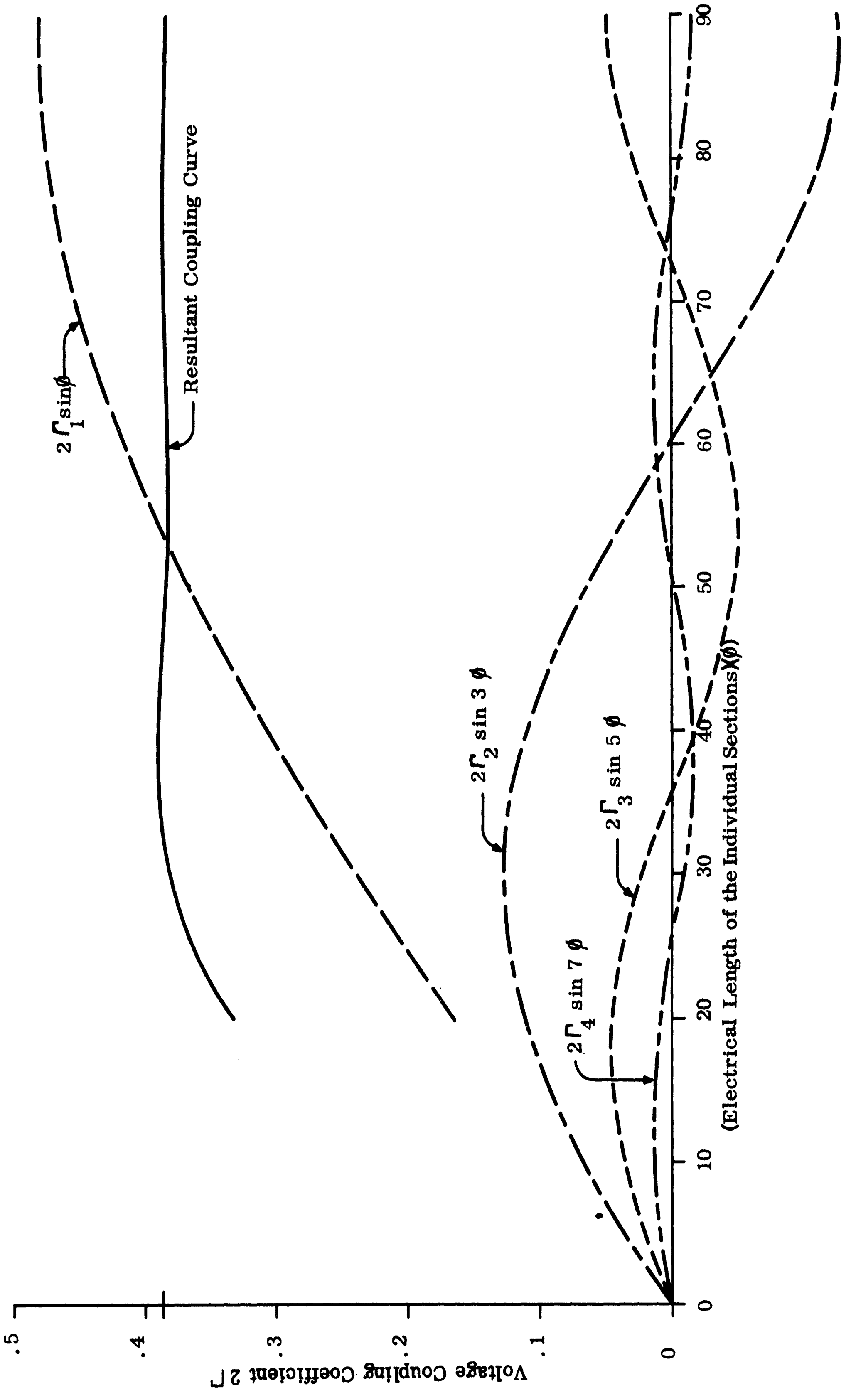


FIG. 2-4: Theoretical Coupling Curve for Seven Section Coupler.
$2\Gamma_1 = .478$, $2\Gamma_2 = .129$, $2\Gamma_3 = .049$, $2\Gamma_4 = .013$

the improper impedance levels in the coupler. The major difficulty at this time is controlling in the coupling level at the extremes of the frequency band. Referring to Fig. 2-4, it becomes apparent that to increase the coupling at the low frequencies the Γ_1 and Γ_2 must be increased. Γ_2 cannot be increased alone since it becomes negative at the center frequency, 1800 MHz, so that the net effect is to cause the coupling to become low at this frequency. Changing the Γ_1 and the Γ_2 will also change the ripple slightly so that the third and fourth coupling coefficient must also be changed to ensure the proper ripple limit across the band. Our past experience has shown that it is difficult to increase the Γ_1 and Γ_2 to improve low frequency coupling without overcoupling at the lower frequencies. This is due to the very tight tolerances that must be obtained in terms of strip width and the overlap for the center sections. Normally the coupler center conductors are etched from double clad dielectric sheets by a process that requires an accurate negative. The original drawing is drawn oversize, generally 5 to 10:1 and this drawing is photo reduced to minimize errors in the original drawing. To reduce costs of the fabrication of trial models, the first models have been made by cutting the conducting center sections out of 1 mil brass shimstock. The difficulty in cutting this brass shimstock within 1 mil tolerance can be appreciated when you consider that the thin material will deform to have a somewhat irregular edge. Another difficulty after the shimstock is cut is the alignment of the two conductors to produce the proper overlap.

Early attempts at etching the center conductors have not significantly improved the tolerances. The material presently being used is Rexolene S which is an irradiated dielectric with a dielectric constant of 2.32. The irradiating process lowers the loss tangent at the higher frequencies but this causes the material to buckle (Fig. 2-6) when etched making it quite difficult to hold close tolerances. The buckled material has air gaps in the coupler even with screws around the edge. These air gaps have an adverse effect since the change in dielectric constant affects the odd modes more than the even modes. Rexolite with a dielectric constant of 2.62 has

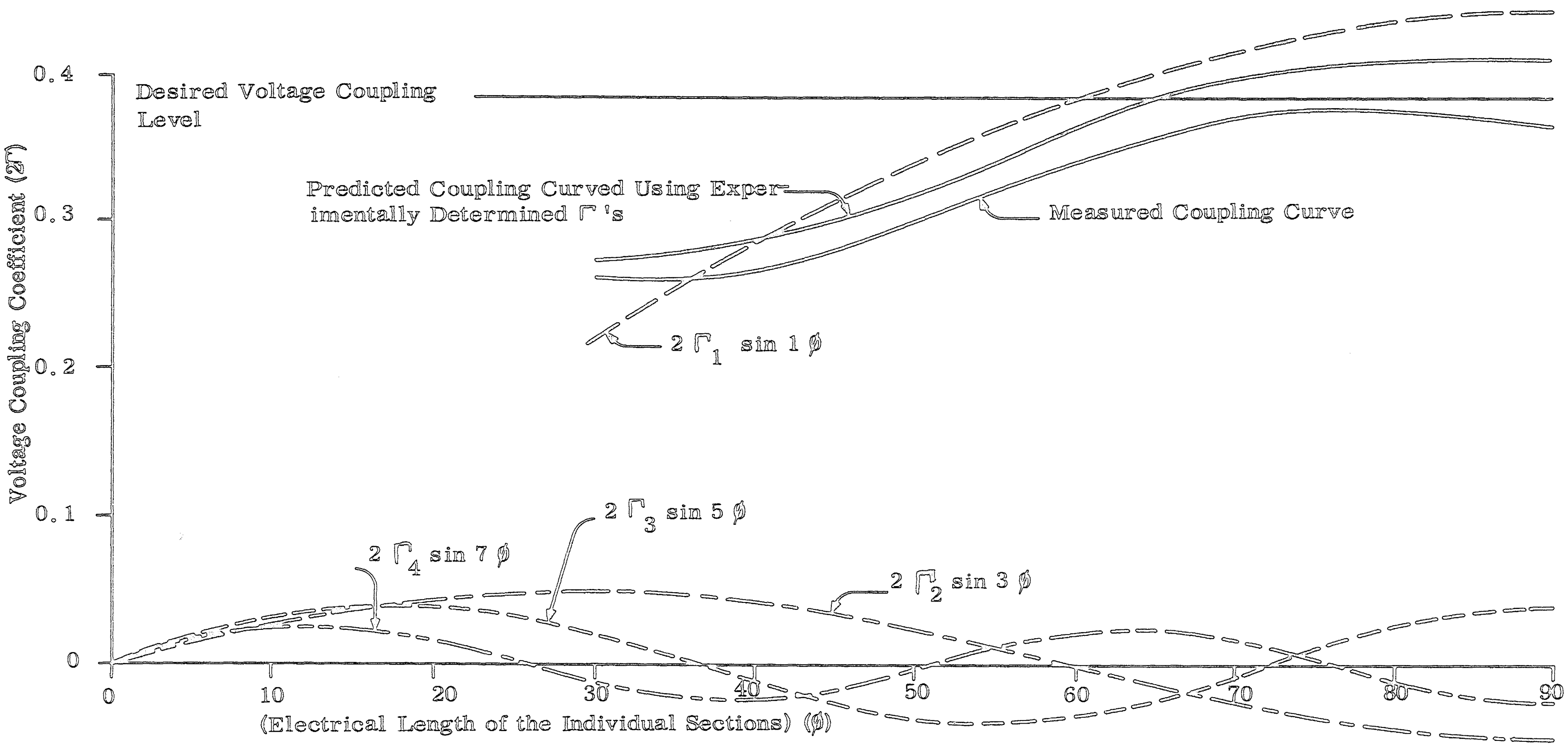


FIG. 2-5: Graph of Measured Coupling Curve and Coupling Curve Predicted From Experimentally Determined Γ's. $2\Gamma_1 = 0.441$, $2\Gamma_2 = 0.050$, $2\Gamma_3 = 0.040$, $2\Gamma_4 = 0.027$

been ordered in an effort to overcome this problem.

The phase shifter is more complex such that its operation is difficult to describe theoretically. The phase shifter is constructed by shorting the two center sections some distance from the input. This distance is an integral number of quarter wavelengths at the center frequency. By shorting the coupled sections, the reflection coefficient is unity and the phase of these coefficients determines the phase shift characteristics of the component.

The basic coupled section configuration for the phase shifter is similar to that of the directional coupler. The mode impedance mismatches still maintain a constant characteristic impedance. The coupled section is terminated by a short between the two center conductors. This termination "appears" as a short circuit to the odd mode, whose major energy components are confined between the two center conductors.

The termination "appears" to be an open circuit to the even mode impedance whose energy is primarily confined between the two center conductors and the ground plane. This oversimplified model can be used to predict the general operation of the phase shifter. Consider the energy in the odd mode. The normalized input impedance for a shorted transmission line is given by

$$Z_{in} = Z \tanh \gamma \ell \qquad \text{where } \gamma = \alpha + j\beta \text{ ,} \tag{2.9}$$

$$Z = \text{odd mode line impedance .}$$

If the line is assumed lossless this becomes

$$Z_{in} = Z \tanh (j\beta\ell) = j\, Z \tan \phi \text{ since } \phi = \text{electrical length of line} = \beta\ell$$

$$\text{and } \tanh\, j\phi = j \tan \phi \text{ .} \tag{2.10}$$

From the transmission line theory, the input reflection coefficient is given by

$$\Gamma_{in} = \frac{Z_{in} - 1}{Z_{in} + 1} \text{ .} \tag{2.11}$$

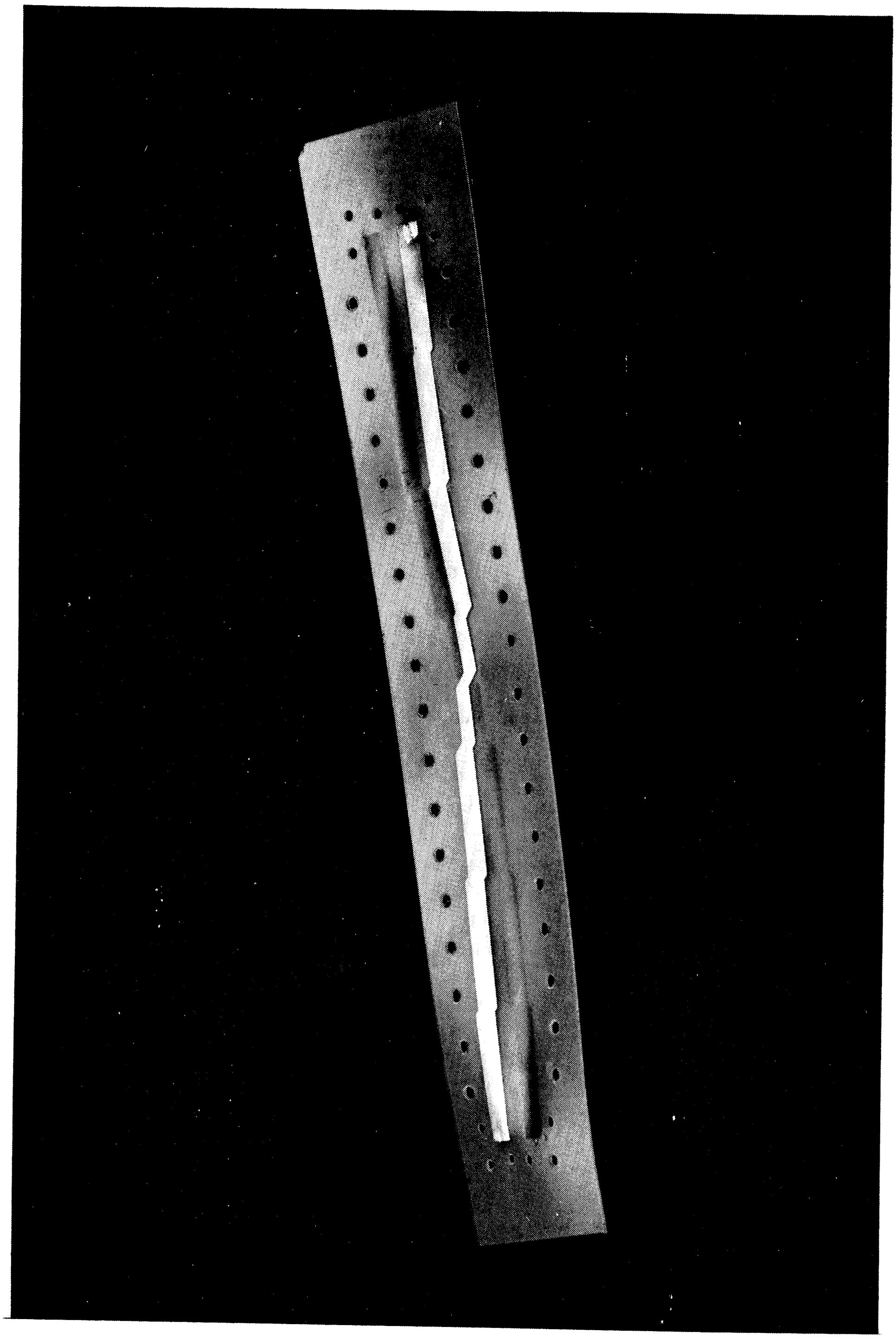


FIG. 2-6: Irradiated Dielectric Stripline after Etching Process

Substituting the value for Z_{in} from equtaion (2.9) the input reflection coefficient becomes

$$\Gamma_1 = \frac{jZ\tan\phi - 1}{jZ\tan\phi + 1} = \text{amplitude } e^{-j\theta}. \tag{2.12}$$

The phase of the reflection coefficient (θ) is

$$\theta = -2\tan^{-1}(Z\tan\phi) \tag{2.13}$$

$$\theta/2 = \tan^{-1}(Z\tan\phi)$$

$$\tan\theta/2 = Z\tan\theta \qquad \theta = \text{absolute phase through coupled section.}$$

Now if a new angle is defined $\mu = \theta/2 - \phi$ and the tangent is taken of both sides the right hand side of the equation can be expanded in a double angle formulation. After some manipulation the equation reduces to

$$\tan\mu = \frac{\left(\frac{1-Z}{1+Z}\right)\sin 2\phi}{1+\left(\frac{1-Z}{1+Z}\right)\cos 2\phi}. \tag{2.14}$$

It is seen that the quantity $\frac{1-Z}{1+Z}$ is the magnitude of the reflection coefficient Γ at the impedance step.

$$\tan\mu = \frac{\Gamma\sin 2\phi}{1+\Gamma\cos 2\phi}. \tag{2.15}$$

If the tangent is taken of both sides of the equation one obtains

$$\mu = \tan\left[\frac{\Gamma\sin 2\phi}{1+\Gamma\cos 2\phi}\right]. \tag{2.16}$$

If the reflections and phase dispersion are small and second order effects are neglected the equation (2.16) can be simplified. If the Γ is assumed to be small the ratio is small and the tan of a small number is approximately the small number.

Further if Γ is small, the Γ cos 2 ϕ term of the denominator can be neglected. Then μ can be expressed as

$$\mu \cong \Gamma \sin 2\phi \tag{2.17}$$

With additional sections the net phase shift μ becomes

$$\mu \cong \Gamma_1 \sin 2\phi + \Gamma_2 \sin 4\phi + \Gamma_3 \sin 6\phi + \ldots\ldots. \tag{2.18}$$

This net phase shift with several terms approaches a straight line slope with frequency. If a reference line with the same frequency slope is used as the other port any phase difference between the two lines is maintained across the linear phase shift region of the coupled region.

III

SURFACE REQUIREMENTS

In the first quarterly (Ferris, et al 1967) it was reported that a hemispherical surface would be required to mount the antennas. Since then efforts have been expended to reduce the undesirable effects of (the natural earth) ground reflections. As a part of this investigation, consideration was given to the antenna system geometry. Although the ground reflections will be discussed in Section VI, one aspect will be presented here. Since we are performing a simple vectorial addition of the relative amplitudes of antennas and no phase information is involved, it is not mandatory for the antennas to be mounted on a hemispherical surface. The surface can be of any arbitrary shape as long as the θ and ϕ pointing direction is known for the normal of each antenna in the radiating system. An alternate configuration would consist of a planar surface with 8 antennas in the outer ring pointed near the horizon. On an inner ring there could be 8 antennas pointed at 45^o in elevation, and a single antenna in the center pointed at the zenith. There may be some antenna blockage in the configuration.

IV

ELECTROMECHANICAL SWITCH

The electromechanical switch has been designed with capacitive coupling to reduce the noise generated by a rotating DC contact. The early model for this switch was a simple laboratory model that had linear motion only. At the present time an engineering model of the final configuration has been received from the shop and is being tested. This model (Fig. 4-1) is 10 inches in diameter with an inner rotating section (rotor) 9 inches in diameter. The 17 inputs are to be located on the stationary section (stator) of the switch on a 3-1/4 inch radius separated by a cord distance of 1-1/4 inches. The rotating port is located on a 3-1/4 inch radius of the rotor, with the energy being extracted from a rotary joint located at the center of the rotor. A stripline is utilized to transfer energy from the rotating port to the rotary joint at the center section shown in Fig. 4-1. In this figure, the top section is the rotor. The stripline has been removed and TNC connectors installed to aid in making experimental measurements of the individual VSWR values for the rotary joint and switch contacts.

At present the engineering model has four input ports located a cord distance of 1-1/4 inches apart as they will be in the final model. Measurements made for these four ports have shown that the decoupling between adjacent ports have at least 20 db isolation. The VSWR of individual parts of the switch have been measured without the stripline from 600 to 3000 MHz and exhibit VSWR's that range from 1.3:1 to 2.0:1 with respect to a 50 Ω line.

Data has been collected for various separations between the input and output ports and the best VSWR seems to occur when the spacing is between 0.003 and 0.004 inches which is felt to be an adequate separation for the 1000 rpm rotational rate. The capacitive coupled rotationg switch is difficult to design since the VSWR is related to: 1) size and geometry of the cavity behind the capacitive probe (Fig. 4-2), 2) the size and geometry of the capacitive probe, and 3) the spacing between coupling probes.

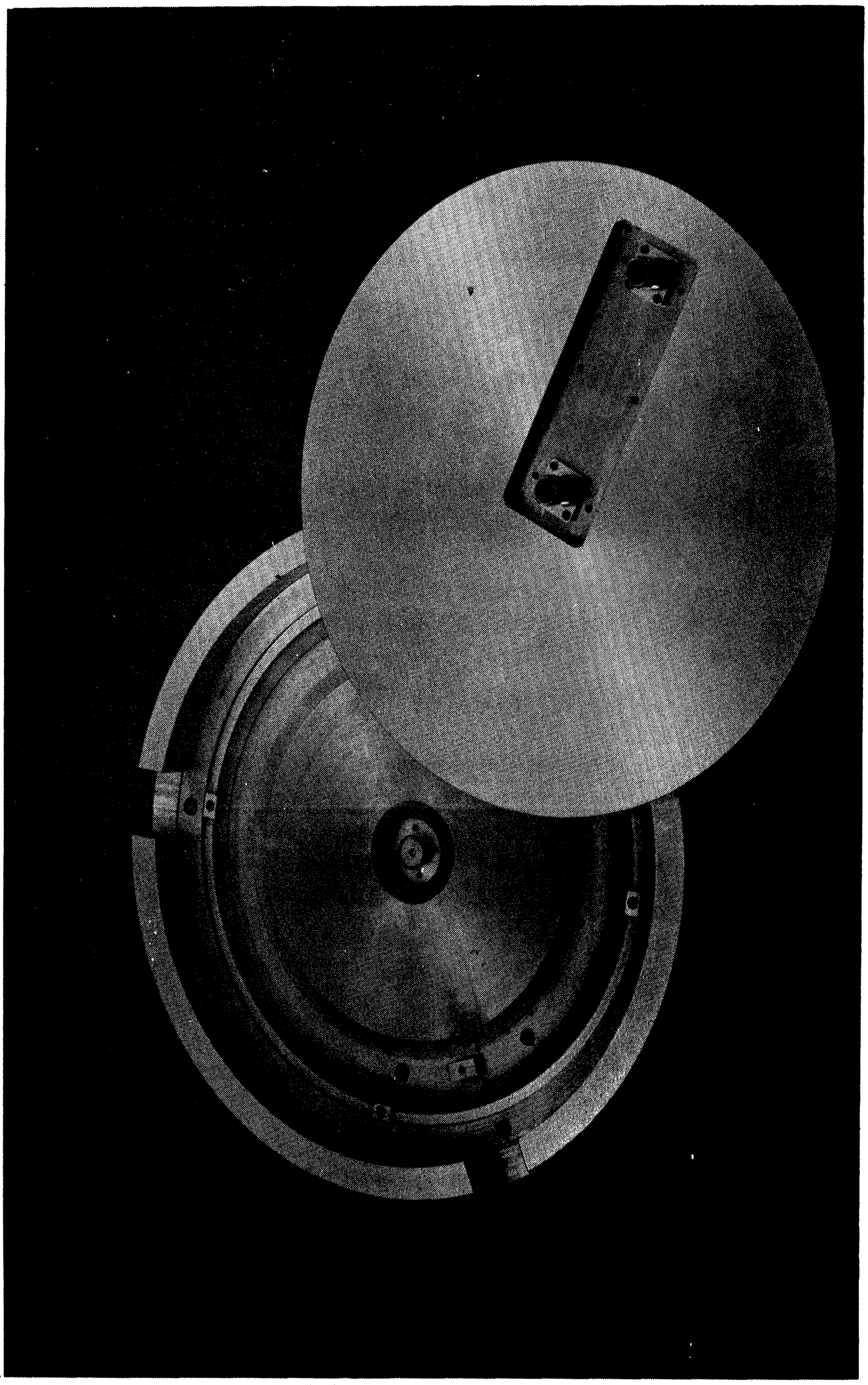


FIG. 4-1: Engineering Model of Electromechanical Switch

FIG. 4-2: Top Section Indicates Geometry of Rotating Capacitive Probe and Rotary Joint.

V

COMPUTER AND DETECTION SYSTEM

Pulsed signals present the most difficult problem in the detection of the various types of communication and radar signals. Analog-to-digital converters typically employ a sampling technique, with a standard sampling rate of 20 KHz/sec. With a 20 KHz sampling rate, there is a possibility that multiple radar pulses could be received in the switch aperture period and not be detected during the sampling intervals. More difficult to detect is a relatively long pulse from a rotating, long range, ground based radar system. If a small portion of this pulse appears in the switch aperture, it would appear to be a very narrow pulse appearing only once in the entire aperture interval.

There are alternatives to reduce the possibility of undetected signals. One method of detecting such small pulse width signals in the aperture is to increase the sampling rate up to 500 KHz. Five hundred KHz would appear to be the upper limit for the sampling rate due to the cycling time of most small, general purpose computers. Any sampling rate higher than 500 KHz would not allow the computer to accept the data from the analog-to-digital converter. This does not assure 100 per cent detection of the signal in the aperture, as there are regions during the conversion of the analog signal where there is no surveillance of the switch aperture. In this method, the sampler converts the voltage level to a digitalized form at all sample points and reads these into the computer, and the computer has to make a decision at the end of the interval as to the largest voltage.

Having the computer search through the voltages at the end of the period is a waste of the computer time, as the computer is not only tied up during the entire switch aperture accepting the data from the sampler, but must spend time after the switch aperture interval sorting information to find a single peak voltage. This method is an example of injudicious application of the computer since it becomes a data storage device with insufficient time allowed for real time computation.

Another analog-to-digital conversion method is the simultaneous method described in the Digital Logic Handbook (1967) in which the voltage is measured by a comparison to a reference. This method involves the construction of several level-detection circuits. Each circuit would be triggered if the amplitude of the voltage peaks in the input signal exceeded that particular level. At the end of the interval the voltage level of the desired peak is indicated by the highest triggered detector circuit and would be determined by the computer through an examination of a single output word. The output word would contain the unique bit position for each level. Thus, the highest bit position containing a bit would indicate the highest amplitude encountered in the interval. With this bit information the computer would correlate the actual voltage reading.

It should be noted that this technique does not require an A to D converter or sample and hold amplifier. It employs a continuous surveillance technique in that the peak voltage appearing during the switch aperture would trigger the comparison circuitry to the closest circuit level of the peak voltage.

The major problem of this technique is the cost associated with attaining accuracy throughout the whole range of voltages that may be expected from the incoming signals. The number of comparison circuits required is apparent when low level signals are considered. With ten detectors designed to cover a 5 volt range, the best accuracy that can be obtained in detecting the peak voltage is $\pm$ 0.5 volt. While this may be acceptable for an interval in which the peak is near the 5 volt level, it is inadequate when the peak voltage is near or below 1 volt. For the 5 volt signal, the 0.5 volt resolution represents a 10 per cent accuracy, but the accuracy of resolution is only about 50 per cent for the 1 volt signal. To improve the low signal resolution perhaps a logarithmic increase in the number of level detectors is required in the low voltage region.

The best method would appear to be a peak reading voltmeter, and this is the method that has been chosen. The peak reading voltmeter will accept all signals from

0 to 5 volts during the switch aperture and record the peak voltage during this time, provided the voltage spikes are at least 1 microsecond in width. The peak reading voltmeter has a field effect transistor (FET) with the control voltage being the input voltage from the receiver. The input voltage allows a certain amount of current through the FET. This current charges a capacitor in the circuit to the value of the peak voltage impressed upon the FET. The FET with its high source impedance will effectively appear to be an infinite impedance during the short switch intervals, thereby assuring no significant discharge of the capacitor. The detection system with the peak reading voltmeter is controlled by three computer generated pulse outputs and an interval timer. The first pulse from the computer is used to re-zero the circuit and must occur at least 50 microseconds before the beginning of the switch aperture. The second pulse output occurs at the beginning of the switch aperture and is used to direct the acceptance of the signal from the multiplexer into the input of the FET circuit. The peak reading voltmeter is switched "on" only after the VSWR of the electromechanical switch is low to prevent transients from producing false readings in the switch aperture. The interval timer in the computer program produces a program interrupt every 100 microseconds and by counting the number of these interrupts the switch aperture interval can be measured to the nearest 100 microseconds. At the end of the switch aperture the computer issues a command with a third pulse to open the FET gate and initiate an analog-to-digital conversion cycle. The A to D converter takes a 200 nanosecond sample of the output of the FET gate and stores this reading in a sample and hold amplifier. The contents of this amplifier are then converted to the 12 bit binary value. The end of the conversion is indicated by a program interrupt request at which time the digital value is available for input to the computer.

There are many advantages to using this type of signal detection. One of these is that during the entire aperture interval, the computer is free to be converting the data from the previous antenna into the rectangular coordinate vectors and summing them with the accumulated sum from all previous antennas in the cycle.

There is a finite lag time in the analog-to-digital conversion after signals have been received and recorded by the peak reading voltmeter. This time occurs during the time interval that lapses between two adjacent antenna switch positions. Two important features associated with this system are: 1) the computer is free to perform calculations, and 2) continuous surveillance of the switch aperture ensures the interception of all signals, within the 1μ sec response time of the peak reading voltmeter.

In choosing the A/D converter, one must take into account the probability of circulating currents. Circulating currents may provide either noise or erronous readings to the computer. A/D converters with transformer inputs convert the data without interference from such circulating currents. For this reason we have chosen a Texas Instrument A/D converter with transformer inputs. This is an economically priced A/D converter with provisions for eliminating circulating currents.

The main function to be performed by the computer is the calculation and display of the azimuth and elevation angular location of the received signal measured from the 17 readings produced by the peak detection system. Although it is expected that most trigonometric functions required can be stored in table form, there are sufficient other calculations to warrant the addition of the multiply and divide hardware to the basic configuration. The addition of the multiply and divide hardware, of course, will increase the speed of the multiplication and division, and also reduce the size of the memory required. Therefore, the 4096 words of memory on the small computer are felt to be adequate. The hardware multiply and divide has a cycle time of approximately 2 microseconds per operation as compared with up to 300 microseconds for complicated multiplication or division in the computer without this option.

Several computer types were considered, as to their ability to perform the task required for this project. Three computers felt to be competitive were the PDP-8, the Varian 620I, and the Hewlett-Packard 2115A. Of the three, the PDP-8 has been exposed to the most years of service in the field. However, it has several limitations which would make it less desirable for the DF problem. These limitations are: 1) its 12 bit word size limits single precision accuracy to 3 digits, 2) its weak command structure of the only 5 standard memory reference instructions, and 3) its absence of hardware index registers substantially increases the time and cost of programming as well as increasing the size of programs. If double precision calculations are required, to achieve the desired accuracy, the size of the program becomes even greater, possibly to the extent that the 4096 word memory would be inadequate for the DF problem.

The Hewlett-Packard 2115A is an integrated circuit computer having the following features: it has a greater variety of basic commands, up to 64I0 slots in addition to being an extremely versatile computer. The Varian 620I computer has comparable calculation capacity of the Hewlett-Packard 2115A at a substantial reduction in cost. The 16 bit word size of the 620I allows for single precision accuracy of 5 digits and contains 32 standard memory reference instructions, thereby reducing the size of the programs. It has hardware index registers that allow direct addressing of all memory as well as facilitating table look up operations. The cost of the Varian 620I ($18,000.00) with hardware multiply divide is comparable to the PDP-8 without this feature. Employing the peak reading voltmeter, which will permit adequate computation time, it is felt the calculations required for the direction finder will be no problem for the Varian 620I with the hardware multiply divide option.

At present the output display is envisioned to consist of an Industrial Electronic Engineers (IEE) unit which employs a 2 digit readout for elevation and a 3 digit readout for azimuth. The control for each unit is provided by two unique output transmission instructions. When executed, the instructions for the two digit display initiates the transfer of the low order 8 bit binary coded decimal (BCD) coded through software, of the accumulator or specified word of core memory to the IEE display interface. The instruction for the 3 digit display performs the same function when executed except that the low order 12 bits (BCD coded through software) of the accumulator or specified word of the core memory and transferred to the digital display. The advantage of this type readout over Nixie tubes is that for each number there is a single light source that illuminates the proper number behind the receiving window. The lens control of the light gives a single readout at the front face. If the light burns out it is replaced by simply inserting a new bulb rather than by replacing the entire assembly as the case of a Nixie tube. Cost of the two types of display, the IEE and the Nixie tube are comparable therefore, the IEE display was chosen for its dependability and ease of maintenance.

VI

GROUND REFLECTIONS

The most difficult problem associated with the azimuth-elevation direction finder is the ground reflections from the broad beamwidth antennas. Antenna patterns have been calculated assuming the DF antenna is in the presence of the earth. The earth is modeled as a uniform, infinitely thick, perfect conductor. Predicted power patterns for antennas above a perfectly conducting earth have been computed by Jones (1948). An antenna with horizontal polarization is oriented with the maximum toward the horizon. These patterns are felt to be a worst case in that they do not take into consideration non-uniformities such as hills, trees, etc. to disperse the reflections so that reflections add to cause deep nulls in the antenna patterns. When the height of the irregularities is greater than the space wavelength (Rayleigh Criterion), the scattered field may come from an extensive region of the surface of the earth, Beckmann (1963). This causes a random scattering with accompanying loss of phase coherence in the scattered wave. As the antenna is raised above the ground, the number of nulls increases in the pattern due to the separation of the antenna and its image. However, the Fresnel region of reflection is farther removed from the antenna, becomes larger, and is more influenced by variations in the terrain, further reducing the effect of the scattered wave upon the free space antenna pattern. There has been a considerable investigation of the affect of the terrain, but it does not appear practical to attempt to calculate the effect of terrain upon the antenna pattern. We have discussed the problem of scattering by a rough earth with several outstanding authorities in the field on this subject. All were in agreement that there would be considerable effect from the irregular terrain in the 600-3000 MHz region, but felt the extent of the effect could not be predicted with a high degree of accuracy employing theoretical methods now available. Based on the work of Beckmann and Spizzichino (1963), it becomes apparent that the patterns cannot be predicted, i.e., the nulls cannot be predicted within 10 db of the actual value.

It has been recommended that a practical solution is to build the direction finder and measure the results experimentally. This becomes a difficult task within the budget allowed for the DF contract. To really find the effect of ground reflections on the DF system, an rf source is required at a sufficient distance to establish ground reflections and this should have the ability to change its elevation angle. The only apparent solution would be to employ an aircraft with a signal source and have the azimuth-elevation direction finder plot the azimuth and elevation direction for this signal as a function of its known position. Its known position could be revealed by a radar system such as the NIKE site which is presently available for use at The University of Michigan.

VII

CONCLUSIONS

During this reporting period emphasis was directed toward developing the antenna feed network required for the quadrifilar log conical antenna. This time was spent gaining a better understanding of the theory and practical characteristics of stripline couplers and phase shifters. We have developed the correction techniques necessary to perform a design iteration to achieve the desired operating characteristics. At present we are waiting on the shipment of stripline material to be employed in the construction of the final model of the stripline feed network. A problem remaining to be solved is a technique to exert a uniform pressure between the ground plane surfaces of the strip transmission line to prevent transmission line discontinuities. Discontinuities affect the phase shifter to a greater extent than the electrical characteristics of the coupler.

An engineering version of the electromechanical switch has been electrically tested. These tests have shown that adjacent ports are adequately decoupled. However, additional effort is required to lower the VSWR in the 600 MHz region. As for the mechanical design the spindle and clutch for the switch are in the final design stage.

The total computer package: 1) the computer, 2) A/D converter, 3) peak reading voltmeter, 4) digital display, and 5) all necessary interfacing has been ordered and is to be delivered within 120 days. In the selection of the computer package, a particular effort was made to 1) obtain an optimum A/D converter, 2) to minimize noise due to circulating currents between the A/D converter and computer, and 3) minimize overall cost while maintaining adequate performance level.

An investigation has been initiated to determine what effect the natural earth will have on the operating characteristics of the DF antenna system. It is apparent the presence of the earth will affect the broad beam antenna patterns but the extent of the influence is not known. The greatest affect on the azimuth-elevation direction finder is reasoned to be on the accuracy of elevation predictions. It is to be noted that an exact solution to the reflection characteristics of the natural earth (including its natural roughness, trees, and variable dielectric constant) is impractical. Therefore, much of this study will require making intelligent assumptions as regards to analytical approximations of the true earth characteristics.

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13. ABSTRACT

This report discusses the construction and operation of the azimuth-elevation direction finder. It considers the proposed antennas and the antenna feed networks required. Emphasis is placed on the theoretical operation of the feed networks and the techniques being developed to achieve the design specifications. While the theory is well understood there are experimental design problems which are discussed along with their effect on the basic feed network. Several methods of signal detection are described with relative merit for each system presented. The report illustrates the function of the computer in the signal detection system and in the computation of azimuth and elevation angles of the monitored signal. Several small computers and analog-to-digital converters were considered during this period and the final choice is described. A brief survey of the effects of ground reflection on the operation of the system is given. However, due to the complexity of the problem no definite conclusions have been reached.

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