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A DIRECTIONAL TRANSPONDER ANTENNA SUBSYSTEM

G. L. Vogt
W. G. Jaeckle

THE BENDIX CORPORATION
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BALTIMORE, MARYLAND 21204

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FOREWORD

This report was prepared by The Bendix Corporation on Air Force Contract No. F33615-72-C-1264A under program 499L, for Air Force Avionics Laboratory, Air Force Systems Command. The Air Force Project Engineer in charge of the Directional Transponder Antenna Subsystem was Mr. Harold Weber (AFAL/TEM). He was assisted by Mr. Paul Springer.

The work on this program was performed at Bendix Communications Division and Bendix Research Laboratories. The contractor design team was directed by Mr. Guilfred Vogt and was performed by personnel of the Air Traffic Control Department, managed by Mr. Joseph L. McCormick. Dr. Wilfried Jaeckle headed the Research Division effort under the direction of Mr. William Harokopus, Manager of the Electromagnetic Technology Department. This report presents the results of the contract effort for the directional transponder antenna.

"Publication of this report does not constitute Air Force approval of the report's findings or conclusions. It is published only for the exchange and stimulation of ideas."

William J. Edwards
Chief, Radar & Microwave Technology Br.
Electronic Technology Division
Air Force Avionics Laboratory

ABSTRACT

The significance of this research and development to the Air Force is the development of an aircraft transponder antenna subsystem which will receive interrogations from a maximum volume of space surrounding each antenna aperture, determine the angle of arrival of each valid interrogation and respond appropriately with a directive beam. The goals include improved receiver coverage to assure response to both air and ground based interrogators and formation of directive transmit beams to limit the volume over which transponder replies are detected. Use of directive beams would significantly reduce the system interference (fruit) levels of interrogation terminals.

The first phase of effort was an antenna subsystem design trade-off study to select an approach to minimize system complexity and antenna installation problems and to maximize performance. A small ring array was selected which can determine angle of arrival and electronically steer a relatively narrow beam for the reply.

The second and third phase of effort constructed and tested the design which evolved from the trade-off study. The transmit array diameter is 1.2 feet, contains 12 elements and produces a 28.5 degree beam width at the 1090 MHz operating frequency. The transmit beam is steered by use of phase shifters at the operating frequency. The receive array is less than six inches in diameter, contains four elements located concentric to the transmit array and provides omnidirectional coverage. The direction finding receiver operates by comparing the relative phase of two receive array output signals.

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1.0 INTRODUCTION

This report discusses the activities and results of a program to develop a directional antenna for use with an aircraft transponder of the variety used to provide the beacon IFF replies for the air traffic control secondary radar system. This development effort was conducted under AIR Force Contract F33615-72-C-1264 for the Air Force Avionics Laboratory, Air Force Systems Command at Wright-Patterson AFB, Ohio by the Bendix Corporation at the Communications Division and Research Laboratories.

The Directional Transponder Antenna Subsystem development is a program which is aimed at improving the air traffic control system by reducing the quantity of "fruit" or unwanted replies received at interrogation terminals and by improving the operation of the aircraft transponder in an environment of heavy interference. It is meant to function in place of present day omni antennas consisting of quarter wave monopole antennas or flush mounted units producing patterns and polarization similar to a monopole. Historically, a single bottom mounted vertically polarized omnidirectional antenna first served the aircraft IFF function and is in widespread use today. To improve this coverage a similar top mounted antenna was added and diversity transponders have been developed to effectively function with the two antennas.

The Directional Transponder Antenna represents a further development of the IFF antenna with a number of objectives. The first involves forming and steering a directional transmit beam at only the interrogator requesting the reply in order that the ground portion of the IFF system may function with less interference or "fruit" than results from today's omnidirectional radiated replies. The second objective seeks to control and improve the aircraft's IFF antenna coverage in an environment of interference in order to raise the probability of obtaining the desired reply at the interrogator. Since the reaction time of the system is in the order of a microsecond, an electronically scanned phased array antenna system is dictated.

The acceptance of such a system will be based on its relative complexity, size, weight, cost, etc., versus its improvement in overall system performance as dictated by the needs of the environment in which it must operate. An additional objective must then be simplicity. There exists the possibility of more than one level of performance for a directional antenna based on the user aircraft requirements.

The installation and aerodynamic feasibility of the antenna design will also affect the acceptance of such a system and must be included in the objectives. A final objective must name compatibility with the existing AIMS system as one which cannot be compromised in any manner.

The program effort was divided into two principal phases. The first phase was a four month study effort to configure and design a system to meet these objectives. The second phase was an eight month hardware development and test effort to implement a laboratory model of the design.

2.0 TECHNICAL INFORMATION

2.1 System Description

At the outset of the program it was necessary to configure a system approach which would provide the required aircraft antenna coverage to meet the IFF system requirements. The selected antenna system must reflect the most simple, straightforward approach to reduce system complexity and provide a practical aircraft installation for ultimate use by a rather large fleet of aircraft. No initial restrictions were placed on the type of elements or antenna arrays to be used nor on the position of the antennas on the aircraft.

The Directional Transponder Antenna Subsystem is required to receive vertically polarized radiation from an interrogator in an omnidirectional mode and is required to transmit vertically polarized energy with a specified beam shape into the direction from which interrogation was received. This implies, ideally, beam steering capability into any given direction. Since "vertical" polarization is defined with respect to the earth's horizon (or the ground interrogator) independent from a coordinate system fixed with respect to the aircraft, the use of linearly polarized receive (or transmitting) antenna elements on the aircraft will necessarily result in a cross-polarized condition under certain conditions. Polarization diversity could, in principle, solve this problem at the expense of having a rather complex system. Another alternative would be the use of circularly (instead of linearly) polarized aircraft antennas.

The cross-polarized condition in conjunction with a linearly polarized interrogator would then be avoided at the expense of 3 dB gain loss. It should, however, be recognized that any energy received from (or transmitted into) any direction tangential to an appreciable portion of the airframe will necessarily be linearly polarized with the electric vector being perpendicular to the airframe. Accordingly, a flat spiral, flush (or semi-flush) with the aircraft skin, radiates with circular polarization into the broadside direction and with linear polarization, perpendicular to the skin, in the plane of the spiral. Thus, when positioning arrays consisting of spiral elements the polarization effects near the plane of the spiral, as well as the element pattern, must be remembered when the array locations are selected to provide coverage for the aircraft.

Approaches employing linear arrays, planar arrays, and circular ring arrays were considered with both linear and circular polarized elements.

The objective was to determine the quantity, position and type of array to maximize coverage in the direction of the horizon at all aspect angles of an aircraft in normal flying altitude. Coverage above and below the aircraft was also considered with the viewpoint that certain high performance aircraft might require a more elaborate system to provide additional coverage in this region.

The selected approach was configured as a ring array of monopole elements as shown in figure 1. Two such arrays would be required, one located on top of the fuselage and one on the bottom. The resulting system would be vertically polarized under normal flying conditions. It would be cross polarized for a horizontal flight position if the aircraft rolls by 90 degrees. Due to the element patterns there exists a lack of coverage for overhead targets or targets directly below the aircraft, just as exists with the IFF antennas in use today.

Each ring array functions in receive mode with an omnidirectional antenna pattern providing 360 degrees azimuthal coverage about the aircraft. The operating frequency is 1030 MHz. In transmit mode a fan shaped beam is formed with 28.5 degrees azimuth beamwidth at the operating frequency of 1090 MHz. This is steerable in 15 degree increments over 360 degrees of azimuth to provide the directional transponder reply. Vertical coverage of the array extends 45 degrees from the plane of each array.

As a system, two arrays would function with a diversity transponder to provide coverage to 45 degrees above and below the aircraft and at all azimuth angles. For high performance aircraft a growth potential exists in that a single element such as a spiral antenna could be added to each array to obtain improved coverage at the Zenith with good system polarization characteristics.

The outer ring of 12 elements shown in figure 1 form the transmit array. These elements are equispaced on a ring diameter of 1.3 λ. The inner ring of 4 elements form the receive array. These are located on a ring diameter of 0.5 λ. The overall antenna diameter is 17 inches with a thickness of 1.6 inches below the ground plane.

A block diagram of the antenna system is shown in Figure 2. Also included in the system are the necessary receiver circuits to determine the angle of arrival of the incoming signals and provide signals to the transponder, beam steering and control circuits to select a transmit beam and

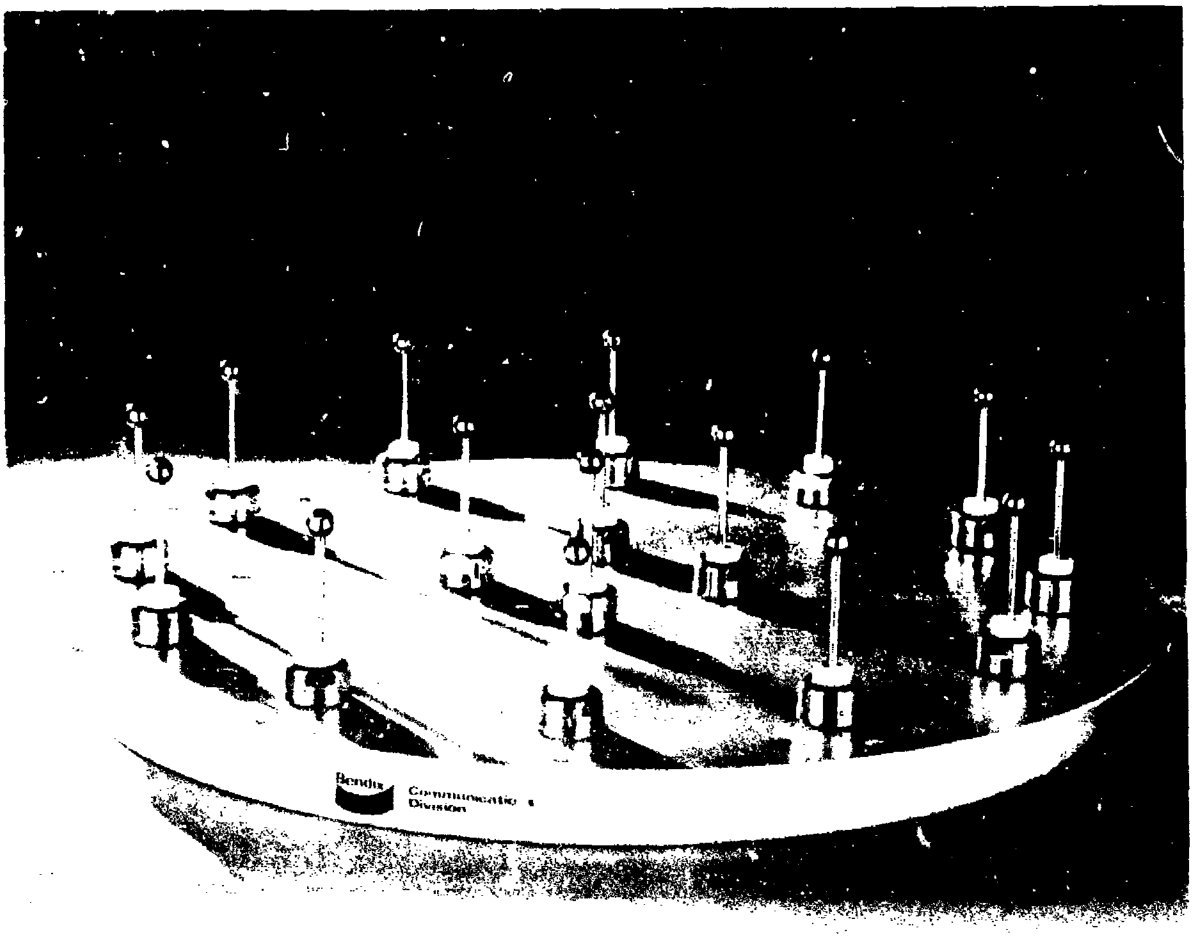


FIGURE 1. DIRECTIONAL TRANSPONDER ANTENNA

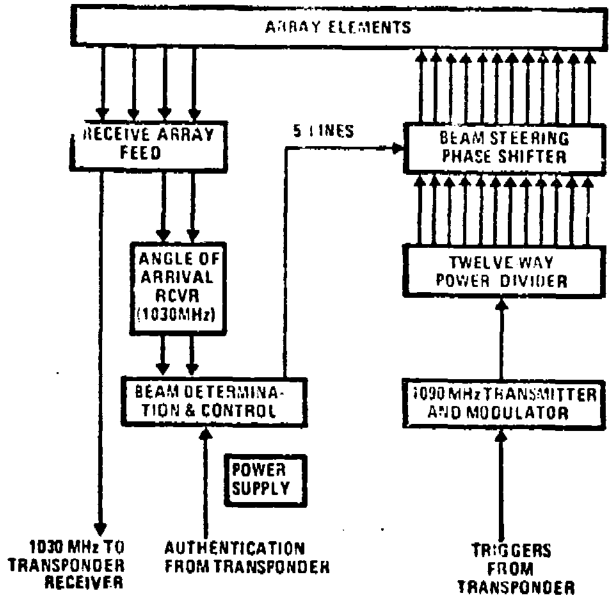


FIGURE 2. BLOCK DIAGRAM OF DIRECTIONAL TRANSPONDER ANTENNA

retain this information until the transponder generates a reply, a transmitter and beam forming system to provide the necessary reply and a power supply.

The receiver is a two channel phase comparing receiver which provides the arrival angle of the pulsed signals by determining the phase difference between two signals from the receiver array feed. It operates at 1030 MHz over a dynamic range from -77 to -27 dBm. Quadrature phase detectors provide two outputs to an A to D converter in the beam determination and control circuits. The phase of each 0.8 microsecond pulse entering the receiver is quantized to one of twenty-four angles corresponding to transmit beam locations. This information is stored for two microseconds to await an authentication from the transponder. In the event the pulse was the second one of a pulse pair generating a transponder decode, a signal from the transponder will result in additional storage of approximately 75 microseconds. Further inputs are inhibited during this time and the beam angle information is sent in digital form to the beam steering system to prepare the transmit array for the transponder replies. Following the authentication signal the reply pulse train, in the form of triggers from the transponder, enters the antenna system modulator and transmitter. The transmitter is a solid state device employing transistors to generate 50 to 100 watt peak pulsed output at 1090 MHz. With the transmit array gain the effective radiated power from the array is 500 watts peak pulse power.

With the use of two such antennas on the aircraft each antenna functions independently in the manner described above to provide azimuth control of the reply. The task of determining the top or bottom array usage is normally provided by a diversity transponder employing Hartlobe techniques.[(8)] The 1030 MHz signal from each array is compared in amplitude in the transponder receivers and a decision made on signal strength. Authentication signals and triggers are then sent only to the appropriate array. It is also possible to eliminate the transponder receiver if desired by adding a logarithmic video amplitude output to the array receivers.

2.1.1 Receive System

2.1.1.1 Approach

The direction sensing system of the directional transponder is based upon a dual mode receiving antenna system. This antenna system is best explained in terms of a transmitting antenna deducing its receiving properties by reciprocity.

One mode, symbolically shown in Figure 3 by a monopole above a ground plane of infinite extent generates a far-field which is omnidirectional in amplitude and has spherical phase fronts concentric with the antenna's phase center. The intersection of the phase fronts with the horizontal plane are concentric circles as shown in the figure.

The second mode of the receiving antenna system would, as a transmitter, also generate an azimuthally omnidirectional beam but with a phase front forming a spiral with a pitch of one wavelength in horizontal planes. The inserts of Figure 3 gives a 3-dimensional picture of the phase fronts of the two modes.

On reception from a common source, the phase of the signal received in the concentric phase mode will be a function of the distance from the source only, independent of the azimuth or elevation of the source of radiation. In contrast, the phase of the signal received in the spiral phase mode will depend upon the azimuth of the source as seen from the receiver, in addition to its dependence upon the distance to the source. The phase difference between the signals received by the two modes is therefore proportional to the azimuth bearing to the source. In fact, there is a one-to-one correspondence between this phase difference and the azimuth position of the source measured from some reference position: the phase difference changes by one (electrical) degree for a movement of the source by one (mechanical) degree. This

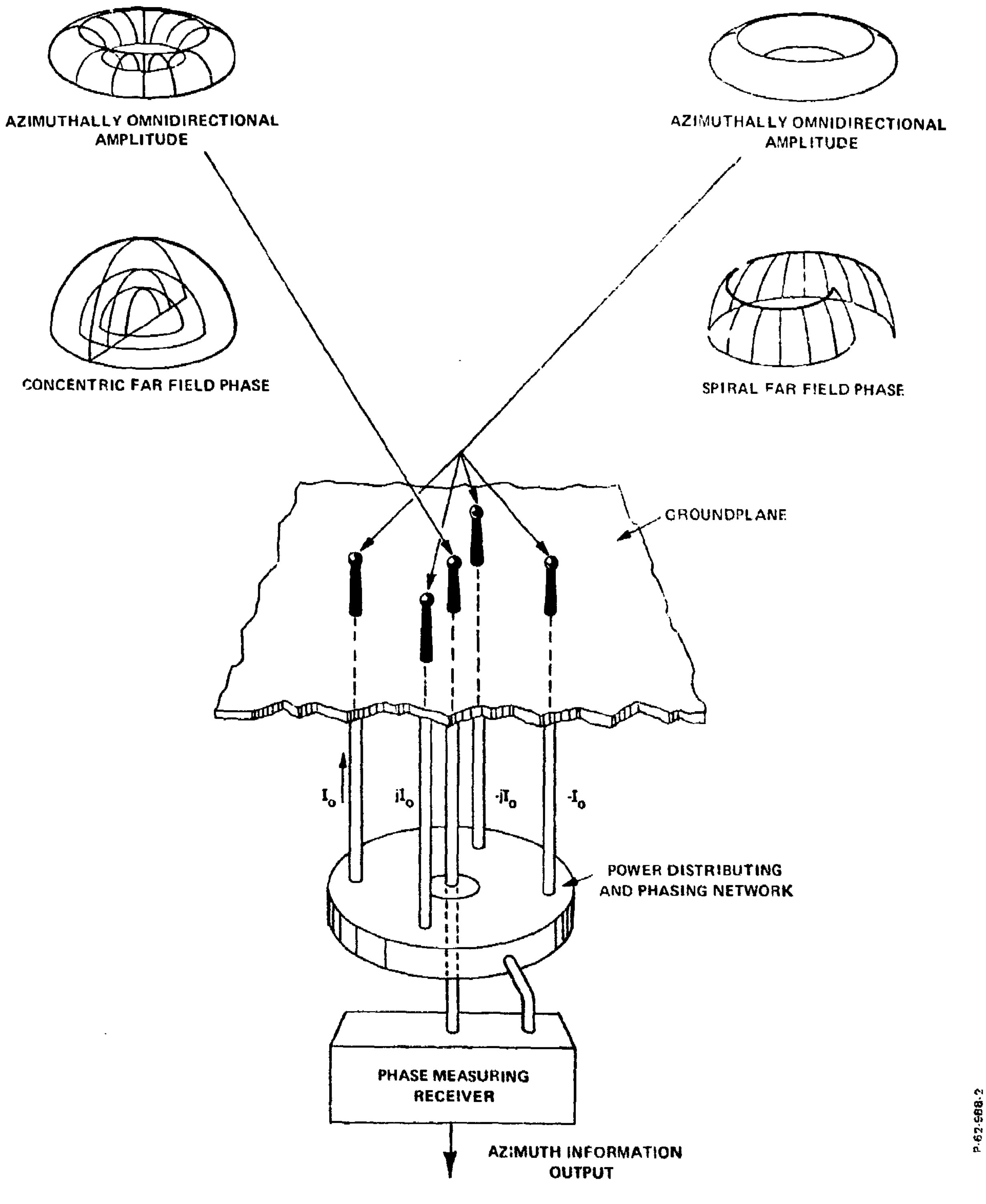


FIGURE 3. DUAL MODE RECEIVE ANTENNA SYSTEM FOR DIRECTIONAL TRANSPONDER

principle is perfectly suited for situations where azimuth direction finding over full 360 degree is desired. In comparison with other methods, it achieves omni-azimuthal coverage without, for instance, sector switching; it has no ambiguities and is, at least ideally, independent of target elevation. The absence of ambiguities is achieved by matching the basic ambiguity of phase measurements by 360 electrical degrees with an azimuth angle of 360 mechanical degrees—therefore, making this ambiguity inconsequential for azimuth determination.

The principle used is not entirely new. It was, for instance, discussed by Sandemann[1] in 1949. Basic information on the generation of spiral phase fields are contained in two papers by Knudsen.[2,3] A comprehensive monograph on circular arrays was written by Tillman.[4] Circular arrays provide a convenient way of generating a spiral phase field. The connection between spiral phase fields and the radiation from spiral antennas (eventually operated in multiple modes) should also be noted.

An ideal way of generating a spiral phase field would be to feed an infinite number of collinear dipoles (or monopoles above a groundplane) arranged in a circle of radius "a" with equal amplitude and progressive phase, such that the total phase progression around the circle is $2n\pi$ with n being an integer. In the usual spherical coordinate system where the circle is in the x-y plane and the radiators are parallel to the z-axis the following pattern results:[2]

$$P(\phi,\theta) = C\, e^{-jn\phi}\, J_n\,(ka \sin\theta)\, P_e(\theta) \tag{1}$$

where C is a normalizing constant independent of ϕ and θ, J_n is the n'th order Bessel function, $k = 2\pi/\lambda$ is the wave number and $P_e(\theta)$ is the elevation pattern of the radiating element used (e.g. $P_e(\theta) = \sin\theta$ for short dipoles). The field expressed by equation (1) is a pure spiral phase field with spiral pitch $n\lambda$.

A finite number m of radiators equally spaced on a circle of radius "a" fed with currents of equal amplitude and progressive phase such that the total phase progression around the circle is $2n\pi$ results in a pattern.[4]

$$\begin{aligned} P(\phi,\theta) = C \cdot P_e(\theta) \Big\{ & j^n e^{jn\phi} J_n(ka \sin\theta) \\ & + \sum_{q=1}^{\infty} \Big[j^{(mq-n)} e^{-j(mq-n)\phi} J_{mq-n}\,(ka \sin\theta) \\ & + j^{(mq+n)}\, e^{j(mq+n)\phi}\, J_{mq+n}\,(ka \sin\theta) \Big] \Big\} \end{aligned} \tag{2}$$

all terms express spiral phase fields with different pitch. The leading term $\exp(jn\phi)\ J_n(ka \sin\Theta)$ has the pitch corresponding to the phase progression of the feed current. The other terms are small under most circumstances. Consider, for instance, 4 radiators (m = 4) on a circle of one wavelength circumference (ka = 1) fed in the n = 1 mode. The largest argument of the Bessel functions is then ka sin $\pi/2 = 1$. The leading term consists of the Bessel function of order 1 and the residual terms contain the functions of orders 3, 5, 7, 9, 11, etc. The relative magnitudes of $J_1(1)$, $J_3(1)$, $J_5(1)$, $J_7(1)$ is 1: 0.044: 0.00057: 0.000003. The terms become rapidly smaller with increasing order of the function. In most cases, all summands can be neglected in comparison with the leading or principal term.

With 4 antennas as used in the directional transponder receive system, the pattern can also be conveniently calculated by direct summation of the field from the set of radiators at radius "a" and azimuth angles $\phi = 0$, $\phi = \pi/2$, $\phi = \pi$ and $\phi = 3\pi/2$ with currents I_o, $-jI_o$, $-I_o$ and jI_o resp. The resulting pattern is:

$$P(\phi,\Theta) = C \cdot P_e(\Theta) \left\{ (\sin(ka \sin\Theta \sin\phi) + j \sin(ka \sin\Theta \cos\phi) \right\} \quad (3)$$

All symbols have been explained previously. The degree of approximation to an azimuthally omnidirectional spiral phase field is not so obvious in this form. Figure 4 shows plots of the amplitude, $|P(\phi,\Theta)|$, of the pattern with respect to a mean value $|P(\phi_o,\Theta)|$

$$\Delta = \frac{|P(\phi,\Theta)|}{|P(\phi_o,\Theta)|} = \left\{ \sin^2(ka \sin\Theta \sin\phi) + \sin^2(ka \sin\Theta \cos\phi) \right\}^{1/2} \quad (4)$$

and the phase deviation, δ, from a true spiral phase field

$$\delta = \arctan \frac{\sin(ka \sin\Theta \cos\phi)}{\sin(ka \sin\Theta \sin\phi)} - \phi \quad (5)$$

for different ka sin $\Theta = 1$ and ka sin $\Theta = 1.57$. Equations (4) and (5) are plotted for $0 \leq \phi \leq 90^\circ$ only. The functions are periodically repeated over the remaining 270 degree azimuth region. It is seen that the theoretical accuracy of the antenna system is ± 2.5 degrees for ka = 1 and ± 7 degrees for ka = 1.57 (antennas on a circle of 0.5 wavelength diameter). The antenna elements of the directional transponder receive system are situated on a circle of 1/2 wavelength diameter at 1030 MHz.

2.1.1.2 Implementation

There are several design approaches for the signal combining and phasing network. The spiral phase mode can be generated by feeding four monopoles

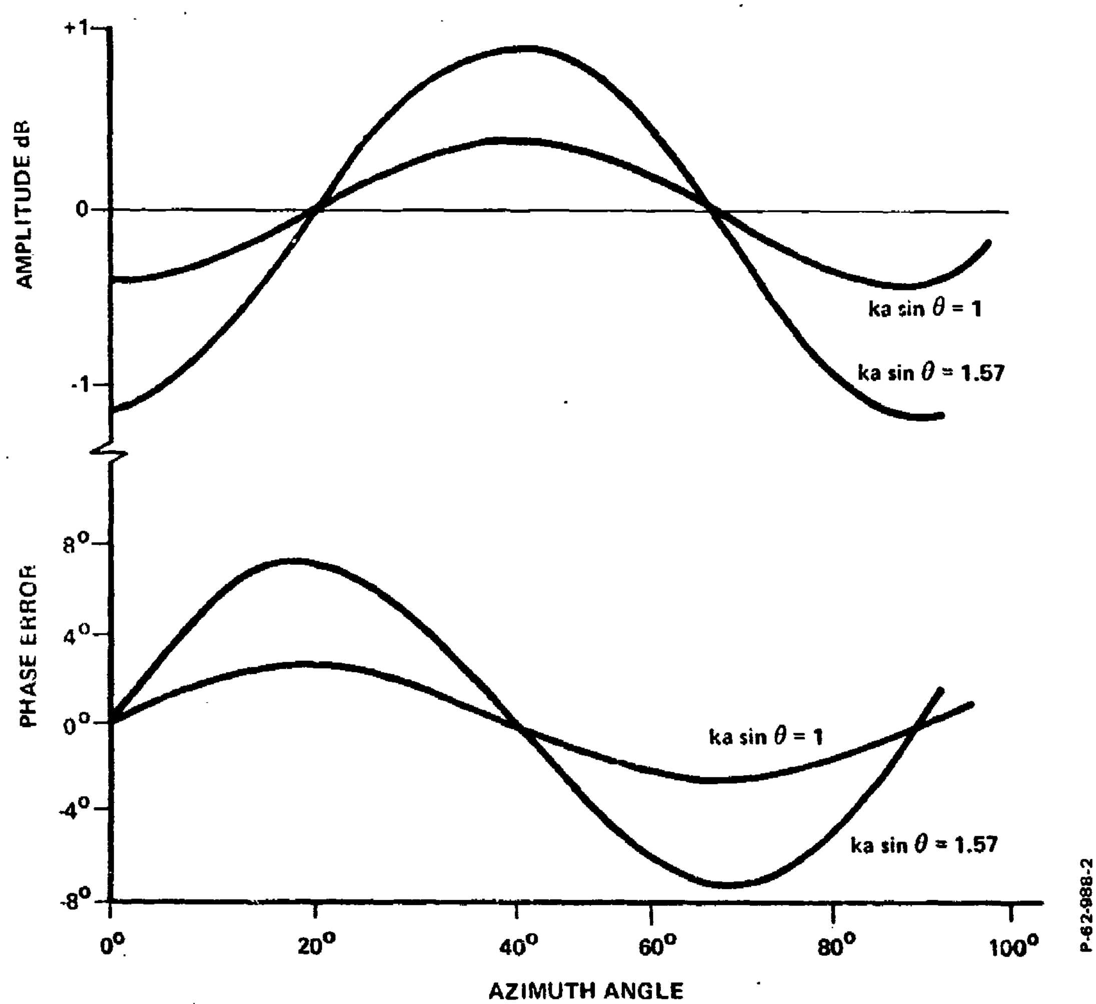


FIGURE 4. THEORETICAL AMPLITUDE VARIATION AND PHASE ERROR OF FOUR ELEMENT SPIRAL PHASE ANTENNA

from three matched Wilkinson power dividers with appropriate phase shifters in the output lines. For small bandwidth operation differential lengths of transmission lines can be used as phase shifters. The reference mode would, in this case be generated by a fifth monopole in the center of the array. This approach was successfully employed in a Bendix-developed proximity warning indicator for small aircraft operating on the same principle.

The directional transponder uses only four monopoles and generates several receive modes with a matrix network[5] of directional couplers and phase shifters similar to a Butler matrix. Figure 5 shows the network in stripline construction. Different lengths of cable between the network and the antenna elements serve as additional phase shifters. One port of the network provides equal amplitude and equal phase excitation of the four monopoles. It therefore provides the reference (concentric phase) mode. Another port delivers the spiral phase mode with 90 degree phase shift between adjacent antenna elements. A third port results in a spiral phase field with opposite sense of rotation to the first. This port with omnidirectional azimuthal pattern can be used advantageously connected to the transponder receiver for interrogation decoding and response initiation. A fourth port of the network results in equal amplitudes of the four monopoles with opposite phase of adjacent antenna elements. This port is not used in the directional transponder application and is therefore resistively terminated. The matrix network when fed from the concentric phase (Σ_o) port or any of the two spiral phase (Σ_{+1}; Σ_{-1}) ports provides equal power to the four antennas to within ± 0.15 dB at 1030 MHz. The average loss in the network is 0.25 dB. Isolation between ports exceeds 30 dB. The transmission phase from the concentric phase input port to the antenna terminals is equal to within ± 2 electrical degrees and the phase from the spiral phase, Σ_{+1}, port to the antenna elements has the desired 90 degree phase progression to within ± 1 electrical degree. If the antenna ports are terminated with matched loads, all remaining ports show a voltage standing wave ratio of less than 1.1:1 at 1030 MHz.

The receiver which is utilized with the array is a two channel phase comparing device fed from the Σ_o and Σ_{+1} ports of the array feed. A superheterodyne, double conversion system was utilized with the phase detectors operating at 10 MHz. The predetection band width is 8 MHz with all out of band signals reduced by -60 dB. Limiting circuits are used prior to the phase detector such that all input signals over the dynamic range of -77 dB to -27dBm

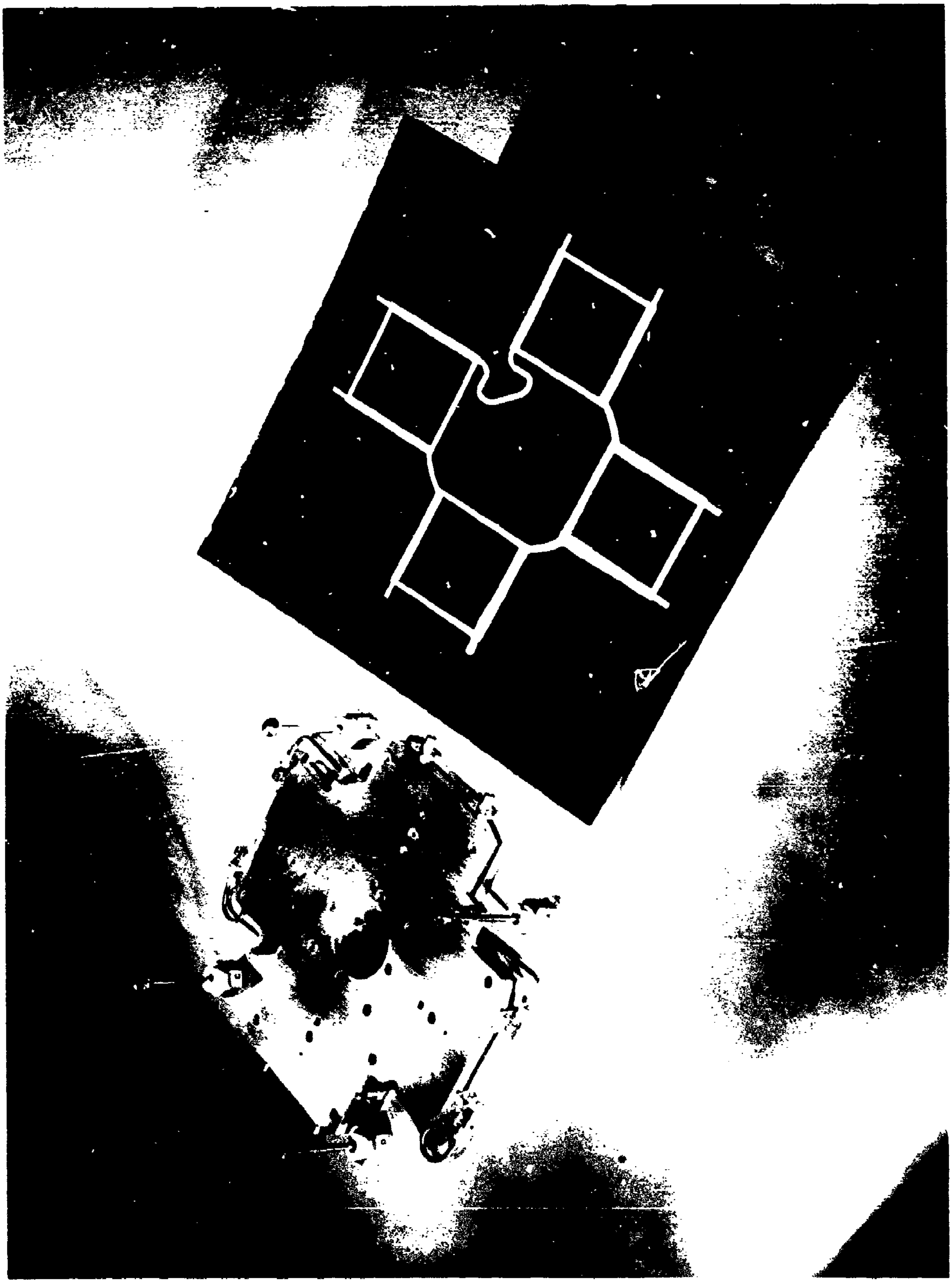


FIGURE 5. RECEIVER ARRAY

are limited. Phase tracking errors between channels are approximately equivalent to the accuracy of the receiver array. One phase detector output provides 0.5 and 0.8 microsecond pulsed signals to an A to D converter which is sampled within the signal pulse width. A second phase detector operating in quadrature to the first provides a similar output to a comparator to remove the 180 degree phase ambiguity of a single phase detector.

2.1.1.3 Performance

The theoretical performance discussed previously was based on certain idealized assumptions as, for instance, using an infinitely large ground plane to support the receive antenna elements and also discounting the influence of the transmit array which is concentric with the receive array. The theory presented does also neglect the mutual effects between the receive elements.

In practice, the performance of the receiving array was evaluated by mounting the array in the center of a finite circular groundplanes of 36 inches and of 48 inches diameter. In most cases the edge of the groundplane was slightly masked with some microwave absorber material. This procedure is motivated by the fact that in an actual installation on an aircraft the groundplane edge would not be present.

Because of the symmetry involved, any circularly symmetric obstacle, concentric with the receive array will not influence the direction finding accuracy at any fixed elevation angle. This means, the field stays spiral-phase or concentric-phase respectively in the presence of the obstacle depending upon which mode is excited (again, talking in terms of a transmitting antenna). The diffracting edge of the experimental groundplane is an obstacle of the required circular symmetry. The circular transmit array with its 12 fold rotational symmetry approximates the circular symmetry fairly well. Neither the groundplane edge nor the transmit array will, therefore, have a major deteriorating effect on the direction finding accuracy in any constant elevation direction. This is not so if one considers the phase at constant azimuth, but variable elevation. Figure 6 presents a heuristic argument why this is so. It is assumed, for this argument, that only the two diametral opposite antenna elements of the transmitting array in the elevation plane considered, contribute significantly to the scattered energy in the respective plane. This is a reasonably realistic assumption. Let the incident field generated by the spiral phase mode of the receive array (if used to transmit)

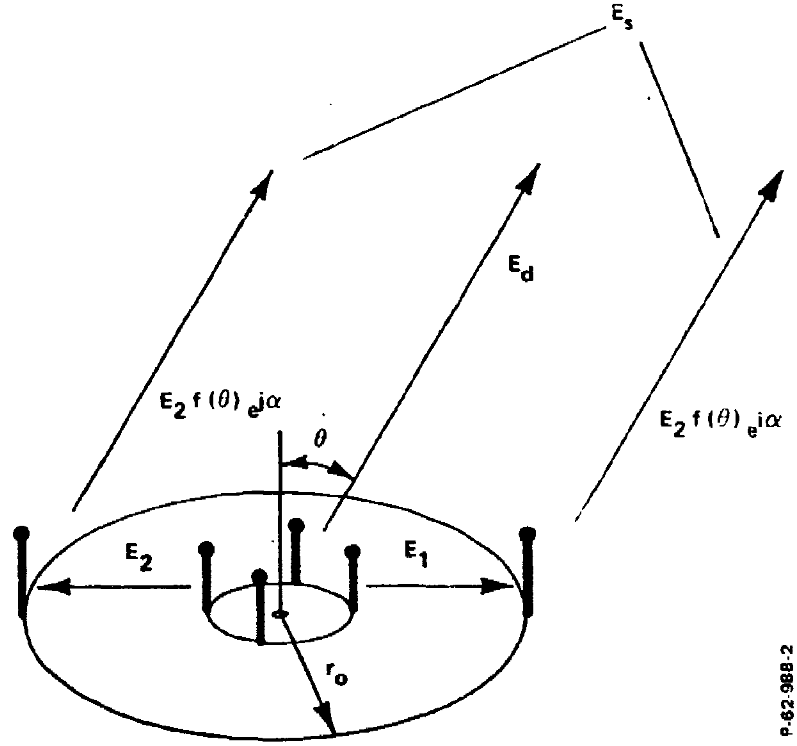


FIGURE 6. SCATTERING OF SPIRAL PHASE MODE BY TRANSMIT ARRAY

be $E_1 = \frac{E_o}{r_o} e^{j(wt + \phi - \beta r_o)}$ and $E_2 = \frac{-E_o}{r_o} e^{j(wt + \phi - \beta r_o)}$ respectively at the location of the two scatterers. Let a fraction $f(\theta)$ of this incident field be scattered into direction θ, undergoing a phase shift α upon being scattered. The scattered field, E_S, can then be represented by

$$E_S = -2j \sin (\beta r_o \sin \theta) \; E_o \frac{f(\theta)}{r} \; e^{j(wt + \phi - \beta r + \alpha)}$$

This scattered field adds to the directly received field, E_d,

$$E_d = \frac{E_o}{r} e^{j(wt + \phi - \beta r)}$$

where r, ϕ, θ are the coordinates of the observation point in a spherical system. It can be observed that the resulting field $E = E_d + E_S$ is still spiral phase for variations in ϕ at any fixed elevation angle but the phase of the resulting field will, in general, change with elevation angel θ. No rigorous analysis of this effect has been attempted. Instead the diameter of the transmit array was experimentally optimized to result in minimum phase perturbation vs. elevation angle in the region from 0 to 45 degrees above the horizon. The optimum diameter turned out to be approximately 14.5 inches diameter. The diameter of the transmit array finally built was chosen to be 14.15 inches which is somewhat smaller than optimum. It was also found that the effect of the transmit array on the receive array depends upon the termination of the elements of the former. In all measurements on the receive array, the transmit elements were therefore resistively terminated and provision was made to establish this same condition in actual operation, as explained in the section on the transmit array. Figure 7 through 12 shows measured performance data.

Figure 7 and Figure 8 are azimuthal amplitude patterns of the spiral phase and reference mode respectively of the receiver antenna system in the plane of the groundplane and at an elevation angle of 20 degrees. The amplitude patterns shown are constant to within approximately ± 0.5 dB. This result is a composite of the theoretically expected pattern variation of ± 1.0 dB under ideal conditions according to Figure 4, plus the previously quoted ± 0.15 dB unbalance in amplitude of the feed network and the influence of the transmit antennas.

Figures 9 and 10 are elevation cuts of amplitude patterns of the spiral phase mode and reference mode. Both patterns show a tilt of the pattern

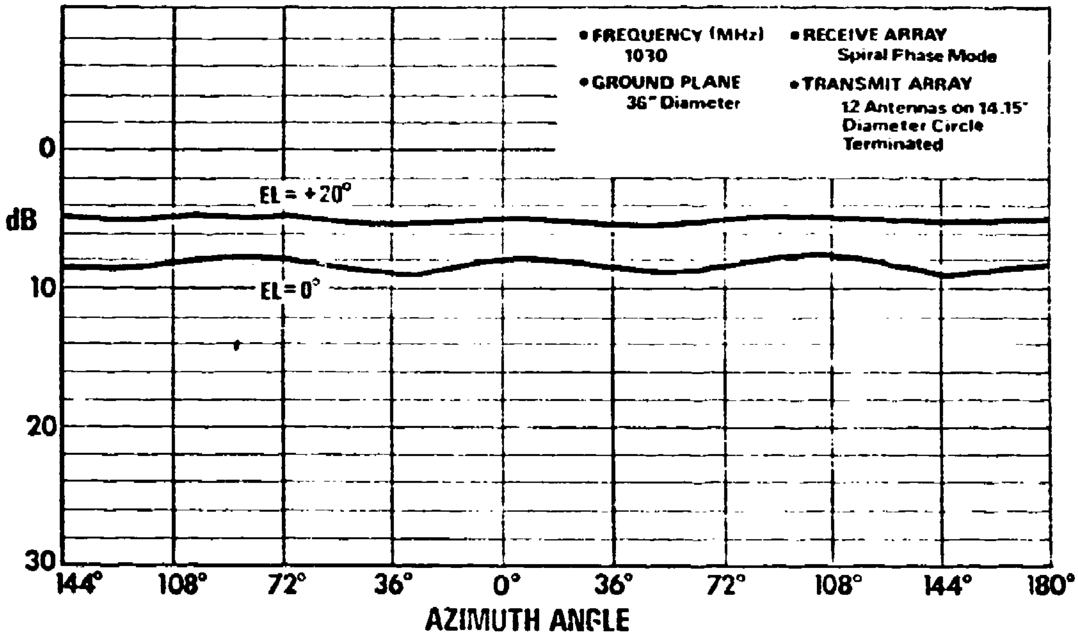


FIGURE 7. AZIMUTH AMPLITUDE PATTERN OF SPIRAL PHASE MODE

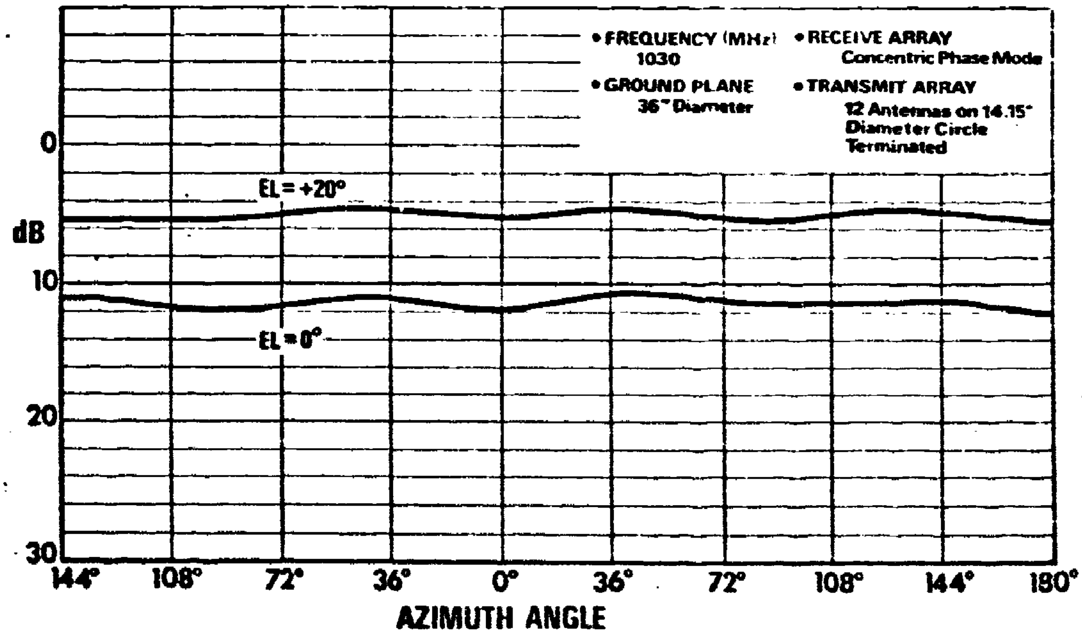


FIGURE 8. AZIMUTH AMPLITUDE PATTERN OF CONCENTRIC PHASE MODE

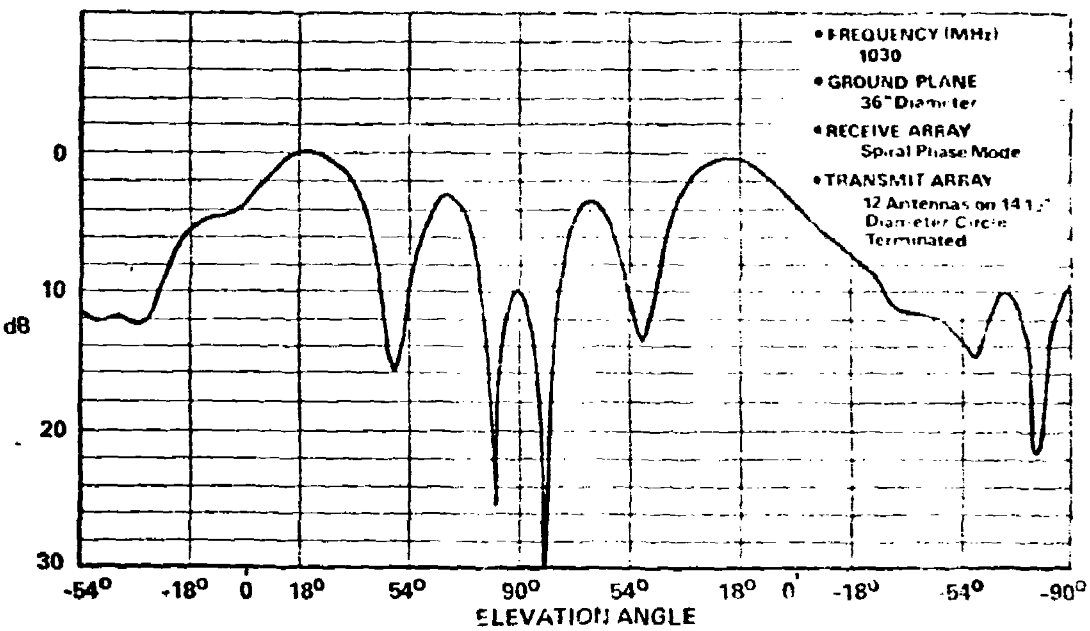


FIGURE 9. ELEVATION AMPLITUDE PATTERN OF SPIRAL PHASE MODE

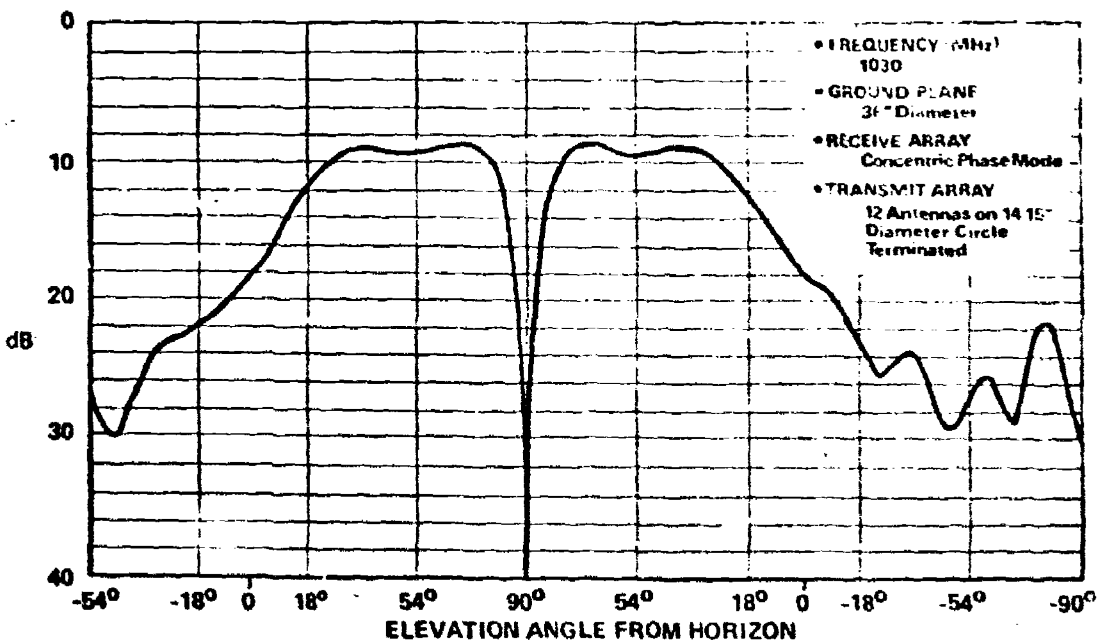


FIGURE 10. ELEVATION AMPLITUDE PATTERN OF CONCENTRIC PHASE MODE

maximum upwards from the horizon where the pattern maximum would be for an infinitely large groundplane. This is characteristic for the pattern of monopole antennas on circular groundplanes of finite size.[7] It is significant that the spiral phase mode pattern has a (relative) maximum at the zenith (center of the pattern recording). Since the element pattern of the monopoles has a null in this direction one can conclude that all energy in the zenith region is scattered energy. Measurements also show that this zenith lobe is circularly polarized.

Azimuth cut phase data are shown in Figure 11 for two different elevation angles. These data are measured with a Hewlett Packard Model 8405A vector voltmeter since the directional transponder receiver was not available. Ideally, the plotted phase difference between the sprial phase mode and concentric phase mode should be a single straight line with a slope of one electrical degree per mechanical azimuth rotation of one degree, independent of elevation. The measured phase response approximates this ideal to within ± 8 degrees. This is close to the theoretical accuracy of ± 7 degrees according to figure 4 and is sufficiently accurate for the directional transponder.

Phase difference between the sprial phase and concentric phase mode versus elevation for different azimuth directions is shown in Figure 12. It also shows the direction finding error within the region from the horizon to 45 degrees above the horizon. For higher elevation angles, large errors result. Targets within this region have to be excluded from the direction finding system by some independent means. A flat spiral antenna located within the receiver antenna circle could serve to detect these targets and identify them by amplitude comparison of the signal received by the spiral antenna with the signal received in, for instance, the spiral phase mode.

2.1.2 Transmit System

2.1.2.1 Approach

At the outset of the program effort the receiver direction finding technique was known. It was necessary to configure a transmit array which also met system requirements for simplicity and good aerodynamic and installation characteristics. The required beamwidth was 30 to 40 degrees with peak side lobe levels of -10 dB or better. The work by Tillman[4] and his associates was considered and the use of a ring array was investigated and selected.

A ring array of vertical radiating elements is shown in figure 13. These elements are located over a flat groundplane extending some distance past the diameter of the ring. The elements are quarter-wave monopoles. When

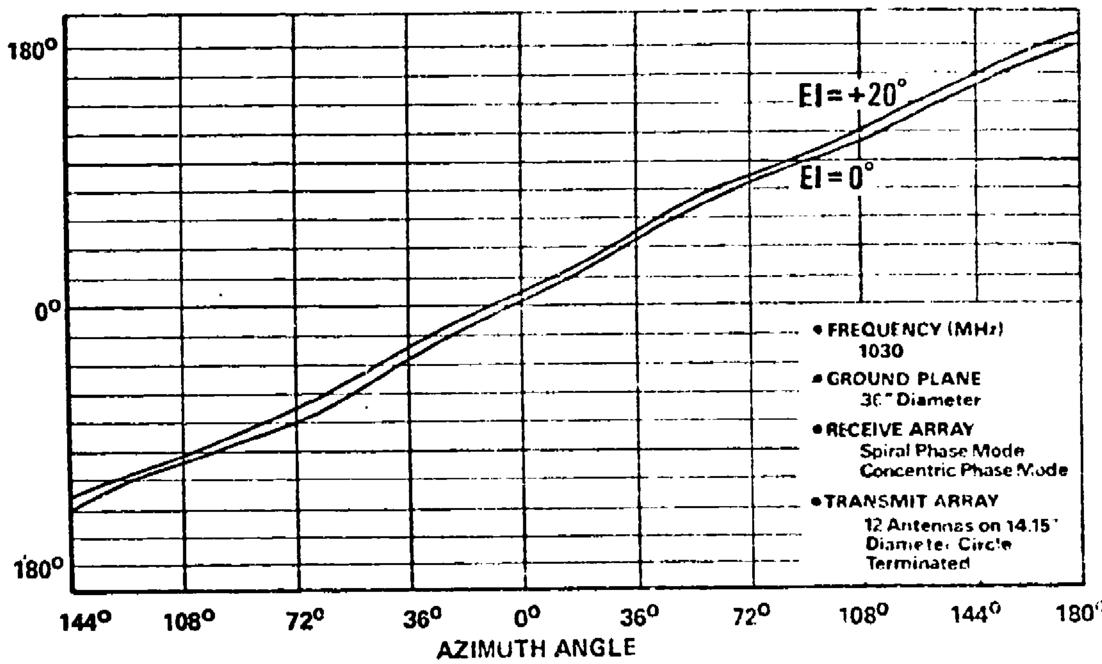


FIGURE 11. PHASE DIFFERENCE BETWEEN CONCENTRIC PHASE AND SPIRAL PHASE MODES VS. AZIMUTH ANGLE

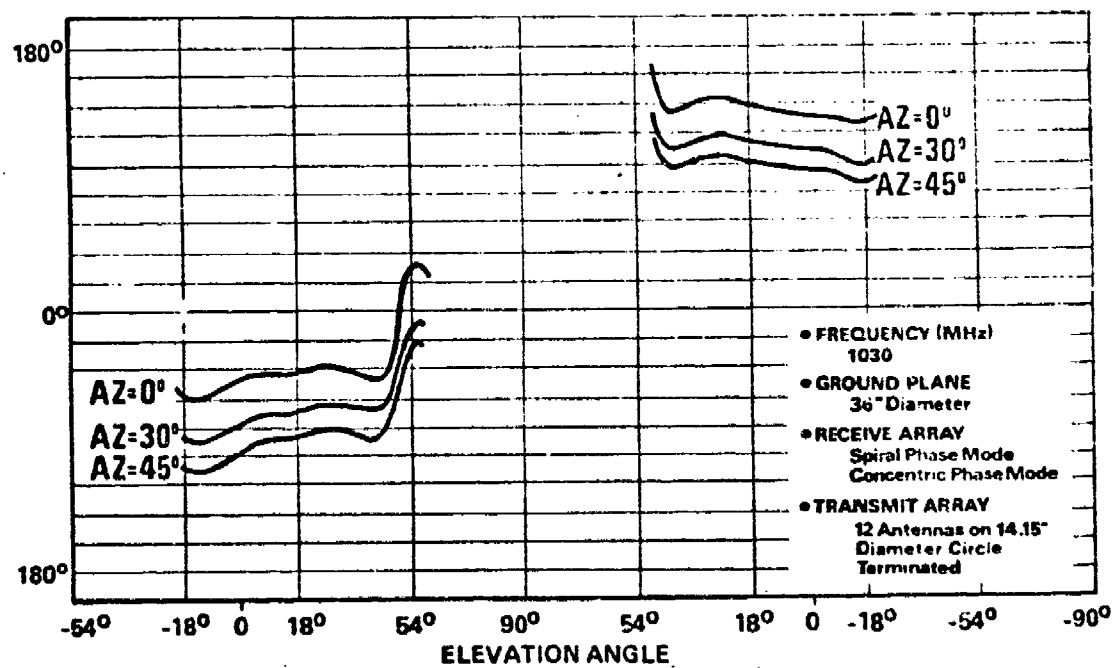


FIGURE 12. PHASE DIFFERENCE BETWEEN CONCENTRIC PHASE AND SPIRAL PHASE MODES VS. ELEVATION ANGLE

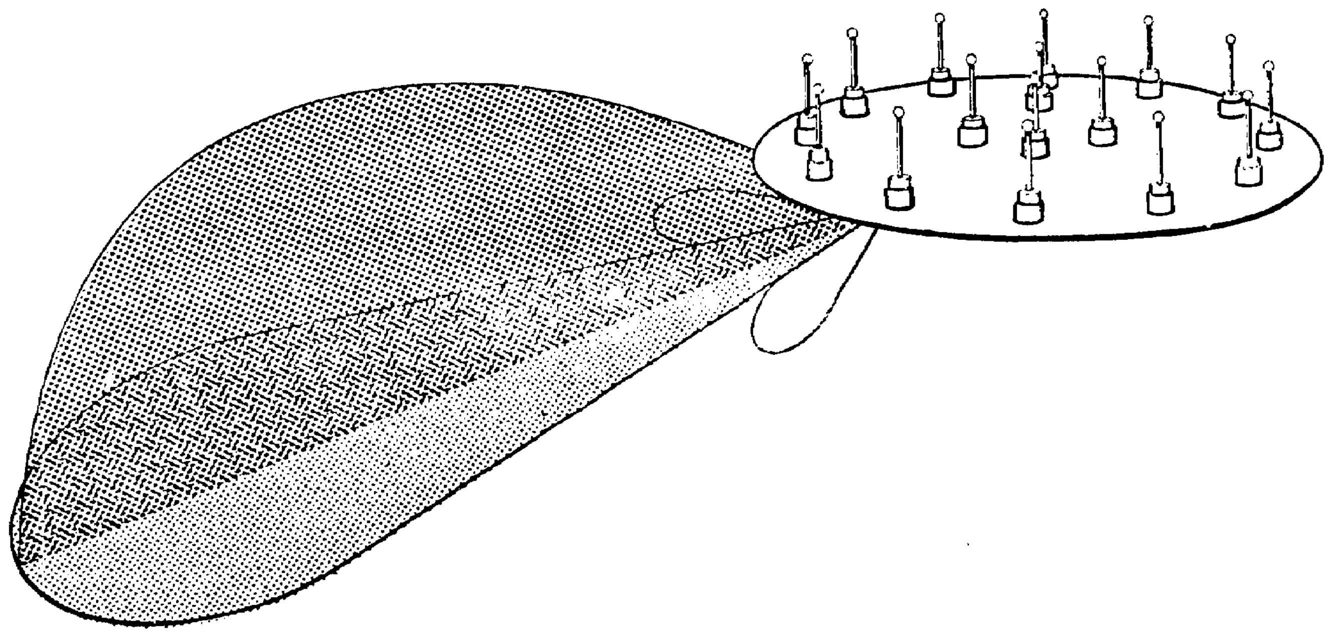


FIGURE 13. MONOPOLE RING ARRAY 12 ELEMENTS

located in a ring array, these elements combine to produce a fan beam with azimuth beam width being a function of the ring diameter and with an elevation 3dB beam width nearly equal to the element pattern. Side lobe levels in both the azimuth and elevation patterns are controlled to a large extent by the ring diameter and element density as well as the element amplitudes and relative phases.

For continuous scanning of the azimuth beam, the type of feed network often considered is the hybrid matrix. When discrete steering angles rather than continuous scanning is a goal, and where the number of beam positions is equal or related to the number of array elements, a considerable reduction in the complexity of the array feed network and steering system can be realized, under certain illumination conditions, by elimination of the hybrid matrix. The resulting feed network can reduce to an N:1 power divider and N phase shifters (where N is the number of elements).

The use of this approach can provide a thin package which can be attached externally to the skin of the aircraft if so desired. The protrusion would be approximately four inches. The diameter of the ring array is set by the desired azimuth beam width and has a dimension of 1.0 to 1.5 wavelengths for the beam widths being considered.

A general formula which gives the ring array azimuth beam width as a function of array diameter is given in equation

$$Bw = \frac{40^{\circ} \times \lambda}{d}$$

where d = diameter of the array expressed in wavelengths

Bw = one way 3 dB beam width

For the selected 1.3λ diameter transmit array the ring diameter is 14.15 inches and the predicted azimuth beam width is 33 degrees at the 1090 MHz operating frequency. This diameter permitted locating the receive and transmit arrays on concentric rings.

2.1.2.1.1 <u>Array Design</u>

The transmit array was designed with the aid of a computer simulated array and verified by experimental measurements on an array built to the dimensions and with the illumination functions predicted by the computer. Through the use of the computer, arrays of various sizes and with different numbers of elements were investigated for the beam patterns and side lobe levels they produced. The influence on the complexity of the beam steering was also investigated to simplify this circuitry.

From the work of Tillman the approximate number of elements to achieve a theoretical side lobe level of 15 dB was determined to be 10. For reasons of symmetry and to aid in reducing the side lobe level, a 12-element array was considered more desirable. Twelve elements can be symmetrically located with respect to the receiver array to minimize the number of different mutual coupling geometries. Also, the work by Tillman to relate element density to side lobe levels did not account for any detrimental effects on side lobe level such as may occur due to the presence of the receiver array. For this reason the minimum acceptable number of elements for the 1.3 λ diameter array is considered to be 12.

The array design approach taken was to first consider only phase optimization of the drive signals to the elements to obtain the desired side lobe levels. It is desirable to feed all elements of the array with equal amplitude signals to simplify the beam steering and feed system. The element positions and uniform amplitude information was entered into a computer which would first calculate the required element drive signal phases to form a beam in the desired direction. The resulting phases are the cophasal set and are related only by the element locations. For ring arrays the cophasal set of drive signals produces side lobe levels approximately 8 dB below the main beam. The desired lower side lobe levels were then entered into the computer which then recomputed drive signal phases in an attempt to reach the inserted side lobe level. To the degree this is theoretically possible for a given ring diameter and element density the side lobe levels are reduced. Peak side lobe levels of 10 to 15 dB were predicted for various arrays. For this initial effort the theoretical pattern of a monopole element on an infinite groundplane was used in the computer to generate the array azimuth and elevation patterns. No mutual coupling effects are included in these calculations.

The computer results were verified on an experimental array with a 1.5λ diameter with good results. The effort with a 1.5λ diameter array produced side lobe levels of 10 to 12 dB as shown in figure 14. Both measured and predicted azimuth beam width is 28 degrees. The illumination function was equal amplitude, only element phasings were controlled to reduce side lobe levels. Figure 15 shows the measured elevation pattern for this array.

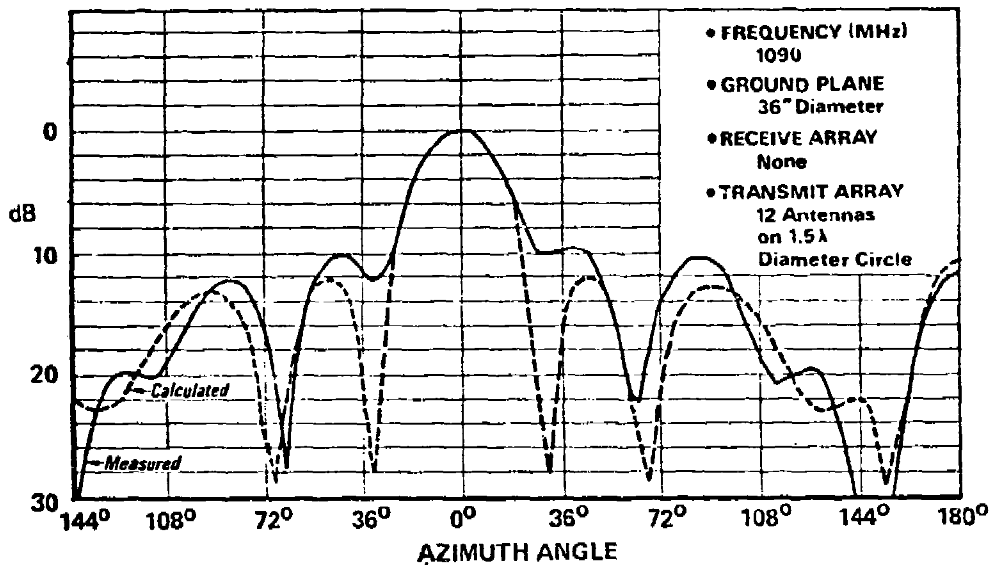


FIGURE 14. AZIMUTH PATTERN 1.5 TRANSMIT ARRAY MEASURED VS. CALCULATED

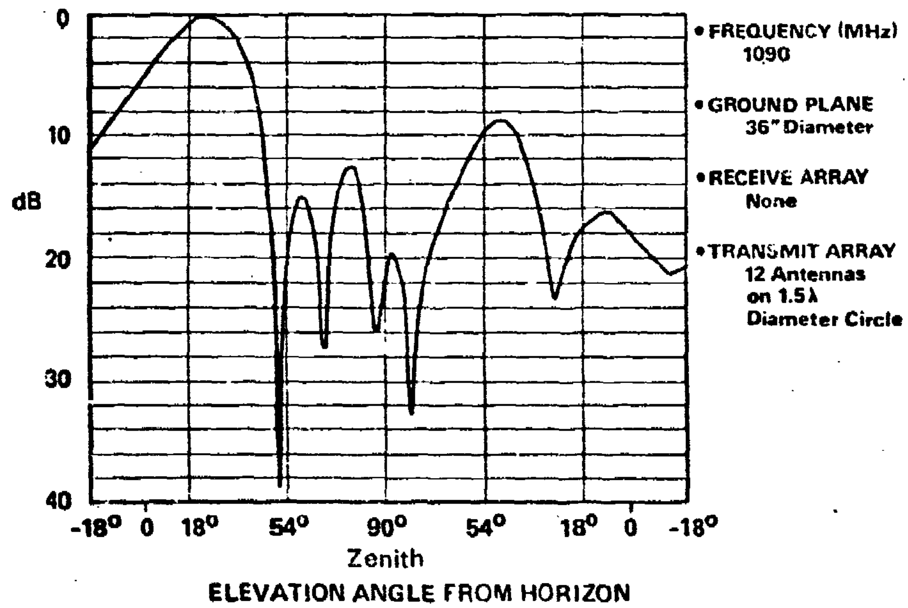


FIGURE 15. ELEVATION PATTERN 1.5 TRANSMIT ARRAY WITHOUT RECEIVE ARRAY

It was found that when the receiver array was added, the side lobe levels degraded due to scattering and blockage effects. To improve the side lobe levels with the addition of the receiver array the element patterns were measured with all transmitter and receiver array elements in position to obtain a first order indication of the mutual effects. This element pattern was then entered into the computer program and the phase optimization repeated. The process was repeated at two steering angles of interest as discussed later. At the beginning of the program the effort was concentrated on arrays of 1.5λ and 1.2λ diameter to determine the feasibility of using constant amplitude illumination. The conclusion was that the specified 10 dB side lobe levels could be met with the simplified feed system.

In the laboratory model it was possible to provide a steerable 3 dB variation in element drive level within the phase shifter circuit to gain some improvement in side lobe level. Very little added complexity was involved and the addition was considered worthwhile.

The final selection of the transmit array diameter of 1.3λ in the model was dictated by the influence of the transmit array on the direction finding accuracy of the receiver array as was discussed. The gain of the array is approximately 11 dB above isotropic.

2.1.2.1.2 Beam Steering Positions

The required beam steering positions for the array was determined as follows. It is apparent that for a noncontinuous or stepped scanned system that the minimum number of steps is equal to the total scanning angle (360 degrees) divided by the beam width (33 degrees) which is 11 positions. With 12 elements it is logical to first think of steering the beam to each of the 12 angular locations of the elements. With this approach the same phase front (or relative phase shifter values) need only be rotated around the 12 element positions. This gives 30 degree steering increments and simplified the steering system.

Such a system would function very well if there were no errors involved in determining the arrival angles of incoming signals or in steering the transmitted beam. Actually this approach can tolerate errors equal (in degrees) to the amount that the beam width exceeds the steering increment.

For a practical system then, the size of the steering increment must be reduced from the 30 degrees discussed above to allow for receiver array phase errors, phase errors in the receivers themselves and quantization errors in the signal arrival angle digital storage circuits.

For the directional antenna system the steering increment has been selected to be 15 degrees which provides 24 beam positions. The 12 positions corresponding to the element locations are used as well as 12 additional positions located midway (at 15 degrees) between these positions. This is illustrated in Figure 16. Such an approach can then tolerate system errors totaling 18 degrees. This is compatible with the measured receiver array errors of 6 degrees and permits receiver phase tracking error and quantization errors in excess of 12 degrees. (Since the phase errors are generally not all in the same direction larger errors can be tolerated, it is the RMS summation of all errors which cannot exceed 18 degrees.)

The total of 24 beam positions can be generated with only 2 distinct sets of phase shifter values, 1 corresponding to steering the beam to an element location and one corresponding to steering the beam to 15 degrees or midway between 2 elements. Each of these two sets of phase shifter values can then be rotated or stepped around the array to 12 different positions to provide the 24 beam locations.

2.1.2.1.3 Beam Steering System

The previous section discussed the relationship between the beam patterns and the number of array elements and also developed the required number of beam positions. This was seen to be two basic beam formations (phase fronts) which are each steered to 12 different locations. For a beam steered in the direction of element number 1 (0 degree) a typical set of element signals phases is shown in Table 1 for a 1.5 diameter array. The values of phase are also shown for a beam steered midway between element 1 and element 12 (15 degrees).

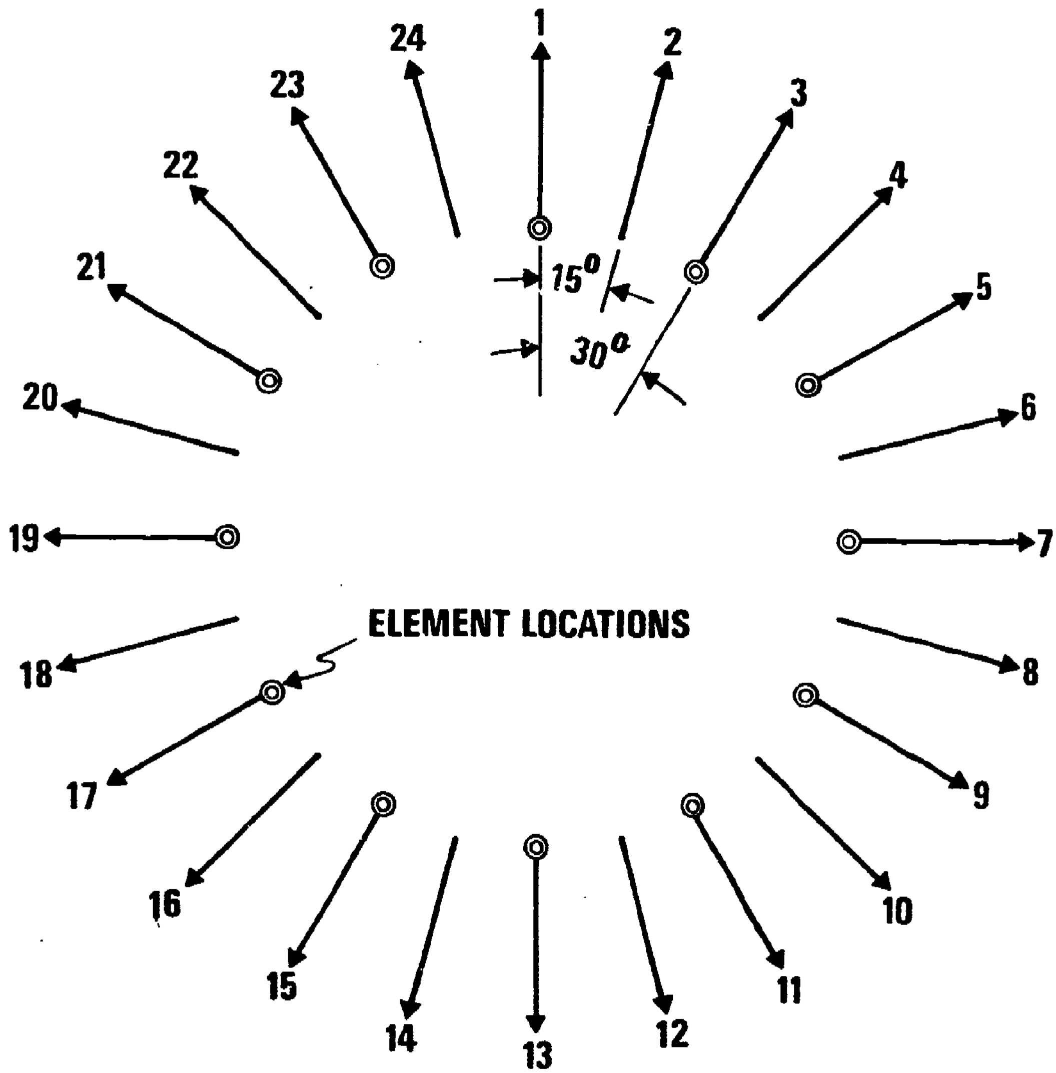


FIGURE 16. BEAM STEERING LOCATIONS

Table 1. Element Phasings

Element Number	0° Bearing Relative Phase	15° Beam Relative Phase	Element Number with Beam Rotated 1 Position
1	$+120^\circ$	$+120^\circ$	2
2	$+108^\circ$	$+155^\circ$	3
3	-149°	-77°	4
4	0	$+77^\circ$	5
5	$+149^\circ$	-155°	6
6	-108°	-120°	7
7	-120°	-120°	8
8	-108°	-155°	9
9	$+149^\circ$	$+77^\circ$	10
10	0°	-77°	11
11	-149°	$+155^\circ$	12
12	$+108^\circ$	$+120^\circ$	1

The phase values shown in the table are the optimized values computed to reduce side lobe levels. To steer the beam to a total of 24 positions the 0 degree and 15 degree set of values are each rotated around the 12-element positions resulting in (for 0 degree set) a beam steered in the direction of element 2, 3, 4, etc., and (for 15 degree set) a beam steered midway between element 1 and 2, 2 and 3, etc. The phase values for the beams rotated one position are also shown in the table.

It is thus necessary for the beam steering phase shifter controlling the phase to each element to be capable of approximating each value of phase shown in Table 1 in order to reach all beam positions. One logical approach is to use a digital phase shifter with discrete bit values of 180 degrees, 90 degrees, 45 degrees, 22.5 degrees, 11.5 degrees, etc. A four-bit phase shifter can approach any value to within $\frac{22.5^\circ}{2} =$ 11.25 degrees and a five-bit shifter can be within 5.625 degrees.

An investigation was performed with the computer to determine the effect on side lobe levels of such a phase quantization by computing patterns with element phases rounded off to the nearest achievable value when using 3-bit, 4-bit, and 5-bit shifters. The results of this effort indicated that for the array model a four bit phase shifter would be required to approximate the desired phase values sufficiently close to produce minimal side lobe level degradation.

The beam steering system has thus been "sized". It is required to quantize the incoming signal angle of arrival to one of 24 beam positions on command of a signal from the transponder indicating a decode has occurred. In turn, it must control 12 phase shifters, each containing 4 discrete values of phase for a total of 48 output signals.

2.1.2.2 Implementation

A photograph of the components comprising the transmit array feed system, before final assembly, is shown in Figure 17. The construction is microstrip. The twelve way power divider is formed with a reactive three way divider and three constant impedance four way dividers. The VSWR of these devices is under 1.2 to 1, the insertion loss is less than .25 dB, and the output phase tolerance is within 1.5 degrees. The isolation of the four way divider exceeds 30 dB. The four bit phase shifter is a series configuration switched line length device employing pin diode switching. A total of 17 diodes are used. Its VSWR is typically 1.5 to 1 and its phase accuracy is ± 3 degrees. The phase shifter includes a resistive 50 ohm termination for the transmitter element which is switched on during receive array operation. Due to transponder characteristics the transmitter array is used a maximum of 3% of the time. The use of a switched resistive termination permits a large reduction in phase shifter current drain since the phase shift control diodes can be unenergized during receiver operation. The resistive loading can also be selectivity used in transmit mode to provide a 3 dB amplitude weighting for a given element.

A phase shifter driver is also shown. It contains a 120 bit read only memory which converts the selected transmit beam number in the form of a 5 bit parallel word (indicating one thru 24) to the desired phase shift value and energizes the appropriate diode driver circuits.

The beam steering control logic which converts the receiver phase detector outputs to the five bit digital word supplied to each phase shifter is contained on a separate PC board (not shown) in the array. This circuitry contains approximately 14 dual in line packaged TTL logic modules.

A photograph of the assembled laboratory model is shown in figure 18.

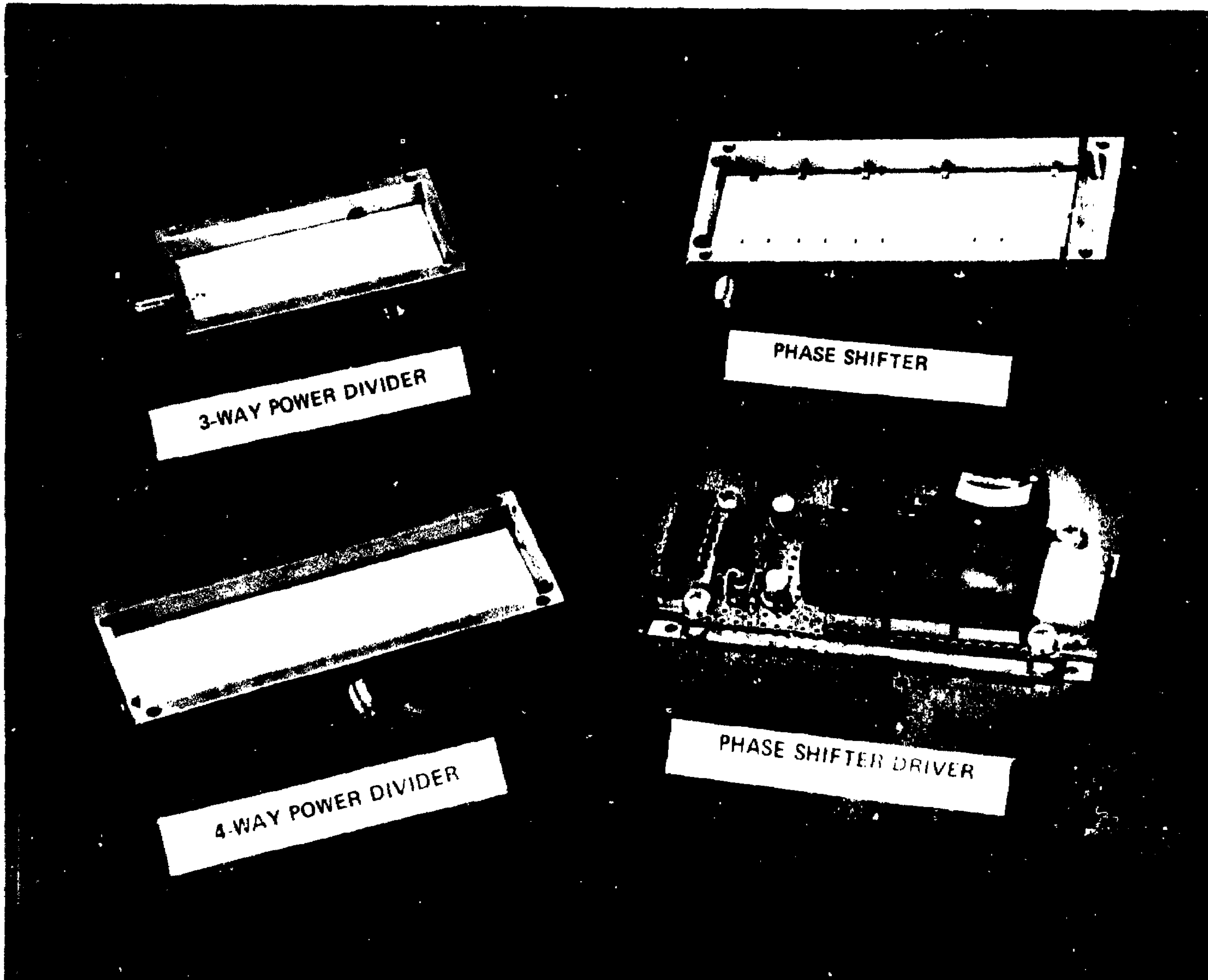


FIGURE 17. TRANSMIT ARRAY FEED COMPONENTS

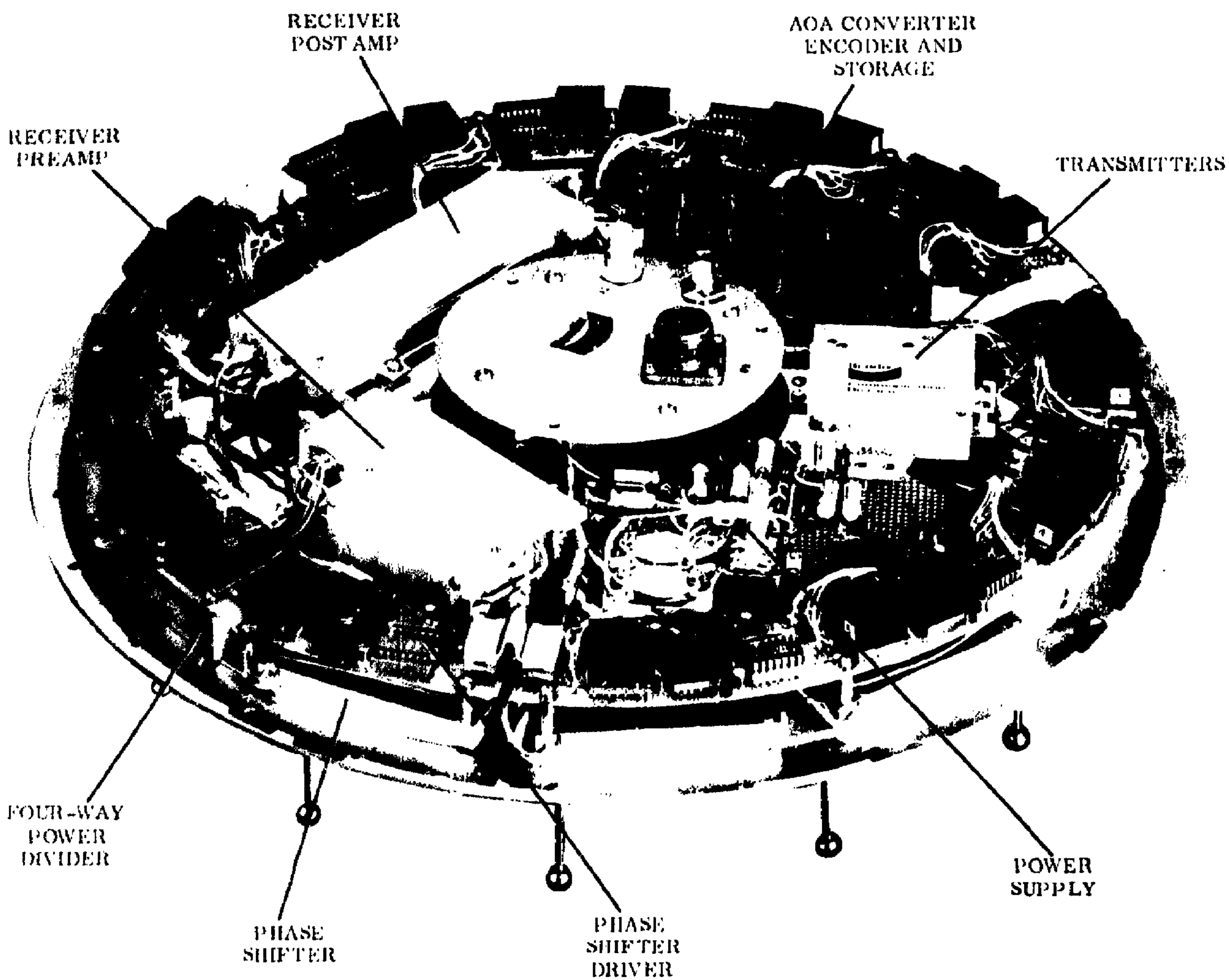


FIGURE 18. PHOTOGRAPH OF ASSEMBLED ARRAY

2.1.2.3 Performance

The transmitter array performance was measured to determine transmit beam pointing accuracy, side lobe level, beam width, and array gain in the completed laboratory model. All measurements were performed with the unit mounted on a 36" diameter ground plane. To accomplish this, external power supplies were used to provide additional power to permit CW operation of the transmit array phase shifters for antenna range measurements. (In a transponder system, for which the internal power supply was sized, the transmitter is limited to a maximum duty factor of 1% which results in a phase shifter duty factor of 3%.) All transmit array patterns were made using a 1090 MHz signal generator with 1000 Hz modulation feeding the transponder input to the transmit array feed (J1 on the unit).

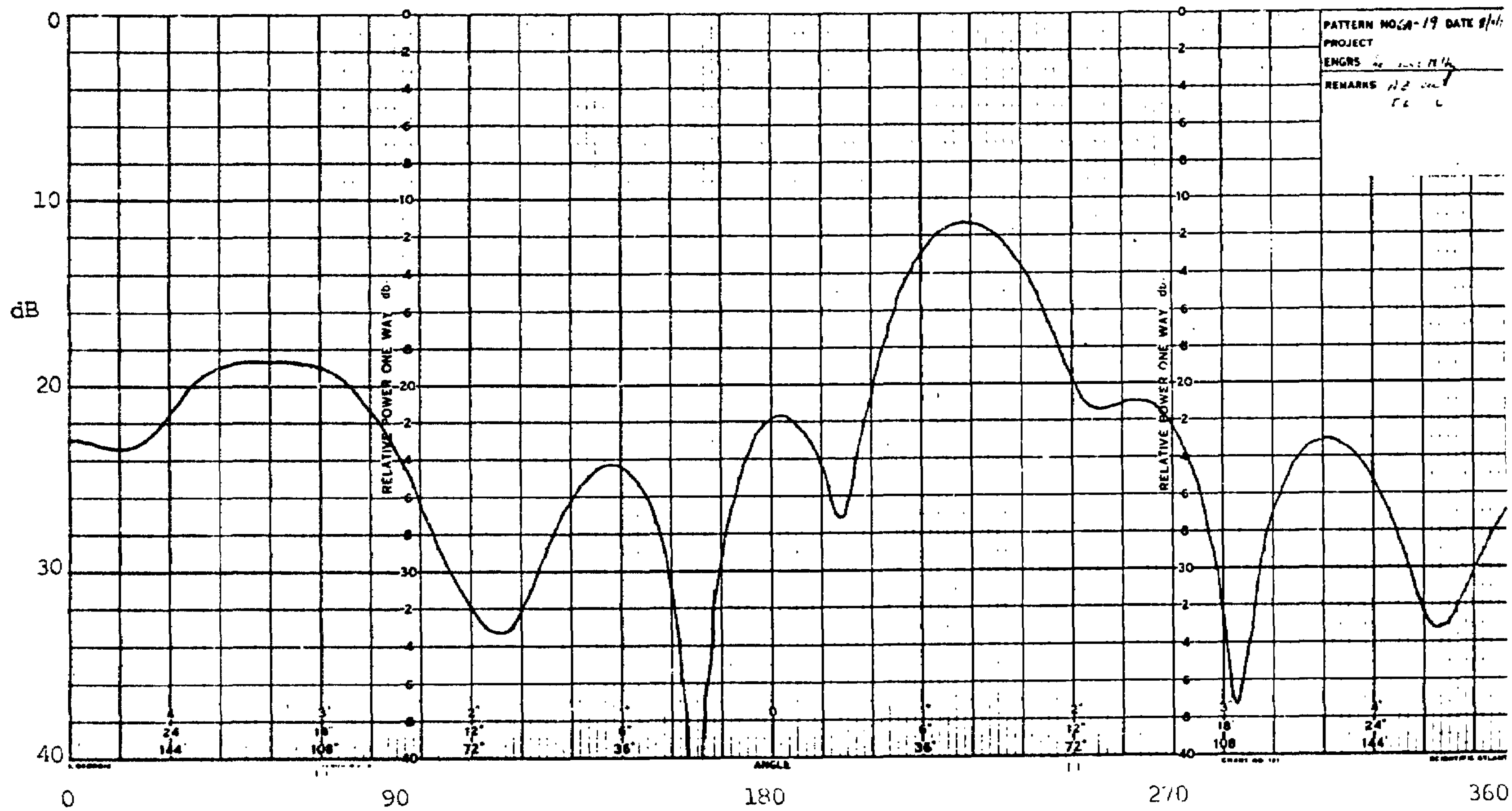
It was also necessary to disable the automatic beam selection circuitry and permit manual beam selection for testing. The microcircuit providing the five wire parallel output binary beam number 1 through 24 (U15 on the AOA converter) was replaced with five toggle switches which could be used to select the desired beam to be recorded.

Figure 19 shows a typical transmit beam azimuth pattern in the plane of the ground plane (Elevation = 0°). The average beam width is 28.5 degrees at the -3 dB point. Over 360° of azimuth there are characteristically six side lobes with two forming a rather large back lobe. Figure 20 shows all 24 transmit beam peaks located at 15 degree intervals. The average crossover occurs at -0.7 dB. Table 2 tabulates the measured characteristics of the transmit beams. From this table it is seen that narrowest and widest beam is 26 and 31 degrees, respectively with 90% of the beam falling in the 28.5 $\pm 1.5^{\circ}$ region. The average beam pointing error is 0.67° with the largest pointing errors being $\pm 1.5^{\circ}$. The variation in peak gain over 24 beams is ± 0.45 dB maximum. The average peak side lobe level over 144 side lobes for 24 beams is 10.3 dB.

Figure 21 is a typical elevation pattern through the azimuth beam center. The beam maximum occurs at an elevation angle of 26 degrees. This angle is a function of the size of the ground plane. The gain increases by 5 dB over the gain at 0 degrees elevation. The elevation angle at which the gain reduces to the 0 degrees elevation gain is 42 degrees. (i.e., the 5 dB vertical beam width is 42 degrees.) The back lobe again is the principal side lobe and is only 7 dB down.

TABLE 2. MEASURED CHARACTERISTICS OF TRANSMIT BEAMS

Beam No.	Azimuth Beam Width in degrees	Beam Pointing Error in degrees	Relative Beam Amplitude in dB	Average of 6 Peak Side Lobe Levels in dB
1	27	-1.5	-.6	-9.75
2	29	0	-.3	-9.42
3	27	-0.5	-.5	-11.92
4	29	-1.0	-.1	-10.08
5	28	-0.5	0	-11.25
6	30	0	-.1	-10.00
7	29	+0.5	0	-9.67
8	29	+1.0	-.1	-10.67
9	27	+0.5	-.9	-11.25
10	28	+1.0	-.3	-9.50
11	26	+1.0	-.9	-10.50
12	28	+1.0	-.9	-10.33
13	28	+1.5	-.9	-11.00
14	30	0	-.9	-9.75
15	28	-.5	-.6	-10.75
16	30	0	-.4	-10.00
17	28	-1.0	-.6	-10.78
18	31	0	-.4	-9.50
19	28	+0.5	-.5	-11.08
20	30	-0.5	-.4	-9.08
21	28	-0.5	-.6	-10.33
22	30	-0.5	-.4	-10.00
23	28	-1.0	-.4	-10.42
24	29	-1.5	-.2	-10.00



AZIMUTH ANGLE IN DEGREES

FIGURE 19. TRANSMIT BEAM TYPICAL AZIMUTH PATTERN

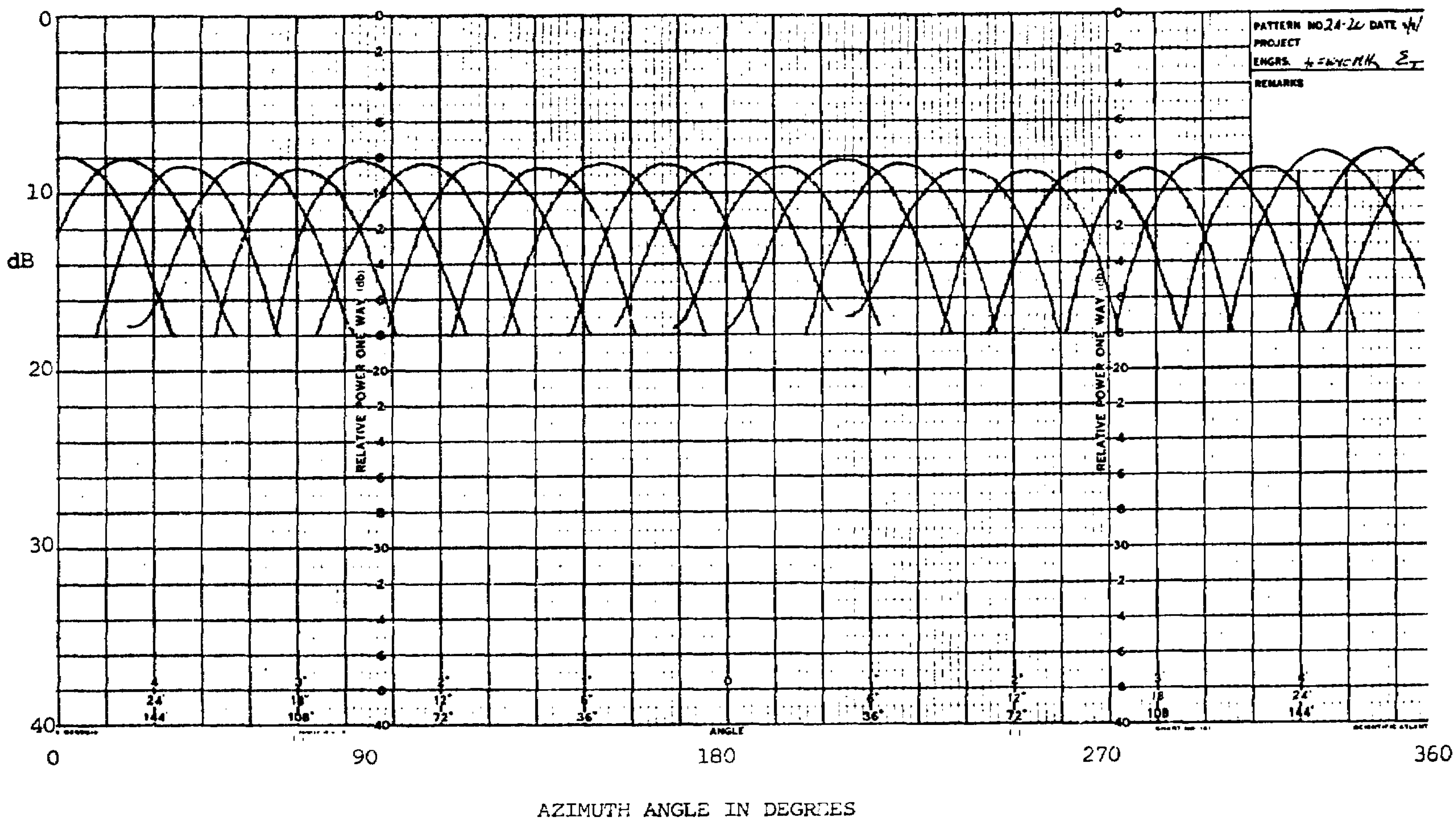
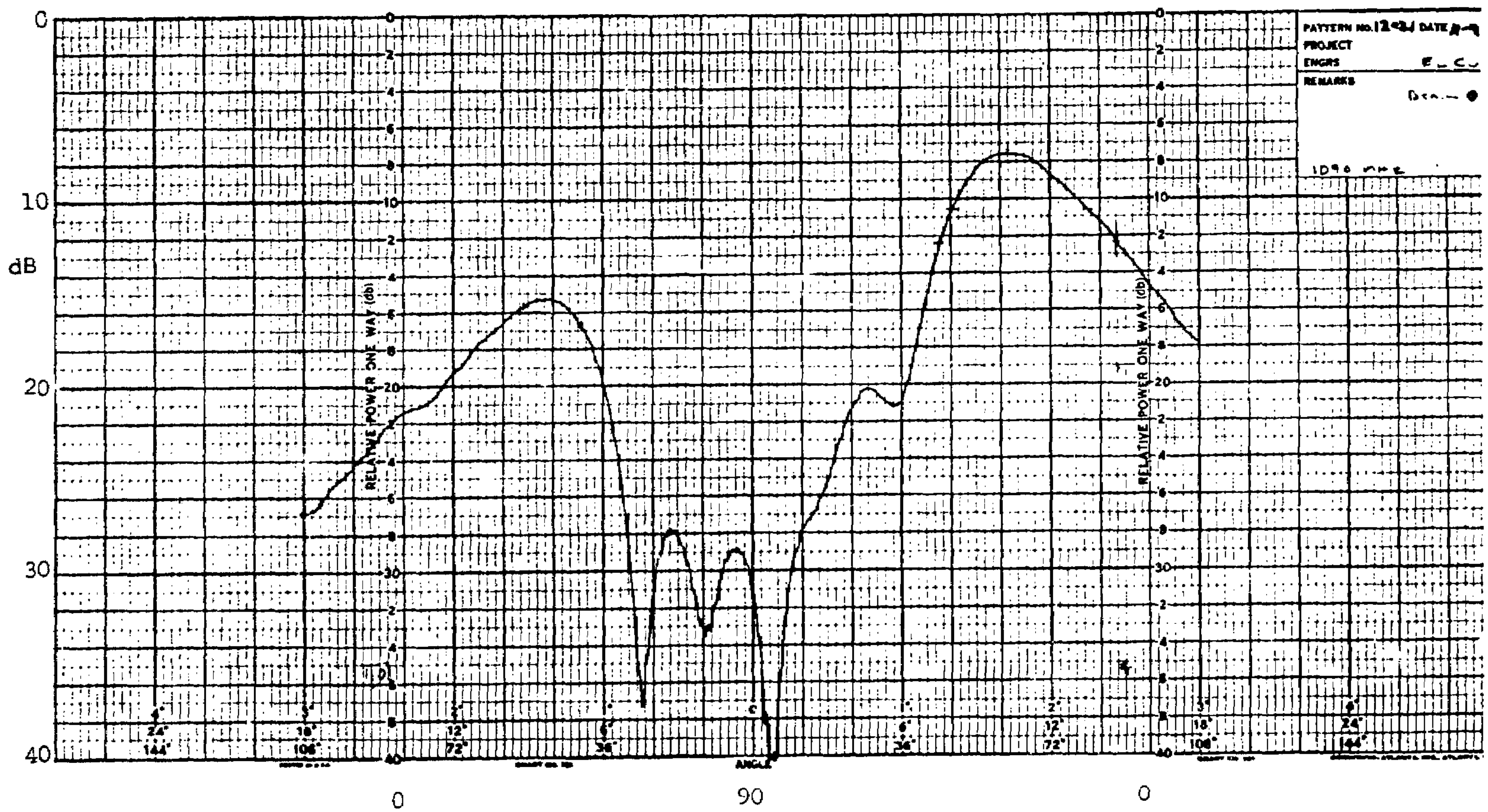


FIGURE 20. TRANSMIT BEAMS, TWENTY-FOUR BEAM LOCATIONS



ELEVATION ANGLE FROM HORIZON IN DEGREES

FIGURE 21. TRANSMIT BEAM TYPICAL ELEVATION PATTERN

The transmit array gain exclusive of the ohmic losses in the feed was found to be 12 dB above isotropic at 26 degrees elevation. The ohmic loss in the feed power dividers and phase shifters was approximately 4 dB. An additional attenuation of approximately 7 dB (6 dB + the loss in a circulator) was added to the array to accomodate existing 500 watt transponder power outputs. The resultant effective radiated power would be 630 watts (12 dB gain -11 dB attenuation, or 1 dB above 500 watts) on the beam nose.

The solid state transmitter which is included in the array is capable of producing 90 to 100 watts. To use this transmitter its output is connected directly to the transmit array feed without the attenuator. The resultant effective radiated power would be 570 watts as follows:

Transmitter Power	=	+19.5 dBw		
Feed Loss	=	-4.0 dB		
Array Gain	=	+12.0 dB		
Total	=	27.5 dBw	=	570 watts

2.2 System Test Results

2.2.1 Equipment Setup

In an operating system the transmit beams shown in the previous section would be automatically selected during the angle of arrival decision performed by the A to D converter and associated logic based on the receiver outputs derived from the receive array feed signals. The direction finding system determines angles which correspond to the theoretical crossover points between transmit beams. To test the complete system, in what has been termed auto-track mode, the array was placed on a rotating mount on an antenna range. Refer to Figure 22 Test Setup. Pulsed signals at 1030 MHz were radiated to the array from the range feed horn. Two 0.8 microsecond pulses were used at 1030 MHz to represent the IFF P1-P3 interrogation pulse pair. The repetition rate of this interrogation was 1000 Hz. A video pulse was also fed to the array logic (at J2) timed at the trailing edge of P3 to simulate the signal from the transponder which signifies a pulse pair decode. The leading edge of this pulse (termed a suppression pulse in the IFF transponder) strobes the analog receiver direction measurement into the beam steering memory system.

To determine the reply beam patterns in auto-track mode it was necessary to use a 1090 MHz, 1000 Hz 100% modulated CW signal as the input to the transmit array (J1 on the array) since the range pattern measuring equipment will not function on the 0.5 microsecond reply pulses at 1090 MHz. The radiated reply from the array was received by the same range feed horn and feed to the pattern recording equipment. The test equipment also synchronized the 1090 MHz "ON" signal to the leading edge of the suppression pulse. As was necessary for the transmit array pattern measurements, the array power supply (output ribbon cable) was disconnected from its input to the array (logic board connector) and external power applied to the system to permit continuous operation of the transmit array phase shifters. (The array supply need only power the phase shifters for a duty factor of 3% when used with a transponder and is appropriately sized.) The dual in line power connector on the logic board has the following connections and power requirements for continuous phase shifter operation:

Pin 1 and 14	-	gnd
7 and 8	-	+5 volts at 3.5 Amps
2	-	-6 Volts at 700 Ma
10	-	+12 Volts at 150 Ma
5	-	+16 Volts at 500 Ma

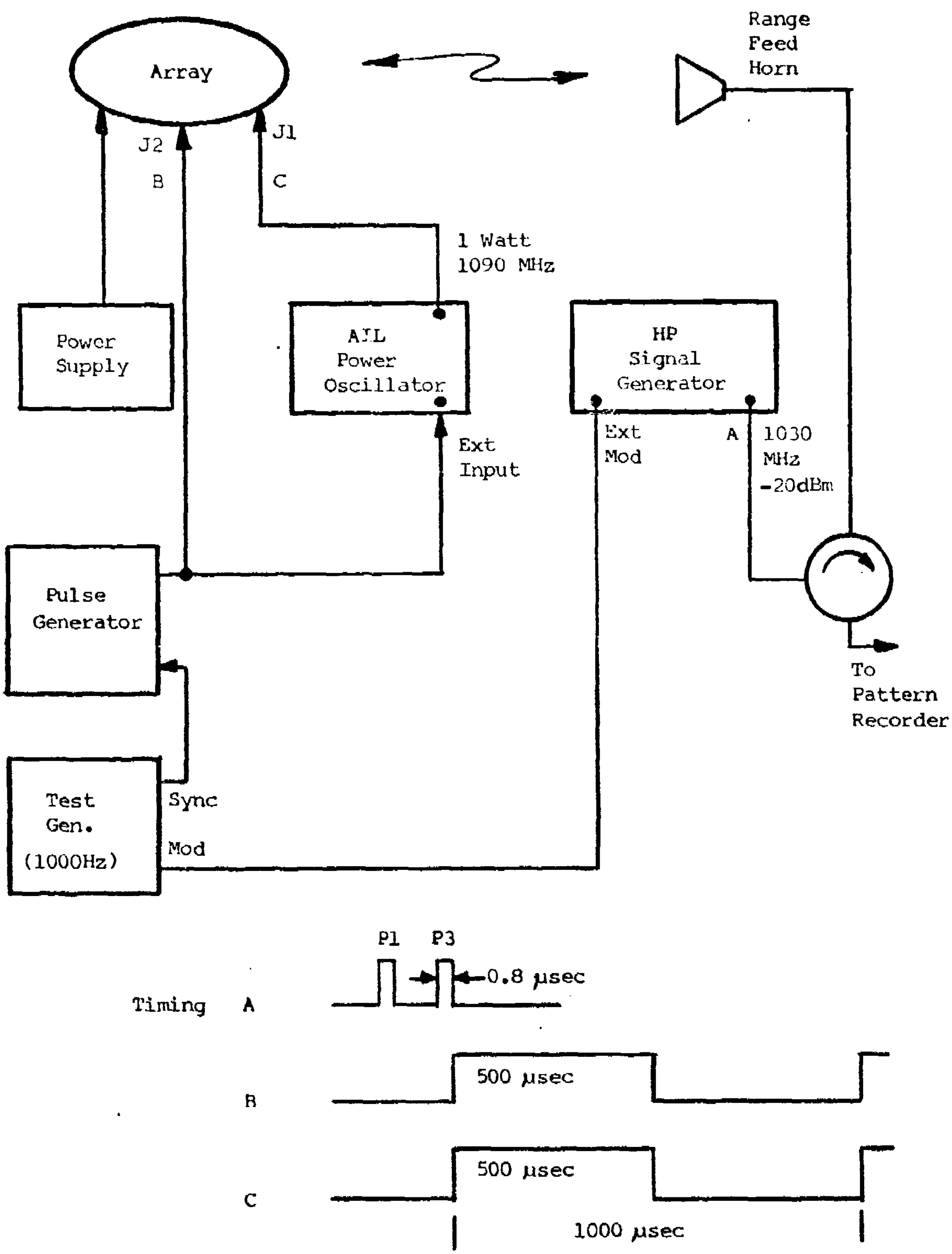


FIGURE 22. TEST SETUP DIAGRAM

2.2.2 Data

During the course of system testing over 50 patterns investigating system operation were recorded and throughout the total program effort several hundred patterns of transmit and receive array operation were obtained. Only those patterns relating to key points of operation are presented in this report.

The auto track testing was performed by rotating the array 360° in azimuth while interrogating at 1030 MHz and recording the resultant patterns at the 1090 MHz reply frequency. The idealized pattern assuming all beam crossover angles were exactly determined would be the peaks of each of the 24 transmit beams with approximately 0.7 dB ripple as shown in previous figure 20. Due to system errors in determining beam selection additional power variations up to 3 dB were allowed. The auto track patterns should then show a selected beam being retained no longer than until the reply power was approximately 3.7 dB down from the peak of a particular beam. Figure 23 is such a pattern taken at an elevation angle of 10° above the ground plane. The abscissa is azimuth angle over 360° rotation and the ordinate is 1090 MHz returned power in dB. The recorded beam peaks occurred at approximately -15 dB relative power level on the charts. The desired pattern would then lie above the -18 to -19 dB level at all azimuths. Eight of the 24 beam crossovers occur below the -19 dB level, 3 are below the -20 dB level and one exceeds -22 dB for 5 degrees of azimuth. At this elevation angle the results are close to the desired results. In fact, some relatively minor adjustments to the A to D converter sensing the receiver outputs would bring all crossover determinations within the desired goal.

Figures 24 and 25 show similar patterns at 0 and 20 degrees elevation angles. The functioning of the system is degraded at these elevation angles with correct operation occurring over 180 to 270 degrees of azimuth. It should also be noted that where degraded performance occurs at one elevation angle correct operation is obtained at the other elevation angle. This operation cannot be adjusted out since the array does not sense elevation angle.

In an effort to improve performance the operation of the array was the subject of a lengthy investigation which centered about the transmit arrays influence on the receive arrays direction finding accuracy. Previous efforts had shown the angle of arrival measurement to be sensitive to transmit array diameter and termination. The investigation even included a microscopic examination of every die bond and connection in the 12 microcircuit diode phase shifters which provide the termination for the transmit array elements during receive mode of

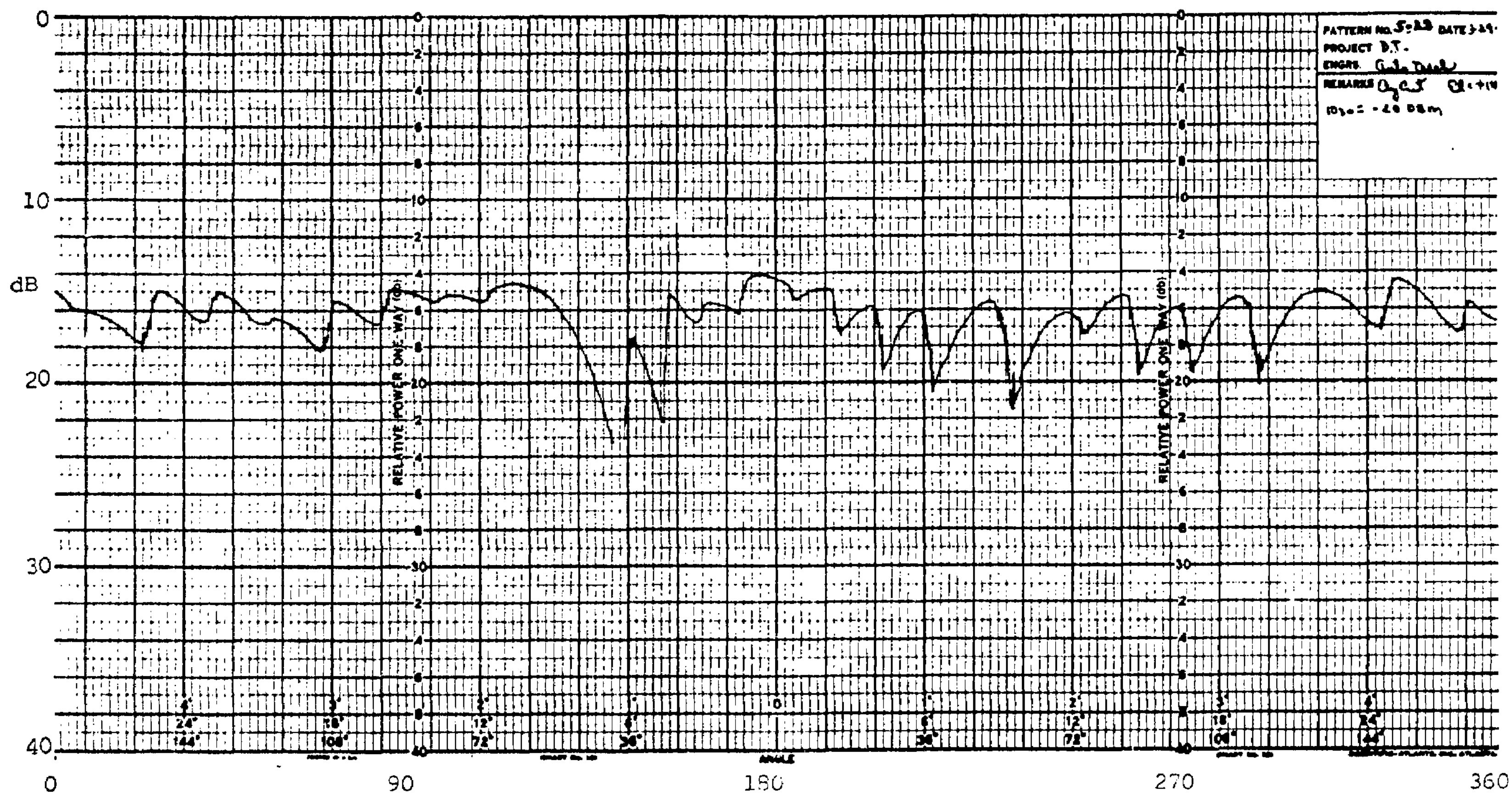


FIGURE 23. AUTO-TRACK REPLY AZIMUTH PATTERN - ELEVATION 10 DEGREES

FIGURE 24. AUTO-TRACK REPLY AZIMUTH PATTERN - ELEVATION 0 DEGREES

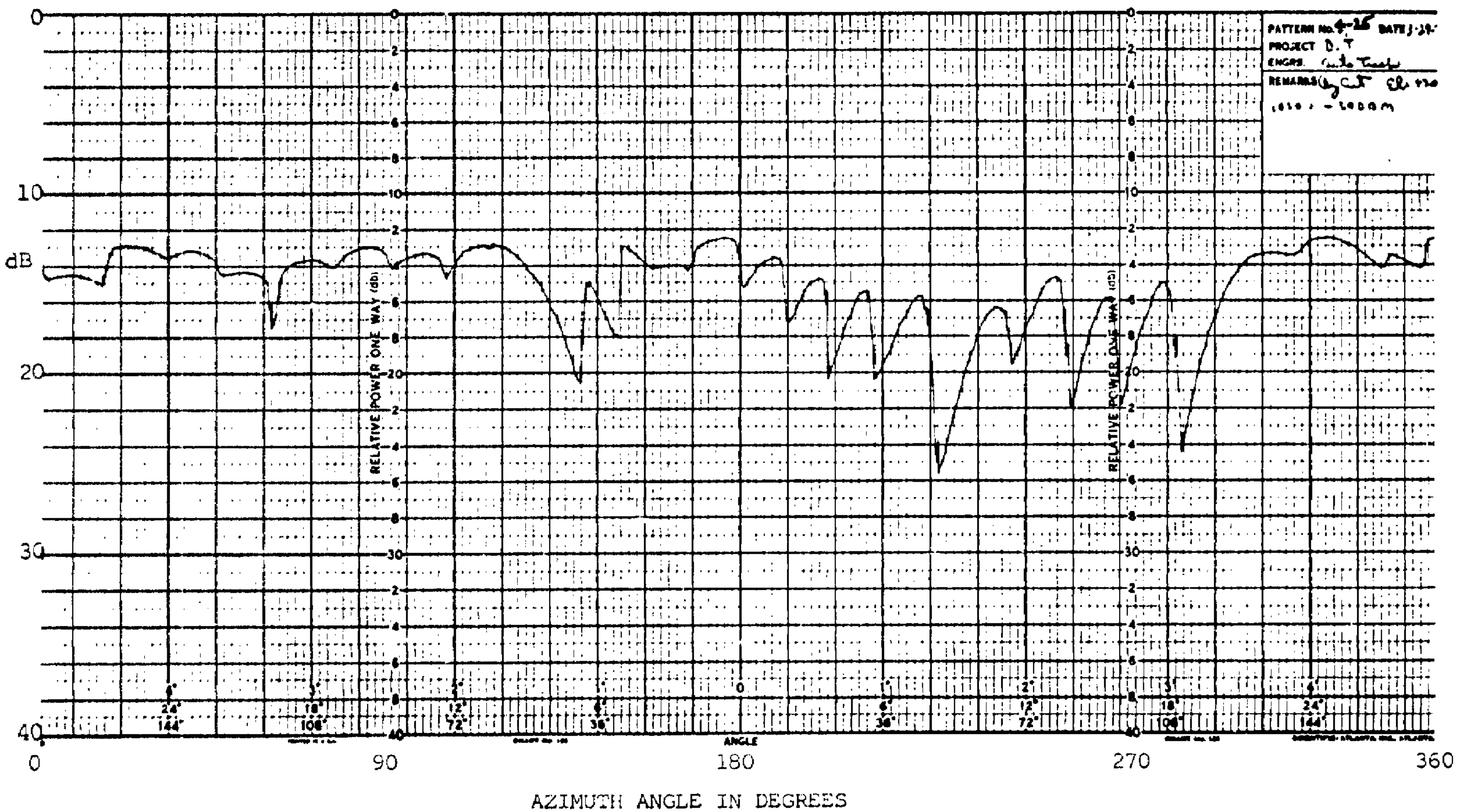
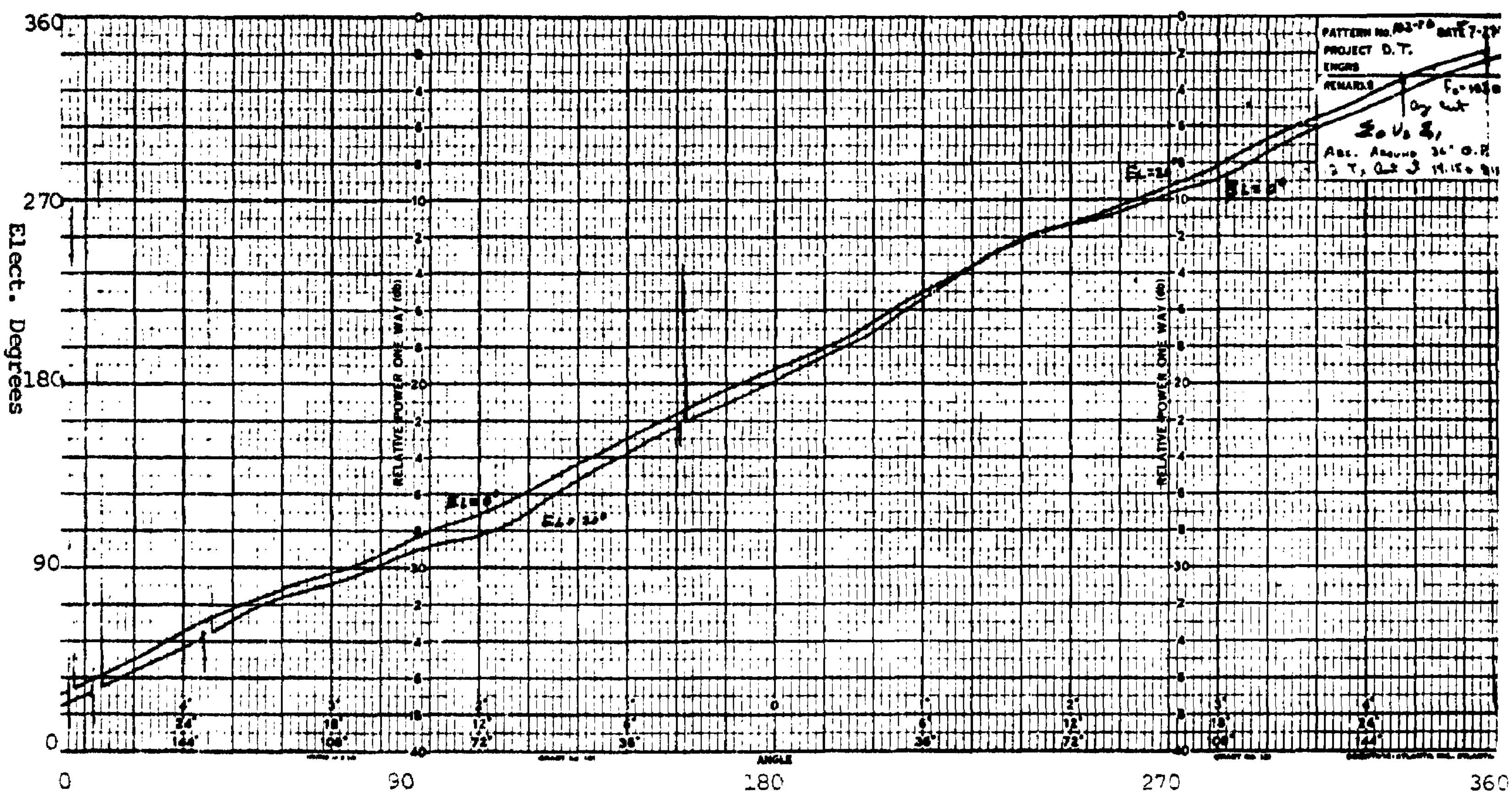


FIGURE 25. AUTO-TRACK REPLY AZIMUTH PATTERN - ELEVATION 20 DEGREES

operation. As a result of the investigation the phase shifters were reworked to incorporate a better internal grounding system and certain units had diode bonded leads and terminations replaced. The VSWR looking back into the transmit feed antenna connections was improved with all ports being under 1.5/1 at the 1030 MHz receive frequency. The linearity of the receiver output improved as a result of this effort. The auto track data shown in figures 23, 24 and 25 is also better than the initial data. The degraded operation at 0 and 20° elevation angles results primarily from the elevation angle influencing the azimuth angle phase output. The presence of the transmit array contributes to this error.

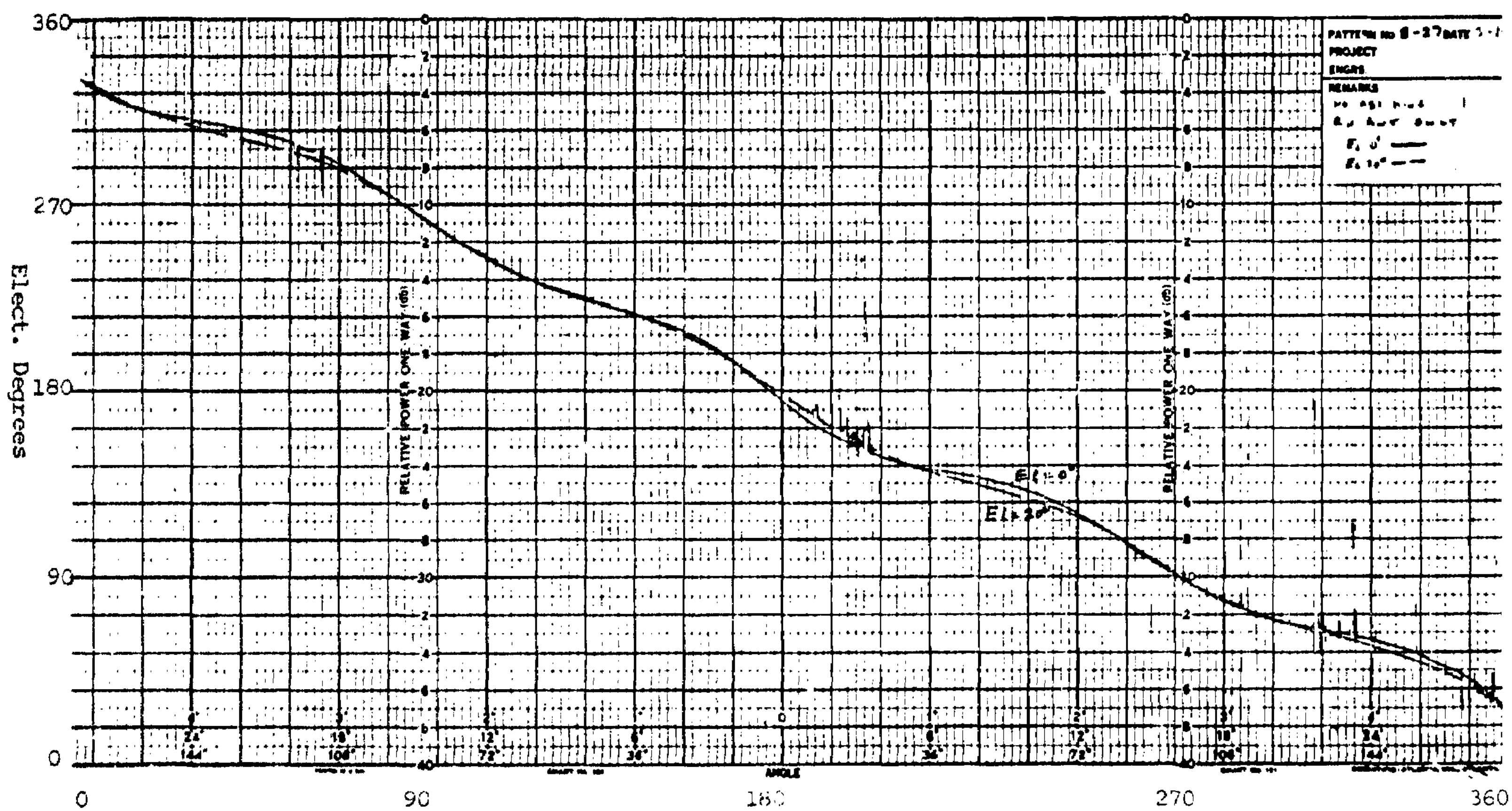
The following patterns illustrate some of these influences on direction finding accuracy. Figures 26, 27 and 28 are plots of the phase difference between the two receiver feed outputs as a function of azimuth angle of incoming signal with a CW 1030 MHz input. The ordinate covers 360 electrical degrees while the abscissa covers 360 degrees of rotation. The ideal curve is a straight line with no displacement as a function of elevation angle. Figure 26 indicates what can be achieved at 0 and 20 degrees elevation with the twelve transmit elements present and terminated with correct loads. Figure 27 is without the transmit array elements present. The elevation error is seen to be very much reduced in the absence of the transmit array. It should be noted that while this particular pattern indicates a very small error, the elevation characteristic in the absence of the transmit array still follows a pattern similar to the one shown in figure 12 for large elevation angles approaching the zenith. Figure 27 also shows a sinusoidal variation and the lack of a beneficial smoothing affect in azimuth due to the presence of the transmit array. (This pattern is measured data confirming the theoretical predicted sinusoidal variation shown earlier in figure 4.) Figure 28 shows the rather large effect of having unterminated elements present (three for this particular case) in the transmit array. System direction finding would not be possible with such a characteristic.

The following pattern (Figure 29) is a plot of the receiver phase detector output voltage (sine output) as a function of azimuth angle at elevation angles of 0 and 20 degrees with a CW 1030 MHz input. The ordinate scale is linear in voltage. This output is ideally a triangular pattern with 180° separation between extremes due to the phase detector ambiguity. This voltage is compared to 12 fixed references in the A to D converter to determine azimuth angle. (A cosine



AZIMUTH ANGLE IN DEGREES

FIGURE 26. PHASE DIFFERENCE BETWEEN CONCENTRIC AND SPIRAL PHASE MODES VS. AZIMUTH ANGLE-COMPLETE ARRAY



AZIMUTH ANGLE IN DEGREES

FIGURE 27. PHASE DIFFERENCE BETWEEN CONCENTRIC AND SPIRAL PHASE MODES VS. AZIMUTH ANGLE-RECEIVE ARRAY ONLY

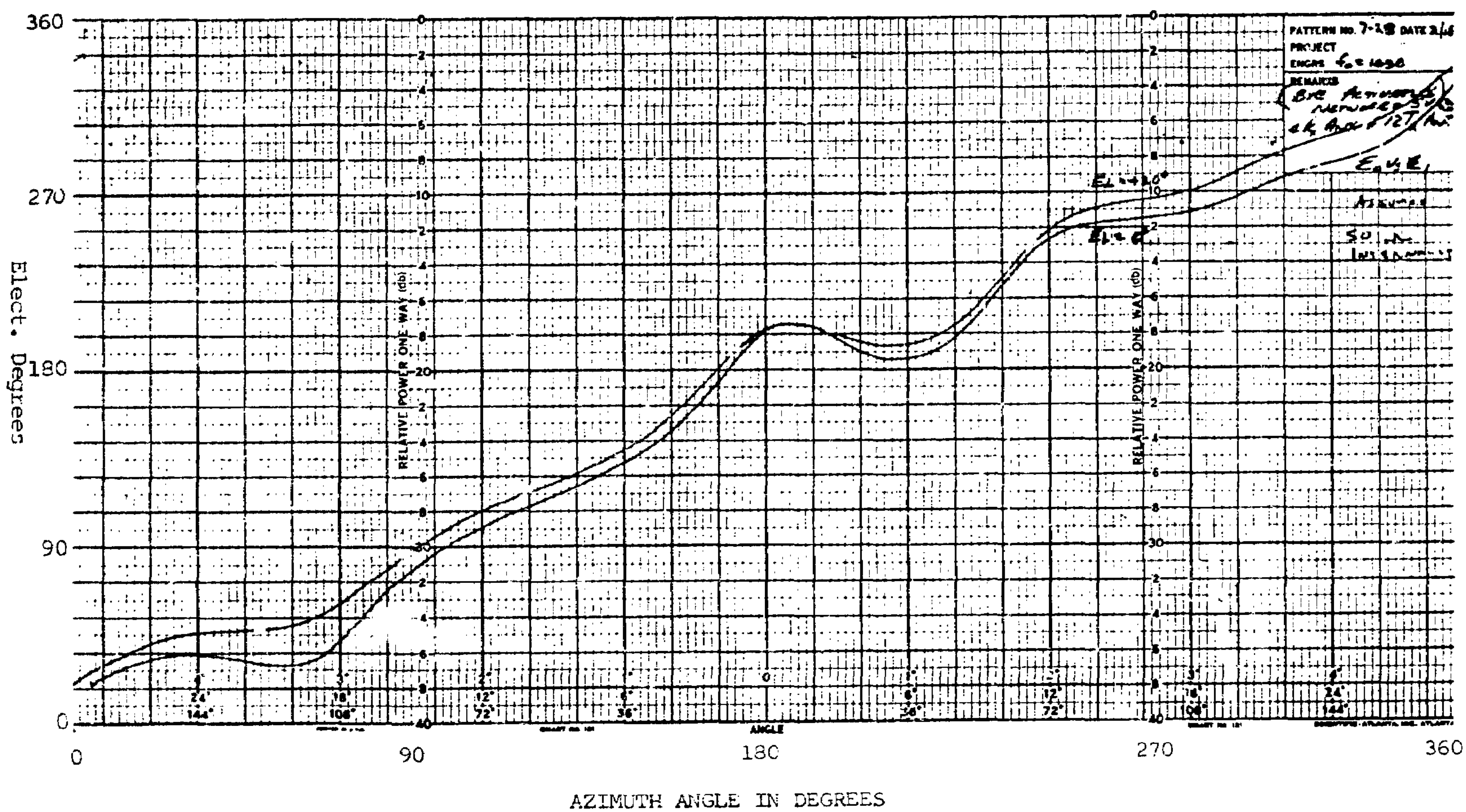


FIGURE 28. PHASE DIFFERENCE BETWEEN CONCENTRIC AND SPIRAL PHASE MODES VS. AZIMUTH ANGLE WITH UNTERMINATED ELEMENTS

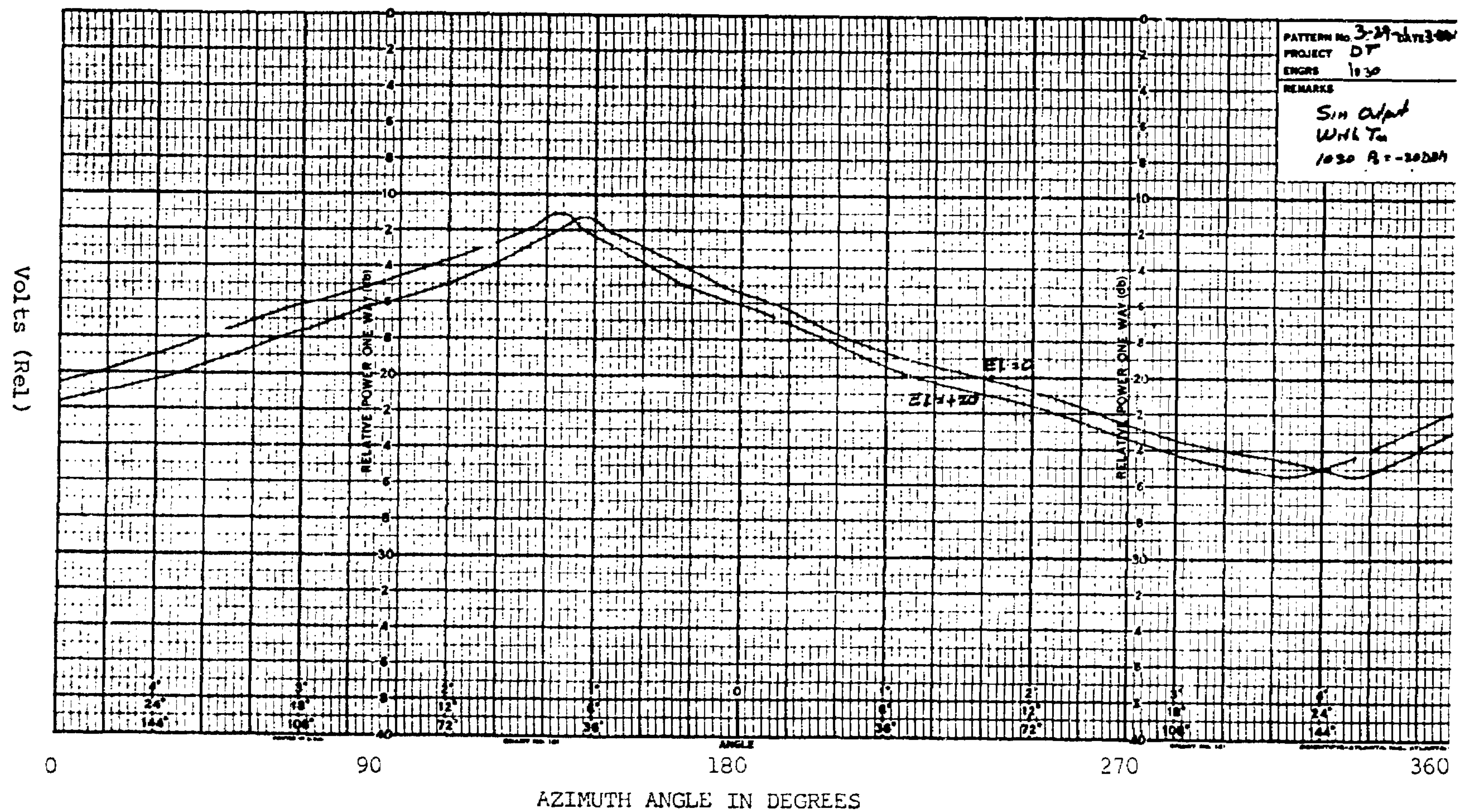


FIGURE 29. RECEIVER SINE OUTPUT VOLTAGE VS. AZIMUTH ANGLE

output is used to resolve the 180° ambiguity.) This pattern illustrates the elevation influence on azimuth direction finding accuracy which was responsible for the auto track operation shown earlier in figures 24 and 25. The magnitude of this effect appears somewhat higher on the completed unit than was obtained in earlier tests on the receive array in the presence of a transmit array with elements terminated with individual coaxial loads.

2.3 Reliability Prediction

A reliability prediction has been made on the Directional Transponder Antenna to quantize the inherent reliability of the design for the purpose of developing a requirement for future procurement contracts. The results of this prediction yield an MTBF of 1846 hours.

The prediction was made according to the techniques presented in Section 5.0 of MIL-HDBK-217A. This document also served as the primary failure rate source. A secondary failure rate source that was used to obtain data on integrated circuits and other state-of-the-art semiconductor devices was the RADC Reliability Notebook (RADC-TR-67-108, Volume II).

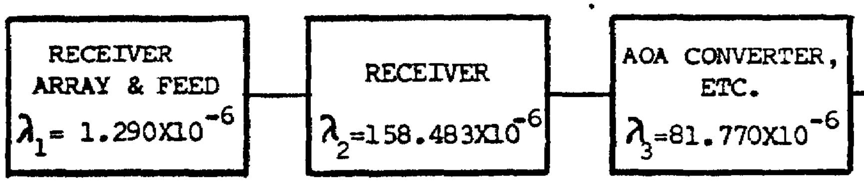
The prediction was conducted for an equipment operating in an avionic environment at an ambient temperature of 25°C. The prediction was based on a parts count prepared from the latest schematics and parts lists. Average stress levels were assigned to each generic part type. These averages were determined from circuit analysis of similar IFF equipments previously developed by BCD.

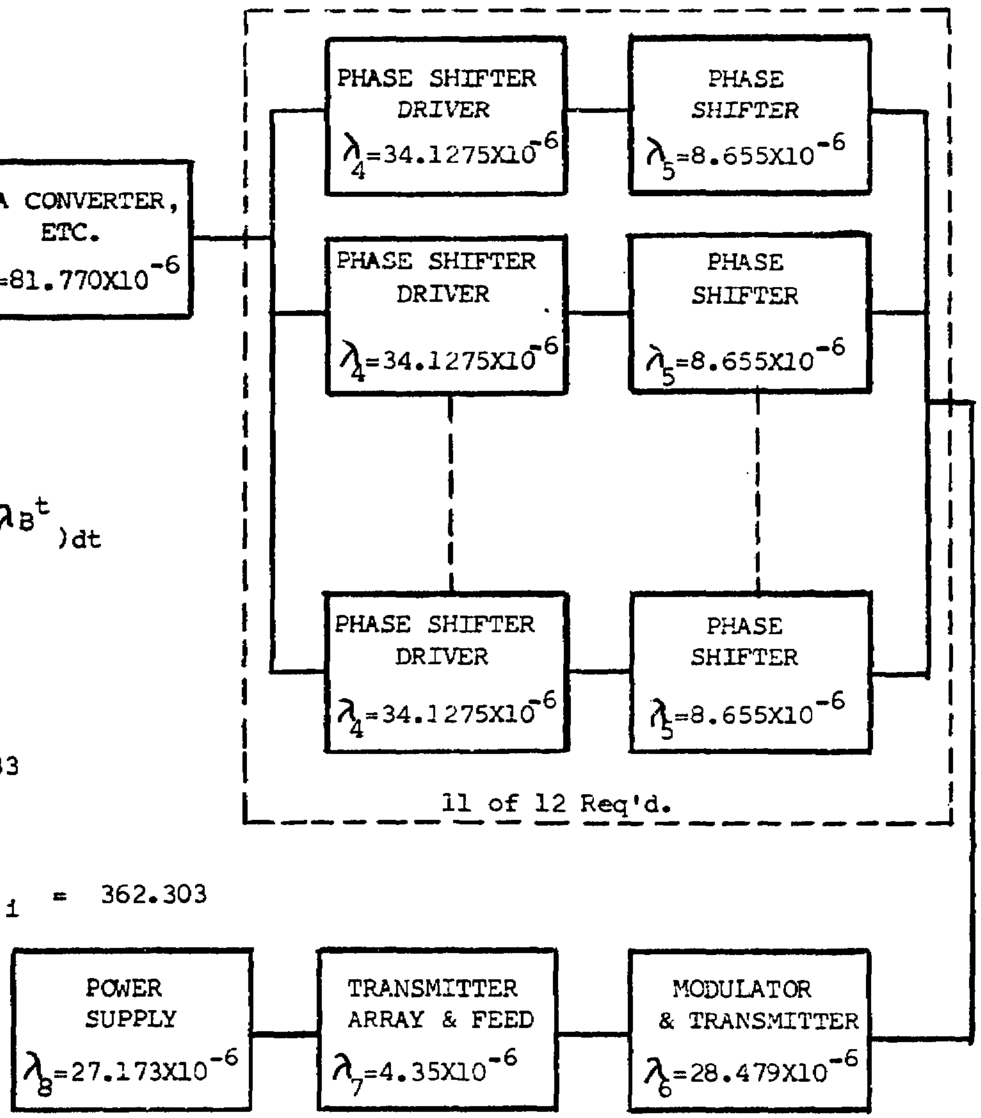
The reliability mathematical model of the Directional Transponder Antenna is given in Figure 30. This model shows the twelve (12) Phase Shifters and Phase Shifter Drivers connected in a parallel configuration with the other series elements of the system. Based on an analysis of the engineering technical data eleven (11) of the Phase Shifter-Driver elements were found to be necessary for adequate performance of the antenna. The resulting reliability equation of the system, which was integrated to determine the system MTBF, therefore reflects the binomial evaluation of an 11 of 12 parallel redundant element and six (6) functional series elements.

During the course of the program difficulties were experienced with the transmit array diode phase shifters which prompted a re-examination of these devices. Some of these difficulties were due to construction techniques which caused failure of conductive epoxy bonds within the devices. The stresses inducing the failures were those incurred by handling and by thermal stresses between materials with dissimilar expansion coefficients. These problems were corrected with improved grounding techniques to eliminate the dependence on epoxy bonds where thermal and mechanical stresses were involved. The examination also showed failure of several die bonds used to connect the diode leads to the substrate circuit. (A total of 204 diodes are used in the 12 phase shifters.)

Due to the delicacy of these wires (5Mil gold leads) the most probable cause of failure is considered to be inadvertant breakage during the testing, inspection and reworking procedures.

The remaining failure noted was in several diodes themselves. These failures cannot be explained by excessive signal levels due to the low operating power level of the phase shifter. A possible source of the failure could have been static discharges into the array during testing. When mounted in the indoor test range the array was grounded and personnel picked up a static change which often discharged to the array ground plane and probably on occasion to an element itself. The diodes are microwave pin diodes (Hewlett Packard #
The stresses experienced by the diodes under such circumstances is unknown.





$$\text{SYSTEM MTBF} = \int_0^\infty R(t)\,dt$$

$$= \int_0^\infty (12e^{-11\lambda_A t} - 11e^{-12\lambda_A t})(e^{-\lambda_B t})\,dt$$

$$= \frac{23\lambda_A + \lambda_B}{(12\lambda_A + \lambda_B)(11\lambda_A + \lambda_B)}$$

where:

$$\lambda_A = \sum_{i=4}^{5} \lambda_i = 42.783$$

$$\lambda_B = \sum_{i=1}^{3} \lambda_i + \sum_{i=6}^{8} \lambda_i = 362.303$$

SYSTEM MTBF = 1846 hours

FIGURE 30. RELIABILITY MATHEMATICAL MODEL

2.4 Transmitter Trade Off Study

For the transmitting array the use of a distributed transmitter amplifier versus a single oscillator amplifier combination was compared to illustrate the advantages and disadvantages of each approach. Three sizes of transmitting arrays were considered containing 8 and 12 elements each to further illuminate the trade offs involved. Cost, quantity of components, operating power levels and reliability was considered as well as phase and amplitude tracking considerations. Referring to the typical block diagram (Figure 31) the two forms of the transmitter being compared are apparent. Approach 1, the distributed power amplifier approach, is shown in the diagram. An array of N power amplifiers, placed between the RF beam steering system and the N elements of the antenna array, is used to form the power output stage of the transmitter. This was the approach described in the statement of work.

Approach II used in the comparison eliminates the array of N power amplifiers by appropriately increasing the power output of the oscillator-buffer amplifier combination and connecting the array antenna elements directly to the output of the RF Beam Steering System.

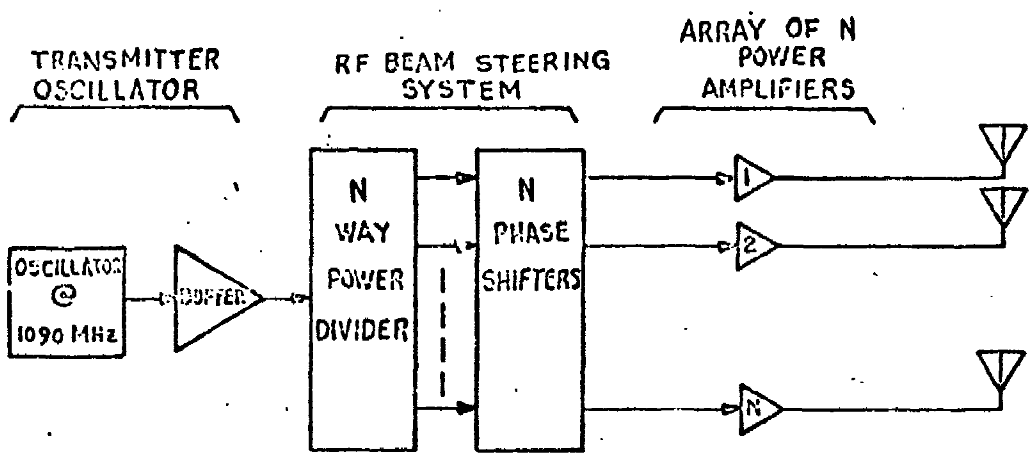


Figure 31. Typical Transmitter Block Diagram

The advantages and disadvantages of the Approaches are summarized below:

Approach I

Advantages

1) For a given solid state device the output power level can be higher for this approach.
2) Failure of the final PA is "graceful" in that performance to a degraded condition is possible when one or more amplifiers fail.
3) Isolation between array elements is increased through the feed network by the forward to reverse signal characteristics through the final power amplifiers.
4) The operating power level of the beam steering phase shifter is reduced.

Disadvantages

1) The phase shift and gain through each of the N amplifiers must be equal and remain stable over temperature variations from -55 C to -125 C and input frequency variation of ± 3 MHz.
2) A large increase in the number of RF transistors is required with the result that reliability is decreased.
3) The control of harmonics produced in the power amplifiers is difficult.

Approach II

Advantages

1) No Phase and Amplitude tracking or stability problems.
2) Simplicity - lower cost.
3) Higher reliability.
4) A single filter can be used to reduce transmitter harmonics.

Disadvantages

1) Higher power required from oscillator buffer amplifier.
2) Higher operating power through the beam steering phase shifters.

To illustrate these approaches, Table 3 Transmitter Trade Offs is presented. The table compares three different array configurations providing beam width of 30 degrees, 45 degrees, and 60 degrees to illustrate the array impact on the transmitter power levels. For purposes of this table a 6 dB RF amplifier gain was used as a logical value achievable with degeneration to provide amplifier gain stability. For the ohmic loss through the n way power divider, phase shifters and cabling a total of 3 dB was used (1.2 to 1.6 dB loss was measured in a 1090 MHz 4-bit phase shifter developed at Bendix).

The oscillator circuitry which would be used in both approaches consists of a 1090 MHz pulsed oscillator followed by a buffer amplifier. The use of isolation between the load VSWR and the oscillator has resulted in good frequency stability and is mandatory. The use of an amplifier is considered more cost effective than a ferrite isolator.

Several important points in Table 3 should be noted:

a. The highest power requirement for any transmitter stage in either approach is 84 watts which is well within the state-of-the-art capability of a single RF power transistor today.

b. The highest power level within each phase shifter in either approach is 10 watts peak (0.1W average at 1 percent duty cycle) which permits the use of low cost pin diodes as the switching element.

c. The use of Approach I, distributed RF power amplification, increases the quantity of RF transistors (and associated circuitry) from a total of 2 (required in Approach II) to 10 or 14 depending on the array size.

Table 3. Transmitter Trade Offs

						Approach I			Approach II		
Beam Width	Array "Size" or Quantity of Power Amplifier	Loss Less Array Gain dBi	Typical Array Gain dBi (Ratio)	Array Input Power for 500W ERP (Watts)	Power Amplifier Output Level (Peak Watts)	Oscillator Output Level (Peak Watts) 1	Phase Shifter Operation Power Level (Peak Watts)	L Band RF Transistors Required	Oscillator Output Level (Peak Watts) 2	False Shifter (Operating Power Level (Peak Watts)	L Band RF Transistors Required
30°	12 monopoles	15.8 dB	13.8 (21X)	20.8 watts	1.75	11	0.0	14	42	3.5	2
45°	8 monopoles	14.0	12.0 (15.8X)	31.6	4.0	16	2.0	10	64	8.0	2
60°	8 mono-	12.8	10.8 (12X)	41.6	5.2	21	2.6	10	84	10.4	2

1 Includes 3 dB feed and 6 dB Amplifier Gain
2 Includes 3 dB feed loss

From the trade-off chart it is apparent the disadvantages of Approach II, without power amplifiers, are not significant. The power levels of the oscillator buffer amplifier output are easily and reliably achieved. A schematic of such an oscillator which has been operated at levels up to 100 watts peak power out is shown in Figure 32. The phase shifter operating power level is also easily handled. A strip-line design utilizing low cost diodes (Unitrode UM 400 PIN Diodes) has been designed and tested at 50 watts peak power.

On the other hand, the disadvantages of Approach I are significant. The requirement that the power amplifiers phase and amplitude track is the most severe. Bendix has had considerable experience with phase tracking receiver and transmitter modules for both L Band and VHF arrays. To achieve phase and gain stability in an L Band transistor power amplifier, will require operating the device at a substantial derating from its maximum capability. Economic factors would also force sufficient control of the device parameters to permit operation in a fixed tuned circuit. An energy storage capacitor will also be required near each power amplifier stage to provide the pulse current for a complete SIF reply train.

The harmonic output of the RF amplifier is also significant. The second and third harmonic output levels from an amplifier are of the order of 30 dB below the fundamental. To meet existing transponder transmitter spurious output requirements it is apparent that filtering is required. Typically, low pass filters with up to 10 elements are used in transponders to reduce the second and third harmonic to 80 to 100 dB below the fundamental power output level. Thus, Approach I also requires 12 low pass filters while Approach II needs only one such device.

The question can be asked, don't the phase shifter diodes produce harmonic outputs and thus also require a low pass filter in each antenna element line. The phase shifter can be expected to produce some low level of harmonic power. The measured harmonic output power level from a strip-line phase shifter discussed in the directional antenna proposal was in excess of 50 dB below a fundamental frequency power level of 50 watts peak. For the operating levels being considered (10 dB below 50 watts) the harmonic output level should decrease even further since it is related to the 2nd power of the incident signal strength. The requirement for a filter can be best determined when measurements are taken at the representative power levels. If needed a very simple filter can be used which can be incorporated into the phase shifter.

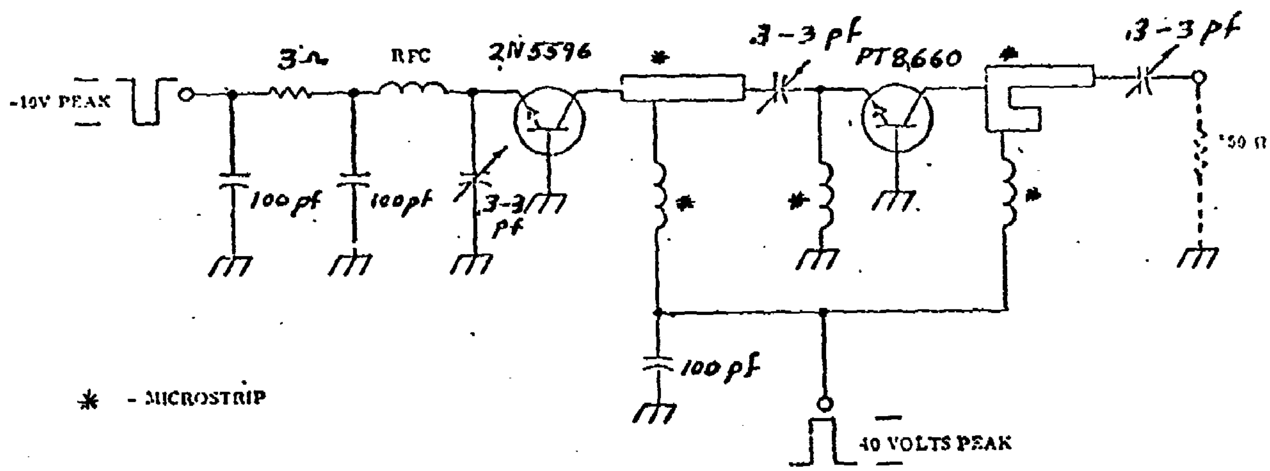


Figure 32. Oscillator-Amplifier Schematic

A cost comparison of the RF devices required by the two approaches has been made to further illustrate the trade-offs. Candidate transistors have been selected for each approach from over 60 transistors (from seven manufacturers) rated in excess of one watt at one gigahertz.

Approach I

Requires one 11-watt oscillator with 12 final amplifiers at 1.75 watt rating.

			Single Unit Cost
Oscillator	CTC D-20-28		$ 70
Amplifier	CTC D-20-28		70
Final Amplifiers (12X)	2N 4430		245
		Total	$ 385

Approach II

Requires one 44 watt oscillator-amplifier.

			Single Unit Cost
Oscillator	CTC D-20-28		$ 70
Amplifier	2N 5595		100
		Total	$ 170

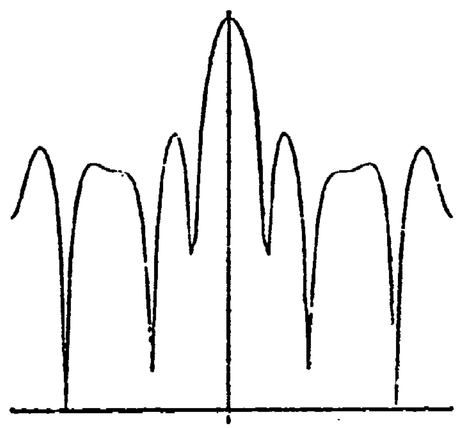
While these costs can be expected to drastically reduce as technology improves they indicate the relative cost of the approaches and favor Approach II. The cost of the circuit components is similarly weighed in favor of Approach II due to the quantities involved in each case.

An evaluation of the "graceful" failure of a distributed amplifier system was performed as part of the trade-off study. The effects of a transmitter power amplifier failure were investigated by computing the resultant pattern as each (in turn) of the 12-element drive amplitudes was allowed to go to zero. The side lobe levels increased to a level of from 8 to 9 dB below the main lobe from their nominal 11 to 12 dB level. These results are tabulated below for the two steering conditions of interest (0 degrees and 15 degrees). The "Standard Condition" used in the comparison is a uniform amplitude illumination with phase optimization from a cophased feed to reduce the side lobe levels.

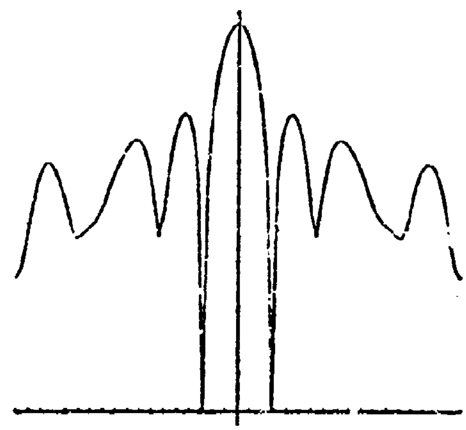
Failed Element	Side Lobe Level	
	Maximum	Location
0 Standard Condition	11.5 dB	1R & 1L
#1	8.8 dB	1R & 1L
#2	8.3 dB	1R
#3	9.2 dB	4L
#4	11.8 dB	3R & 3L
#5	9.2 dB	4R
#6	8.3 dB	1L
#7	9.0 dB	1R & 1L
15 Standard Condition	11.0 dB	4R & 4L (Back lobe)
#1	8.3 dB	1R
#2	8.2 dB	4R & 4L (Back lobe)
#3	9.1 dB	2R
#4	9.1 dB	2L
#5	8.2 dB	4R & 4L
#6	8.3 dB	1L

The resulting patterns are shown in Figures 33 and 34 which illustrate the effect of a power amplifier failure on array operation.

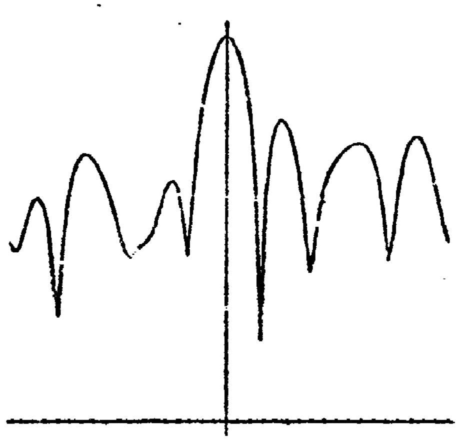
In conclusion, it was shown that the distributed power amplifier approach is undesirable and that the single oscillator, buffer amplifier approach should be followed in the development effort.



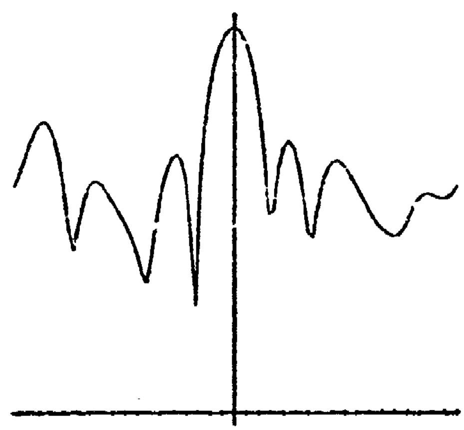
Standard Conditions



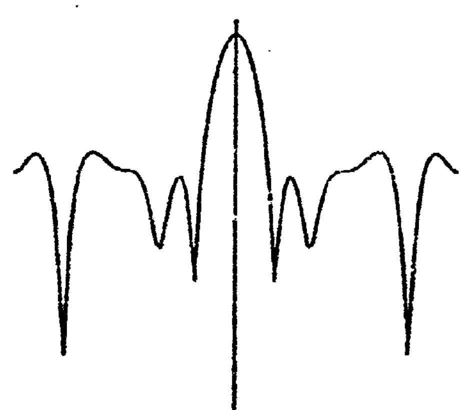
No. 1 (or No. 7) Element Failed



No. 2 (or No. 6) Element Failed

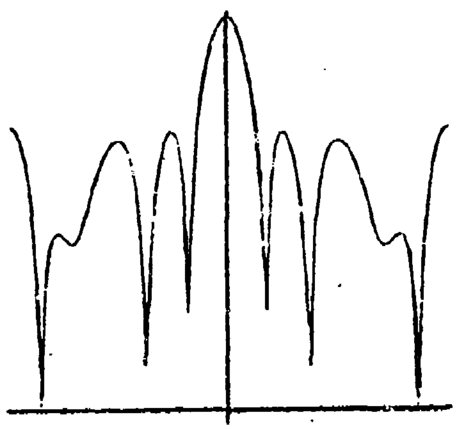


No. 3 (or No. 5) Element Failed



No. 4 Element Failed

FIGURE 33. 0° PATTERNS SHOWING EFFECTS OF A FAILED ELEMENT



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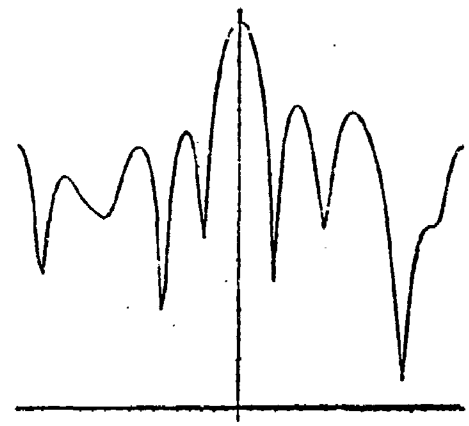
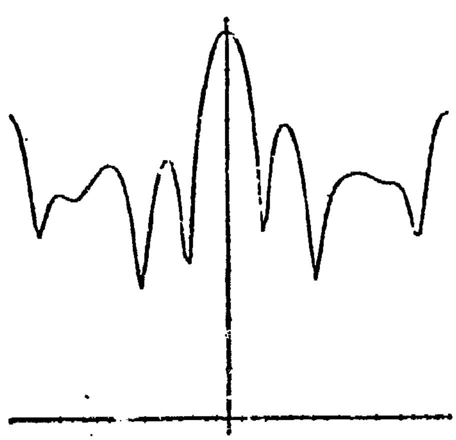
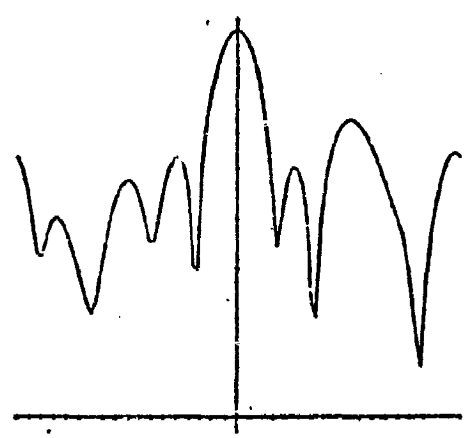


No. 1 (or No. 6) Element Failed



No. 2 (or No. 5) Element Failed



No. 3 (or No. 4) Element Failed

FIGURE 34. 15° PATTERNS SHOWING EFFECTS OF A FAILED ELEMENT

3.0 SUMMARY

A directional transponder antenna subsystem was configured in the form of a ring array of vertically mounted monopole antennas. Such an antenna would mount on top and bottom of aircraft similar to the existing omnidirectional IFF antennas. The antenna array provides direction finding on received interrogation signals and steers a directional transponder reply beam (approximately 30° azimuth beamwidth by 40° elevation beamwidth) at the azimuth of the interrogator.

The system benefits which would accrue from wide spread use of such antennas would be a reduction in the interference now present at the ground interrogator sites as well as a reduction in detectability of the radiated replies in a military environment. The quantity of this interference will certainly increase as more transponder equipped aircraft are put into service.

The goals of system compatibility and good aerodynamic and installation characteristics were met with the design approach. The antenna array was kept small in size and simplified as much as possible within the goals established for the reply beamwidths.

The array includes the RF circuitry required by the transponder which reduces aircraft cabling losses. Improved receiver sensitivity and decreased transmitter power output requirements are additional benefits of the design approach.

A laboratory model of the array was designed and constructed and the feasibility of the concept was investigated. As a result of the program the concept of a ring array to provide the directional antenna capability with the minimum of system complexity was reinforced.

The testing of the laboratory model revealed some limitations of the system which basically relate to direction finding accuracy and the beamwidth of the directional reply. It is, of course, necessary that the reply beamwidth be wide enough to tolerate all sources of direction finding error in order to ensure a proper power level for the reply to the interrogator. The test results showed that with 15 degree steering increments and a 28.5 degree beam the adjacent transmit beam crossover is approximately 0.7 dB down from peak. An error of 10 degrees in determining when to switch to the adjacent beam would result in approximately 3 dB additional power reduction by virtue of the transmit beam shape. Three dB is considered the maximum allowable system error. The average transmit beam pointing error is only 0.67 degrees, however, the A to D angle storage conversion, receiver and receive array contributed more than 10 degrees error for approximately one third of the beam crossover determinations. While a more elaborate A to D converter with more adjustments to compensate, or tune out, some of these errors would improve performance a basic limitation exists which is explained below.

It was also found in the system testing that the azimuth direction finding accuracy has some errors introduced (approximately 15 degrees) depending on the elevation angle of the incoming signal. This error contribution was smaller in some earlier tests of the receive array with a simulated transmit array terminated with more precise loads than provided by the diode controlled terminating resistance in the transmit array. The sensitivity of the receive array to the terminations provided by the outer ring of transmit elements was known (the transmit elements act somewhat as a passive reradiator) and was at first considered beneficial since the transmit array would tend to isolate the receive array from ground plane effects such as would be encountered when mounted on an aircraft.

In summary it must be concluded that future efforts take the direction of employing wider transmit beams. Two principle benefits can be expected from such an approach. One, system complexity (and cost) to improve accuracy is reduced. Two, it is to be expected that additional direction finding errors would result from an aircraft installation which could be tolerated by such a design approach.

4.0 RECOMMENDATIONS FOR FUTURE WORK

In light of the experience gained on this effort to provide a directional transponder antenna capability, the use of a ring array is considered to lead to the simplest system and should be retained. A logical change from the previous effort which is recommended for future consideration would be to combine the transmit and receive arrays in a single ring of elements and eliminate the interaction and sensitivity between separate transmit and receive arrays. Experience has shown this would lead to improved direction finding accuracy. Also, the detrimental effects of the receive array on transmit beam sidelobes would be eliminated. An array such as the following was considered and would appear to have several desirable characteristics.

Table 4 Single Ring Array

Array Circumference	2 Wavelengths
Array Diameter	0.64 Wavelengths
Quantity of Elements	8
Directional Beamwidth	60°
Quantity of Tx Beams	16
Location of Tx Beams	22.5° increment
Allowed Error	$\pm 20^{\circ}$ for -3 dB
Beam Steering Phase Shifter	4 Bit (22.5, 45, 90 and 180°)

Some advantages of this array over the previous model include:

1. Eliminates interaction between transmit and receive ring arrays.
2. Doubles the allowed direction finding error.
3. Improves theoretical accuracy of receiver array by approximately 5 degrees.
4. Requires only 8 phase shifters in the feed compared to 12 in the previous model.
5. Possibility of using one feed system for both transmit and receive, i.e., the spiral phase receive feed could be obtained by suitable settings of the phase shifters.
6. Number of antenna elements reduced from 16 to 8.
7. Complexity of A to D conversion and angle storage circuits reduced by 50%.

The only system penalty incurred is the widening of the directional beam. This, however, would seem to be desirable in light of expected aircraft influence on direction finding and beam pointing.

Early in the program a growth potential was identified to provide improved coverage in the zenith region. Such coverage was found to be best obtained through the use of an additional element with a pattern selected to complement the existing vertical coverage from the ring array. The polarization of this element should be circular in the zenith region to avoid cross polarization with respect to the interrogator antenna for a highly maneuverable aircraft. The incorporation of this feature in any future effort should also be considered.

Another recommendation concerning the use of directional transponder antennas is involved with the effects of the aircraft on the direction finding and beam pointing accuracy. Typically antenna siting on the aircraft is a compromise between areodynamic and installation considerations, system priorities and other related influences. The ring array antenna approach was selected with these factors in mind and its capabilities should be investigated in its intended environment. Related experience with a direction finding system has indicated a distortion in apparent arrival angle of incoming signals in the order of 10 to 20 degrees at certain azimuth angles when operating on an aircraft. An investigative approach employing scale model aircraft and frequency scaled arrays is perhaps the most cost effective means of obtaining the needed information to determine the practical limits to system accuracy and the feasibility of available compensating techniques.

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13. ABSTRACT

The significance of this research and development to the Air Force is the development of an aircraft transponder antenna subsystem which will receive interrogations from a maximum volume of space surrounding each antenna aperture, determine the angle of arrival of each valid interrogation and respond appropriately with a directive beam. The goals include improved receiver coverage to assure response to both air and ground based interrogators and formation of directive transmit beams to limit the volume over which transponder replies are detected. Use of directive beams would significantly reduce the system interference (fruit) levels of interrogation terminals.

The first phase of effort was an antenna subsystem design trade-off study to select an approach to minimize system complexity and antenna installation problems and to maximize performance. A small ring array was selected which can determine angle of arrival and electronically steer a relatively narrow beam for the reply.

The second and third phase of effort constructed and tested the design which evolved from the trade-off study. The transmit array diameter is 1.2 feet, contains 12 elements and produces a 28.5 degree beam width at the 1090 MHz operating frequency. The transmit beam is steered by use of phase shifters at the operating frequency. The receive array is less than six inches in diameter, contains four elements located concentric to the transmit array and provides omnidirectional coverage. The direction finding receiver operates by comparing the relative phase of two receive array output signals.

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