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A STUDY OF HIGH PERFORMANCE ANTENNA SYSTEMS FOR DEEP SPACE COMMUNICATION

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FOREWORD

In the interests of compactness, for this Status Report III for the period from November 15, 1968 to June 15, 1969, it was decided to exclude the sections which appear in the last Status Report II for the period from February 15, 1968 to November 15, 1968. Only new material and revised sections are included in this report. The following is a list of the sections omitted in this report and their pages in Status Report II.

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I. INTRODUCTION

The objective of this program is to study the most recently defined parameters for a high data rate of communication system which can operate between an earth station and a vehicle in space over great distances. An effort will be made to describe and delineate the characteristics of radiating subsystems and their internal sub-divisions which can satisfy the requisite performance criteria for an S-band system. Considerations will be given to the advance technology concerned with the ground based antenna, and where pertinent, to the spacecraft antenna as well. An effort will be made to determine the feasible design approaches for the ground antennas and its component parts. Appropriate design criteria will be investigated analytically, and where possible a comparison will be made with empirically determined results in an effort to define areas of research and development which need long term attention. The data rates of long term interest are 10^6 to 10^7 bits/second for a Mars mission and 10^4 to 10^6 bits/second for a Jupiter mission.

The ground-based antennas are discussed in this program as components of a link designed to fulfill the specific function of providing uninterrupted communication from a spacecraft to the earth at planetary distances. For obvious reasons, the most attention is given the down link aspects using a carrier frequency of 2.3 GHz, since a frequency in this region has advantages for an all-weather ground station and is presently in use in the NASA Deep Space Instrumentation Facility. It is assumed also that future mission plans will require information rates of the order of 10^4 to 10^7 bits/second with a given probability of error, 10^{-2} to 10^{-5}. These parameters imply a specific system performance in terms of bandwidth and signal-to-noise ratio. When the characteristics of the available transmitter and receiver are evaluated or assumed, the required performance characteristics of the overall radiating system are determined either directly or by implication. The overall radiating system is taken to include the combination of the spacecraft and the ground or relay station antenna equipment in their inevitable environment. Thus, for this study, certain gain and aperture requirements will be assumed nominal as parameters to satisfy a variety of space missions.

There are two general areas of concern that must be investigated relative to the ground-based receiving system which of necessity must be large compared to wavelength to achieve the desired performance characteristics. The first involves questions about the received signal to noise level or the gain that must be provided to handle it. Consideration must be given to methods by which it may be enhanced, and the limitations that may be encountered during the various phases of a mission. The second areas embraces questions about the contributions made to the noise of the communications link, the manner in which these are introduced, and methods by which they may be minimized. These questions are, of course, interrelated, and the limitations encountered are intensely practical and economic, as well as theoretical. For this study, emphasis will be given to the first area and when necessary, results of other investigations into the questions involved in the second area will be used.

The requirement of a minimum signal to noise level forces the sum of the gains of the space and ground antennas to be of some value that can be specifically determined for a particular mission. It is important to be able to allocate the antenna gains at each end of the link according to reasonable expectations concerning the practical designs and performance characteristics that can be accomplished in the next ten to fifteen years. An optimum allocation of these gains is difficult although some progress has been made along these lines. For this study nominal values shall be used as parameters in an effort to establish quantitative relationships between pertinent dimensions and techniques. It has been shown that at 2.3 GHz, dimensions on the order of 600 to 1000 ft or more are probably realistic aperture sizes to consider for the high data rates and low error probabilities listed above. Using the plans of the communication link characteristics for the 1971 Voyager Mission at 1AU as a basis for comparison, the sum of the gains on future missions can be estimated to be about 110 db to achieve a data rate of 10^6 bits/second or a 20 db increase over the gains specified in the Voyager link for which a spacecraft transmitter of 50 watts has been postulated. If the spacecraft antenna is postulated to be capable of 30-40 db of gain using a transmitter with 50-100 watts of power, then the ground based receiving system must be studied for the following range of parameters:

Antenna Gain-- 60 to 80 db
Data Rate -- 10^4 to 10^7 bits/second
Error Probability -- 10^{-2} to 10^{-5}

Final results will be given for this entire range of parameters although nominal values will be used to illustrate and expedite the discussion of various techniques during the intermediate phase of this program.

Because of the significance of the noise level in determining the overall gain requirement, many studies have been directed to a consideration of the noise that competes with the signal and is collected and introduced at the ground end of the down link. The convention of treating the noise as resulting from an equivalent antenna temperature has followed in this program. Since the noise level is highest when the antenna beam is directed at or near a noise source, attention is being paid to techniques which can be used to mitigate these deleterious effects in certain special mission circumstance.

The characteristics of high gain techniques, either electrical or mechanical, form essential parts of tradeoffs in system accuracy, reliability, and cost. Of course practical compromises must be made for certain aspects of a particular mission. These compromises will depend on the techniques available for directing or steering the receiving beam on the ground as compared with those for controlling the vehicle attitude. Three types of steering mechanisms are possible for spacecraft antenna systems: mechanical (as for large appendage antennas); electromechanical; and electronic or inertialess. Electronic techniques offer the greatest versatility with regard to communications between a vehicle in space and earth. These are two generic types: those that require external controls to phase the elements properly and those that are self-steering. The externally controlled systems, such as the conventional phased array, need an external sensor (IR, RF, or ground station) to point the beam, and a computer, a phasing network, and an attitude sensing device to point the beam appropriately. In the self-steering system, however, attitude information is presented to the antenna system by a pilot beam from a ground station, and electronic circuitry senses the phase of incoming pilot signals to position a beam in the direction of these pilot signals. Multiple beam systems may be accommodated by the use of diplexers and multiple electronic channels. Each of these spacecraft systems is being worked on by various research and development groups through the country and abroad. Appropriate results of these efforts will be used to achieve stated objectives of this program.

II. PROGRAM DESCRIPTION

As has been discussed in earlier reports on this program, there are basically two fundamental kinds of antenna systems that can be used in applications requiring large apertures. The first is a large mechanically steerable paraboloidal reflector or a number of smaller reflectors of this type which are connected and fed as an array and mechanically steered as individual radiators. The second is a phased array with stationary or fixed apertures composed of subapertures whose relative phasing controls the direction of the antenna beam. Thus, this program considers the various aspects and organizations of the following generic types of large ground based antenna systems:

A. A SINGLE LARGE APERTURE -- mechanically steerable.

A system of this type will be discussed in this study only to provide a basis for the comparison of performance characteristics with the other systems listed below.

B. AN ARRAY OF LARGE DISHES -- each of which is mechanically steerable.

The appropriate organization of a system of this type is considered herein with respect to the element spacing and their interaction.

C. A PHASED ARRAY OF SMALL CLOSELY SPACED ELEMENTS ORGANIZED INTO SUBAPERTURES-- electronically steerable.

Most of the effort in this program will be concerned with the various organizations, the feeding techniques, and the elements appropriate to this type of system.

D. A SELF-STEERING ARRAY -- rapidly switched multiple beams or adaptive systems.

Systems of this type can be used to mitigate the effects of high intensity noise sources and employed in conjunction with the type C systems (above) to accomplish optimum mission performance. The feasibility of application of these techniques for a high data rate communication system is being investigated during the course of this program.

Consideration is being given to the capabilities and limitations of each of the above types during the course of this study and a report made in the above listed categories.

During this report period, the activities performed were a result of a cooperative effort between the personnel of the Center for Research (CRES) at the University of Kansas, and the Electroscience Laboratory (ESL) at the Ohio State University. Although some of the results and information described herein were obtained in one institution and some in the other, this report, as have previous reports, will be written with the idea of integrating the results of various research efforts and techniques. Results of this investigation will be described in such a way as to implement the objectives of the program without regard to the actual source of the material, whether it be obtained by the above mentioned institutions or by reference to activities outside this specific program. It will be the purpose of this report to glean as much pertinent material as possible and to organize it into a form which permits a quantitative comparison of the various high performance antenna systems.

In an effort to expedite the activities of this program, it was decided by the program manager at CRES and the technical monitor at ERC/NASA that a slight reorganization of the duties and responsibilites of the various groups involved was necessary. As has already been indicated, the major portion of the work in this program is being done as a cooperative effort between CRES and ESL. At the conclusion of the study phase, a final report will be made of the entire study and submitted to a group of expert consultants as described in the original proposal and modified below. This group will serve as an evaluation team and will assist with the interpretation of the various results obtained during the course of the program, including the details of the over-all system problems and the individual problems associated with the antenna subsystems. The outcome of this study will be a series of recommendations to ERC concerning the pacing technology which needs long term research and development. Appropriate design approaches and performance criteria will be suggested, primarily as they pertain to the ground based antenna subsystems and the subsystems on the space vehicle in an effort to optimize the overall performance characteristics of the down link (toward the earth) portion of the communication channel.

This program has been active for the past twenty-four months and has uncovered a number of technical details that need further consideration and more recent fundamental data. Data are now becoming available that concern the performance characteristics and production costs of low loss transmission lines, radiating elements, and other subsystem components. Since these relationships are the primary factors which govern the establishment of criteria for a large scale antenna design, this program has been extended until 15 November 1969 for the following work statement and personnel organization.

REVISED STATEMENT OF WORK

The Center for Research in Engineering Science at the University of Kansas proposes to extend its present study program with Electronic Research Center of NASA. This extension will continue and update four of the seven items listed in the original work statement to include considerations and estimates of the component costs involved in various array configurations. In addition, a fifth item is added which is pertinent to the processing techniques necessary to limit the external noise or interference in certain portions of deep space missions. A sixth item is included to provide a quantitative method for studying and optimizing the overall cost of the various types of ground antennas which appear to be most promising for a high data rate system. This program is to be accomplished as a cooperate effort between the personnel from Ohio State University and the University of Kansas. The extended program will include but not necessarily be limited to the following items as revised. The underlined portions of these tasks are revisions of those listed in the original proposal.

1. A continuing effort will be devoted to an intensive review and assessment of the research programs and techniques studies in progress or recently completed which may have influence on the objective of this program. This additional study is to assist ERC/NASA in assuring that no significant matters or techniques on electronic beam shaping and steering have been slighted or overlooked.

2. A study will be made of various types and sizes of radiating elements and their associated control circuitry in an effort to evaluate their potential in a large ground based array with a very large number of elements. <u>This investigation will be concerned primarily with low noise circuitry to provide the phase and amplitude control of the elements of the array. The circuit may include mechanical or ferrite phase shifters or the use of integrated semiconductor devices and hetero-</u>

dyning techniques. Consideration will be given to the state-of-the-art in techniques for phase controlling individual elements and groups of elements. An assessment will be made of their adequacy in providing control sufficient to satisfy the requirements of the system.

3. A study will be made of methods for arranging, grouping, exciting and interconnecting the requisite number of elements to provide the appropriate radiation characteristics from array antennas. Particular consideration will be given to the investigation of novel feeding and phasing techniques which would either significantly reduce array costs or increase their flexibility.

4. A study will be made of methods of achieving a capability to handle several satellites at planetary distance. The study also includes evaluation of the feasibility of providing rapidly switched multiple beams for communication with near earth orbital satellites.

5. A study will be made of the feasibility of switching from a self steering or adaptive array where the pattern is determined by the size of the subaperture to one where the steering is accomplished by externally controlling the phase between elements so that the pattern is determined by the entire radiating aperture. This switching is to be accomplished by an appropriate signal processing scheme which is actuated by the externally generated noise or interference level. Such a scheme will produce a system capable of more efficient performance in the presence of high external noise and interference levels. For example, mission problems presented by the occultation of the sun and other high intensity noise sources will be minimized. The capability of switching to such an externally scanned accurately boresighted system must always be maintained.

6. The Electronic Reseach Center of NASA is currently developing the capability for simulation of communication systems. It is desired to expand this capability to include array antennas.

The objectives of this study will be to provide the following:

a) To identify and to describe by analytical means the pertinent parameters which should be considered in the analysis of array antennas such as element type and configuration gain, beam-scan angles, noise temperatures, data rate and line loss. Also included should be the associated computer parameters.

b) The inputs to the analysis will be in the form of discrete point inputs. Data will be generated for array antennas relating weight and costs to the pertinent parameters which will have been established in a).

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REVISED ORGANIZATION

III. ACTIVITIES DURING THE PERIOD

During this report period several aspects of the Program Description (Section II) were pursued. These items are to be summarized briefly in a qualitative manner in this section and reported in detail in the appropriate portion of the Technical Summary of Section IV.

1) A continuing effort is being made to determine quantitatively the performance characteristics of an array of independently steerable paraboloids by mechanical means. Consideration is being given to the proper size and separation of large disk antennas to achieve the requisite high performance characteristics over a $\pm$ 60° angle of scan. A minimum separation distance must be determined in order to utilize a given aperture size most efficiently. However, as the separation is decreased, the interference between adjacent paraboloids becomes important, especially at large scan angles. This interference phenomenon is being investigated by several theoretical approaches in an effort to determine quantitatively the pattern degradation of closely spaced paraboloidal antennas which can be mechanically scanned. As the separation is increased, the formation of grating lobes in a large array of parabolic reflectors presents a problem which requires a detailed study and a quantitative assessment of the results of overall system performance.

2) During the period, a continuing effort has been made to uncover components and techniques that would provide a low loss system for a large phased array antenna as described in subsection IV-C. Thus, subsection IV-C-(5) entitled "Circuit Components" has been used to discuss various aspects of the need for a low loss system. This discussion considers the individual components, the feeding and distribution systems, as well as the possibilities for combining other scanning techniques with phase shifters to achieve optimum system performance. Some of these problem areas and components will need further study and some quantitative improvement before a large phased array can satisfy the basic objectives of this program.

3) The most active effort in this report period was concerned with a study of possible application and the experimental implementation

of adaptive antenna arrays. An "adaptive antenna" may be defined as one that modifies its own pattern, frequency response, or other parameters, by means of internal feedback control, while the antenna is operating. Such automatic control of the antenna characteristics may be used: (a) to exclude interfering signals from the output of the antenna by filtering in both the space and the frequency domains, thus reducing the sensitivity of the receiving system, (b) to maintain antenna performance in the presence of a changing near-field environment. Various applications of these characteristics have been considered and are discussed in subsection IV D.

Most of the work in this period has been devoted to the practical aspects of implementing a feedback control scheme based on an algorithm for the minimization of the least mean square (LMS) error. Utilizing this method, the weighting coefficient for each element of the array is continuously adjusted (in a feedback loop) in a way that forces the output from the antenna to be equal to a "desired response" in at least mean square error sense. The desired output is specified by either the expected angle or arrival or/and the spectrum of the communication signal.

IV. TECHNICAL SUMMARY

The requirement of a constant information rate of the order of 10^6 bits per second with a given probability of error implies a specific system performance in terms of bandwidth and signal-to-noise ratio. In any communications link, the data rate system parameter, R_D, can be given as the product of the following three factors

$$R_D \; \alpha \quad P_T G_T(f) \quad \frac{\lambda^2}{4\pi R^2} \quad \frac{G_r(f)}{T_r(f)}$$

where the constant of proportionality directly involves such factors as data quality which is determined by the information coding method employed, and inversely the various loss factors in the transmission link. The bracketed terms list the design system and mission parameters as follows: the first bracket contains the transmitter parameters; the second bracket contains the transmission media or free space loss characteristics; the third factor involves the receiver parameters which are the primary concern in this study. Based on Shannon's work, the limiting value of the data rate in terms of signal-to-noise ratio and bandwidth is given by the expression

$$R_D \leq B \log_2 (1 + R_o/B)$$

where

$$R_o = \frac{S}{N} B = \text{information rate parameter}$$

The maximum data rate can be approached with negligible error by a proper choice of coding technique.[2, 3, 4] A simple and fairly efficient technique, for example, is coherent biphase coding. The characteristics of this code in terms of signal-to noise and bandwidth-to-data-rate ratios, and its relation to the Shannon limit are shown in Figure IV-1.

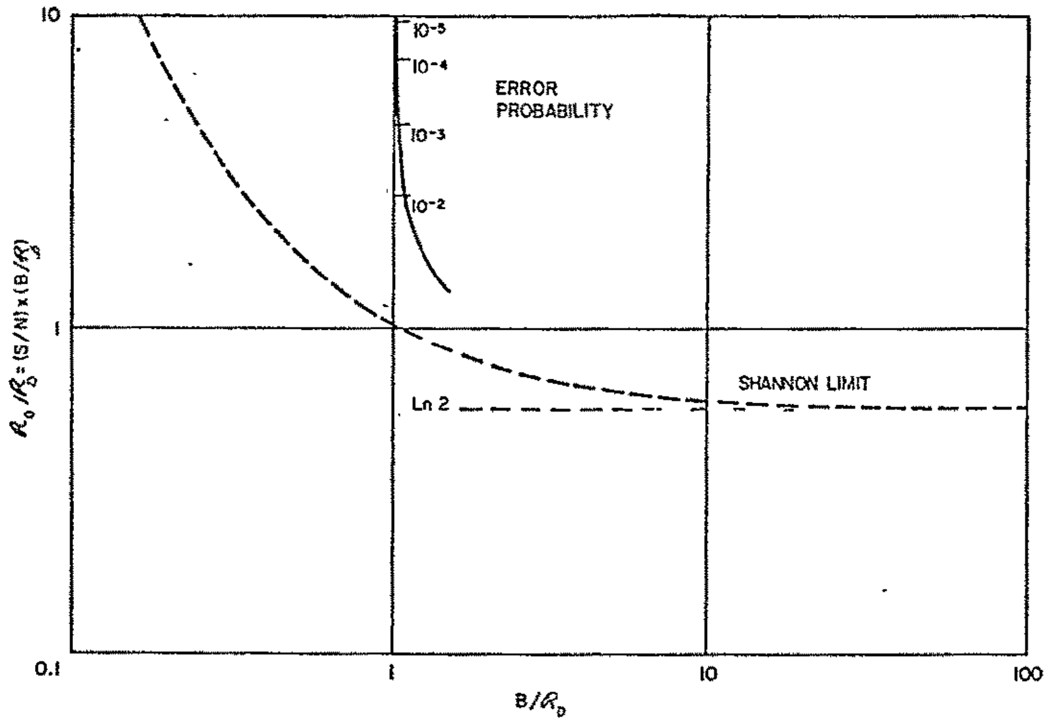

Figure IV-1 Efficiency of Biphase Coding

For small error probabilites, the figure indicated that the bandwidth required must be comparable with the data rate ($B/R_D = 1$). Thus an increase in SNR as measured by R_o/R_D is serving to reduce the error probability without appreciable effect on the bandwidth requirements. Tolerable error or probabilities range from 10^{-5} to 10^{-2} depending on the type of data.[5] Thus the practical limit for the product of signal-to-noise and bandwidth, even with a simple code, need not exceed the ideal limit by more than an order of magnitude to provide acceptable performance. The expression $R_o = (S/N)B = 10R_D$ will therefore be taken to represent a practical relationship between signal and bandwidth and the limiting noise level. (The actual relationship for a specific system design will depend on the particular coding scheme adopted as well as on error-rate requirements). Thus,

$$\frac{S}{N} = 10\frac{R_D}{B} = 10 = 10 \text{ db}$$

correspond to the requisite error probabilities.

In view of the background material discussed above, it is possible to make some general assessments of the gain and associated aperture size required to provide nearly continuous communication between the ground and the spacecraft of future mission. It can be anticipated that a gain of 60 to 80 db will be needed for the ground antenna. The diameters of circular apertures corresponding to these gain values at 2.3 GHz are 200 and 2,000 feet, respectively. This is based on the supposition that the beam formed is always perpendicular to the aperture during the steering processes and that an allowance is made for taper and other losses inherent to the antenna type. The 3-db beamwidths are on the order of 2.2×10^{-3} and 2.2×10^{-4} radians, respectively. In this section consideration is given to problems associated with satisfying the aperture and gain requirements with various types of ground based antenna systems. Each of the candidate types is discussed on the basis of its suitability to long range communication systems whether these antennas or their essential components have been developed, are in the experimental form, or are only in the conceptual or planning stages. Thus, each system is presented in terms of its capabilities and limitations even though some of the crucial component devices and techniques are still being developed. In some cases, the expected ultimate performance must be discussed in terms of a series of competing parameters whose final value is as yet unavailable.

In deep space communication systems requiring high data rates it is necessary to have a very large receiving antenna in order to achieve a SNR which will yield the error rates described above. Ultimately, as the distance or data rate increases, the required aperture may become too large to be constructed as a single antenna element as described in subsection A, and it is necessary to array several smaller apertures as described in subsection B and C. The upper limit on the subaperture size may be imposed by such factors as atmospherically induced wavefront distortion or unobtainable phase tolerances. An additional advantage of subdividing the large aperture into smaller subapertures is the possibility of arraying and processing them in a manner which will give improved performance over that of a single antenna. For example, the weighting factors on the subapertures as elements of the larger array might be adjusted to place a null or region of low sidelobes in the direction of an interfering source, thus reducing the effective array noise temperature. This process, however, requires sophisticated techniques and will be discussed in subsection D.

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B. AN ARRAY OF LARGE DISH ANTENNAS

1) Introduction

As has been mentioned before, an array of independently mechanically steerable paraboloids with proper size and separation may be one of several workable approaches capable of achieving the high gain requirement for the DSCS. To provide the requisite scanning angle of $\pm 60^{\circ}$ without interference between adjacent paraboloids, the spacing between reflectors must be kept at a reasonable distance which is larger than the diameter of the paraboloids. Thus, a minimum separation distance must be determined which utilizes a given aperture size most efficiently. As the separation is increased, the formation of grating lobes in a large array of parabolic reflectors constitutes a serious difficulty for which no generally satisfactory solution has yet been developed. The problem can be visualized if the array pattern is considered as the product of an element pattern and an array factor. The element pattern consists of the radiation pattern produced by a parabolic reflector, while the array factor is the pattern of an array of isotropic radiators which is a two-dimensional grating lobe pattern. The array factor can be steered electronically by shifting the phase between elements while the element pattern is directed by the mechanical movement of the individual dishes. In the ideal case, the element pattern and a single lobe of the array factor will both point in the desired direction. Multiple beams appear, however, when more than one grating lobe falls within the main beam of the element factor; this condition occurs when the array spacing is substantially greater than the diameter of the subapertures.

It can be easily shown that the spacing of the grating lobes from the main beam can be increased by a decrease in the separation of the parabolic reflector antenna elements. However, if this spacing is decreased, the diameter of the reflectors must also be decreased so that the effective scan range can be maintained, while at the same time more array elements must be added to meet the gain requirement. The end result will be a broader element pattern which in turn will ensure that the grating lobes will have essentially the same amplitude relative to the main beam. The beamwidth of both the main beam and the grating lobes will, for all practical purposes, remain the same as long as the overall array dimensions remain unaltered.

The fine grain structure around the various lobes will change, however, as more elements are added. Similarly, if the spacing between the elements is increased, and the diameter of the reflectors is increased correspondingly, the grating lobes will be moved in closer to the principal beam. Once again the relative amplitude and beamwidth of all the grating lobes should remain essentially constant.

There are some esoteric techniques available to suppress the size of the grating lobes. A possibility exists that the grating lobes adjacent to the principal beam may be reduced in amplitude by the use of random spacing among the array elements. However, it is anticipated that the selection of such a design will prove to be an extremely difficult problem. Another means of suppressing the grating lobes might involve the use of an auxiliary array that could be steered and phased to cancel out any given lobe. A major difficulty that might be anticipated from such a scheme would be the obtaining of sufficient gain from the auxiliary array.

The juxtaposition of spacing and reflector size discussed above is predicated on little or no interaction between the elements as a function of scan angle. When this interaction effect is taken into account an entirely different solution may be obtained for the competing parameters. Thus, it shall be the purpose of this section to study the problems associated with being able to analytically determine a spacing and antenna size which is optimum between the interference effects at minimum separation, and the grating lobe effects at a maximum distance commensurate with high aperture efficiency. Since the theory and manipulation of the array factor and element pattern is available elsewhere, the effort herein shall be concerned with methods and techniques for analyzing the interaction effects between large parabolic reflectors in a relatively closely spaced array.

An analysis of the blocking effect of a closely spaced array obtained by the consideration of the geometric optics only has been done in paragraph (3). First, the field in Fraunhofer region for an antenna system of two closely neighboring paraboloids has been formulated; then the field for a linear array of N-paraboloid is obtained. In these expressions, they show clear evidence of the interaction between neighboring paraboloids due to the close separation between them.

3) The Blocking Effect of a Closely Spaced Array

a) Consideration of the Coordinate Systems - The fixed coordinate system (x, y, z) with origin at point 0 will be used to define the observation point in space. The paraboloid coordinate system (x' , y' , z') with origins at the vertex of each paraboloid will be used to define the source points in space. The condition of the paraboloid coordinate system is specified in such a way that when the axis of the paraboloid (z'-axis) points in its zenith direction (in the direction of z-axis) the remaining x' and y' axes coincide with the fixed x and y axes respectively. That is, when paraboloid is at its zenith direction, the coordinates x', y' and z' coincide with the fixed coordinates x, y and z respectively. In order to define uniquely the pointing direction of the paraboloid in the direction (θ', ϕ') in the fixed coordinate, the axes of the paraboloid are being rotated as follows: first, x'-axis is rotated by an angle ϕ' in azimuth direction with z-axis as the axis of rotation. Hence, the angle between axes y' and y is ϕ'. Next, z'-axis is rotated by an angle θ' with y'-axis as the axis of rotation. Thus, the angle between axes z' and z is θ' and the angle between x-axis and the projection of x'-axis on the xy plane is ϕ'.

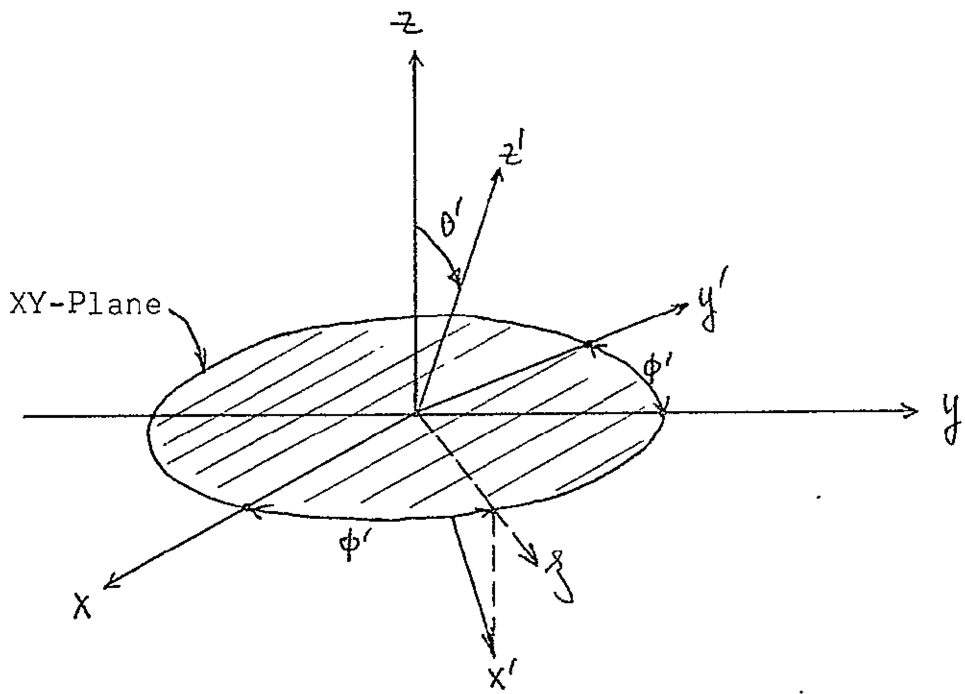
Let the direction of the projection of x'-axis on the xy plane be ξ then

$$\bar{a}_y, \bar{a}_{y'} = \phi'$$

$$\bar{a}_x, \bar{a}_\xi = \phi'$$

$$\bar{a}_z, \bar{a}_{z'} = \theta'$$

$$\bar{a}_{x'}, \bar{a}_\xi = \theta'$$



By these two rotations, the paraboloid coordinates have been uniquely defined in the fixed coordinate system. Hence,

$$
\begin{aligned}
\bar{a}_{x'} &= \bar{a}_x \cos\theta'\cos\phi' + \bar{a}_y \cos\theta'\sin\phi' + \bar{a}_z \left(-\sin\theta'\right) \\
\bar{a}_{y'} &= \bar{a}_x \left(-\sin\phi'\right) + \bar{a}_y \cos\phi' + \bar{a}_z \, 0 \qquad \text{(B-17)} \\
\bar{a}_{z'} &= \bar{a}_x \sin\theta'\cos\phi' + \bar{a}_y \sin\theta'\sin\phi' + \bar{a}_z \cos\theta'
\end{aligned}
$$

b) Fields in Fraunhofer Region for an Antenna System of Two Neighboring Paraboloids

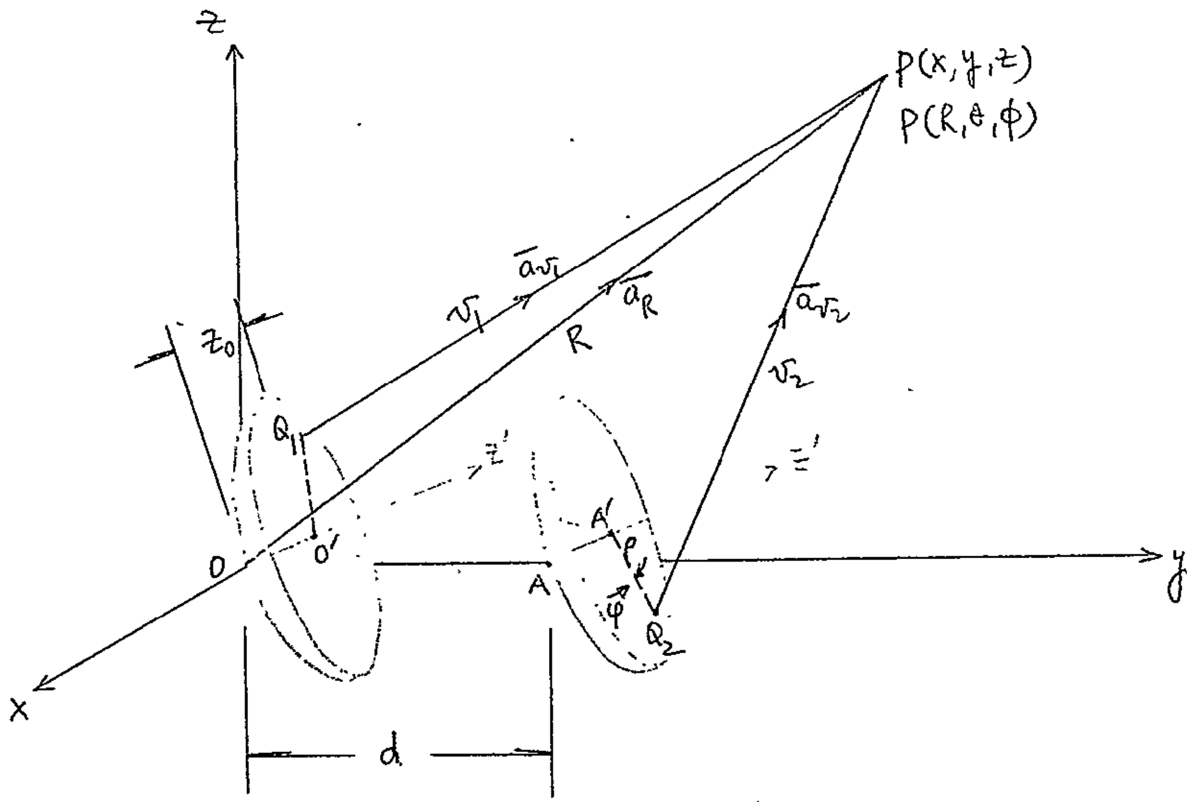


Fig. B-4

A Two Neighboring Paraboloidal Antenna System is shown in Fig. B-4, where $\bar{a}_{v_1}$, $\bar{a}_{v_2}$ and $\bar{a}_R$ are unit vectors in the direction of v_1, v_2 and R respectively. Both paraboloids point in the direction of z'-axis. Let the field distribution over the circular aperture be designated by

$$F(\rho , \psi) = A(\rho , \psi) e^{-j\Psi (\rho , \psi)} \tag{B-18}$$

with amplitude distribution $A(\rho , \psi)$ and phase distribution $\Psi(\rho , \psi)$; where ρ and ψ are the variables for the polar coordinates on the aperture.

For the far-zone region, the field due to a single aperture is given by

$$U_p = \frac{j}{2\lambda} \frac{e^{-jkR}}{R} (1+\gamma)\, e^{jkz_0\gamma} \cdot \int_{\psi=0}^{2\pi} \int_{\rho=0}^{a} F(\rho,\psi)\, e^{jk\rho[\alpha\cos\psi + \beta\sin\psi]} \rho\, d\rho\, d\psi \qquad \text{(B-19)}$$

where

$$\gamma = \sin\theta \sin\theta' \cos(\phi-\phi') + \cos\theta\cos\theta'$$
$$\alpha = \sin\theta\cos\theta'\cos(\phi-\phi') - \cos\theta\sin\theta' \qquad \text{(B-20)}$$
$$\beta = \sin\theta\sin(\phi-\phi')$$

For the configuration in Fig. B-4, the total field at observation point due to the identical aperture distribution $F(\rho, \psi)$ on apertures No. 1 and No.2 is

$$U_p = U_{p_1} + U_{p_2}$$
$$= \frac{j}{2\lambda} \frac{e^{-jkR}}{R} (1+\gamma)\, e^{jkz_0\gamma} \cdot \int_0^{2\pi} \int_0^{a} F(\rho,\psi)\, e^{jk\rho[\alpha\cos\psi + \beta\sin\psi]} \rho\, d\rho\, d\psi$$
$$+ \frac{j}{2\lambda} \frac{e^{-jkR}}{R} (1+\gamma)\, e^{jkz_0\gamma}\, e^{jkd\sin\theta\sin\phi} \cdot \int_0^{2\pi} \int_0^{a} F(\rho,\psi)\, e^{jk\rho[\alpha\cos\psi + \beta\sin\psi]} \rho\, d\rho\, d\psi \qquad \text{(B-21)}$$

Equation (B-21) is the total field at observation point without considering the blocking effect. In the case that the separation between the neighboring paraboloids is not large enough, the blocking effect due to the geometric optics obstacles has to be taken into account when the system scans away from its zenith direction. In the latter case, the first aperture of the paraboloid with vertex at origin is partially blocked by the presence of the second aperture of the second paraboloid with vertex at A in Fig. B-4. Considering the blocking effect due to the geometric optics obstacles, the field in far-zone region can be taken care of as follows: looking back along z'-axis toward the vertices, the apertures overlap due to the scanning away from its zenith direction as shown in Fig. B-5.

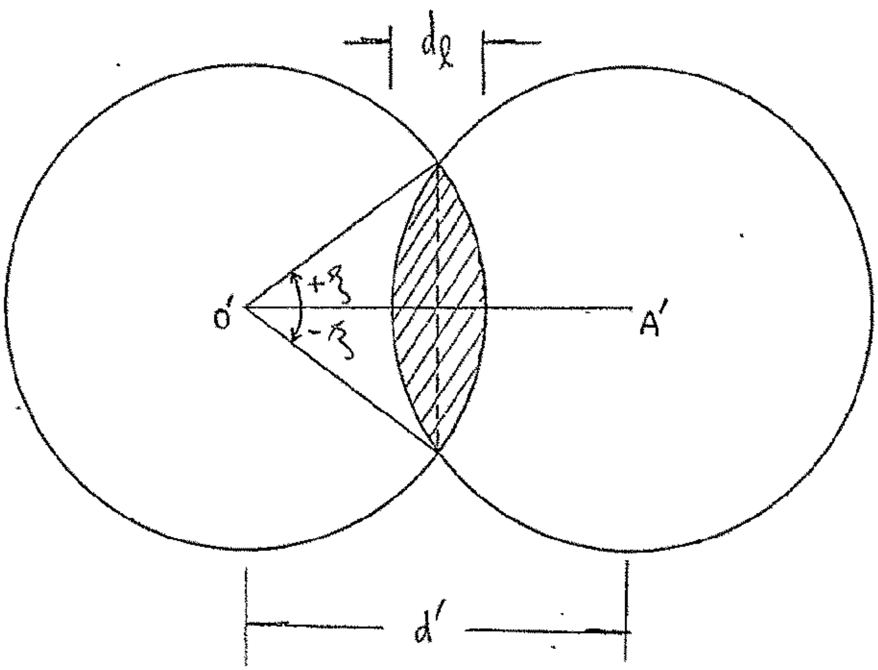


Fig. B-5

$$d' = d\sqrt{1-(\sin\theta'\sin\phi')^2}$$

$$d_\ell = d - d' = d\left[1-\sqrt{1-(\sin\theta'\sin\phi')^2}\right] \tag{B-22}$$

where d' is the distance between the axes, which is the projection of the separation d of the vertices of the paraboloids on the plane perpendicular to z'-axis, when the axes point at (θ', ϕ') direction and d_ℓ is the overlap distance along this projection. Hence, the overlap angle ξ is given by

$$\xi = \cos^{-1}\left(1 - \frac{d_\ell}{2a}\right) \tag{B-23}$$

with $\underline{a}$ being the radius of the aperture. The blocked area A_b which is the shaded area in Fig.B5 can be found as

$$A_b = 2\left[a^2\xi - \left(a - \frac{d_\ell}{2}\right) a \sin\xi\right] \tag{B-24}$$

Hence, the blocking effect can be taken care of by subtracting the part of contribution due to the blocked aperture; thus

$$U_{P1} = \frac{j}{2\lambda}\frac{e^{-jkR}}{R}(1+\gamma)\, e^{jkz_0\gamma} \cdot \left\{ \int_0^{2\pi}\int_0^{a} F(\rho,\varphi)\, e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]}\, \rho\, d\rho\, d\varphi - \int_{-\xi}^{+\xi}\int_{a-d_\ell}^{a} F(\rho,\varphi)\, e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]}\, \rho\, d\rho\, d\varphi \right\} \tag{B-25}$$

Therefore, the total field at observation point due to apertures No. 1 and No. 2 with blocking effect is

$$U_p = U_{p_1} \text{ (partially blocked)} + U_{p_2} \text{ (unblocked)}$$

$$= \frac{j}{2\lambda} \frac{e^{-jkR}}{R} (1+\gamma) e^{jkz_0\gamma} \cdot \left\{ \int_0^{2\pi} \int_0^{a} F(\rho,\varphi) e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]} \rho\, d\rho\, d\varphi - \int_{-\xi}^{+\xi} \int_{a-d_\ell}^{a} F(\rho,\varphi) e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]} \rho\, d\rho\, d\varphi \right\}$$

$$+ \frac{j}{2\lambda} \frac{e^{-jkR}}{R} (1+\gamma) e^{jkz_0\gamma} e^{jkd\sin\theta} \cdot \int_0^{2\pi} \int_0^{a} F(\rho,\varphi) e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]} \rho\, d\rho\, d\varphi \qquad \text{(B-26)}$$

c) Fields in Fraunhofer Region for a Linear Array of N-Paraboloid

The array is composed of N identical paraboloids and it is assumed that they point in the same direction simultaneously without delay.

The total field at observation point due to a linear array of N-aperture with the arrangement in Fig. B-6 will be the sum of the contribution of the first (N-1) partially blocked apertures and the last unblocked aperture, thus

$$U_p = \sum_{j=0}^{N-2} U_{p_j} \text{ (partially blocked) } + U_{p_{N-1}} \text{ (unblocked)}$$

$$= \frac{j}{2\lambda} \frac{e^{-jkR}}{R} (1+\gamma) e^{jkz_0\gamma} \left[\sum_{n=0}^{N-2} e^{jknd\sin\theta} \right]$$

$$\cdot \left\{ \int_0^{2\pi} \int_0^a F(\rho,\varphi) e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]} \rho \, d\rho \, d\varphi \right.$$

$$\left. - \int_{-\xi}^{+\xi} \int_{a-d_\rho}^{a} F(\rho,\varphi) e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]} \rho \, d\rho \, d\varphi \right\}$$

$$+ \frac{j}{2\lambda} \frac{e^{-jkR}}{R} (1+\gamma) e^{jkz_0\gamma} e^{jk(N-1)d\sin\theta}$$

$$\cdot \int_0^{2\pi} \int_0^a F(\rho,\varphi) e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]} \rho \, d\rho \, d\varphi \qquad \text{(B-27)}$$

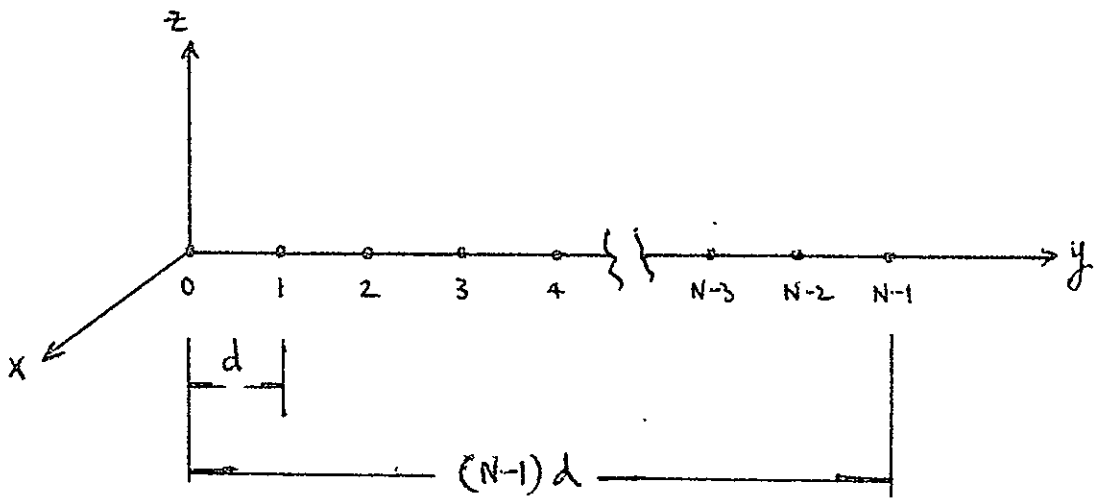


Fig. B-6

or

$$U_p = \frac{j}{2\lambda}\,\frac{e^{-jkR}}{R}(1+\gamma)\,e^{jkz_0\gamma}\left[\sum_{n=0}^{N-1} e^{jknd\sin\theta}\right]$$

$$\cdot\int_0^{2\pi}\int_0^{a} F(\rho,\varphi)\,e^{jk\rho[\alpha\cos\varphi+\beta\sin\varphi]}\,\rho\,d\rho\,d\varphi$$

$$-\frac{j}{2\lambda}\,\frac{e^{-jkR}}{R}(1+\gamma)\,e^{jkz_0\gamma}\left[\sum_{n=0}^{N-2} e^{jknd\sin\theta}\right]$$

$$\cdot\int_{-\xi}^{+\xi}\int_{a-d\rho}^{a} F(\rho,\varphi)\,e^{jk\rho[\alpha\cos\varphi+\beta\sin\varphi]}\,\rho\,d\rho\,d\varphi \qquad \text{(B-28)}$$

Let

$$I = \int_0^{2\pi}\int_0^{a} F(\rho,\varphi)\, e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]}\, \rho\, d\rho\, d\varphi \qquad \text{(B-28a)}$$

$$I_b = \int_{-\xi}^{+\xi}\int_{a-d_\rho}^{a} F(\rho,\varphi)\, e^{jk\rho[\alpha\cos\varphi + \beta\sin\varphi]}\, \rho\, d\rho\, d\varphi \qquad \text{(B-28b)}$$

Hence,

$$U_p = \frac{j}{2\lambda}\,\frac{e^{-jkR}}{R}\,(1+\gamma)\, e^{jkz_0\gamma}\, I \left[\sum_{n=0}^{N-1} e^{jknd\sin\theta}\right]$$

$$-\frac{j}{2\lambda}\,\frac{e^{-jkR}}{R}\,(1+\gamma)\, e^{jkz_0\gamma}\, I_b \left[\sum_{n=0}^{N-2} e^{jknd\sin\theta}\right] \qquad \text{(B-29)}$$

Where the factors I and $\sum_{m=0}^{N-1} e^{jkmd\sin\theta}$ are the element factor and the array factor respectively for a linear array of N-paraboloid without blocking effect; the factors I_b and $\sum_{m=0}^{N-2} e^{jkmd\sin\theta}$ are the element factor and the array factor respectively for taking into account the blocking effect; where α, β, γ, are given in Eq. (B-20).

d) <u>Consideration of a Simple Case</u>

In order to observe the pattern of the system in Fraunhofer region, first we consider a simple case in which the array lies along the y-axis and the scanning will perform in the right corner sector of the yz-plane. For this given condition, ϕ is $\pi/2$, and ϕ' is $\pi/2$. Thus from Eq. (B-20)

$$\gamma = \cos(\theta - \theta')$$
$$\alpha = \sin(\theta - \theta')$$
$$\beta = 0$$

Hence from Eqs. (B-28a) and (B-28b) we have

$$I = \int_0^{2\pi}\int_0^{a} F(\rho,\psi)\, e^{jk\rho \sin(\theta-\theta')\cos\psi}\, \rho\, d\rho\, d\psi \tag{B-30}$$

$$I_b = \int_{-\xi}^{+\xi}\int_{a-d_\ell}^{a} F(\rho,\psi)\, e^{jk\rho \sin(\theta-\theta')\cos\psi}\, \rho\, d\rho\, d\psi \tag{B-31}$$

For the aperture distribution $F(\rho, \psi)$, it is assumed that the feeds are normally designed to illuminate the paraboloidal reflectors with an intensity at the reflector edges that is approximately 10 db below that at center. Thus

$$F(\rho, \psi) = 1 - (1-\delta)\frac{\rho^2}{a^2} \quad \text{for } \rho \leq a \tag{B-32}$$

For this 10db tapered illumination, the value of δ has to be equal to 0.1.

To obtain the desired aperture distribution, in the present case, the 10db tapped aperture illumination, is itself an attractive problem namely aperture synthesis. For the purpose of analyzing the blocking effect of the closely spaced linear array of N-dish it is assumed that the desired aperture distribution has been achieved without worrying about the actual technique to obtain it. The effect of tapering the illumination down toward the edge is: reduction in gain, increasing beamwidth, and reduction in side lobes as compared with the uniform aperture distribution, and reduction of the energy spilled over the edge.

To perform the integrations in Eq. (B-30) and (B-31), a change of variables is done as follows: Let

$$r = \frac{\rho}{a} \qquad \text{(B-33)}$$

$$u = ka \sin(\theta - \theta')$$

Then the aperture distribution becomes

$$F(r, \psi) = 1 - (1 - \delta) r^2 \qquad \text{(B-34)}$$

with $\delta = 0.1$ for 10 db tapered illumination and the factor

$$e^{jk\rho \sin(\theta - \theta') \cos\psi} = e^{jur \cos\psi} \qquad \text{(B-35)}$$

Thus, Eq. (B-30) and (B-31) become

$$I_b = a^2 \int_{-\xi}^{+\xi} \int_{1-\frac{d_\ell}{a}}^{a} r e^{jur\cos\psi} dr d\psi - a^2(1-\delta) \int_{-\xi}^{+\xi} \int_{1-\frac{d_\ell}{a}}^{a} r^3 e^{jur\cos\psi} dr d\psi$$

$$= a^2 \left\{ \frac{2\xi}{u^2} \Big[z J_1(z) \Big]_{z_1}^{z_2} - (1-\delta) \frac{2\xi}{u^4} \Big[z^3 J_1(z) - 2z^2 J_2(z) \Big]_{z_1}^{z_2} \right\}$$

$$+ a^2 \sum_{n=1}^{\infty} \frac{(j)^n 4 \sin(z\xi)}{n} \left\{ \frac{1}{u^2} \int_{z_1}^{z_2} z J_n(z) dz - \frac{(1-\delta)}{u^4} \int_{z_1}^{z_2} z^3 J_n(z) dz \right\} \qquad \text{(B-36)}$$

$$I = a^2 \int_0^{2\pi}\int_0^1 e^{jur\cos\varphi}\, r\,dr\,d\varphi - a^2(1-\delta)\int_0^{2\pi}\int_0^1 r^2 e^{jur\cos\varphi}\, r\,dr\,d\varphi$$

$$= 2\pi a^2 \delta\, \frac{J_1(u)}{u} + 4\pi a^2 (1-\delta)\, \frac{J_2(u)}{u} \qquad \text{(B-37)}$$

When integration is being performed, I_b will be a complex number, hence I_b may be denoted by its real part I_{br} and imaginary part I_{bi}; thus

$$I_b = I_{br} + j\, I_{bi} \qquad \text{(B-38)}$$

With

$$I_{br} = a^2 \left\{ \frac{2\xi}{u^2}\Big[z J_1(z) \Big]_{z_1}^{z_2} - (1-\delta)\frac{2\xi}{u^4}\Big[z^3 J_1(z) - 2z^2 J_2(z) \Big]_{z_1}^{z_2} \right.$$

$$+ \sum_{m=1}^{\infty} \frac{(-1)^m 4 \sin[2m\xi]}{2m} \left[\frac{1}{u^2}\int_{z_1}^{z_2} z J_{2m}(z)\,dz \right.$$

$$\left.\left. - \frac{(1-\delta)}{u^4}\int_{z_1}^{z_2} z^3 J_{2m}(z)\,dz \right] \right\} \qquad \text{(B-38a)}$$

$$I_{bi} = a^2 \sum_{m=1}^{\infty} \frac{(-1)^m 4 \sin[(2m+1)\xi]}{(2m+1)} \left[\frac{1}{u^2} \int_{z_1}^{z_2} z J_{2m+1}(z)\,dz - \frac{(1-\delta)}{u^4} \int_{z_1}^{z_2} z^3 J_{2m+1}(z)\,dz \right] \tag{B-38b}$$

where

$u = ka \sin(\theta - \theta')$

$\xi = \cos^{-1}(1 - \frac{d_\ell}{a})$

$d_\ell = d\,(1 - \cos\theta')$

$z = ur$

$z_1 = u(1 - \frac{d_\ell}{a})$

$z_2 = u$

$k = \frac{2\pi}{\lambda} = \frac{w}{c}$

a = The radius of the circular aperture

d = The separation between the adjacent paraboloids

and the array factors

$$\sum_{n=0}^{N-1} e^{jknd\sin\theta} = \frac{\sin\frac{kNd\sin\theta}{2}}{\sin\frac{kd\sin\theta}{2}} e^{j\frac{k(N-1)d\sin\theta}{2}}$$

$$\sum_{n=0}^{N-2} e^{jknd\sin\theta} = \frac{\sin\frac{k(N-1)d\sin\theta}{2}}{\sin\frac{kd\sin\theta}{2}} e^{j\frac{k(N-2)d\sin\theta}{2}}$$

Let

$$F_1(\theta) \doteq \frac{\sin \frac{kNd\sin\theta}{2}}{\sin \frac{kd\sin\theta}{2}} \tag{B-40a}$$

$$F_2(\theta) \doteq \frac{\sin \frac{k(N-1)\sin\theta}{2}}{\sin \frac{kd\sin\theta}{2}} \tag{B-40b}$$

Therefore, the total field U_p at observation point in Eq (B-29) becomes

$$U_p = \frac{j}{2\lambda}\frac{e^{-jkR}}{R}(1+\gamma)\,e^{jkz_0\gamma} \cdot \left\{ F_1(\theta)\, I\, e^{j\frac{k(N-1)d\sin\theta}{2}} - F_2(\theta)\, I_b\, e^{j\frac{k(N-2)d\sin\theta}{2}} \right\} \tag{B-41}$$

The angular distribution $g(\theta,\phi)$ of U_p is

$$g(\theta,\phi)=(1+\gamma)\left\{\left[F_1(\theta)I-F_2(\theta)\left(I_{br}\cos\frac{kd\sin\theta}{2}+I_{bi}\sin\frac{kd\sin\theta}{2}\right)\right]^2+\left[F_2(\theta)\left(I_{br}\sin\frac{kd\sin\theta}{2}-I_{bi}\cos\frac{kd\sin\theta}{2}\right)\right]^2\right\}^{1/2}$$

$$\cdot\, e^{jkz_0\gamma+j\frac{k(N-1)d\sin\theta}{2}}$$

$$\cdot\, e^{j\tan^{-1}\frac{F_2(\theta)\left(I_{br}\sin\frac{kd\sin\theta}{2}-I_{bi}\cos\frac{kd\sin\theta}{2}\right)}{F_1(\theta)I-F_2(\theta)\left(I_{br}\cos\frac{kd\sin\theta}{2}+I_{bi}\sin\frac{kd\sin\theta}{2}\right)}} \qquad \text{(B-42)}$$

Let the amplitude and phase distributions of $g(\theta,\phi)$ be denoted by $A(\theta,\phi)$ and $\Psi(\theta,\phi)$ respectively, then

$$g(\theta,\phi)=A(\theta,\phi)\,e^{j\Psi(\theta,\phi)} \qquad \text{(B-43)}$$

with

$$A(\theta,\phi)=(1+\gamma)\left\{\left[F_1(\theta)I-F_2(\theta)\left(I_{br}\cos\frac{kd\sin\theta}{2}+I_{bi}\sin\frac{kd\sin\theta}{2}\right)\right]^2+\left[F_2(\theta)\left(I_{br}\sin\frac{kd\sin\theta}{2}-I_{bi}\cos\frac{kd\sin\theta}{2}\right)\right]^2\right\}^{1/2}$$

(B43a)

$$\Psi(\theta,\phi)=kz_0\gamma+\frac{k(N-1)d\sin\theta}{2}+\tan^{-1}\frac{F_2(\theta)\left(I_{br}\sin\frac{kd\sin\theta}{2}-I_{bi}\cos\frac{kd\sin\theta}{2}\right)}{F_1(\theta)I-F_2(\theta)\left(I_{br}\cos\frac{kd\sin\theta}{2}+I_{bi}\sin\frac{kd\sin\theta}{2}\right)}$$

(B43b)

where $F_1(\theta)$ and $F_2(\theta)$ are defined in (B-40a) and B-40b) respectively; I_{br} and I_{bi} are defined in (B-38a) and (B-38b) respectively and I is defined in (B-37).

A computer program has been developed to calculate the pattern of this simple case and is currently being tested. Up to the present time, no numerical result has been obtained. The result will be included in the next report.

C. A PHASED ARRAY OF SMALL CLOSELY SPACED ELEMENTS ORGANIZED INTO SUBAPERTURES

1) Introduction

Although the present state-of-the-art in extremely large phased arrays, especially at S-band, is behind that for large dishes, there is no fundamental reason that limits the size of an array except the questions of signal to noise ratio, availability of low loss transmission line, and the basic cost of the individual components. At present these questions concerned with the fundamentals of organization versus economics is one of the problems to which this program has been addressed during its entirety. There will be more discussion of this point at a later date after some of the results obtained in the section can be analyzed and compared with the corresponding results from the other types of antenna systems. These problems coupled with the practical problems of distribution and feeding techniques, element type, and scanning techniques require some special consideration when the array is divided into an appropriate number of subapertures. It is the purpose of this report to delineate some of the studies and to present the information that has been uncovered in the area of phase array technology which must be advanced to make such an array feasible for the DSCS program. An additional purpose is to relate the problems areas of various phased array techniques and to establish avenues for the solution in each of the problem areas to have the highest probability of success.

An important consideration in the design of such a large array is how the system should be organized; i.e., how the individual elements should be combined, phase shifted and detected to obtain the required specifications at the minimum cost. In order to quantitatively study this problem and obtain some numerical results, a dense array of dipoles over a ground plane was chosen as a receiving antenna model; this choice of a model was made partly because it could be analyzed rather easily and partly because it represents a practical high gain element which could be economically mass produced by depositing or photoetching techniques. All the calculations reported here were made for uniform distribution broadside condition (equal amplitude and constant phase) and linear polarization. Phase shifters were included in the models, however, so that the results could validly be extended to the beam steering mode of operation and used for problems in adaptive systems.

It was assumed that for large arrays or subarrays with fixed interelement spacing the effective collecting aperture is proportional to the number of elements and, in fact, is equal to the physical array size. This assumption is verified in Appendix I. Thus an interelement spacing was fixed at $\lambda/2$ (center to center) in both directions and elevated $\lambda/4$ over a ground plane; this choice was made because it represents a model commonly used in practice, and because it avoids any spurious or grating lobes.

In order to make some quantitative evaluation of the merits of the different organization schemes some numerical values were established for the communication link. These are

Frequency	2.3 GHz
Transmitted power	50 watts
Transmitter antenna gain (30' parabolic dish with 55% aper. eff.)	44 dB
Data rate	10^6 bits/sec
Maximum bit error probability	10^{-5}
Modulation	Biphase modulation 70°

During this six-month period, an economic analysis was developed for the array of dipoles. A computer program was written that calculates the required system cost as a function component characteristics cost using the component cost subarray size parameters. Several choices and values for each component can be analyzed simultaneously; the program determines how to construct the array with the minimum total cost and also tabulates the cost and size of the remaining possible system configurations. Of course, the results are highly dependent upon component characteristics and costs which require frequent review and update. However, the technique for this economic analysis can be easily applied at any time to new date points since the computer program is listed in its entirety in Appendix II.

3) Predection vs. Postdetection Combining

There are two basic ways in which the detection process can be performed. The first, as shown in Fig. C-2, consists of summing the properly adjusted IF ouputs from each subarray and then detecting the resultant to obtain a series of ones and zeros at the modulation rate. The second scheme, as shown in Fig. C-3, consists of detecting the output of each subarray at the IF level and then using a majority count to make the final decision as to whether a one or a zero occurred. The first, the coherent addition scheme, will obviously be more efficient than the second, but the latter system has several advantages which merit closer consideration; for example, the time delay can be a digital device such as a shift register. The summation is also done digitally at the base band frequency rate, rather than at IF.

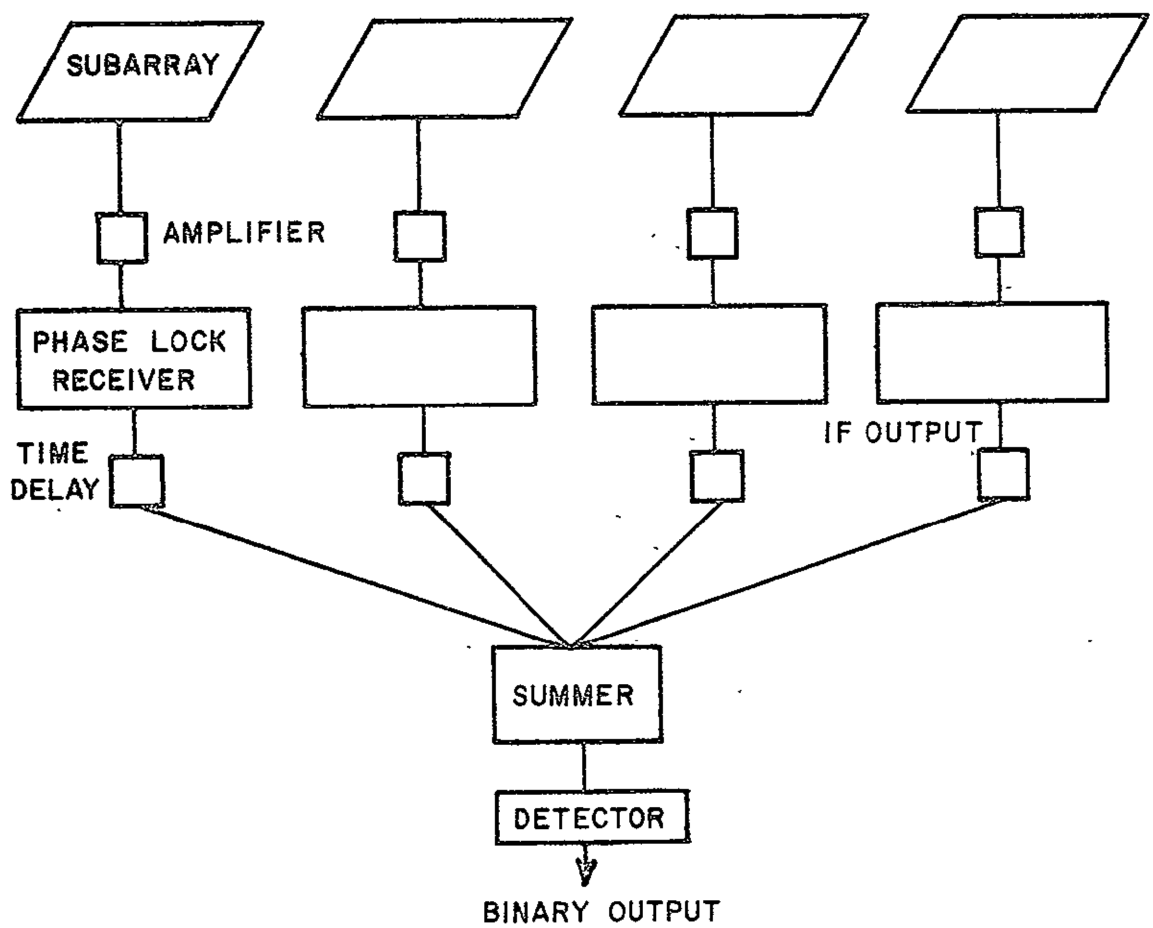


Fig. C-2 Predetection combining program

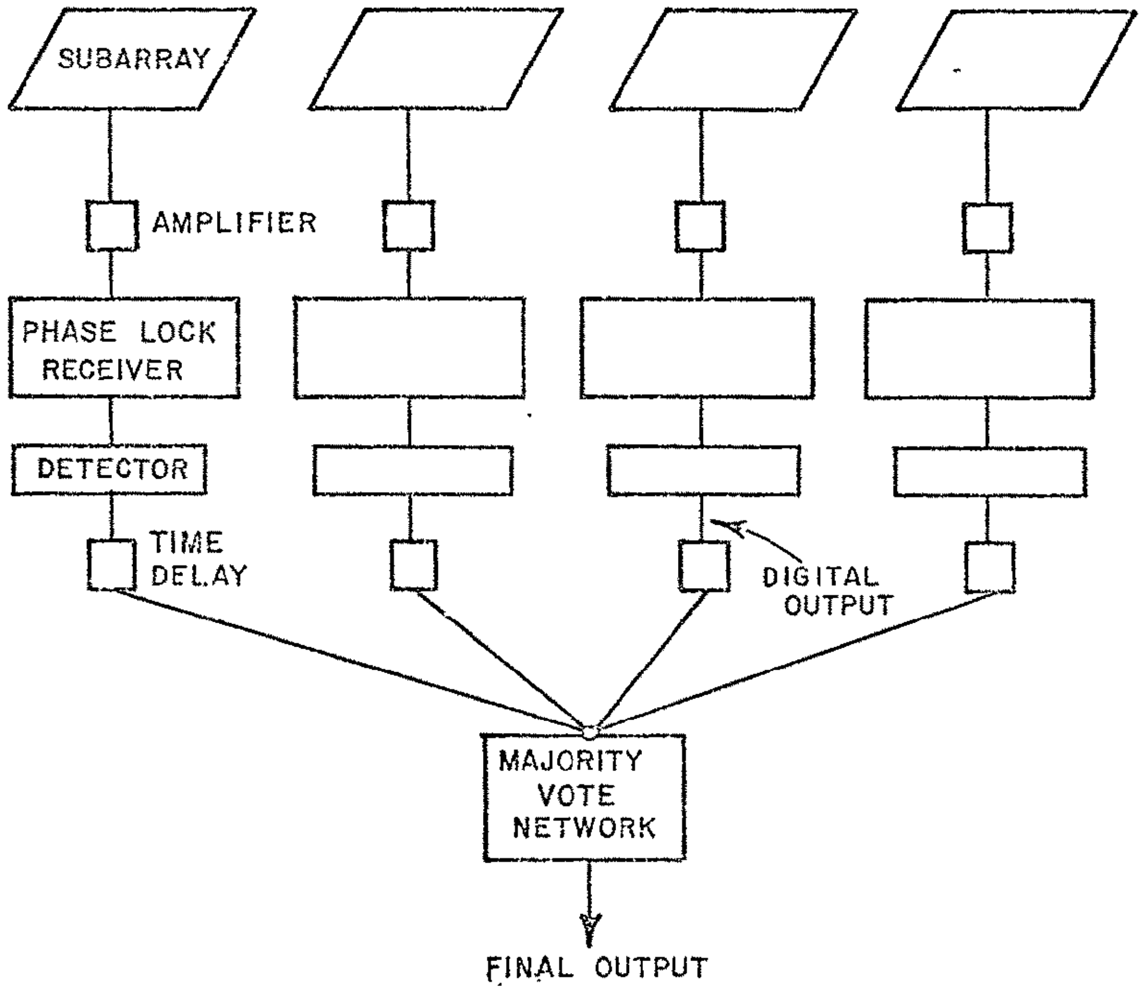


Fig. C-3 Postdetection combining diagram.

An analysis has been done on these two schemes (C-5) which showed that for the limiting case where the SNR of each subarray is very small, but the SNR of the combined subarrays is large, the postdetection summing requires a total SNR $\pi/2$ (2dB) greater than the predetection combining in order to produce the same bit error probability. Since this establishes the relationship between the two processes the remainder of this report will be concerned with coherent predetection combining system.

4) Array-Subarray Organization

The subarray model consists of dipole elements which are phase shifted and combined to form a single output at the RF frequency. Due to the relatively large beamwidth of a single subarray, it is expected that the proper phase adjustment can be performed with a special purpose computer using a priori knowledge of the source location.

The number of elements required to achieve the specified 10 dB SNR will, in general, be a function of the phase shifter loss and temperature, feed line losses, amplifier noise temperature, and subarray size.

a) Maximum Subarray Size If phase control is used for combining, rather than time delay compensation, the total time delay across the subarray must be less than the modulation period in order that each element simultaneously receives the same information bit. For an information rate of 10^6 bits per second this time delay must be much less than 1 μ sec, which limits the maximum subarray size to about 30 meters (1μ sec has spatial length of 300 meters) if the system is required to operate at low elevation angles. This does not represent a stringent limitation; for the antenna model considered here a subarray of this size would contain about 200,000 elements.

b) Minimum Subarray Size For any adaptive scheme each subarray must produce a SNR which is sufficient to lock on the signal during the acquisition mode and maintain lock during the information transfer mode. For a typical phase lock system using coherent addition the following equations can be used to obtain a comparison between different organizational parameters:

$$SNR_{TOT} = N \quad SNR_{SA} = 10 \tag{C-8}$$

$$CNR_{PLL} = K \quad SNR_{SA} \frac{B_{IF}}{B_{PLL}} \tag{C-9}$$

where

SNR_{TOT} = total numeric signal-to-noise power ratio taken to be 10 in order to produce a bit error probability of 10^{-5}.

N = number of subarrays

CNR_{PLL} = carrier to noise ratio in the phase lock loop of each subarray receiver

K = fraction of power transmitted at the carrier frequency

B_{IF} = bandwidth of the IF, taken to be 0.5×10^6 Hz to receive 10^6 bits/sec using matched integrate and dump detection

B_{PLL} = bandwidth of phase lock loop, taken to be 10 Hz.

During the acquisition time all the power can be transmitted at the carrier frequency (K = 1) so that

$$CNR_{PLL} = SNR_{SA} \quad \frac{B_{IF}}{B_{PLL}} = \frac{10^6}{2N} \tag{C-10}$$

From experience (9),(10) it has been shown that about 6-7 dB CNR_{PLL} is required for acquisition; using this criteria and the above constants yields the minimum SNR_{SA} = -40 dB to obtain lock. However, during normal operation of this subarray, when most of the power is contained in the modulation components, the CNR_{PLL} would drop to -3 dB which is not sufficient to maintain phase lock. Hence the actual minimum SNR_{SA} is not set by the acquisition requirement but rather by having to maintain lock during the signaling. Requiring a 3 dB SNR_{SA} during normal operation constrains the minimum SNR_{SA} to be -34 dB.

c) <u>Feeding Techniques</u> Two types of feed systems were considered; the commonly used modified series-series shown in Fig. C-4 and the equal length corporate feed shown in Fig. C-5. The effective noise temperature and SNR at the subarray output are now calculated for both feeding systems.

<u>Series-series model</u> -. Consider an arbitrary unit of power delivered to this subarray (Fig.C-4). The fraction of power delivered to the phase shifters is

$$\Gamma = \frac{\alpha}{N^2} \left(\sum_{n=1}^{N} \alpha^{n-1} \right)^2 \tag{C-11}$$

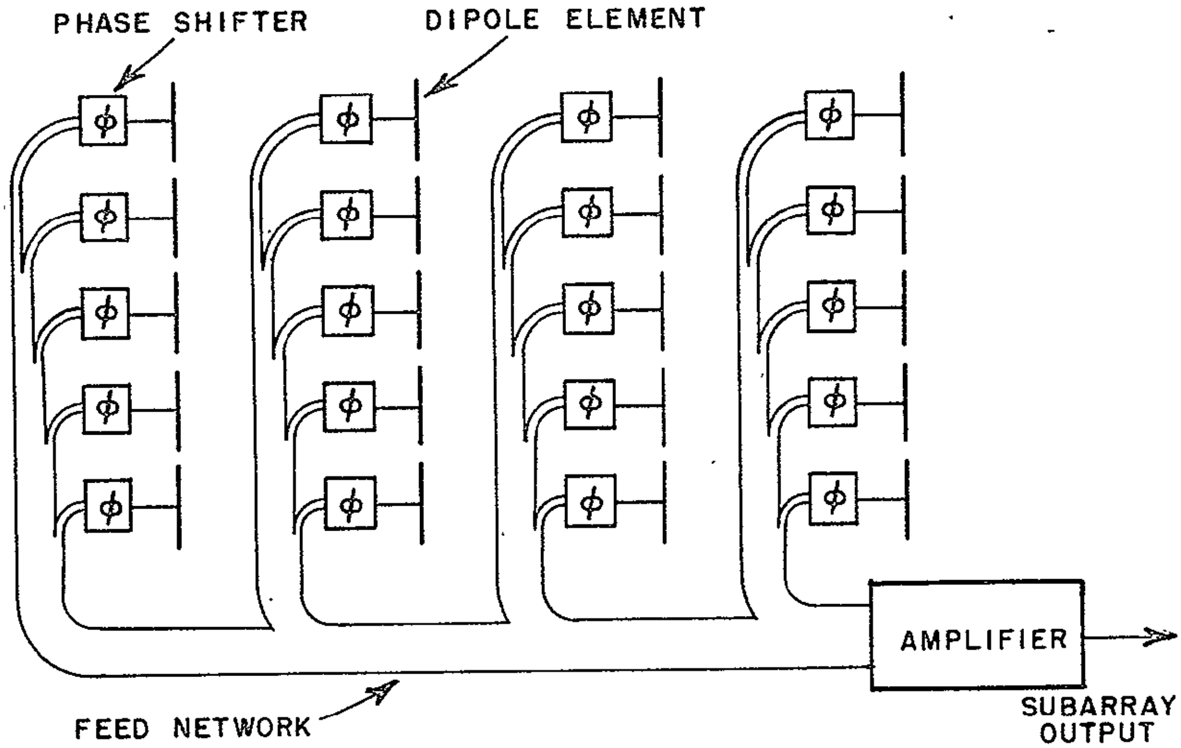


Fig. C-4 Series-series feed system.

where

N^2 = number of elements in the subarray

α = transmission coefficient for a $\lambda/2$ section of the feed line.

Hence the fraction of power absorbed by the feed system is $1 - \Gamma$. The fraction absorbed by the phase shifters is $(1 - \alpha_\phi)\,\Gamma$, where α_ϕ is the transmission coefficient of the phase shifters, and the fraction of power is delivered to the dipole antennas is $\alpha_\phi\,\Gamma$.

Finally, the expression for the total effective noise temperature of the subarray is

$$T_{eff} = \left[1 - \Gamma\right] T_o + T_\phi \left[1 - \alpha_\phi\right] \Gamma + T_a\,\alpha_\phi\,\Gamma + T_{amp} \qquad \text{(C-12)}$$

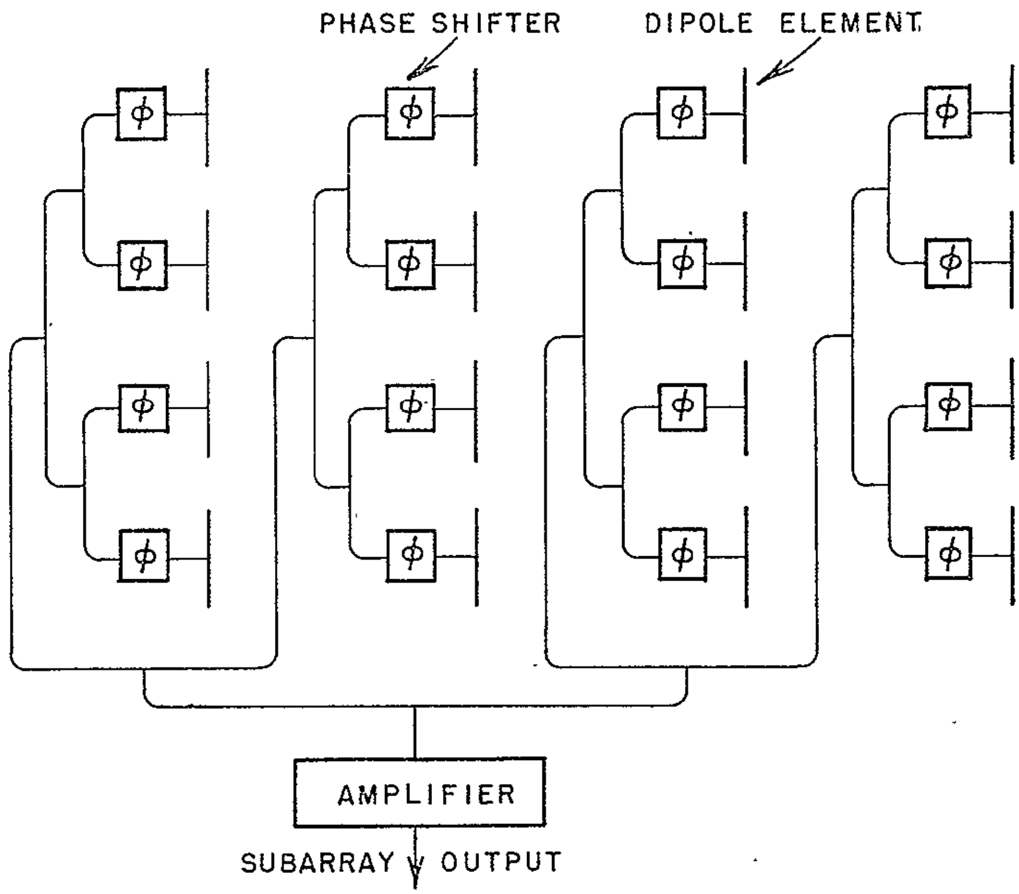


Figure c-5 Equal length corporate feed system

where

T_o = physical temperature of the feed structure assumed constant at 290°K

T_ϕ = physical temperature of the phase shifters

T_a = antenna temperature = 9°K for the dipole model

T_{amp} = effective amplifier noise temperature.

For a transmitted power of 50 watts and a thirty-foot transmit antenna two Au from the array the resulting expression for the subarray SNR in dB is

$$SNR_{SA} = -144 - PSLDB + 10 \log_{10} N^2 - 10 \log_{10} \left(\frac{1}{\Gamma}\right) \quad \text{(C-13)}$$
$$-10 \log_{10} (k \; T_{eff} \; B)$$

where

PSLDB = phase shifter loss in db

k = Boltzmann constant

B = bandwidth = 0.5×10^6 Hz

Equal length corporate model A similar analysis yields the effective temperature for this subarray model :

$$T_{eff} = LT_\phi + L_\phi \left[T_a - T_\phi\right] + \left[1-L\right] T_o + T_{amp} \tag{C-14}$$

where

$$L = \exp\left[-2.3 LDB/10\right]$$

$$LDB = LPF\,(\lambda/2)\left[2\log_2 N + \sum_{i=0}^{\log_2(N-1)} 2^i\right]$$

LPF = Attenuation of the feed in db per foot

The resulting SNR in db at the subarray is:

$$SNR = -144 - LDB - PSLDB + 10\log_{10} N^2 - 10\log_{10}\left[k\,T_{eff} B\right] \tag{C-15}$$

As shown in the numerical results the equal length system is slightly less efficient than the series-series system; it has the advantage of not requiring any phase shifting devices if the subarray panels are to be mechanically pointed.

5) Circuit Components

The optimum antenna system for the ground terminal is one that maximizes the signal-to-noise ratio under the practical constraints imposed by tolerance, reliability, noise environment, and cost. The antenna must have a low equivalent noise temperature and must provide a high-gain pattern which is steerable through a wide angle ($\pm$ 60°). It will be the purpose of this section to consider the circuit components and techniques appropriate to the design of a large phased array and to delineate their characteristics as parameters in determining sub-aperture size and performance characteristics. A phased array consists of radiating elements, a power distribution or collection network, a beam-steering or phasing system, and an optimal number of low noise preamplifiers. Each of these antenna components plays an important and interdependent role in the determination of the overall antenna performance. There exist a variety of beamsteering techniques applicable to a large antenna of phased array type; these include the use of a phase shifter at each element, and the use of a mixing scheme that translates a phase shift from the operating frequency to a convenient frequency band. Those areas in phased-array distribution and component technology that must be advanced to make the large arrays practical are to be discussed and delineated in this section. In addition, some consideration is being given to other types of scanning techniques in an effort to provide an optimum response to communication signals under a wide variety of environmental conditions.

a) Distribution Networks The distribution network collects the signal from each of the radiating elements and phase shifters of the array and brings them to a common receiving port so that they combine in phase with a minimum of loss or distributes the energy to the individual radiating elements from the signal generator with proper phases and minimum loss in order to obtain a desirable radiation pattern. The distribution network largely and sometimes wholly determines the antenna aperture distribution; hence, it determines the antenna pattern, sidelobe level, and directivity. In the present study where the applicability of any particular overall system technique is determined by the various loss factors discussed above, the nature of the distribution network is most critical since it can shift the balance of effectiveness from one type of ground based system to another;

a few tenths of db/100' of loss in a transmission line can change the desirability of a particular technique since there are many hundreds of feet involved in the overall signal distribution. Distribution systems to be considered herein will include those which are essentially optical and the several types of transmission lines as shown in Table C-II (See Ref. C-11). The various types of distribution networks to be evaluated in this phase scanned system can also apply to multiple-beam system where low-noise is an essential feature. At this stage in this study, it is already obvious that performance figure of merit of a large phased array will be largely determined by the characteristics of the distribution network and that further study and development beyond the present state-of-art in low loss transmission lines will be needed to satisfy the requirements of this program.

There are several distribution networks for feeding a phase array. The basic principles of each is briefly described as follows:

Constrained Series. Figure C-6 shows several types of series feeds. In all cases the path length to each radiating element has to be computed as a function of frequency and taken into account when setting the phase shifters. The series feed lends itself to simple assembly techniques. Figure C-6a is an end-fed array. It is frequency sensitive and leads to more severe bandwidth restrictions than most other feeds. Figure C-6b is center fed and has essentially the same bandwidth as a parallel feed network (Ref. C-14). Sum and difference pattern outputs are available, but they have contradictory requirements for optimum amplitude distribution that cannot both be satisfied. As a result, either good sum or good difference patterns can be obtained, but no reasonable compromise seems possible that gives good sum and difference patterns simultaneously. At the cost of some additional complexity the difficulty can be overcome by the method shown under Fig. C-6c. Two separate center-fed feed lines are used and combined in a network to give sum and difference pattern outputs (Ref. C-15). Independent control of the two amplitude distributions is possible. For efficient operation the two feed lines require distributions that are orthogonal within each branch of the array, that is, in each branch the two feed lines give rise to patterns where the peak value of one coincides with a null from the other and the aperture distributions are respectively even and odd.

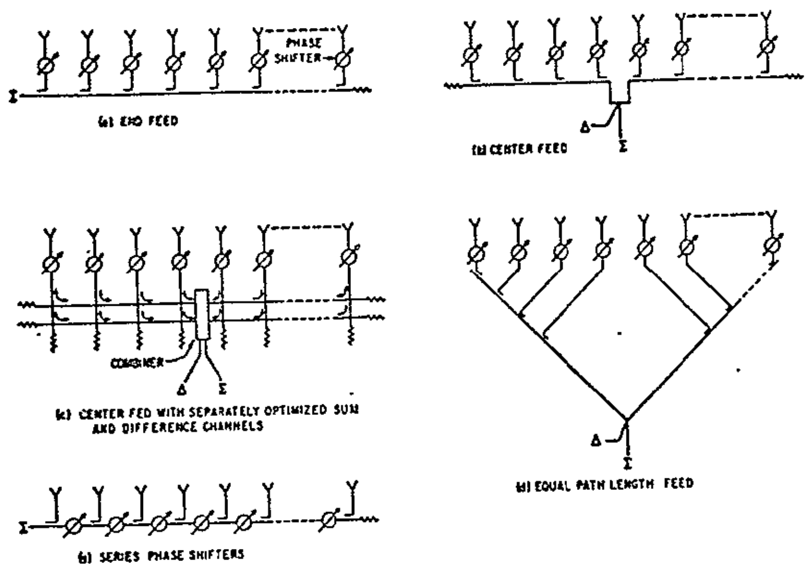


Fig. C-6 Series Feed Networks

A very wide band series feed with equal path lengths is shown in Fig. C-6d. If the bandwidth is already restricted by the phase shifters at the aperture, very little advantage is obtained at the cost of a considerable increase in size and weight. The network of Fig. C-6e permits simple programming since each phase shifter requires the same setting. The insertion loss increases for successive radiators and the tolerances required for setting the phases are high. A modified series phase shifters technique, series-series feed system, has been investigated in Sec. 3c for feeding an array of subarrays. The signal to noise ratio (SNR) of individual subarray in terms of number of elements in the subarray, phase shifter loss, the fraction of power delivered to the phase shifters and the total effective noise temperature of the subarray has been obtained in Equation C-13. Curves of subarray SNR versus phase shifter temperature, which were computed from Equation C-13, for a 100-element subarray for each range, 1AU and 2AU are shown in Fig. C-10.

Parallel Feeds. Figure C-7 shows a number of different parallel feed systems. They would usually combine a number of radiators into subarrays and the subarrays would then be combined to form sum and difference patterns.

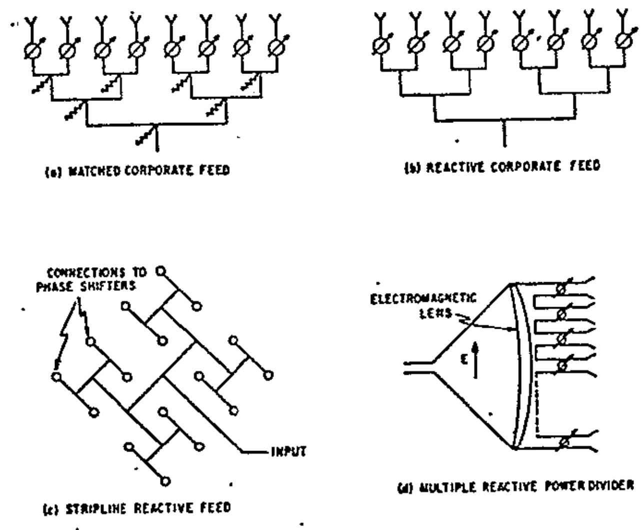


Fig. C-7 Parallel Feed Networks

Figure C-7a shows a matched corporate feed which is assembled from matched hybrids. The out-of-phase components of mismatch reflections from the aperture and of other unbalanced reflections are absorbed in the terminations. The in-phase and balanced components are returned to the input, and no power reflected from the aperture is re-radiated. To break up periodicity and reduce peak quantization lobes (Ref. C-14), small additional phase shifts may be introduced in the individual lines and compensated by corresponding readjustments of the phase shifters. An equal length corporate feed system has also been investigated in Sec. 3c for feeding an array of subarrays. The signal to noise ratio of individual subarray has been obtained in Equation C-15.

With nonreciprocal phase shifters the two-way path length is a constant, independent of the phase shifter setting. Under these conditions the performance of a reactive corporate feed is similar to that of the matched corporate feed. However, if additional phase shifts are added to the individual arms or if reciprocal phase shifters are used, then the out-of-phase components of the reflections due to the aperture mismatch will be reradiated (Ref. C-14). Figure C-7b shows a schematic layout for a reactive power divider which may use waveguide. A stripline power divider is shown under Fig. C-7c. A constrained-optical power divider using an electromagnetic lens is shown under Fig. C-7d. The lens may be omitted and the correction applied at the phase shifters. With nonreciprocal phase shifting, a fraction of the power reflected from the aperture will then be re-radiated rather than returned to the input. The amplitude distribution across the horn is given by the wave-guide mode. It is constant with an E-plane horn as shown.

In this section of the report, transmission line feeding systems have been considered which to date are deemed appropriate for large phased arrays. From a manufacturing viewpoint strip-line is by far the most desirable type of transmission line because it is readily adaptable to mass producing techniques. However, its extremely high loss relative to coax and waveguide is due to dielectric losses rather than ohmic conductor loss. One of the most useful low loss high frequency dielectrics is Teflon (polytetrafluoroethylene). Because pure Teflon has such a poor coefficient of thermal expansion it is usually mixed with glass or quartz; it is this additive which seriously degrades its attenuation properties. It is expected that considerable improvements will be made in dielectric materials and will make stripline devices more desirable.

TABLE C-II

	Attenuation in db/100' at 2 GHz
Brass Waveguide	0.6
Rigid and Semi-rigid coax	1-2.5
Flexible coax - RG 20	6
Flexible coax - RG 9	12
Flexible coax - RG 58	35
Microstrip	19
Stripline (Triplate)	18

All subsequent calculations will be made using nominal values of feed line loss ranging from a lossless line to that of coax.

Optical Feed Systems. Phased array apertures may be used in the form of lenses or reflectors, as shown in Fig. C-8, where an optical feed system provides the proper aperture illumination. The lens has input and output radiators coupled by phase shifters. Both surfaces of the lens require matching. The primary feed can be optimized to give an efficient aperture illumination with little spillover (1 to 2 db), for both sum and difference patterns. If desired the transmitter feed can be separated from the receiver by an angle α, as shown. The antenna is then rephased between transmitting and receiving so that in both cases the beam points in the same direction. The phasing of the antenna has to include a correction for the spherical phase front. To the first approximation this correction is

$$\frac{2\pi}{\lambda}\left[\sqrt{f^2+r^2}-f\right]=\frac{\pi}{\lambda}\frac{r^2}{f}\left[1-\frac{1}{4}\left(\frac{r}{f}\right)^2+\cdots\right]$$

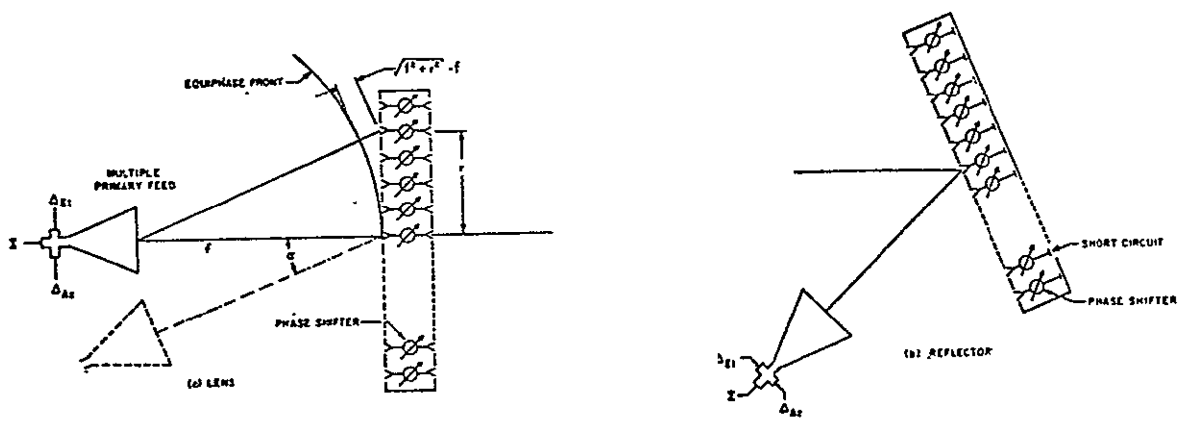


Fig. C-8 Optical Feed Systems

With a sufficiently large focal length, the spherical phase front may be approximated by that of two crossed cylinders, permitting the correction to be applied simply with row and column steering commands. Correction of the spherical phase error with the phase shifter reduces peak phase quantization lobes (Ref. C-14). Space problems may be encountered in assembling an actual system, especially at higher frequencies, since all control circuits have to be brought out at the side of the aperture.

Multiple beams may be generated by adding further primary feeds. All the beams will be scanned simultaneously by equal amounts in sin θ.

The phased array reflector shown in Fig. C-8b has general characteristics similar to those of the lens. However, the same radiating element collects and re-radiated after reflection. Ample space for phase shifter control circuits exists behind the reflector. To avoid aperture blocking, the primary feed may be offset as shown. As before, transmit and receive feeds may be separated and the phases separately computed for the two functions. Multiple beams are again possible with additional feeds.

The phase shifter must be reciprocal so that there is a net controllable phase shift after passing through the device in both directions. This rules out nonreciprocal phase shifters.

b) Scanning Techniques and Systems There are several techniques for electronically scanning a beam presently being employed in various systems for diverse applications in both radar and communications. Since the basic objectives of this program may require implementation of a combination of these techniques, the basic principles of each is described briefly below:

Phase Scanning. This is the principle technique discussed in this sub-section. Here the beam of an antenna points to a direction that is normal to the phase front. In phased arrays this phase front is adjusted to steer the beam by individual control of the phase of excitation of each radiating element. This is indicated in Fig. C-9a. The phase shifters are electronically actuated to permit rapid scanning and are adjusted in phase to a value between 0 and 2π. With an inter-element spacing s, the incremental phase shift ψ between adjacent elements for a scan angle θ_o is

$\psi = \frac{2\pi}{\lambda} s (\sin \theta_o)$. If the phase ψ is constant with frequency, then the scan angle θ_o is frequency dependent such that $\frac{\sin \theta_o}{\lambda}$ is constant.

Time Delay Scanning. The simple discussion above indicates that phase scanning is fundamentally frequency sensitive. Time delay scanning is independent of frequency. Delay lines are used instead of phase shifters, as shown in Fig. C-9b, providing an incremental delay from element to element of $t = \frac{s}{c} \sin \theta_o$. Individual time delay circuits (Ref. C-14) are normally too complex to be added to each radiating element. A reasonable compromise may be reached by adding one time delay network to a sub-array of elements that have phase shifters. This type of compromise may provide a lower loss factor for the entire system.

Frequency Scanning. Frequency rather than phase may be used as the active parameter to exploit the frequency sensitive characteristics of phase scanning. Figure C-9c shows the arrangement. At one particular frequency all radiators are in phase. As the frequency is changed, the phase across the aperture tilts linearly, and the beam is scanned. This type of scanning may be used for "fine tuning" of the scan angle.

IF Scanning. When receiving, the output from each radiating element may be heterodyned (mixed) to an IF frequency. All the various methods of scanning are then possible, including the beam switching system described below, and can be carried out at IF where amplification is readily available and lumped constant circuits may be used. Equivalent techniques of mixing may be used for transmitting.

Beam Switching. With lenses or reflectors, a multiplicity of independent beams may be formed by feeds at the focal surface. Each beam has substantially the gain and beamwidth of the whole antenna. Allen (Ref. C-16) has shown that there are efficient equivalent transmission networks that use directional couplers and have the same collimating property. A typical form after Blass (Ref. C-17) is shown in Fig. C-9d. The beams may be

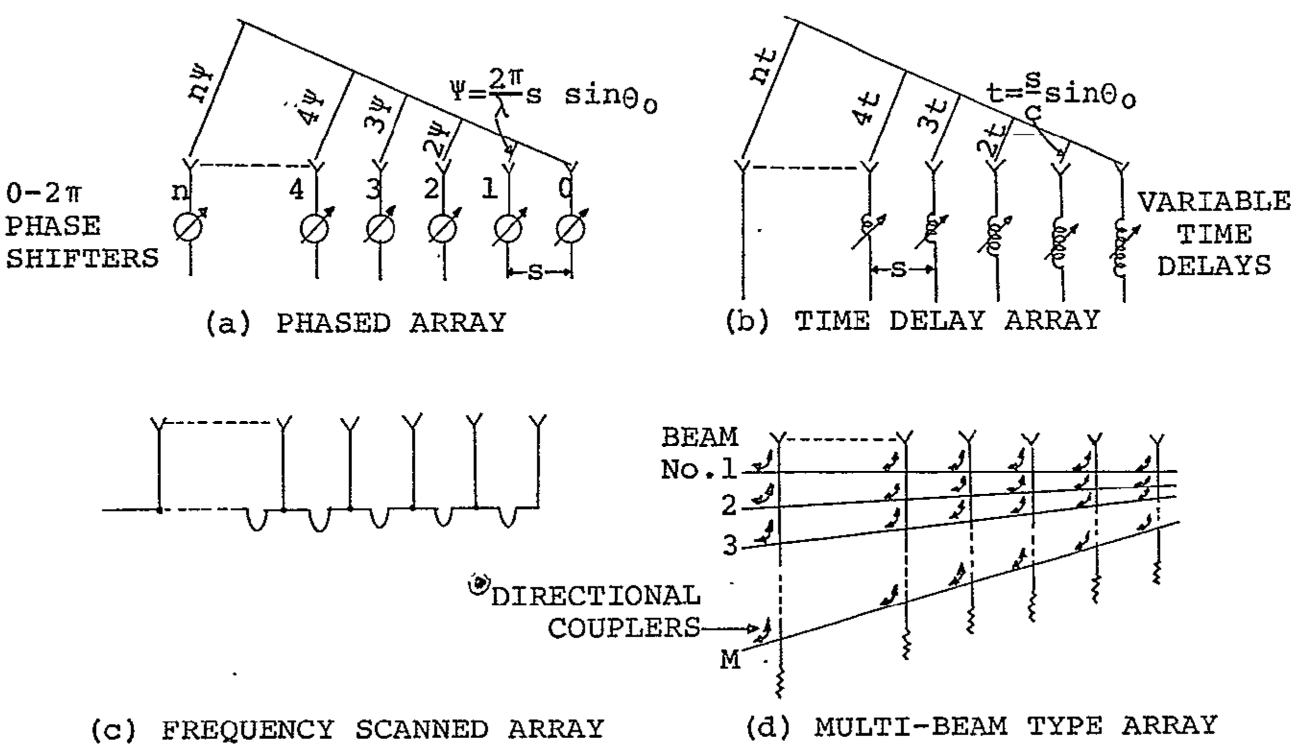


Fig.C-9 Generation of Scanned Beams

selected through a switching matrix requiring (M-1) SPDT switches to select one out of M beams. The beams are stationary in space and overlap at about the 4 db points. This is in contrast to the previously discussed methods of scanning, where the beam could be steered accurately to any position. The beams all lie in one plane. Much more complexity is required for a system giving simultaneous beams in both planes.

c) RF Phase Shifters Beam steering for a conventional phased array requires some type of phase shifting device at each element. The primary requirements for such a device are that it be capable of 360 degrees of phase shift and that it has an extremely low insertion loss, preferably less than 0.1 db. In addition these devices must be relatively inexpensive since their requisite number is proportional to the total aperture size, be capable of being packaged to fit within the array element spacing, and be temperature insensitive to ambient environments.

At present, there is no phase shifting device that will meet all of these requirements. Typically, electronic phase shifters such as ferrite and diode devices have insertion losses on the order of 0.5 db. While this loss does not greatly reduce the incoming signal strength, it does contribute considerable noise and consequently seriously degrades the SNR which influences the required aperture size. As shown in Figure C-10 which was computed from Equation C-13 (series-series model) considerable improvement in SNR is possible by cooling the device. This seems like particularly feasible approach for the diode type of phase shifters where a Peltier cooling device could be incorporated as an integral part of a semiconductor chip. Several commercial manufacturers are presently developing and manufacturing Peltier cooling devices for inclusion in a diode phase shifter and for direct attachment to the semiconductor.

Since the objectives of this present program are completely dependent upon adequate phase shifting devices and techniques, a continuing effort will be made to assess the performance parameters of the present state-of-art devices as well as to evaluate the potential of newly discovered structures. The listing shown in Table C-III describes the nominal performance parameters of various generic types of phase shifters at X-band frequencies since these devices are readily available and extensively used in radar systems. As may be seen from the table, many of the devices

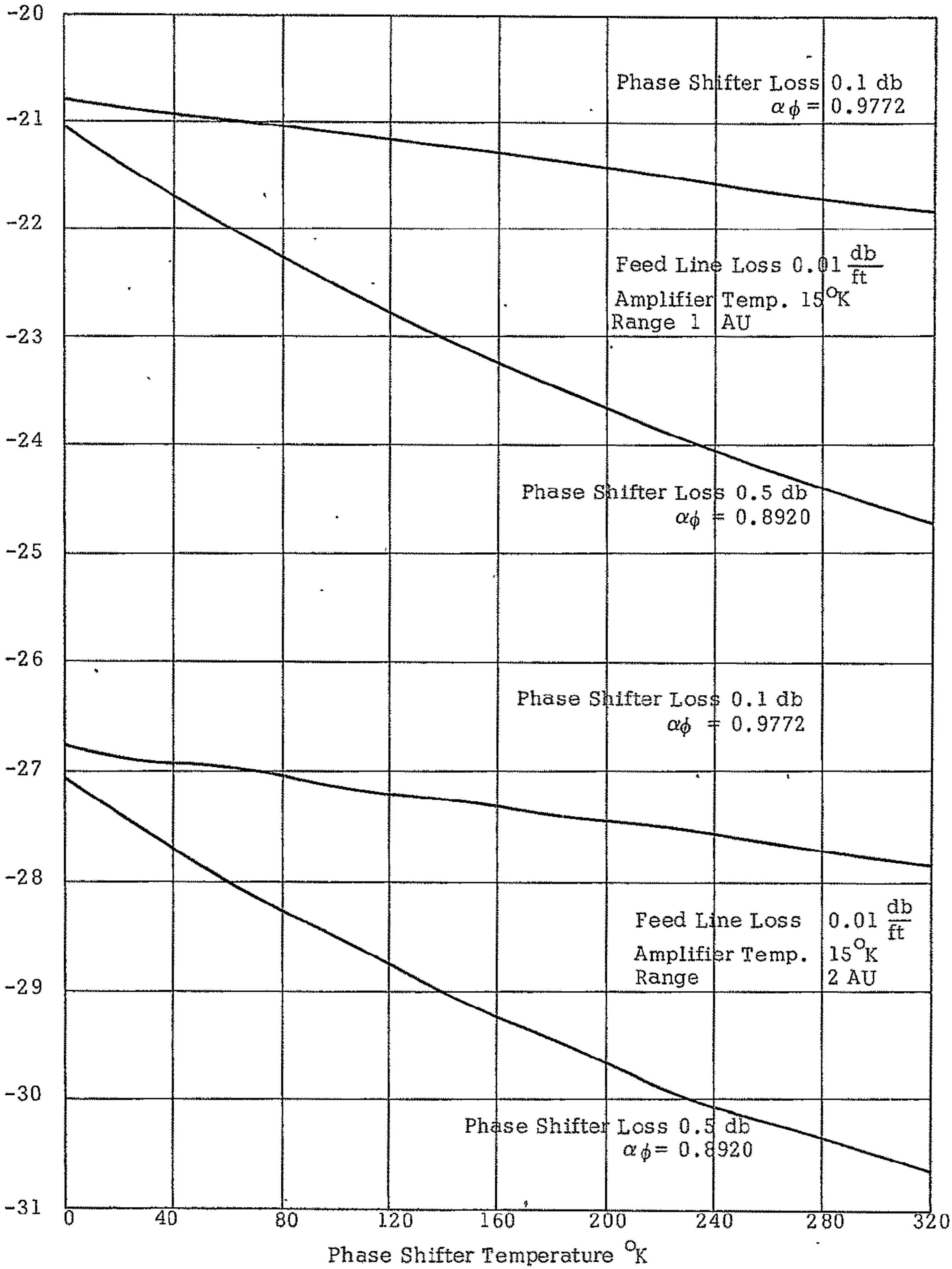


FIGURE C-10 SNR vs PHASE SHIFTER TEMPERATURE
FOR A 100 ELEMENT SUBARRAY

have relatively high insertion losses for the present communication application. These large loss values are due partly to the universal requirements of fast switching speed and high power handling capacity as dictated by radar application. Neither high-speed nor high-power capability are necessary for a ground based communication system, and consequently it can be expected that special designs of the above devices may be available with a substantially lower loss than the values of .6 to 3.0 db for 360° of phase shift as shown in Table C-III. However, for the present studies a nominal insertion loss value of .5 db shall be used until analysis and the appropriate experimental hardware are available to reduce the insertion loss to the desired value.

The results for X-band phase shifters are given only as a temporary expedient until precise descriptions of the corresponding devices operating at S-band frequencies can be obtained. In Table C-IV, a listing of commercially available S-band phase shifters is given with only some of the pertinent performance characteristics. More information will become available as these devices are employed in various array applications.

The devices that are presently available for phased arrays fall into three general groups which require consideration and some critical observation. A preliminary discussion of these groups, their advantages and disadvantages is given below and will be updated as new pertinent information becomes available:

c1) Diode Phase Shifters - Digital diode phase shifters are small, light-weight devices that are insensitive to temperature and can be switched from one phase setting to another in a few nanoseconds. Two types of digital phase shifters are in current use. One uses a transmission line structure in which different susceptances are switched across the line to produce incremental phase shift. The other design configuration is a reflection structure that may be converted to a transmission component by the employment of a 3 db coupler or a circulator. Diode phase shifters, are at present, somewhat costly because of the cost of the diodes and their mounting structure. P-i-n diodes are typically used as the control elements because of their high power handling capability. Since high power is not of prime concern in a receiving system, other arrangements of solid state materials may be more desirable although to date there has been no

Microwave Phase Shifters (May 1968)

Characteristic	Ferrite Longitudinal Field, Digital, Reciprocal	Ferrite Transverse Field, Digital, Nonreciprocal	Ferrite Longitudinal Field, Digital, Reciprocal	P-I-N Diode Transmission Digital (Stripline)	P-I-N Diode Reflection Digital - One Port Device	P-I-N Diode Reflection Digital - with 3-db Coupler
Frequency	X-band	X-band	X-band	X-band	X-band	X-band
Phase Shift (Maximum)	360° continuous	360° 22.5° steps (4-bit)	360° 22.5° steps (4-bit)	360° 22.5° steps (4-bit)	360° 90° steps (2-bit)	360° 90° steps (2-bit)
Figure of Merit	600°/db	500°/db	350°/db	100°/db	100°/db	200°/db
Temperature Sensitivity	5°/°C	1°-3°/°C	5°/°C	Negligible	Negligible	Negligible
Excess Noise Temperature	0°K	0°K	0° K	Negligible	Negligible	Negligible
Control Power	0.1 watt	3 by 10^{-4} watt-sec.	10^{-3} watt-sec	0.1 watt	0.05 watt	0.1 watt
Time Constant	100 μsec.	2-10 μsec.	2-10 μsec.	0.2-10 μsec.	0.2-10 μsec.	0.2-10 μsec.
Size (inches)*	6 by 1 by 1/2	5 by 1 by 1/2	7 by 1 by 1/2	10 by 1 by 1/2	1 by 1 by 0.025	2 by 1 by 0.025
Weight	2-1/2 oz + waveguide	2-1/2 oz + waveguide	5 oz. + waveguide	1-1/2 oz.	1/2 oz.	1 oz.
Disadvantages	1. Requires magnetic shielding 2. Requires temperature stabilization 3. Requires continuous holding power	1. Nonreciprocal 2. High current driver	1. High current driver 2. Requires temperature stabilization	1. Requires large numbers of diodes 2. Difficult to package 3. Susceptible to burnout	1. Susceptible to burnout 2. High loss for large number of phase bits	1. Susceptible to burnout 2. High loss for large number of phase bits
Range of frequencies at which practical devices can be built	2 to 50 GHz	2 to 40 GHz	2 to 20 GHz	0.1 to 10 GHz	2 to 18 GHz	2 to 18 GHz

* Assumes device slips into solid-state array.

Table C-III

Characteristic	MA*-8356 05X Diode Digital	MA-8356 151X Diode Digital	MA-8356 251X Diode Digital	MA-8356 451X Diode Digital	MA-8356 851X Diode Digital	MA-8356 8525 Diode Digital
Frequency	S-Band	S-Band	S-Band	S-Band	S-Band	S-Band
Phase Shift (Maximum)	22.5° 22.5°steps	45° 22.5°steps	90° 22.5°steps	180° 22.5° steps	360° 22.5° steps	360° 5.6° steps
Figure of Merit (Maximum VSWR) in db)	2.3	2.3	2.3	2.3	2.3	1.75
Temperature Sensitivity	Negligible	Negligible	Negligible	Negligible	Negligible	Negligible
Excess Noise Temperature	Negligible	Negligible	Negligible	Negligible	Negligible	Negligible
Control Power	+5V at 200mA -200V at 1mA	+5V at 400mA -200V at 1mA	+5V at 800mA -200V at 1mA	+5V at 1.6A -200V at 1mA	+5V at 3A -200V at 1mA	+5V at 3.4A -200V at 1mA
Time Constant	0.4 μs	0.4 μs	0.4 μs	0.4 μs	0.4 μs	0.5 μs
Size						
Weight						
Disadvantages						
Range of Frequencies at which practical devices can be built	2.9-3.1GHz	2.9-3.1GHz	2.9-3.1GHz	2.9-3.1GHz	2.9-3.1GHz	2.0-4.0GHz

* MA - Microwave Assiciates

TABLE C-IV

stimulus for such analysis and design. The engineers of the Texas Instrumental Corporation who are involved in the MERA module and system design report that they have been able to produce IC phase shifters with 1.2 db insertion loss as the average value of a large group with 1.5 db as a maximum value.

c2) Ferrite Phase Shifters - Ferrite phase shifters (Ref.C-12) are typically waveguide size, moderate in weight, somewhat temperature sensitive, can be switched from one phase setting to another in a few microseconds, and require significant drive energy. They are somewhat costly because of the cost of the ferrite material. Two general configurations are available. One uses a transverse magnetizing field; the second uses a longitudinal magnetizing field. The former is reciprocal only for certain configurations while the latter is intrinsically reciprocal, a property desirable in arrays to be used for both transmission and reception. Phase shifters that use longitudinal magnetization also produce greater phase shifts at lower levels of applied magnetic field than do those that use transverse magnetization. General characteristics of ferrite phase shifters that affect spacecraft scanning applications are reciprocity impedance matching, frequency dependence of phase shifts, temperature sensitivity, and hysteresis effects. Weight can also be a great problem with ferrite phase shifters for a spaceborne array with large numbers of elements. However, weight is only a secondary problem in a ground array compared to the temperature effects.

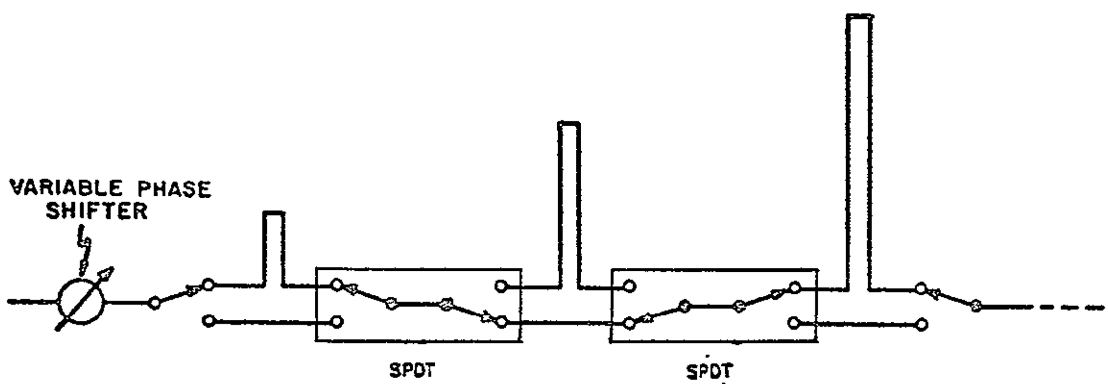
c3) Novel Devices - There are several new devices which are now being developed whose progress bears some observation. Ferroelectric phase shifters are quite small and light weight. They are, at present, extremely temperature sensitive, due to the sensitivity of the ferroelectric crystal, and they have very high insertion loss characteristics. Since they are still in the experimental stages, production costs are unknown. At present, it appears that a major improvement will be required in the basic crystal before these devices can be considered for use in an array. As in the case of the ferroelectric phase shifter, the plasma phase shifter is still in the experimental state. It is moderate in size and weight with a negligible temperature sensitivity. The insertion loss is comparable to that of the ferrite and diode phase shifters, but a significant reduction may be possible. At the present time, it is not a low cost device and requires significant drive energy; both factors are due to the need for the generating and sustaining of a plasma.

The high loss associated with the electronic phase shifting device can be eliminated or reduced by either mechanically scanning the subarrays, by using mechanical phase shifting devices such as a line stretcher, or by some form of simple air filled guide which may employ a multi-moding technique to properly gather the signals from numerous input ports. Each of these schemes needs further study and experimentation to develop the low loss feed system required by a high data rate communication link.

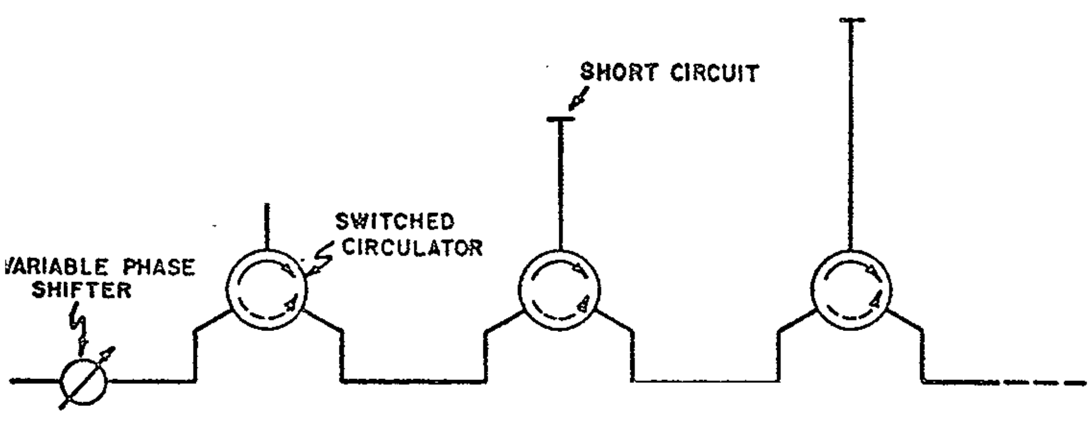
From the preceding equations it can be shown that one of the most important components which influence the required aperture size is the phase shifters. Electronic phase shifters such as ferrite and diode devices typically have insertion losses in the order of 0.5 db instead of the more desirable 0.1 db. While this loss does not greatly reduce the incoming signal it does contribute considerable noise and consequently seriously degrades the SNR. As shown in Fig. C-10 for a typical set of parameters, considerable improvement in SNR is possible by cooling; this seems particularly feasible for the diode type phase shifters where a Peltier cooling device could be an integral part of the semiconductor chip. The Peltier cooling effect is a thermo-electric phenomenon in which heat is absorbed or generated by current passing through a semiconductor junction. Several companies (Ref.C-12) are presently developing and manufacturing Peltier cooling devices for inclusion in the diode case and for direct attachment to the semiconductor chip. These problems will require further study and work is now in progress to examine the results using parameters that are more closely related to values which are possibilities for the future.

d) Time Delay Networks. Figure C-11a shows a time delay network that is digitally controlled by switches. The total delay path length that has to be provided nondispersively amounts to 'a $\sin\theta_{max}$,' where θ_{max} is the maximum scan angle for the aperture 'a'. The smallest bit size is about $\lambda/2$ or λ, with the precise setting adjusted by an additional variable phase shifter. A 1° beam scanned 60°, for example, requires a time delay of 6 or 7 bits, the largest being 32 wavelengths, as well as an additional phase shifter. The tolerances are tight, amounting in this case to a few degrees out of about 20,000, and are difficult to meet. Problems may be due to leakage past the switch, to a difference in insertion loss between the alternate paths, to small mismatches at the various junctions, to variations in temperature or to the dispersive characteristics of some of the reactive components. Painstaking design is necessary. The switches may be diodes or circulators. Leakage past the switches may be reduced by adding another switch in series in each line. The difference in insertion loss between the two paths may be equalized by padding the shorter arm. The various problems are comprehensively assessed and analyzed by Temme and Betts (Ref. C-18).

Figure C-11b shows another configuration that has the advantage of simplicity. Each of the switchable circulators connects either directly



(a) TIME DELAY BY CHOOSING UPPER OR LOWER PATHS
(SPDT = SINGLE-POLE, DOUBLE-THROW SWITCH)



(b) TIME DELAY USING SWITCHED CIRCULATORS

Fig. C-11 Time Delay Configurations

across (counterclock wise) or via the short-circuited length. Isolation in excess of 30 db is required, and the higher insertion loss of the longer path cannot easily be compensated. Each time delay network would therefore, precede a final power amplifier on transmitting and follow a pre-amplifier on receiving or a special design which is as yet unavailable.

Only the edge elements or edge subarrays of the antenna require the full range of time delay. The center does not need any time delay, only a biasing line-length. The amount of delay required increases as the edge of the aperture is approached. This is shown in Fig. C-12.

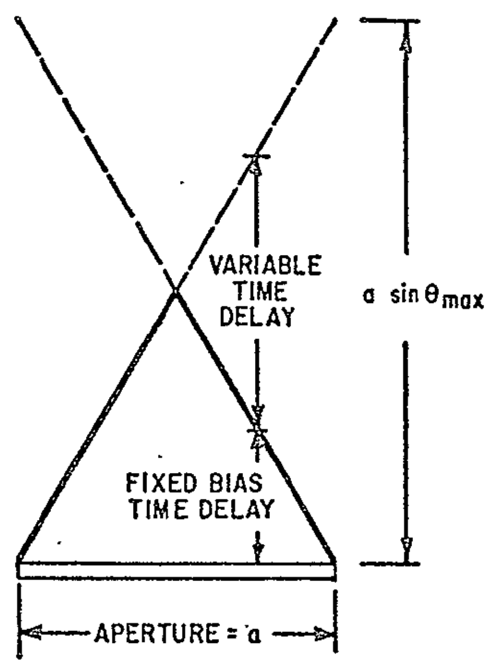


Fig. C-12 'Variable' and 'Fixed Bias' Time Delay for an Aperture

A further method of providing delay is possible by translating the problem from the microwave domain and delaying at IF since the insertion loss of time delay circuits is usually too high for most practical systems at RF.

e) <u>I-F Phase Shifting Techniques</u>. Because of the modular nature of the electronically steered systems being considered for this study, it will be possible to employ I-F phase shifting techniques. These techniques (Ref.C-1) offer several advantages as compared with R-F phase shifting techniques. First, requirements on the phase-shifting components may be relaxed as compared with requirements on corresponding R-F components. In addition, since the phase shifting for reception is performed after frequency translation and I-F amplification, losses in the phase shifters do not degrade system noise figure nor do they contribute to reduced system gain as would R-F phase shifters without individual R-F preamplifiers. The phase shifting for transmission can be done at low power levels with I-F phase shifters so that the power handling capabilities and losses of the phase shifters do not present problems. Typically, each complete module includes an antenna element, an R-F diplexer, a mixer, an I-F amplifier, and a phase shifter for reception; for transmission a similar set of components is required with the addition of a high-power R-F source. A number of configurations are possible to accomplish the desired performance characteristics but each requires I-F phase shifting devices. These devices are discussed in the following paragraphs.

Delay Line Phase Shifter. The simplest type of I-F digital phase shifter is that composed of discrete sections of delay lines that can be switched in and out with electronically controlled single-pole, double-throw switches. Such a device is illustrated in Figure C-11a. The various delay lines could be distributed or lumped parametric types depending on the particular frequency ranges being used. The 180° phase step is obtained merely by reversing the polarity of the line connections at that point.

f) Solid-State Components During the past years, technical literature has reported significant improvement in solid-state devices and circuitry for electronically steered arrays. Typically, improvements have been effected in phase shifters, I-F amplifiers, microwave power sources, mixers, filters, and circulators.

Filters. Excellent filters are commercially available in the frequency range up through X-band and beyond. These include filters employed in communication systems; for example, bandpass (nominally flat), band rejection, diplexers, and high Q stabilizing cavities. In these higher frequency ranges the structures may be waveguide, strip transmission line, coaxial, or microstrip; but for space applications, the small, lightweight strip transmission line coaxial devices, or microstrip, are most attractive. The performance of the latter, in terms of loss, needs improvement to be competitive with waveguide filters.

Preamplifiers. There are two possibilities for the preamplifier that lend themselves to microstrip application: tunnel diode amplifiers (TDA's) and transistor amplifiers. With the present state-of-the-art at 2 to 10 GHz and above, the TDA is slightly lower in noise figure than available transistors. Since a TDA must use a circulator, a 0.5db insertion loss must be added to the noise figure to give a value of 4.5db and perhaps 30db of gain. In comparison present day transistors can give a noise figure of 5.2db and 20db of gain.* At present, a 1 GHz, transistors have 3.5db noise figures, but manufacturers (KMC Corporation and NEC) anticipate that devices with better noise figures will be available within a year. Such devices would give a receiver noise figure of 4.4db at S-band. A transistor amplifier can be fabricated into a smaller package than the TDA due to the use of microcircuit lumped elements. The TDA uses at least one circulator which, with present technology, has a minimum size of about 1 inch square. Thus, on a size and weight and future performance comparison, the transistor amplifier is the preferred device.

At X-band a tunnel diode amplifier will give the best noise figure. However, because a mixer is simpler, lighter in weight, and lower in cost and has a competitive noise figure, it is anticipated that it will remain the preferred component at the higher frequencies for several years.

Mixer. The element that most determines the design of the receiver is the mixer. Present conventional balanced mixers have produced single sideband noise figures of less than 5db at S-band. However, this value represents carefully matched low loss conditions which may be hard to achieve in mass production in microstrip.

An alternative design for the conventional mixer with a low-noise

* Nippon Electric Co., SM153 Gallium Asenide Schottky Barrier Diode. I-F amplifier noise figure assumed to be 1.5 db.

preamplifier is the image enhancement mixer. Recently at MIT* an S-band image enhancement mixer was measured with less than a 3 db single sideband noise figure and 0 dbm saturation level. The local oscillator power and complexity of this device is greater than that of the conventional mixer. A local oscillator drive of 50 mw was reported; this figure compares with 1 or 2 mw for normal operation. This type of mixer will need further development before its merits can be fully evaluated.

As the integrated circuit technology advances, solid-state devices are being developed for integration into array antennas to form and phase beams and also for amplification. Microstrip transmission circuits have been developed that contain various microwave circuit elements such as circulators, switches as well as amplifiers, mixers, and multipliers (Ref.19). With these devices, systems become possible where many relatively low power transmit amplifiers are used and distributed over the aperture with each amplifier connected to a radiating element. The expected advantages of integrated antennas include high reliability and low cost, simple low voltage power supplies for the RF amplifiers and a system which is simple and light in weight and yet capable of operating with relatively high RF power.

* R. P. Rafuse and D. Steinbrecher as reported in Sprint, MIT Quarterly Progress Report and by private communication.

D. A SELF-STEERING ARRAY

1) Introduction

This portion of the report is concerned with the problems associated with self-phased and adaptive arrays which can be employed to follow the relatively weak signal from a distant spacecraft. These arrays are also called self-focusing antennas since they use the incident RF energy to phase the elements so that a beam is formed in the direction from which the energy is received. The arrays may be contrasted with the usual electronically steerable arrays that require external sensors and information to do the steering. Here, no external commands are necessary to adjust the illumination across the aperture, since, in principle, the self-steering array automatically steers the beam in the desired direction. By the inclusion of appropriate signal processing circuitry, an adaptive array can perform filtering in both the space and the frequency domains, thus reducing the sensitivity of the receiving system to interfering directional noise sources. Thus the problems associated with pointing a narrow beam in a specified direction or with atmospheric scintillation effect may be handled in a self-phasing mode and those associated with periodic interference effects may be handled with adaptive array techniques as an alternative mode of operation. This section will consider the feasibility of switching from a system where the steering is accomplished by externally controlling the phase between elements to a self-steering or adaptive array whenever a high external noise or interference level is present in the angular region subtended by the receiving beam.

At present, the behavior of a two-element adaptive array and a four-element adaptive array has been studied on the digital computer using the basic algorithm. A number of computer programs were written and used to simulate the operation of these arrays under a variety of conditions (namely, as a function of the feedback loop parameters, the power levels of the signal, the interfering signal,and the noise, and the arrival angles of the signal and interfering signal). In all cases (once the computer programs were properly debugged) the arrays performed admirably. The weighting coefficients converged, and the resulting antenna patterns were such as to reduce the noise from the interfering signal to the minimum possible, given the number of elements in the array. An experimental adaptive array (at S-Band) based on this principle is presently under construction and is described below. It is planned to use this array to learn what limitations are placed on the adaptation process by the idiosyncracies of the actual electronics.

2) Theoretical Description of an Adaptive Array

An "adaptive antenna" may be defined as one that modifies its own pattern, frequency response, or other parameters, by means of internal feedback control, while the antenna is operating. Such automatic control of the antenna characteristics may be used (1) to exclude interfering signals from the output of the antenna, or (2) to maintain antenna performance in the presence of a changing near-field environment, as explained below. Other uses for such antennas are also discussed in subsection IV-D.

The work to date in this area has concerned an adaptive array as shown in Fig. D-1 and is based on a feedback algorithm for least mean square error (LMS) as discussed originally by Shor (Ref. D-1) and also by Widrow, et. al (Ref. D-2, D-3). In the basic form of such an antenna, $x_1(t), \ldots, x_i(t), \ldots, x_n(t)$ represent the signals received from the individual elements of the array. These signals are multiplied by weighting coefficients, $w_1, \ldots, w_n$ and then added together to produce the array output $s(t)$. In order to control the weighting coefficients w_i, the output signal $s(t)$ is compared with a "desired signal" $d(t)$ to produce an error signal $\varepsilon(t) = d(t) - s(t)$. (The question of how $d(t)$ is obtained is discussed below.) The error signal $\varepsilon(t)$ along with the signals $x_1(t), \ldots, x_n(t)$ are used as inputs to a feedback system which adjusts the weighting coefficients w_i. The feedback operates in such a way as to minimize $\varepsilon^2(t)$, i. e., to make the output of the array approximate the desired signal $d(t)$ as closely as possible, in a minimum squared error sense. The operation of the feedback loop may be described as follows: since the output from the array is (see Figure D-1)

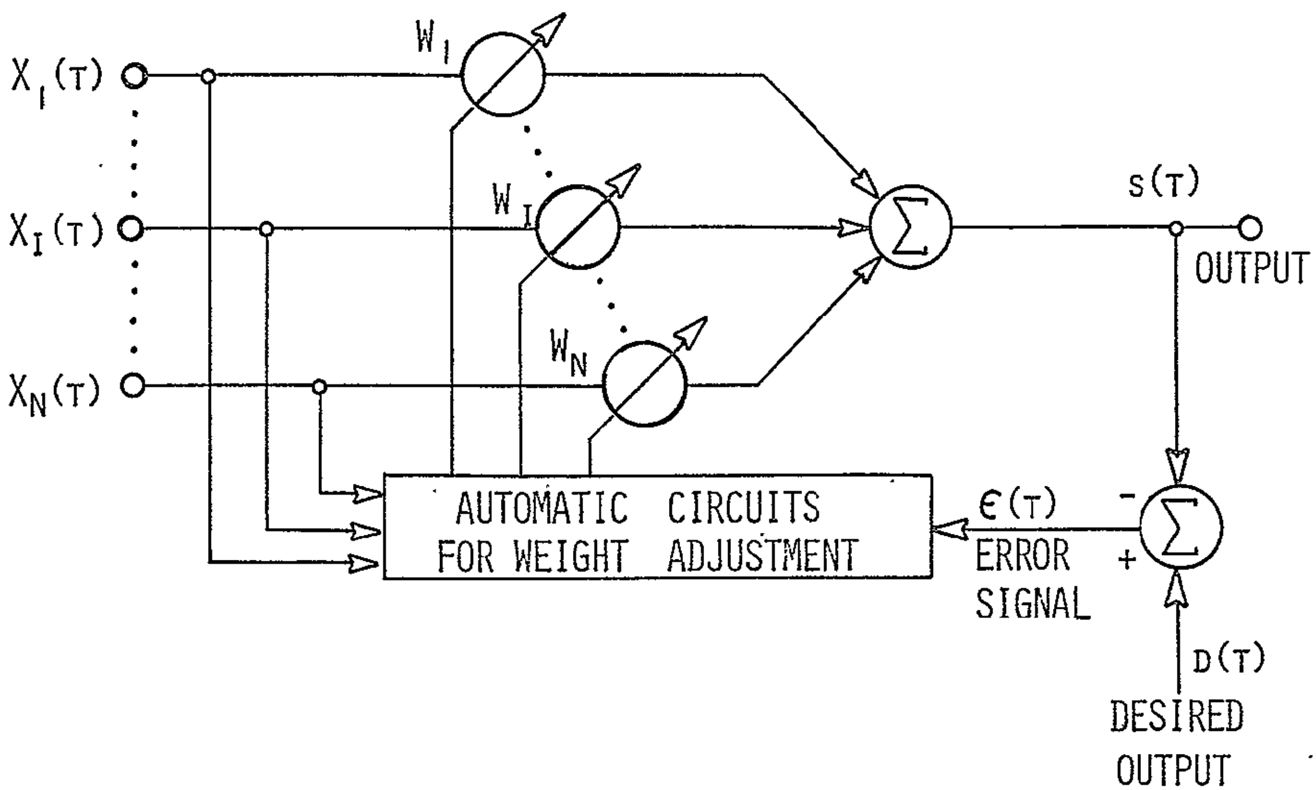
$$s(t) = \sum_i w_i x_i(t)$$

the error signal is

$$\epsilon(t) = d(t) - \sum_i w_i x_i(t)$$

and hence the "squared error" is

$$\epsilon^2(t) = d^2(t) - 2d(t)\sum_i w_i x_i(t) + \sum_i \sum_j w_i w_j x_i(t) x_j(t)$$



BASIC ADAPTIVE FEEDBACK SYSTEM

FIGURE D-1

$\varepsilon^2(t)$ is always a positive quantity and may be used as a performance criterion for the array. The lower ε^2, the better is the array adjustment. At any given time, ε^2 is a quadratic function of the weights w_i, so the surface defined by plotting ε^2 versus the w_i's is a "bowl-shaped" surface with a well-defined minimum. The value of w_i can be adjusted to keep the array operating near the bottom of the bowl, i. e., to minimize ε^2. To do this, w_i is adjusted according to a steepest descent method by computing the gradient of ε^2 with respect to the w_i, and moving the w_i in the maximum downhill direction.* Specifically, compute $\nabla(\varepsilon^2)$ from

$$\nabla(\epsilon^2) = \frac{\partial \epsilon^2}{\partial w_1}\hat{w}_1 + \frac{\partial \epsilon^2}{\partial w_2}\hat{w}_2 + \ldots + \frac{\partial \epsilon^2}{\partial w_n}\hat{w}_n$$

and then adjust each w_i so that

$$\frac{dw_i}{dt} = k_s \nabla_i(\epsilon^2) = k_s \frac{\partial \epsilon^2}{\partial w_i}$$

where k_s is a negative constant. Thus, if $\frac{\partial \epsilon^2}{\partial w_i}$ indicates a large sensitivity of ϵ^2 to w_i, w_i is changed quickly to move toward the bottom of the bowl. If $\frac{\partial \epsilon^2}{\partial w_i}$ is very small, w_i changes very slowly.

Since

$$\nabla_i(\epsilon^2) = 2\epsilon\nabla_i\epsilon = -2\epsilon X_i$$

* Comparison of the work of Shor (Ref. D-1) and Widrow (Ref. D-3) also shows that this is identical to a steepest-ascent optimization of signal-to-noise ratio.

the feedback rule is actually

$$\frac{dw_i}{dt} = -2k_s \quad (t)\, X_i(t)$$

or

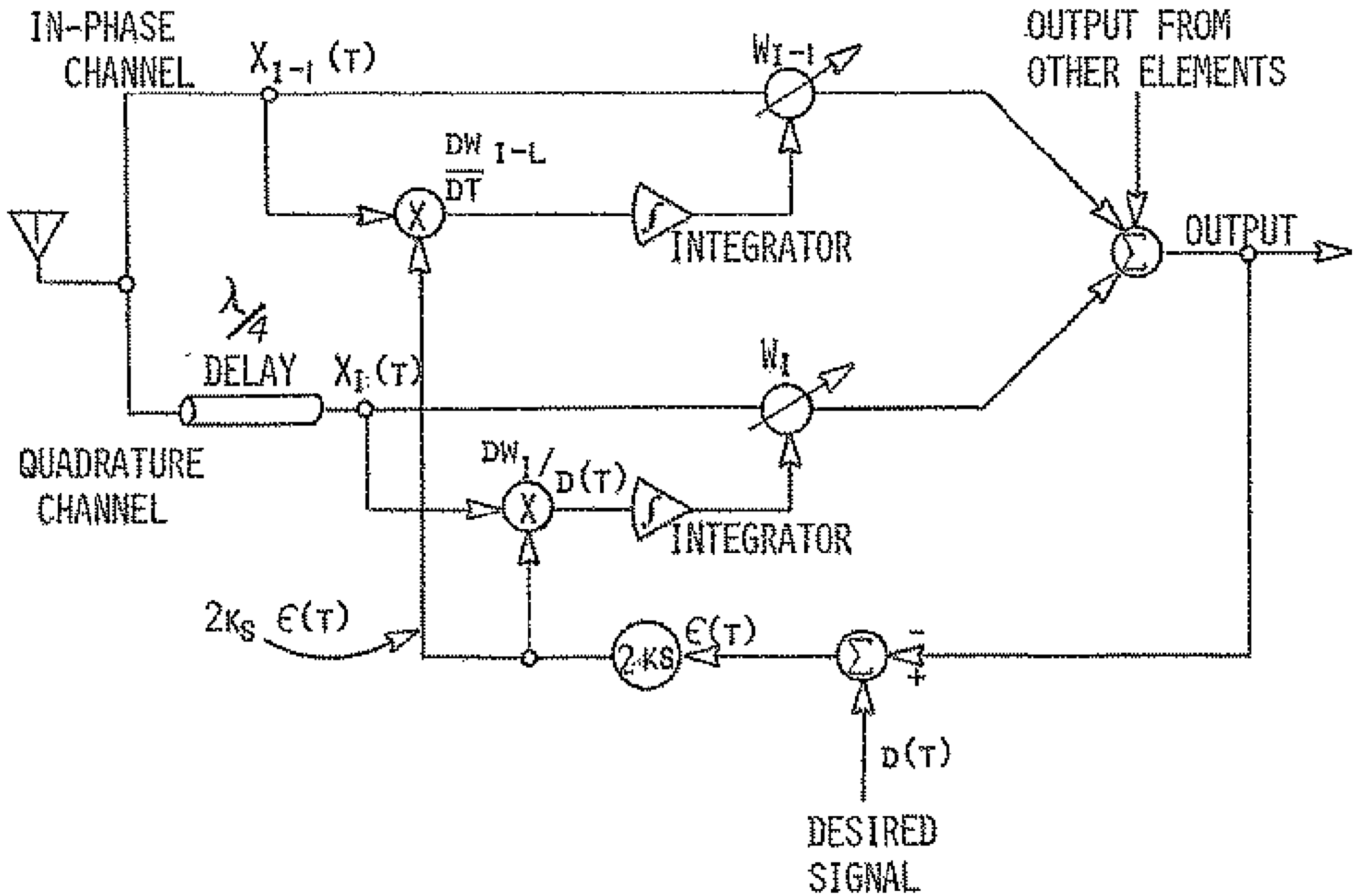
$$w_i(t) = -2k_s \int_o^t \epsilon(t')\, X_i(t')dt' + w_i(o)$$

3) Experimental System

The feedback system described above has been implemented using the system shown in Figure D-2. This diagram shows the feedback arrangement for one element of the array. To obtain arbitrary magnitude and phase for each element, the signal from each element is split into two channels, an "in-phase" channel and a "quadrature" channel, with the signal in the quadrature channel delayed 90^o with respect to the in-phase channel. The gain in each channel then operates in a feedback loop of the type described above.

Most of the work during the last semi-annual period has been devoted to the practical aspects of implementing this feedback scheme. The hardware has been designed and constructed. Final adjustments have been made on two complete units for two elements in the adaptive array. A photograph of the electronics package for one element is shown in Figure D-3 and a complete block diagram of the functions contained in this unit is shown in Figure D-4. Note that four amplifiers (A_1, A_2, A_3, A_4) are actually used, two for the in-phase channel and two for the quadrature channel. This arrangement is used so that both positive and negative gain may be obtained for each channel, i. e., so that a full 360^o range is available for the phase angle of the signal from the element.

At present, experimental tests on a two-element adaptive array are in progress. When these are completed, the program plans call for the building of four such units altogether, and to perform tests on a four-element adaptive array before September, 1969. In addition to the experimental program, theoretical studies are in progress concerning the behavior of this feedback system for more general types of inputs (non-stationary signal, signals that are part deterministic and part random, etc.). The paper by Widrow (Ref.D-3) treats only the case where the input signals are stationary uncorrelated



FEEDBACK CIRCUIT FOR EACH ELEMENT

FIGURE D-2

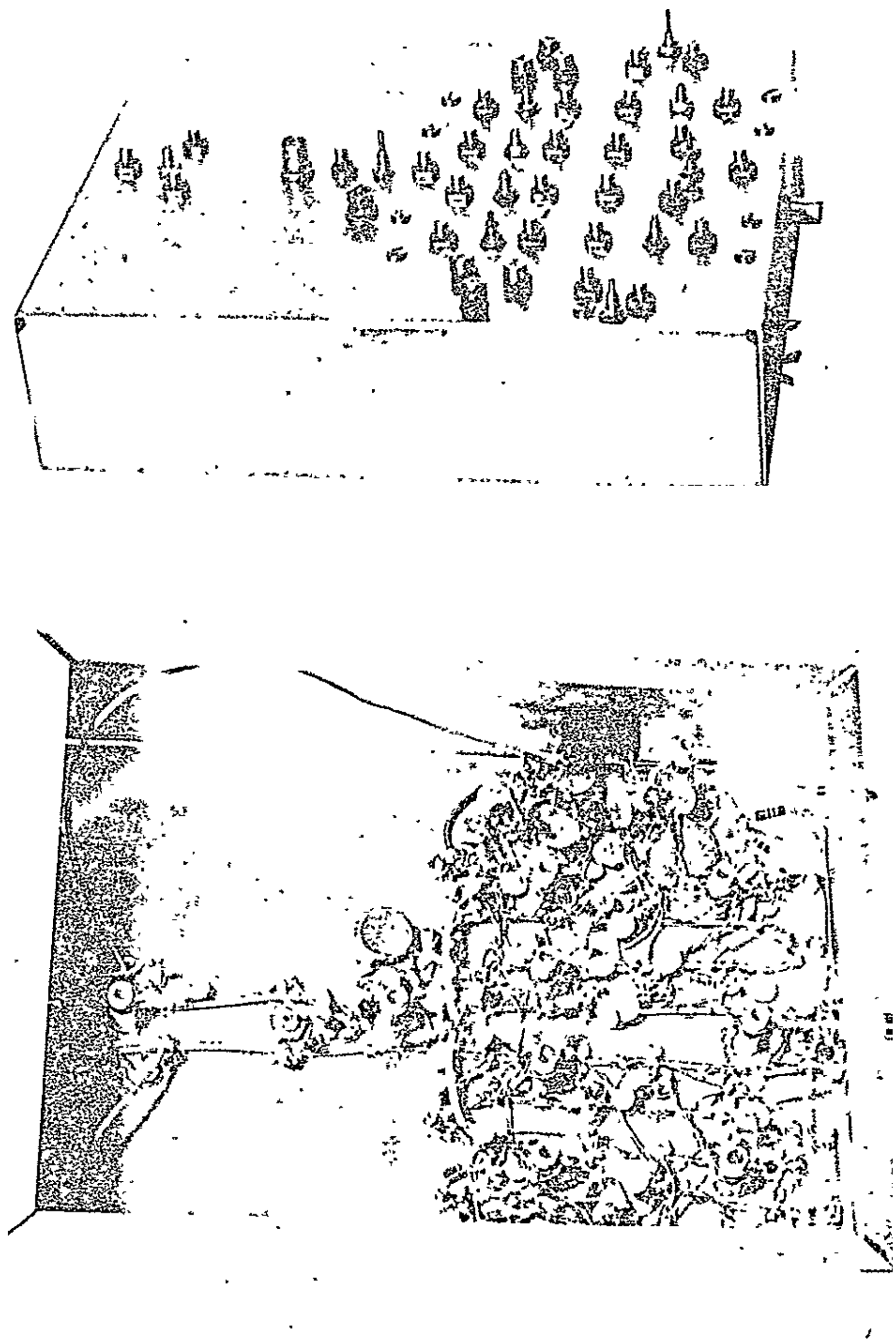


Figure D-3

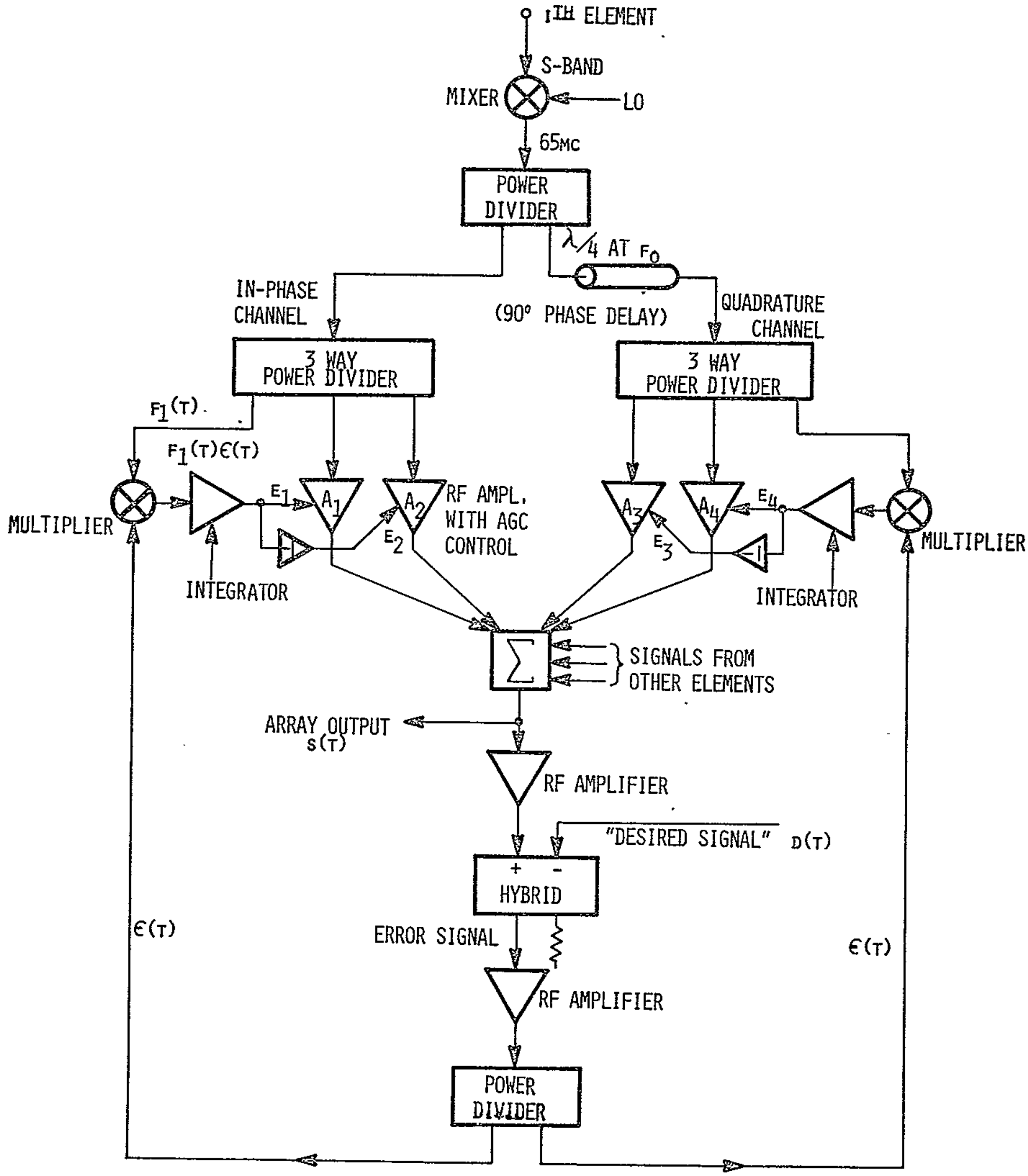

I^TH ELEMENT
S-BAND
MIXER
LO
65MC
POWER DIVIDER
λ/4 AT F0
(90° PHASE DELAY)
IN-PHASE CHANNEL
QUADRATURE CHANNEL
3 WAY POWER DIVIDER
3 WAY POWER DIVIDER
F1(T)
F1(T)ε(T)
MULTIPLIER
E1
A1
A2
RF AMPL. WITH AGC CONTROL
E2
-1
INTEGRATOR
A3
A4
E3
E4
-1
INTEGRATOR
MULTIPLIER
Σ
SIGNALS FROM OTHER ELEMENTS
ARRAY OUTPUT S(T)
RF AMPLIFIER
"DESIRED SIGNAL" D(T)
+ -
HYBRID
ERROR SIGNAL
ε(T)
ε(T)
RF AMPLIFIER
POWER DIVIDER

FIGURE D-4

random variables. A report describing both the theoretical and experimental results on this subject is to be published in the near future.

4) Possible Applications

Although the prime motivation for studying adaptive arrays with respect to high performance communication systems has already been mentioned, it is certainly worthwhile to consider how such an antenna type can be used in other applications. Since it is a versatile technique, it needs some further consideration before its total value is clearly understood. A number of applications have occurred to the workers on this program, and as the behavior of these antennas becomes clearer and the performance characteristics become more definitive, more applications become apparent. These applications are worth discussing becuase they provide the motivation and justification for working in this portion of the program. However, it has now become apparent that adaptive arrays have versatility which can be employed successfully in other areas. Consequently, this subsection will be devoted to a qualitative description of some of the areas where adaptive antennas appear to be useful.

The first and perhaps most important application of adaptive antennas will be as a "design tool" for conformal arrays-- for arrays whose elements must be placed on a curved surface. In practice, it is difficult to design the phasing networks for a conformal array (particularly if the antenna beam must be electronically scanned) because the element patterns and mutual impedances are different for each element. Adaptive antennas offer a possible solution to this problem. A conformal array can be built on the surface on which it is to be used, and then by going through a special test procedure, while the antenna is operated in an adaptive mode, the optimum set of weighting coefficients can be found. The test procedure would consist of illuminating the antenna with a test signal from various directions in space, while using the same test signal for the desired signal d(t). The adaptive feedback will find, for each direction of illumination, the best set of weighting coefficients. In other words, the antenna will design its own aperture distribution. The coefficients found can be stored and used later in a normal scanning mode.

A second use for adaptive antennas is in situations where the antenna is subject to a changing near-field environment, and it is necessary to recalibrate or readjust the pattern of the antenna during such changes. There are many examples of this. For instance, antennas used for aircraft control around airports must have patterns which are accurately known. After such an antenna is installed and operating, it may happen that at a later date airport officials would like to put up a new building, but are unsure what effect the presence of the building may have on the antenna pattern. If the antenna could be operated in an adaptive mode, it could be recalibrated periodically as the building is being constructed, using a test procedure similar to the one outlined above.

A third use for adaptive antennas is for communication antennas that are resistant to jamming and other forms of interference. By providing a "desired signal" d(t) and a test signal on each element of the array with a phase corresponding to a given "look angle", an adaptive antenna has the property that it receives signals from the desired direction, but tends to form nulls on signals arriving from other directions. (This behavior is described by Widrow, et. al., Ref. D-3). If an interfering signal appears from a certain direction, the weighting coefficients in the array readjust themselves to form a null in that direction.

As a fourth possibility, the same technique as used for anti-jamming antenna above can also be used to eliminate low-angle clutter from a radar antenna. The requirement for operating an antenna at low elevation angles results in a difficult pattern synthesis problem, namely, the synthesis of a main beam with a nonsymmetric sidelobe structure. The sidelobes on the "ground" side must be minimized at the expense of the "sky" sidelobes. The adaptive array achieves the desired characteristics in an optimum way by minimizing the undesired power return from wherever it may arrive and at the same time maximizing the desired signal.

Many other possibilities exist. It is clear that the applications for adaptive arrays, although of a special nature, are sufficiently numerous that study in this area is worthwhile.

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